THE RADIO AND ELECTRONIC ENGINEER

The Journal of the British Institution of Radio Engineers

FOUNDED 1925 INCORPORATED BY ROYAL CHARTER 1961

"To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering."

VOLUME 26 SEPTEMBER 1963 NUMBER 3

INTERNATIONAL CO-OPERATION IN SPACE RESEARCH

ONE of the most encouraging results of the development of space technology has been the increasing collaboration between scientists and engineers of different nations in the planning and execution of research and other projects.

The first joint space research venture, in April 1962, was the launching by the United States National Aeronautics and Space Administration of a satellite containing four experiments designed and built by British University groups—*Ariel I.* A second satellite, at present known as UK-2, is to be put into orbit later this year and will carry a payload of three experiments designed respectively by Cambridge University, Manchester University and the Meteorological Office. The satellite structure and ancillary equipment of UK-2 are being provided by the U.S.A., but UK-3—scheduled for launching by N.A.S.A. in two to three years' time—is being designed and built entirely in Great Britain. The five experiments in this satellite are being prepared by the Universities of Cambridge, Sheffield and Birmingham, and by the Meteorological Office and the Radio Research Station of D.S.I.R.

An ionospheric topside sounder satellite—*Alouette*—was launched by N.A.S.A. for the Canadian Defence Research Board last year and has provided much useful data on the structure of the ionosphere. It is also interesting to record that, at the request of N.A.S.A., a mass spectrometer probe, of the type designed by University College, London, and used successfully in *Ariel I*, will be fitted in the N.A.S.A. topside sounder satellite, S-48.

Recently N.A.S.A. invited United Kingdom scientists to propose, in competition with U.S. scientists, experimental instruments to be carried in a number of N.A.S.A. scientific satellites during the next two or three years. Some of the first proposals have already been selected for installation in a series of large 'observatory' satellites which will carry numbers of different instruments into orbit. Experiments designed by University College, London, and the University of Leicester to study solar x-ray emission, and solar helium II resonance emission are to be carried in an orbiting solar observatory (OSO-D). Another joint U.C.L./Leicester experiment, for studying stellar x-ray emission will be carried in an orbiting astronomical observatory (OAO). Other British space research techniques are being considered for subsequent 'observatories'.

International collaboration in space is not confined to pure research and the exploitation of satellites for communications has called for the combined work of engineers from several countries in the experiments with *Telstar I* and *II*, *Relay* and *Syncom II.* These satellites have been completely American in design, but the ground stations have employed techniques devised by the countries in which they are sited.

Even more far reaching results could come from the agreements which have recently been made between the United States and the Soviet Union for joint effort in the design, launching and use of three types of satellite: weather watching, communications, and magnetic survey. The first will involve transmission of pictures of cloud cover between the satellites and Washington and Moscow to aid weather forecasting. The communications project will use a passive reflector satellite, *Echo II*, and Jodrell Bank radio telescope will take part as a terminal point. The magnetic survey satellites will play an important part in the programme of the World Magnetic Survey during the second of the International Years of the Quiet Sun (1965).

INSTITUTION NOTICES

International Symposium on Cold Cathode Tubes

The Institution is arranging an International Symposium on "Cold Cathode Tubes and their Applications", which will be held in Cambridge from 17th to 19th March 1964. Meetings will be in the Cavendish Laboratory and there will be residential accommodation in Colleges of the University.

It is intended that the papers read at the Symposium will primarily provide information for equipment designers on all aspects of cold cathode tubes. The types of devices falling within the theme of the Symposium are as follows:

Stabilizers and Diodes	Counting Tubes
Trigger Tubes	Indicator Tubes

The following broad subjects will be covered:

Physics of operation	Circuit design
Tube development	Application
Manufacturing techniques	Reliability

Offers of papers for consideration for inclusion in the Symposium are invited, particularly from overseas. Titles and synopses of papers should be sent as soon as possible to the Secretary, Programme and Papers Committee, Brit.I.R.E., 9 Bedford Square, London, W.C.1.

The Indian Proceedings of the Brit.I.R.E.

Following the establishment of the Indian Divisional Council of the Institution, arrangements have been made for the regular publication of *Indian Proceedings*. The first issue has now been sent to all members in India.

The contents of No. 1 of the *Indian Proceedings* include a Foreword by the President, a report of the first meeting of the Indian Divisional Council, a 'preview' of a paper describing a radio installation for an important pipeline in India (this paper, "A Telecommunications and Telecontrol System for a Crude Oil Pipeline" by W. T. Brown, appears in full in this issue of the *Journal*), notices of special interest to Indian members and the question papers set for Part A of the Institution Graduateship Examination in November 1962, together with the examiners' reports and answers.

Although primarily intended for members in India, copies of the *Proceedings* may be obtained by members resident elsewhere price Rs. 2 (3s.) per single copy or Rs. 10 (15s.) for one year's subscription. Subscriptions from non-members will be accepted at the following rate: Rs. 25 (37s. 6d.) per year, single copies Rs. 5 (7s. 6d.). Remittances in all cases should be addressed to Captain G. Swaminathan, Administrative Secretary, Brit.I.R.E. Indian Division, 33/3 Infantry Road, Bangalore.

Circulation of the *Journal*

The Audit Bureau of Circulations has certified that for the period January to June 1963, the average circulation of *The Radio and Electronic Engineer* was 10338 per month. This represents an increase of 423 over the previous half year and is 758 more than for January–June 1962.

With the achievement of a monthly five figure circulation, the *Journal* further strengthens its position as one of the leading British publications in the field of radio and electronics.

International Conference on Microwaves, Circuit Theory and Information Theory

An International Conference on Microwaves, Circuit Theory and Information Theory will be held at Akasaka Prince Hotel in Tokyo, Japan, from 7th to 11th September 1964. The Conference is sponsored by the Institute of Electrical Communication Engineers of Japan with the support of the Science Council of Japan and of the International Scientific Radio Union.

There will be sessions on the following major topics:

Microwave Theory and Techniques and Electron Devices;

Microwave Antennæ and Propagation;

Microwave Communication Systems;

Circuit Theory;

Information Theory.

Original papers in these fields will be considered for presentation at the Conference, and abstracts and summaries of papers should be sent to the Chairman of the Papers Committee:

Dr. Kiyoshi Morita, c/o The Institute of Electrical Communication Engineers of Japan, 2–8 Fujimicho, Chiyoda-ku, Tokyo, Japan.

Abstracts and summaries must be written in English; a one-page abstract of about 100 words and a two-page summary of 80–1200 words should be sent to Tokyo not later than 31st March 1964. Authors will be notified of the acceptance of their papers by the end of May 1964.

Programme of Meetings in Great Britain 1963/64

Members are reminded that full details of Institution meetings in London and of Local Sections in Great Britain are now published in the *Proceedings*. In addition to a diary of meetings for successive months, synopses of papers with information on the meeting arrangements are also given.

U.D.C. 621.375.4 : 621.382.3

A High Power Audio Frequency Push-pull Transistor Amplifier to Drive an Electro-magnetic Vibrator

By

R. F. C. BENNETT

(*Associate Member*)†

Presented at a meeting of the Electro-Acoustics Group in London on 19th December 1962.

Summary: A 1 kVA electro-magnetic vibrator, having a nominal impedance of 10 ohms and requiring a 150 V peak excitation source, is taken as a model for studying the problems involved in driving large vibrators.

The output transformer, usually associated with audio power amplifiers, determines the amplifier low frequency response. A design for a high power transformerless transistor push-pull amplifier is considered, with emphasis in particular on the problems associated with transistor series/parallel arrangements, and the power handling requirements of an amplifier driving a load of low power factor.

1. Introduction

Electro-magnetic vibrators, of various sizes and ratings, are used extensively throughout industry. The excitation for these devices, which is derived in most instances from a valve power amplifier with an appropriate matching transformer, range from low powers to greater than 75 kVA.

To a great extent the frequency performance is determined by the matching transformer, the primary inductance determining the low frequency response, while the maximum frequency is governed by leakage inductance and inter-winding capacitances. Although specially-designed transformers will improve the bandwidth, nevertheless at the higher powers this problem is more difficult because of the large physical dimensions of the transformer.

To make an appreciable improvement to the frequency response it is necessary to eliminate the transformer and to excite the coil of the vibrator directly from a suitable source. Since the coil impedance is normally low, and therefore requiring high current and relatively low voltage, a transistor push-pull amplifier may be used to advantage.

An amplifier was designed to deliver 1 kVA to a typical vibrator, having a nominal coil impedance of 10 ohms, to assess the various design problems which would be encountered.

2. Design Considerations

Before considering the amplifier design it is essential to obtain a complete picture of the load which is to be driven. The frequency characteristic of the vibrator is given in Fig. 1. The load is complex, and dependent on the frequency, may be either inductive, capacitive, resistive or resonant. The impedance curve is only representative of the load under one set of mechanical conditions.[1] The effective power factor and the resonance are determined by the mechanical loading of the table.

The maximum current rating of the vibrator coil is 15 A but at the low frequencies the maximum current is determined by the permissible excursion of the coil. The maximum voltage excursion was limited to ± 150 volts.

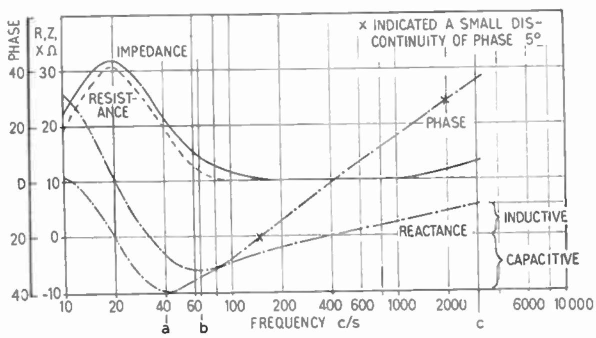


Fig. 1. Series equivalent of the vibrator—unloaded.

2.1. *Output Amplifier Power Handling Capacity*

The power output stage will of necessity take the form of a push-pull system in order to ensure no direct current will flow through the coil (Fig. 2). Figures 3(*a*), (*b*), and (*c*) illustrate a number of typical elliptical load lines suitably scaled to the maximum voltage and current ratings for the vibrator coil.[2] These load lines are shown as a theoretical class B push-pull arrangement. The single-ended power dissipation curves for 560 and 1000 W are also included.

With a purely resistive load of 10Ω the maximum dissipation in the transistors will be 560 W, but, with a complex load, the maximum dissipation will be

† U.K. Atomic Energy Authority, Atomic Weapons Research Establishment, Aldermaston, Berkshire.

increased. Figure 3(*c*) illustrates the complex impedance condition when the maximum dissipation is as high as 1000 W. It should be noted that this figure, although adequate for this particular vibrator, would not be applicable for lower power factors, e.g. in the extreme condition of a purely reactive load the maximum dissipation of the output stage will be 3000 W.

The maximum rated collector dissipation of a power transistor cannot be exceeded even for a short period without damage. This is particularly true when the period of the driving frequency is long compared to the time-constant of the junction. In order to achieve reliable operation it is essential that a reasonable power margin is allowed.

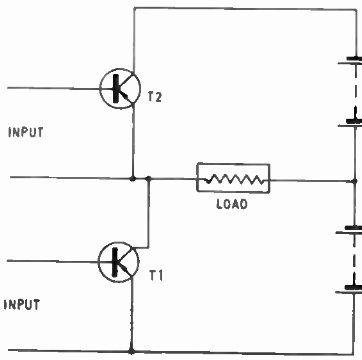


Fig. 2. Push-pull power output stage.

Referring again to Fig. 3(*c*), the 1000-W dissipation is the theoretical optimum for a perfect class B push-pull system.[3] This mode of operation is not ideal or practical since low distortion is required, and voltage sharing would prove troublesome. A class A mode of operation is desirable to fulfil these requirements.

Figure 4 shows the composite curves for an operating condition approaching a class A push-pull system with an elliptical load line representing the most stringent conditions. It will be seen that the individual single-ended load line crosses the 1000-W dissipation curve and to allow a reasonable design margin the power dissipation of the output stage must be increased to nearly 1500 W.

2.2. *Choice of Transistors and General Configuration*

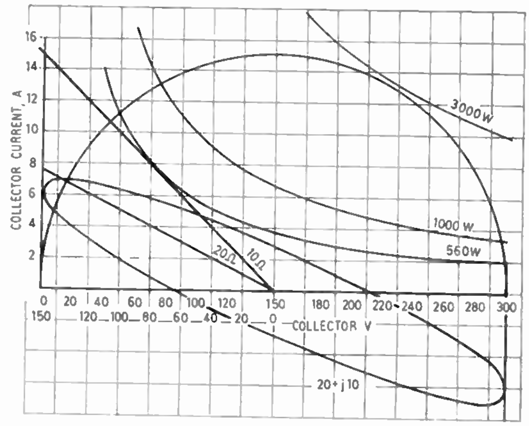
The diagram of Fig. 2 illustrates the typical push-pull transformerless output stage using transistors which was the basic circuit used for this application.

Taking into consideration the points referred to in Sect. 2.1, each transistor in the diagram must meet the following conditions:

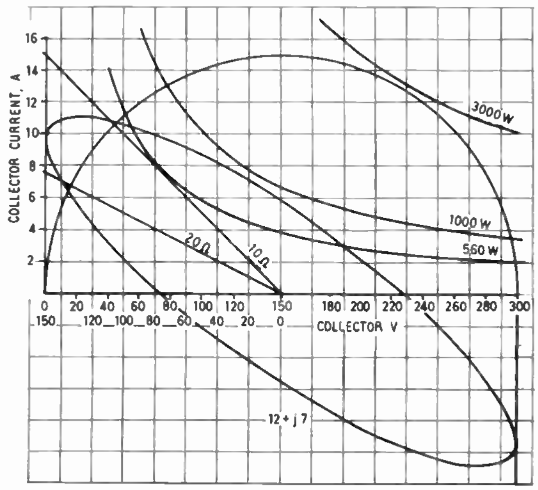
(1) Have a maximum power dissipation of 1500 W.

(2) Handle a peak–peak collector voltage of 300 V.

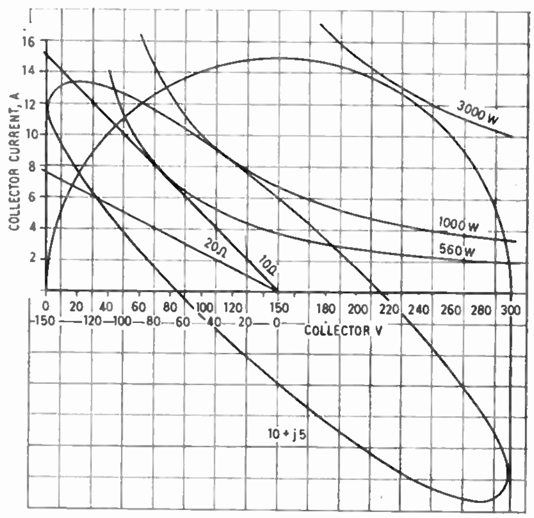
(3) Deliver a peak current of 15 A.



(*a*) Elliptical load line 20 + j10



(*b*) Elliptical load line 12 + j7



(*c*) Elliptical load line 10 + j5

Fig. 3. Load lines and power dissipation curves for theoretical class 'B' operation.

(4) Have a frequency response flat from 2 c/s to 5 kc/s.

These four design criteria determine the design of the output stage and the transistors to be used.

The choice of a suitable transistor to fulfil the requirements was governed not only by its characteristic but by the cost of a suitable device. The silicon transistor would meet the specification more fully than a germanium one, but the cost becomes prohibitive at the present time. Even with germanium transistors, the prices of the different types vary considerably and again a compromise must be realized between this and the specification. The OC 28 offered the best compromise at the time of the investigation and was used for the pilot model. This transistor has a maximum dissipation of 30 W and a maximum collector voltage rating of 60 V.

A total of fifty of these transistors are required per side of the push-pull system, to give the required 1500-W dissipation. They may be arranged in parallel, series or any combination of series/parallel providing the voltage and current supplied by each transistor has been shared to within reasonably close limits, i.e. ± 5%.

2.3. *Maximum Voltage Rating*[4]

The final transistor amplifier is required to provide a peak–peak output voltage of 300 Vs. In other words, the transistor must withstand the full 300 V under low current conditions (Fig. 4). The OC 28 is only capable of withstanding 60 V between collector and emitter. A minimum of five could, therefore, be connected in series, but in practice to obtain reliable operation and to allow a margin for voltage-sharing, six transistors were used.

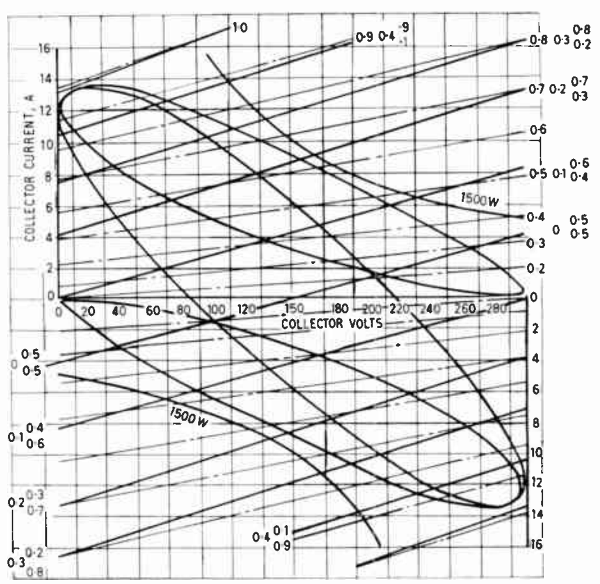


Fig. 4. Composite load lines and power dissipation curves class 'B' operation.

The optimum configuration to provide the collector dissipation and the voltage swing was a series/parallel arrangement of six blocks in series, each block composed of eight transistors connected in parallel. Obviously other combinations are possible (i.e. less in parallel, more in series), but it is simpler to connect transistors in parallel than in series.

The maximum current requirement is not a serious problem, since the parallel connection of eight transistors give a greater current capacity than is required.

The *required frequency response* is readily achieved. The typical 'cut-off' frequency of these transistors is approximately 10 kc/s which, although adequate to give the required frequency response, may cause an increase of the distortion at high frequencies.

3. Circuit Description

The complete amplifier is built up from a number of basic units (Fig. 5). This method of construction facilitates the addition of units to extend the voltage and power output. The basic units are the power stages and their associated driving amplifiers.

An output stage consists of eight power transistors connected in parallel and mounted on a water-cooled copper plate. An efficient cooling system is essential to utilize fully the maximum power dissipation of the transistors. For the same reason it is also necessary to ensure that each transistor dissipates the same power and this is achieved by including resistance in the emitter circuit of each transistor to improve current sharing. Any degree of current sharing may be achieved at the expense of loss of power. For this application a resistance value is chosen to equalize the current to better than ± 5%.

Twelve of these basic units are connected in series to give a typical push-pull arrangement (Fig. 5).

The voltage appearing across each series element must be equalized not only for the d.c. working conditions but also for the a.c. signal conditions from 2 c/s to 5 c/s. At frequencies above this range concern is only with switching transients which will be discussed later (Sect. 4).

To achieve the required degree of equalization the outputs from each element must be essentially a constant voltage source (i.e. low impedance). Voltage negative feed-back, therefore, is applied from the output, to the input of the driver amplifier A2. The d.c. and a.c. voltage gain from the points AA′ to the output is defined by the resistances of R1 and R2 to within 5%.

Another feature of this series arrangement, when using negative feed-back to equalize each unit, is the

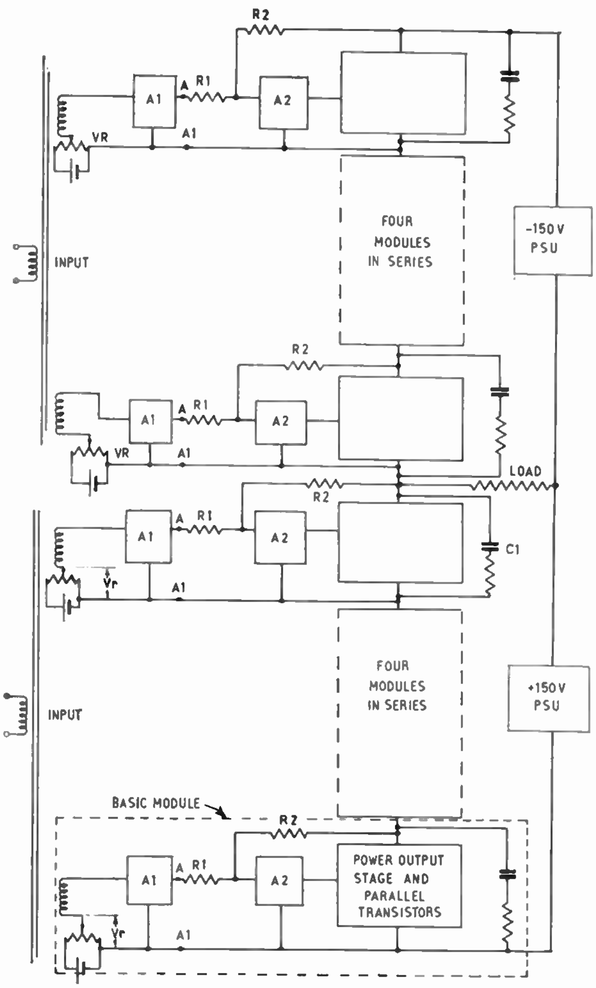


Fig. 5. Complete amplifier.

need to maintain an adequate phase margin to prevent high-frequency loop oscillations. The dynamic load of any one of the output stages consists of the a.c. impedance of one or more of the other output units. This impedance is complex and will tend to increase at high frequencies so increasing the loop gain of the stage, thus creating an unstable condition. To overcome this problem it is necessary to introduce a suitable low pass filter C1 R3 across each output unit to maintain a constant output impedance.

The third basic unit is the amplifier A1 which is an impedance matching amplifier with a low output and a high input impedance. The latter is necessary to facilitate the design of the transformer to give the required low frequency response.

The accuracy of the voltage equalization is determined by the stability of the predetermined d.c. working conditions and the a.c. gain from the input of the transformer to the output of each element relative to each other element. The voltage across each element is set up by the adjustment of the reference voltage V_R which will also control the standing current of the output stages. Variation with time of the output voltage is determined by the drift associated with the input stage, the gain of the amplifiers and the stability of the reference voltage. The circuit is designed to minimize these variations.

The a.c. gain of each amplifier is accurately controlled by negative feed-back to give an overall gain stability, including the transformer, of better than $\pm 5\%$.

4. Transient Conditions

When the h.t. is applied to the series units the voltage distribution will be quite arbitrary for the recovery time of the amplifiers. During this period avalanche may take place, resulting in the destruction of the transistors. To prevent this occurring a current limiting resistance is connected in series with the output transistor and is short-circuited after the amplifiers have settled down.

5. Power Supplies

The two main power supplies are each designed to deliver the required 150 V at a peak current of 15 A. The essential requirements of these power supplies are good regulation, which must be better than $\pm 5\%$, and low ripple output. This is achieved by using a twelve-phase rectifier configuration with a suitable three-phase transformer. The output ripple is less than 6 V peak–peak, i.e. $\pm 2\%$.

6. Amplifier Performance

Figure 6 gives the typical distortion achieved with the amplifier when driving either a resistive load or the complex vibrator load. Under these conditions total distortion remains below 1% over the whole frequency range. The increased distortion at high frequencies is caused by the decrease of the loop gain at these frequencies.

The collector efficiency of the output stage, when driving the 10Ω resistive load at the maximum output of 1 kW, is 42%. In all cases the maximum power output, or the maximum VA output for reactive

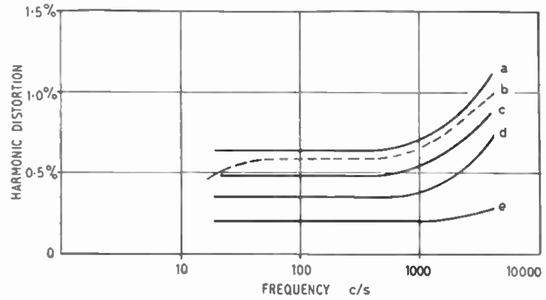


Fig. 6. Typical distortion characteristics.

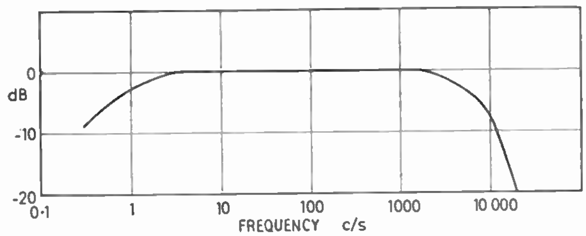


Fig. 7. Frequency response characteristic of the amplifier.

loads, is determined by the maximum voltage excursion of $\pm$ 150 V, and the impedance or resistance of the load at a particular frequency.

The frequency response of the amplifier which is shown in Fig. 7 is sensibly flat from 2 c/s–4 kc/s. The low-frequency droop is determined by the input transformer and may be extended with an alternative design. On the other hand, the upper frequency limit results from considerations of the loop stability of the output stages.

7. Conclusions

The paper has dealt specifically with the design of a power amplifier which is capable of delivering 1 kW into a resistive load of 10Ω. It has been shown that, with the use of suitable electronic circuits, voltage sharing across a number of series banks of transistors may be readily achieved.

To increase the power output or VA rating of an amplifier of this nature either more parallel or series units may be added. The relative numbers of each will be determined by the impedance of the load.

In order to drive the 75 kVA vibrator it would appear that the optimum basic amplifier would be designed to deliver an output of 10 kVA at the correct voltage level. Eight units may be added in parallel to give the maximum output. The basic 10 kVA amplifier may be used to drive smaller tables requiring an output of this order of magnitude.

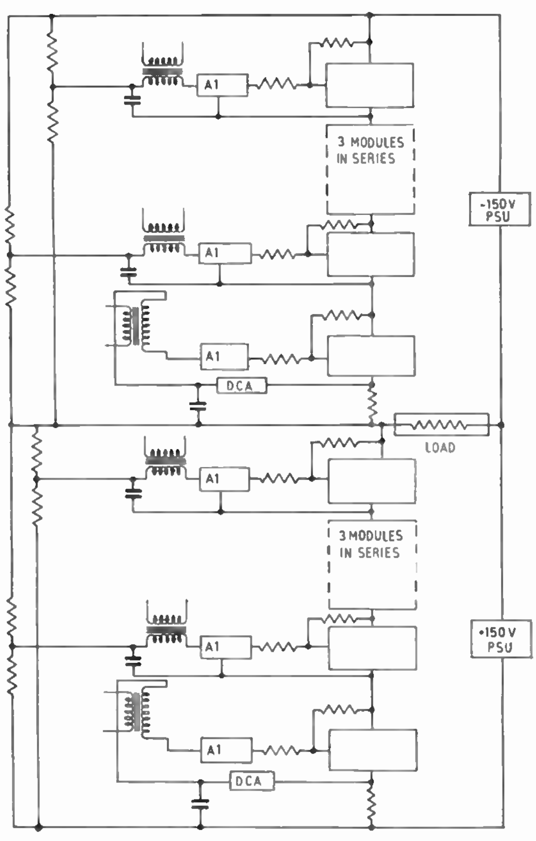


Fig. 8. Amplifier circuit showing a method of d.c. stabilization.

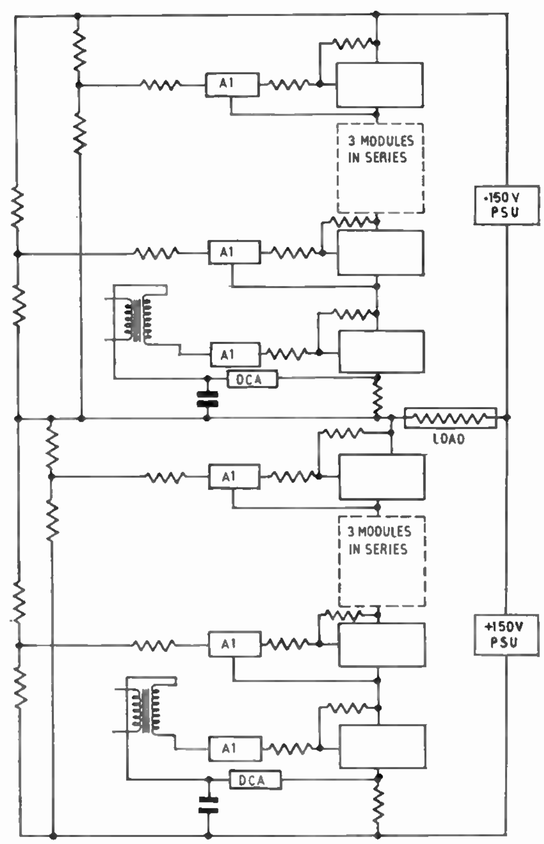


Fig. 9. Development of Fig. 8 eliminating the driving transformers from all except two units.

There are a number of additional points which are not mentioned in this paper, but which would require particular attention in the higher power amplifier. Care in the design of the driving stage to ensure adequate d.c. stability is most important and this is particularly so because of the class A working conditions. An elegant method to ensure good stability is indicated in Fig. 8. The overall d.c. voltage feedback is arranged so that the voltage across each unit is compared with, and controlled to equal, the voltage across the resistance divider. Direct current feed-back is applied to a single unit and will maintain the standing current to predescribed limits. A further development of this arrangement would be the elimination of the driving transformer to all units with the exception of two (Fig. 9). Whether or not this method would achieve the accurate voltage sharing at the higher frequencies could only be determined by further investigation.

Although class A operation has many advantages, such as low distortion and the relative simplicity of the voltage sharing, the one major drawback is the collector dissipation and the resulting cooling problems. The theoretical maximum efficiency is 50% and therefore under no-signal conditions the collector dissipation would be twice the maximum power output required. The cooling system will be required to remove this quantity of heat without allowing the temperature to rise above 40°C.

A mode of operation approaching class B would alleviate this problem, but the percentage distortion of the output would increase. Nevertheless, this may well remain within tolerable limits and by far the greater problem would be ensuring adequate voltage sharing between units with this mode of operation. Again this would require further investigation to determine the feasibility of this system.

8. Acknowledgments

The author acknowledges the assistance of a number of colleagues, particularly Messrs. J. N. Neilson and V. Ashton, who carried out much of the development and engineering design.

The paper is published by permission of the Director of the Atomic Weapons Research Establishment.

9. References

1. R. C. Lewis and D. J. Fritch, "The electrical characteristics of shaker systems used for vibration testing of equipments," *Trans. Inst. Radio Engrs* (*Audio Engineering*) No. PAGE 2, p. 1, December 1961.
2. F. Langford Smith, "Radio Designers' Handbook," (Iliffe, London, 4th edit. 1953).
3. R. A. Hilbourne and D. D. Jones, "Transistor power amplifiers," *Proc. Instn Elect. Engrs*, **102**, Part B, No. 6, November 1955 (I.E.E. Paper No. 1861 R, April 1955).
4. L. P. Hunter, "Handbook of Semiconductor Electronics," (McGraw-Hill, New York, 1956).

Manuscript first received by the Institution on 22nd September 1962 and in final form on 12th February 1963. (Paper No. 843/EA9.)

U.D.C. 621–52

Hybrid Digital/Analogue Servomechanisms

By

G. B. KENT, B.Sc.(Eng.)
(*Associate Member*)†

Presented at the Convention on "Electronics and Productivity" in Southampton on 18th April 1963.

Summary: The paper is an introduction to hybrid digital/analogue servo systems utilizing the application of digital techniques to conventional closed-loop servo systems. The term hybrid is used to describe systems where the error signal is derived digitally and then converted into an analogue signal for the operation of the power source. This results in an improved system accuracy compared with that obtainable using conventional analogue methods, and avoids many of the difficulties associated with the pure digital approach.

An introduction consisting of a comparison of the salient features of analogue and hybrid servo systems provides a background against which the underlying principles and design features of both velocity and position servo systems are discussed. The constituent elements of each system are derived in block diagram form with parallel references to their analogue counterparts. The paper concludes with an assessment of the advantages of hybrid servomechanisms and a brief review of the various fields of application of such systems.

1. Introduction

In a conventional servo system as typified in the simple closed loop system the error signal providing the input to the servo amplifier results from the difference between the input and feedback signals and may be positive or negative depending on their magnitudes (Fig. 1). The error signal may provide the input to a servo amplifier driving the motor. The amplifier gain is generally quite high and is dependent on the servo stiffness required. At equilibrium the input demand signal, proportional in this case to the desired speed, is nearly equal in magnitude to the feedback signal which is proportional to actual speed. The error signal is therefore almost zero and approaches zero as the amplifier gain approaches infinity. A practical system with a high gain amplifier and a corresponding small error signal is relatively insensitive to drift introduced into the system after the derivation of the error signal. Obviously the drift introduced prior to the subtraction process results in a velocity error.

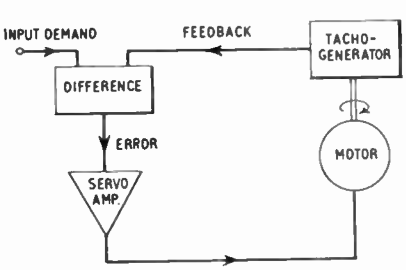


Fig. 1. Simple velocity servo system.

† Newman Electronics, Yate, Bristol.

A hybrid servo system of the type under discussion utilizes digital methods to undertake the subtraction process so producing a digital error signal which, after a digital to analogue conversion, forms the input to the servo amplifier as in a conventional servo system. The resolution of such a servo depends on the capacity of the differential element.

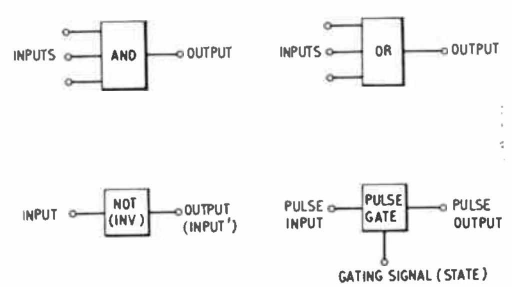


Fig. 2. Basic logical elements.

The digital error-producing elements used depend on the nature of the input and feedback signals, the type of servo (e.g. velocity or position), the nature of the application and the desired response. The concept of basic logical elements is briefly introduced and arrangements of these are shown to form the larger system elements as indicated in the block diagrams. Hybrid servomechanisms are not limited by the dictates of any particular circuit approach, and indeed, they need not be electronic. These systems are discussed in terms of the philosophy of basic logic elements.

The AND and OR gates usually refer to the AND-ing and OR-ing of states, for example, presence or absence of a voltage. The NOT function implies the inverse and may therefore be shown as an inverter stage. It is necessary for some operations to control the transmission of 'pulses' as distinct from 'states' in order to operate bi-stable elements. The control signals are usually 'states' and the element performing this function is termed a PULSE gate (Fig. 2). Thus the presence or absence of a gating voltage on one terminal of the element controls the transmission of pulses to the following stage, usually a bi-stable circuit thereby defining the mode of operation. Thus a series of bi-stable elements can be interconnected to form a bi-directional counter as used in some hybrid servo-mechanisms (Fig. 3). Other digital elements such as shift registers may be similarly arranged.

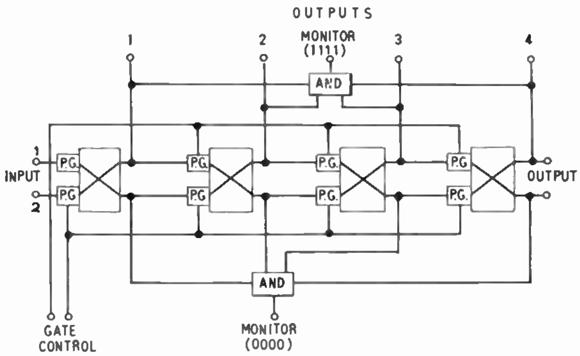


Fig. 3. Bi-directional counter.

The more complicated system elements, which will be described later, are arrangements of the above basic logic elements and use transistor circuits, but other techniques, such as magnetic core methods, could be used.

2. Velocity Servo System

The velocity servo system has an input demand signal, in the form of a serial pulse rate or frequency, and a feedback signal derived from a magnetic transducer, operating on the 'phonic' disc principle, from the output shaft (Fig. 4). The input and feedback signals are gated into a bi-directional binary counter in such a way that the magnitude of the count is increased by the input and reduced by the feedback. The magnitude of the binary number in the counter is therefore proportional to the difference of these signals and provides the error signal directly. This error signal is then converted into a d.c. voltage by means of a set of summing resistors connected in a binary progression to each counter stage. A reverse d.c. voltage may also be required as a 'set zero' control and is added to the input signal. Having achieved the digital to analogue conversion in this manner the system follows that of a conventional d.c. servo, the power amplification being provided by a transistor or valve power amplifier or a transductor.

The input signals to the gate circuit must not be permitted to operate the counter directly since the effect of coincident pulses on each input would not be reliably interpreted by the counter. These input signals are therefore gated alternately in the appropriate direction into the counter. Consequently the gating frequency chosen must be appreciably higher than the maximum information rate. Furthermore the gating unit must have a digit store on both demand and feedback channels so that information is not lost should a pulse arrive on one channel whilst the counter is accepting information from the other channel. The phasing of the gating unit and the direction of count of the counter is achieved by means of a four phase oscillator providing four outputs in quadrature. The system is therefore independent of the relative phases of the demand and feedback signals, since these signals are stored until the counter is appropriately phased to accept and clear the stored information.

If the instantaneous error signal in a conventional d.c. servo is sufficiently large to saturate the servo amplifier the system limits to an extent corresponding to either of the maximum limits of applied torque. If the hybrid system is examined in this light it is evident that if the counter exceeds full scale in a positive direction, or zero in a negative direction, the error signal appears at the input to the servo amplifier as a saw-tooth waveform. To avoid this problem additional control signals, derived when the counter is either full or empty, are used to control the gating units so that the appropriate signal is inhibited. Loss of information now occurs but this is analogous to the conditions obtaining when a d.c. servo is saturated.

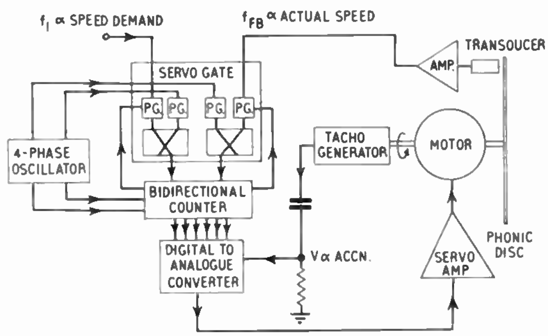


Fig. 4. Velocity servo system.

The resolution required determines the number of counter stages adopted for a particular system, and whether the system is reversible or unidirectional. The half scale count is made to correspond to the centre of the velocity scale. Thus a constant speed servo would find equilibrium with the counter half full whereas this would correspond to zero speed for a reversible zero. A reversible servo would also require a second feedback transducer, or an alternative method of defining the direction of rotation, and a modified gating unit.

In addition to the velocity error signal the derivative of this is generally required to maintain stability and prevent 'hunting'. This derivative may be obtained either by differentiating a tacho-generator signal or directly from a 'drag-cup' generator with d.c. excitation. It is generally more economic to generate the derivative in an analogue manner since to effect this digitally would require a much higher frequency for a given servo system. This derivative is proportional to the angular acceleration and is fed as a d.c. signal into the servo amplifier. Hence it does not require digitizing. The normal techniques of error signal shaping can therefore be readily applied if required.

3. Position Servo System 1

The first position servo system to be examined has a slow response and is very similar to the previous velocity system but the demand and feedback signals are in the form of a magnitude of count from a datum. The block diagram of the system is shown in Fig. 5.

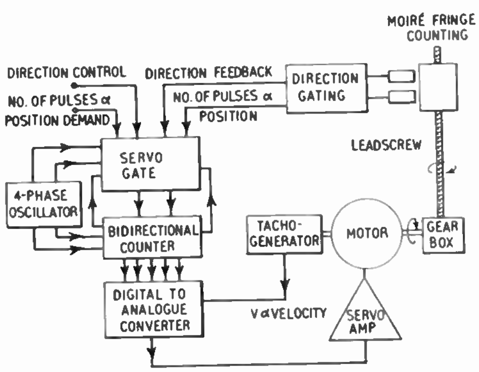


Fig. 5. Position servo system 1.

The arrangement of the counter and its gating control is similar to that of the velocity system and the form of position feedback shown in this example is that derived from moiré fringe counting. By the use of four photocells both the count and direction of travel can be generated. The capacity of the counter is defined by either the maximum value of the count corresponding to full scale or the maximum error possible. In the former case the positional information could be completely stored in the system before the servo is permitted to move. Stability in this case is obtained from a tacho-generator providing a velocity signal directly into the servo amplifier. The servo amplifier may drive a motor directly or operate a hydraulic or pneumatic servo valve for use with larger systems.

Such a system can be used on a co-ordinate positioning device for machine tool applications in which case the magnitude of count is derived from a counter/shift register. This enables the basic information to be inserted in either a serial code, for instance from a digital computer, or alternatively in a parallel code as derived from a set of control switches or a punched-card reader. When the counter/shift register has been 'loaded' the register control is arranged for counting. The set-up number is counted out by means of an oscillator signal fed into the digital servo system. In order to prevent any loss of information, if the bi-directional counter becomes full, the 'counter full' signal closes the gate in the signal path from the oscillator rather than inhibiting the servo gating unit. The system can be extended to cover more than one axis and control signals, confirming that the positional servos are balanced, are readily obtained since at equilibrium the bi-directional counter will be half full.

4. Position Servo System 2

Whilst combining accuracy with simplicity the previous system demands that the input information should be in the form of a discrete number of serial pulses and this may not always be readily obtainable. A more sophisticated system which overcomes this difficulty is that based on a scheme proposed by Porter and Stoneman† in which the input is in the form of a parallel binary code as obtainable from a digital computer. The feedback is obtained from a shaft encoder or digitizer in a cyclic progressive binary or Gray code. In order to study such a system an experimental comparator unit called a 'digital subtractor' was made. Since the system was designed for 360 deg positioning, minimum sector discrimination was essential. This implies that the direction in which the servo is caused to move is such that the angle described is a minimum. This feature arises from the code adopted; zero and full scale angles are the same. A digit 1 in the most significant position is used to define a negative number. Thus equal angles each side of the datum position correspond to the binary number and its complement. Furthermore the system will provide an integral of error in addition to the error signal into the servo amplifier so reducing the dynamic following error when heavy external servo loads are applied. The integral error signal achieves this result without increasing the natural frequency of the system. A fast response can be achieved which permits several such systems to be run sequentially from one computer

† A. Porter and F. W. Stoneman, "A new approach to the design of pulse monitored servo systems", *Proc. Instn Elect. Engrs*, Part II, **97**, No. 59, p. 597–611, October 1950.

producing demand signals at discrete intervals of time. By the adoption of a high scanning rate relative to the servo response time each servo system suffers no degradation from the discontinuity of the information.

4.1. *Principle of Operation*

The block diagram of the complete system is shown in Fig. 6. The prototype system used a word length of 10 bits thus achieving a resolution of approximately 1 part in 10^3. Increasing the word length would enable the resolution to be improved.

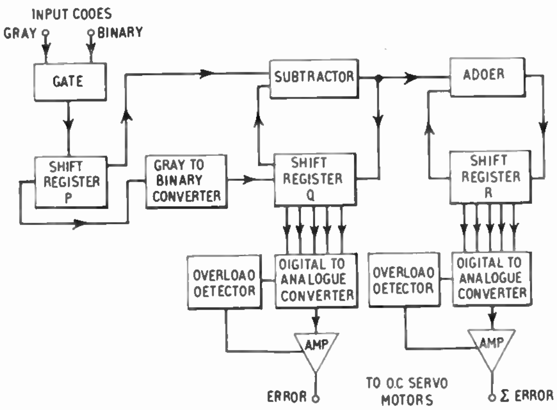


Fig. 6. Position servo system 2.

A parallel cyclic progressive binary code is presented by the input digitizer to the shift register P via the gate. At a specified instant of time the computer presents a parallel pure binary code, representing the required position, to the input—B and passes a 'ready' pulse to the digital servo control (omitted for clarity). This gates the cyclic progressive binary word into shift register P. Having set up this register it is then unloaded serially via the Gray/binary converter; the resulting pure binary signal is loaded serially into shift register Q. Features of the Gray/binary conversion will be described later. The control then gates the computer pure binary signal into shift register P. Shift registers P and Q are then both serially unloaded whilst the signals are gated via a subtractor. The resultant serial difference word is fed back into shift register Q behind the outgoing word. Shift register Q now holds the difference or error signal resulting from subtracting the input demand and feedback positions. A digital/analogue converter consists of a binarally arranged set of summing resistors which add the binary signals from each stage of the shift register into a d.c. amplifier, providing a d.c. output voltage. This voltage can be readily handled by standard analogue servo techniques. During the subtraction process shift register Q momentarily stores the position feedback signal, but this is of such a short duration that it does not significantly affect the error signal output.

The integral error signal is derived digitally by successive summation of the error signal to the previous summation of error from earlier computing cycles. This summation must be sign-conscious and the presence or absence of the digit 1 in the most significant position achieves this. A digital/analogue converter connected to shift register R enables a d.c. output voltage to be obtained. The converters used on shift registers Q and R are identical. In the event of a digitally derived signal being large corresponding to a saturation level in an analogous d.c. servo system, a limiting voltage is fed out from the overload detector to over-ride the digital/analogue converter.

The overload detector defines whether or not the digital number in the shift register is greater than its preset level e.g. $+31$ bits or -32 bits. Thus if any of the five most significant digits in the ten stage shift register contains a 1 with the remainder at 0 then the number is in excess of 31 and a clamping voltage is transmitted. Similarly the complement of five 1's corresponding to -32 and the appearance of a 0 constitutes an overload. This function is achieved by means of two AND gates feeding into an OR gate.

Having detected an overload it is necessary to determine if it is positive or negative so that the correct d.c. clamping voltage may be applied. By earlier definition a negative number is indicated by the presence of a 1 in the most significant digit so that the clamping voltage polarity operates from the presence or absence of the most significant digit.

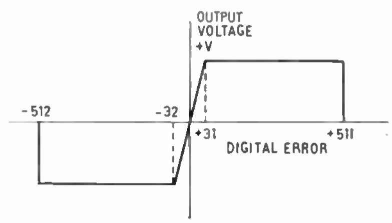


Fig. 7. Error signal characteristic.

The graph (Fig. 7) shows the analogue output voltage equivalent to the digital error. For the error output the scaling is as shown whereas the summated error has a saturation level corresponding to ± 127 bits. The linear portion through the origin consists of small steps, each step being the voltage increment of a least significant digit. It is necessary to reset the cumulative error in shift register R prior to the commencement of a servo run. In order to avoid a large initial overshoot the connection of the integral error signal into the system may be delayed.

To return to the principle used in the Gray/binary code conversion, a feature of the present instrument

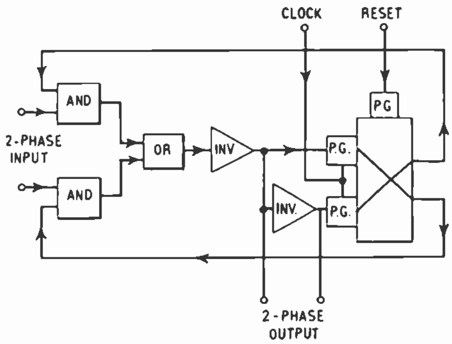


Fig. 8. Gray/binary converter.

makes use of the fact that if the code conversion is operated serially with the most significant digit leading, i.e. opposite to the conventional practice, then the converter simplifies to the circuit shown in Fig. 8. Thus it is necessary to unload the most significant digit of shift register P through the Gray/binary converter, but during the subtraction process, shift registers P and Q must both operate conventionally, i.e. least significant digit first. This is achieved by the use of bi-directional shift registers. The alternative method of conversion which can operate with the least significant digit leading requires a memory capable of storing a full length word, since only on the receipt of the most significant digit can conversion commence. This will become evident by stating the conversion rule:

> To convert from Gray code to pure binary code commencing with the most significant digit, copy the first digit; repeat if the next digit is a 0 but change if the next digit is a 1.

The following table gives some conversion examples:

Gray Code	Binary Code
0000	0000
1111	1010
0011	0010

5. Conclusions

The first two systems although much less complex than the third system described are capable of quite a high resolution although their dynamic responses are inferior. Nevertheless a considerable field of application is found where the rate of change of input demand is low. For example, the velocity servo system of Fig. 4 is ideally suited to a system requiring a preset velocity to be accurately maintained and where the fluctuations in load are not excessive. The speed setting is conveniently arranged by controlling the frequency of a stable oscillator. Two or more such servo systems can therefore be arranged to run in synchronism even when displaced remotely, so that a synchronous link is effected. The velocity ratios need not be unity and by the insertion of binary divider stages the equivalent of an 'electronic gearbox' can be effected. Figure 9 shows such a system where 11 speed ratios from 1 : 1 to 32 : 1 can be achieved.

The simple position control system finds many applications in accurate servo positioning as met in the machine tool and steelmaking industries. The example described of a co-ordinate positioning device is typical of such an application where a fast response may be unnecessary.

The volume occupied by the necessary circuit components can be much less than that required for a conventional d.c. servo circuit of the same resolution. Many miniaturizing techniques applicable to transistor digital equipments can be utilized. Furthermore if several such systems are required to be combined the gating oscillator may be common thereby effecting economy.

The third system typifies the approach to the larger fast response position servo systems as met in many industrial and military applications. Since this system utilizes information at discrete intervals, time sharing of a central computer, providing information for a set of such servo systems, is an attractive proposition. This permits the individual systems to be readily correlated and run asynchronously or synchronously as required.

The hybrid digital servo-mechanism has been applied in a variety of fields of which a few have been mentioned and is proving to be an invaluable adjunct to the technique of control engineering. The economics of a high resolution hybrid servo system are very favourable in comparison with those of its all analogue counterpart. Analogue systems with resolutions better than 0·5% are not normally obtainable without excessive complication and cost whereas a 0·1% resolution hybrid servo system can be readily achieved.

6. Acknowledgments

The author wishes to thank the Directors of Newman Industries Limited for permission to publish this paper and also to thank the Admiralty for permission to describe the digital subtractor made for them.

Manuscript first received by the Institution on 6th March 1962 and in final form on 14th March 1963. (Paper No. 844.)

(For Discussion on this paper turn to next page)

DISCUSSION

Under the chairmanship of Professor A. D. Booth

Mr. P. Wood: Referring to the velocity servo system (Fig. 4) the servo shown is not directly comparable with the simple case shown in Fig. 1 since the bi-directional counter integrates the error signal rather than provide an error signal directly proportional to the speed error. A frequency difference of $f_{FB} - f_{I}$ will be applied to the counter and it is in fact the rate of change of the counter that is proportional to the speed error.

The author also states that the counter will stabilize at the mid-position when the motor is running at the correct speed. In fact under steady state conditions the contents of the counter will be at such a level that, when converted into an analogue form, the motor will be running at the correct speed, and will therefore depend upon the load and speed requirements.

The Author (*in reply*): It is agreed that an integration of the error signal is involved in the hybrid system described. The verbal comparison with the simple analogue velocity system was used to assist the explanation of the hybrid. The resultant phase lag of the error signal must be borne in mind in the design of such a system.

Referring to the second part of the question the counter half-full refers to the 'design centre' for a constant speed servo working on normal load. Variation in load demand causes the counter average signal to adjust itself accordingly.

Mr. J. J. Hunter: Digital servomechanisms for speed control can achieve very high accuracies. It is necessary that the differential frequency comparator part of these systems can handle coincident pulse inputs on the add and subtract lines. A simple solution to this problem is to detect and cancel overlapping pulses before the counting stages. The method adopted in this equipment seems relatively complicated as it requires an auxiliary oscillator.†

The Author (*in reply*): The method of cancelling coincident pulses mentioned is an alternative to the sequential gating of the demand and feedback signals as described. I would not care to comment on the relative complications without studying the alternative circuitry.

† W. H. P. Leslie, "Precise control of shaft speed". *Electrical Energy*, **1**, p. 1, September 1956.

W. H. P. Leslie and D. Nairn, "A fast counter for adding and subtracting". *Electronic Engineering*, **34**, No. 410, pp. 227–33, April 1962.

The N.I.R.N.S. 7 GeV Proton Synchrotron

The National Institute for Research in Nuclear Science has announced the first operation of the 7 GeV proton synchrotron *Nimrod*, built at the Rutherford High Energy Laboratory, Chilton, Berkshire. After three weeks of intense commissioning tests, the first fully accelerated beam was observed at 5.20 p.m. on 27th August. The energy achieved at that stage was 6·5 GeV with an intensity of 4×10^9 protons per pulse. Within an hour, an energy of 8 GeV had been achieved, 1 GeV above the design energy, following which operation was continued at 7 GeV with intensities up to 10^{10} protons per pulse. Further development will be carried out during the next few months with the object of increasing intensity to 10^{12} protons per pulse, while at the same time preparations for the programme of high energy physics experiments on the machine will be intensified.

Nimrod which has cost approximately £11M, is the second accelerator to be operated by the N.I.R.N.S. This organization was formed in 1957 with the object of providing for common use by Universities and similar institutions large and costly equipment needed for fundamental research, which is beyond the scope of individual universities.† The main activity of the Institute has been in the field of nuclear and high energy physics where the main tool of research is the particle accelerator. Accelerators vary in complexity and size depending on the energy and intensity of the accelerated beam of the particle it is required to produce. The most powerful type of accelerator developed so far is that known as the proton synchrotron and *Nimrod* is of this class.

† *J. Brit.I.R.E.*, **18**, p. 174, March 1958.

The earliest design studies which led to *Nimrod* were started in 1955 by a group engaged on accelerator development at the Atomic Energy Research Establishment at Harwell. In view of the considerable university interest it was decided to transfer the project to the newly-formed N.I.R.N.S. and construction work began in August 1957 on a new Institute site adjacent to Harwell and subsequently named the Rutherford High Energy Laboratory. The U.K.A.E.A. retained close links with the project, being responsible for supervising construction on behalf of the Institute and supplying staff and services, although most of the staff involved transferred to Institute employment in 1961.

The main feature of *Nimrod* is a large ring-shaped electro-magnet, 155 ft in diameter and weighing 7000 tons. A toroidal-shaped evacuated chamber made from a glass-fibre reinforced epoxy resin is situated between the poles of this magnet. A pulse of protons given an initial acceleration of 15 MeV in a linear accelerator is injected into this chamber and the protons are forced by the magnetic field into a circular orbit in which they receive an acceleration from a radio-frequency electric field once in each revolution. When after approximately 10^6 revolutions the protons have reached their maximum energy, they are either extracted from the vacuum chamber or allowed to bombard internal targets and the resulting secondary particles channelled into an adjoining building where they will be used for experiments. During the acceleration period, lasting 0·72 seconds, the magnetic field strength and frequency of the electric accelerating field have both to be increased steadily to confine the proton orbits to the magnet ring, and in such a manner as to maintain the

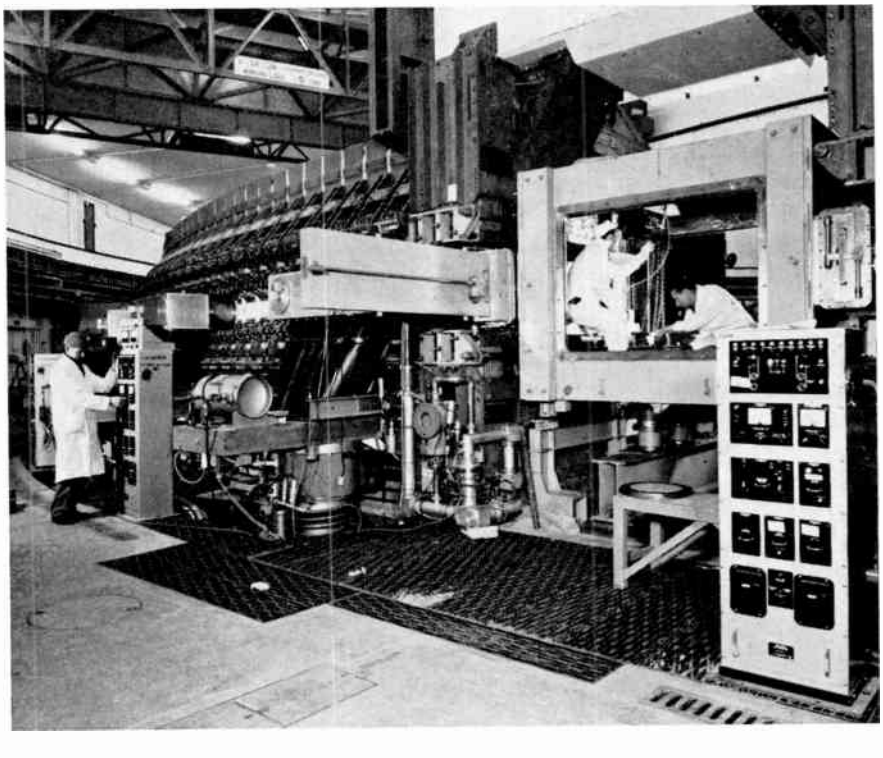


Fig. 1. A general view of one of the 'octants' of *Nimrod*, showing the installation of the accelerating cavity. A vacuum pump can be seen on the left.

Rutherford High Energy Laboratory photograph

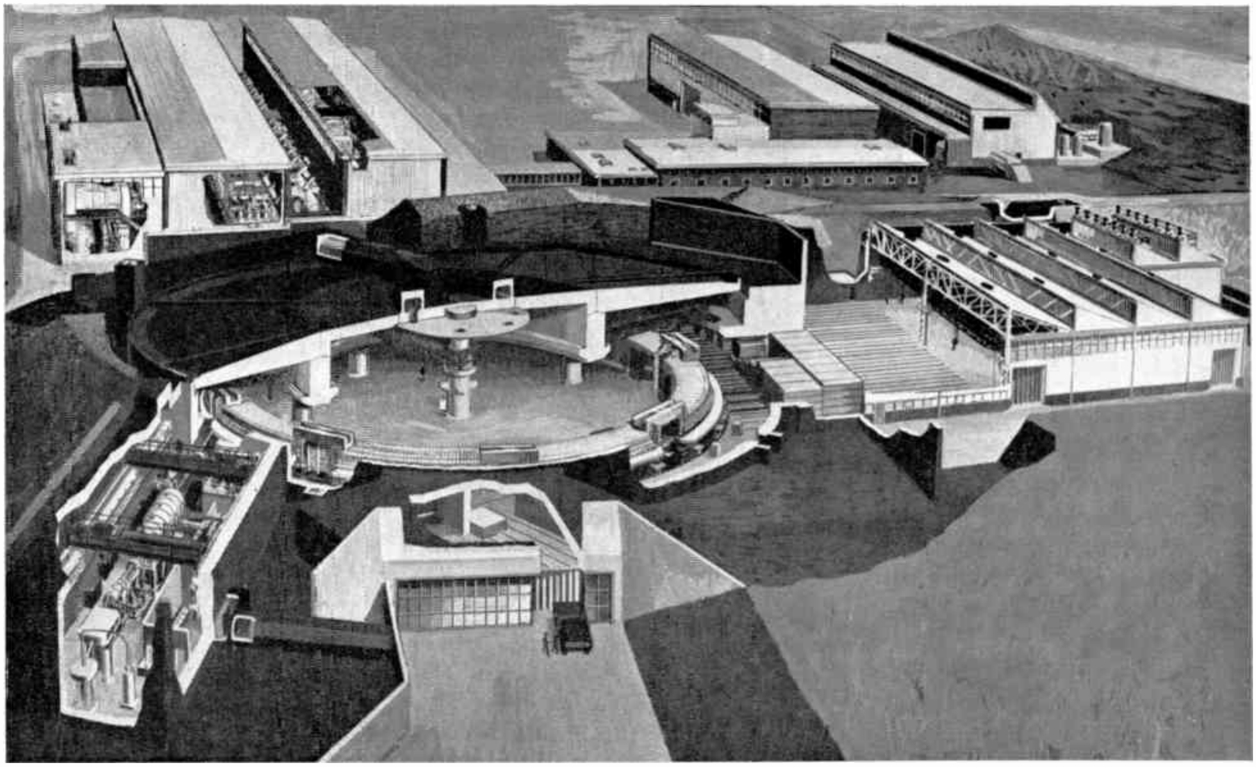


Fig. 2. An artist's impression of the 7 GeV proton synchrotron *Nimrod.*

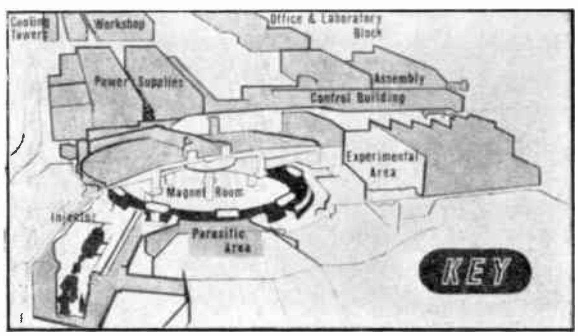
Rutherford High Energy Laboratory photograph.

delicately balanced stability in the motion of the protons.

The whole machine is housed in a semi-underground circular building of reinforced concrete 200 ft in diameter with 6 ft concrete roof on which a 20 ft layer of earth is placed as additional radiation shielding. The artist's impression above shows the extent of the installation.

Heavy currents up to 9000 amperes with an applied voltage up to 15 kV are needed to energize the electromagnet during the short acceleration time. The power requirements are intermittent since the magnetizing current is only required for the duration of the pulse while the protons are being accelerated, so some form of energy storage is required. This is provided by heavy flywheels incorporated in a motor alternator set with a combined rating of 100 MW, connected to the magnet through transformers and a bank of rectifiers. This equipment supplies direct current of gradually increasing strength during the pulse, followed by a period of decay. Energy is thus stored in the inductive windings of the magnet coils (made from 250 tons of extruded copper bar) during the current-rise period and is subsequently returned to the flywheels in the intervals when power is no longer required.

The machine is designed to produce at least 10^{12} protons per pulse at a repetition rate of 28 pulses a minute: this is equivalent to about 1/16th of a microampere which although small in absolute terms represents a very high current compared with those obtained from similar machines elsewhere. *Nimrod* will be used for fundamental research into the physics of the 30-odd elementary particles now known. The study of the modes of creation and decay (most of them are unstable) of these particles, the way in which they interact with other particles and the interpretation of such observations in terms of unifying theory is the object of present-day energy research.



This article is based on a communication to the Institution by the United Kingdom Atomic Energy Authority on behalf of the National Institute for Research in Nuclear Science.

U.D.C. 534.88: 621.396.677

A High-resolution Electronic Sector-Scanning Sonar

By

V. G. WELSBY, Ph.D.
(*Member*)†

AND

J. R. DUNN, B.Sc.†

Presented at the Symposium on "Sonar Systems" in Birmingham on 9th–11th July, 1962.

Summary: The success of sea-trials of an experimental pulsed sonar system using 'within-pulse' electronic scanning has led to the development of a new version using a considerably larger number of channels and a shorter wavelength in the water. The new system, which has 30 channels and works at a carrier frequency of 500 kc/s in the water, is designed for a maximum range of 100 yards and a scanned sector of ± 15 deg. Pulse lengths down to 100 μs can be used, corresponding to a range resolution of about 6 in. Facilities are provided to permit a rapid change-over from normal additive signal processing to a multiplicative process which is capable of approximately doubling the angular resolution of the system, at the expense of some reduction in effective maximum range. Initial trials have confirmed that the performance of the equipment meets the design specification.

1. Introduction

The sonar system to be described uses a pulsed sinusoidal carrier and electronic sector scanning, the rate of scan being normally chosen so that its periodic time is equal to the duration of the transmitted pulse. Although the idea of scanning the far-field by electronically deflecting the directional response of a fixed receiving array is well known, there are several ways in which the necessary time-varying delays can be introduced into the outputs of the transducer elements.

The particular method used was developed in the Electrical Engineering Department of the University of Birmingham and includes certain novel features. The success of sea-trials of an experimental model of the system led to the design of a new version using a considerably larger number of channels and a shorter wavelength in the water. One of the objects of this work was to confirm certain theoretical predictions about the maximum range resolution, bearing resolution and scanning rate which could be obtained with this kind of system.

A considerable amount of information about the theory has already been published in the Institution's *Journal*[1–5] and it is not proposed to give here a great deal of detail. What is presented refers to the higher order of performance expected of the new version and some details of its construction. A fuller description and further experimental results will be published separately later.

† Electrical Engineering Department, University of Birmingham.

2. Specification

The receiving transducer consists of an array of 30 lead-zirconate plates and the scanning system thus has 30 signal channels, one associated with each element of the transducer array. The 3-dB beamwidth of the receiver is nominally $\pm\frac{1}{2}$ deg, giving a useful scanned sector angle of ± 15 deg. There is a separate transmitting array, using nine elements identical with those of the receiver, but these are combined and fed from a single transmit channel. The array is curved to enable the transmitted power to be radiated practically uniformly over the ± 15 deg sector angle. The 'within-pulse' output of the transmitter is 200 W, corresponding to a radiated power of about 80 W into the water.

The carrier frequency is 500 kc/s in the water and pulse lengths down to 100 μs (i.e. scanning speeds of up to 10 000 per second) can be used, corresponding to a range resolution of about 6 in.

Facilities are provided to permit a rapid change-over from normal additive signal processing to a multiplicative process which is capable of approximately doubling the angular resolution of the system, at the expense of some reduction in the effective maximum range. The latter naturally depends on the type of target concerned but is nominally about 100 yards.

The output of the system is displayed on a television-type tube, with a special long-persistence screen. Horizontal deflections represent range and vertical deflections represent angular bearings relative to a line drawn perpendicular to the transducer face. In

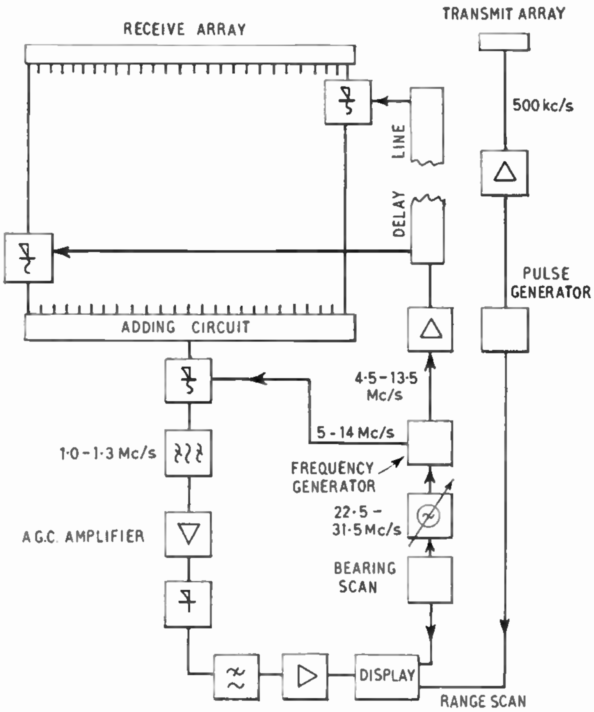


Fig. 1. Block schematic of the equipment set up for additive operation.

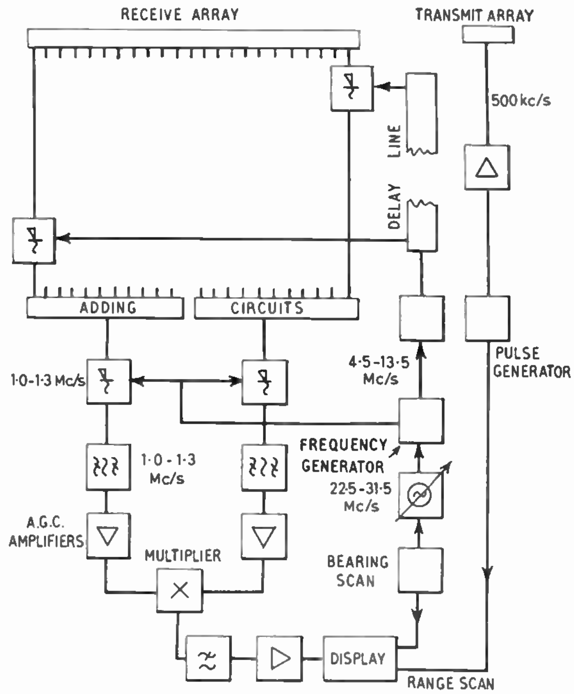


Fig. 2. Block schematic of the equipment set up for multiplicative operation.

addition to the display unit, there is a cubicle (5 ft × 2 ft × 2 ft overall) containing the delay line and channel modulators, a power-supply unit (3 ft × 2 ft × 1 ft overall) and additional standard rack-mounted panels occupying a total rack height of 5 ft. This particular set uses thermionic valves and no attempt has been made to restrict its size. Obviously the use of specialized miniaturization techniques would have greatly reduced both the overall dimensions and the power requirements but the application of these techniques is regarded as a separate problem.

The transmit and receive transducers, together with 30 2-stage transistor receiving pre-amplifiers, form a separate unit (12 in × 10 in × 7 in overall), which is connected to the rest of the system by an 80-ft flexible cable made up of twin coaxial leads with an outer metallic screen. The depth of water at which the transducer unit can operate is limited only by the length of the connecting cable. A streamlined fibre-glass housing has been provided, into which the transducer unit can be fitted. The housing is designed so that it can be towed by the ship, with the axis of the transducer array directed either vertically downwards or at some other pre-determined angle.

3. Principle of Operation

Consideration of the method of producing a series of time-varying delays can be simplified if the pulse envelope is ignored for the moment and each delay regarded merely as a phase-shift of the carrier waveform. The scanning process then requires time-varying phase-shifts which are obtained in the following way.

The local oscillator signal is frequency-modulated with a linear saw-tooth time-waveform and then fed to a delay line with tapping points at regular intervals along its length. The delay line produces a phase-shift which is proportional to frequency. Since the oscillator has a time-varying instantaneous frequency, it can be seen that, in principle, the various outputs at the tapping points along the delay line will have time-varying phase-shifts proportional to the respective lengths of line traversed. These outputs are used as the carrier supplies to frequency-changers, one in each signal channel. After the addition of the signal channel outputs has taken place, a process of 'remodulation' is used to translate the resultant frequency-modulated carrier back to a fixed frequency. It is explained in ref. 3 how the remodulation technique simplifies the design of the various filters in the circuit and helps to minimize unwanted phase-shifts in the channel paths.

3.1. *Multiplication*

An important feature of this sonar set is that, if desired, the outputs of the 30 receiver channels can be added together in two separate groups of 15 each; the resultants are then multiplied together and passed through a low-pass filter. The principles involved are discussed in ref. 1. It is shown that, not only is the accuracy of bearing determination for a single target doubled by multiplication, but also, provided a certain condition is met, it can be expected to provide a 'fineness of detail' of an unknown far field which is comparable with the best obtained with an additive array having twice the aperture length. The condition is that the phase relationships between the signals from targets on the same range must vary relatively rapidly with time, while the amplitude/bearing distribution of the targets changes only relatively slowly. This situation is one which is likely to be fulfilled in most practical applications.

4. Theoretical Limitations

Although, in the previous section, the presence of the pulse envelope was ignored, it does have to be taken into account in the design of the system. Since the time-shifts of the modulation envelopes of the signals in the various channels do not always coincide with the time-shifts of the corresponding carrier waveforms, it is easy to see that the modulation envelope of the resultant output after all the channel signals have been combined, will tend to be distorted.

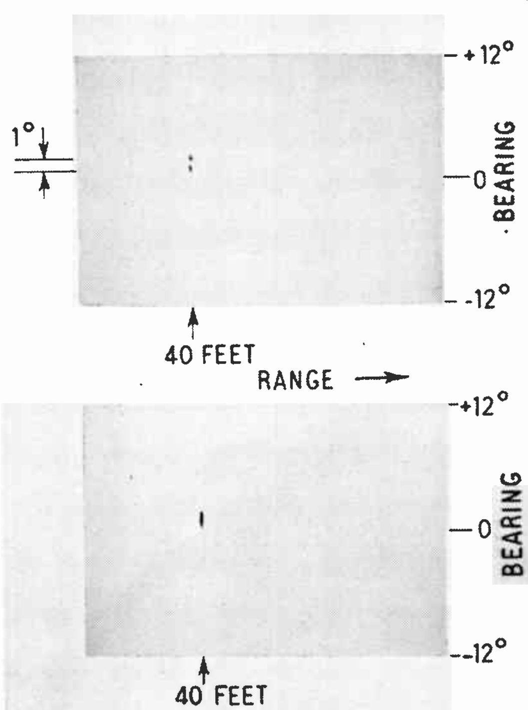
Another less obvious kind of distortion is inherent in the method used to obtain the time-varying carrier phase-shifts. It can be shown theoretically[2] that an unwanted phase-shift component is introduced which is proportional to the square of the length of delay line associated with a given channel, instead of being linearly related to the length of delay line. The relative magnitude of the resulting 'quadratic' distortion of the scanning pattern is proportional to the speed of scan. Although 'quadratic' distortion could be reduced by the insertion of suitable phase equalizers, these would have to be separately adjusted for a particular scanning rate and would lead to a considerable additional circuit complexity. The present system has been designed to work up to the practical limit, without compensation, and an acceptable amount of pattern distortion is therefore tolerated at the maximum scanning speed. A much more detailed discussion of the problem of distortion in electronic scanning systems will be found in ref. 2. Since one scan occurs within the duration of one transmitted pulse, the maximum scanning speed also determines the maximum range resolution.

5. Experimental Results

The main objective of the first sea-trial, carried out on board F.R.V. *Mara* in December 1962, was to confirm that the design specification had been met, particularly with respect to the angular resolution of multiple targets. Earlier attempts to do this at the University of Birmingham's underwater acoustic research station at Belvide Reservoir had been inconclusive owing to the exceptionally high reverberation level, caused by the presence of dense shoals of small fish in the vicinity of the transducers. Suitable low-reverberation conditions for this purpose were obtained in Loch Ness by using the transducer with its axis vertical. In this way surface reverberation was avoided and the test was made at a spot where the depth of water was sufficient to ensure that bottom echoes were negligible. Six-inch hollow plastic net-floats, attached to a suitable framework suspended below the ship, were used as improvised targets. The floats were attached to the frame by short lengths of line so that they floated clear but were held at the required spacing from one another.

It was shown (see Fig. 3) that, with multiplicative operation, two such floats could easily be distinguished at the same range when their centres subtended a bearing separation angle of one degree. The test was carried out with the targets at a range of 40 ft. The intention had been to repeat this test with various numbers of floats, various ranges and various angular separations. Unfortunately, prevailing weather con-

(*a*) MULTIPLICATIVE



(*b*) ADDITIVE

Fig. 3. Typical display, taken during resolution tests in Loch Ness (December 1962). The targets were two net floats, supported so that their angular separation was 1 deg at a range of 40 ft. Note that, as predicted theoretically, the targets were resolved by multiplicative processing but not by additive processing. (For the purpose of these tests the scale of the bearing display was expanded so that the sector angle displayed was 24 deg instead of the full 30 deg.)

ditions made it necessary to limit the time available for the tests in Loch Ness and the planned programme could not be completed.

A spot check of the range limit of the system, used as an echo-sounder in Loch Ness, showed that reliable bottom echoes could be displayed at depths down to about 80 fathoms.

Tests made in the canal basin at Crinan, with the axis of the transducer horizontal, indicated that a single 6-in net-float could be distinguished in the presence of a number of fixed scatterers (unidentified) on the bottom of the basin at a range of 40 yards, when the float was moving parallel to a vertical dock wall, and at a distance of about 2 yards from it. This result is of some significance in connection with the possible application of apparatus of this type to the display of fish near the bottom of the sea and lying some distance ahead of a ship.

Some fish observations were made, both with the transducer unit lowered over the side and under towing conditions. The appearance of the display screen confirmed that the fineness of detail of the picture was generally as expected.

More precisely controlled tests are planned to take place in the near future but there is already little doubt that this sonar set will prove to be a valuable tool for fishery work. The results of some further successful tests in a large sea-water tank are described in ref. 6.

6. Acknowledgments

The authors thank the Scottish Fisheries Laboratory at Aberdeen for valuable support and encouragement given to this project, the National Institute of Oceanography and the D.S.I.R. for financial support, and colleagues in the Electrical Engineering Department of the University of Birmingham for their advice and assistance.

7. References

1. V. G. Welsby, "Multiplicative receiving arrays—the angular resolution of targets in a sonar system with electronic scanning", *J. Brit.I.R.E.*, **22**, p. 5, July 1961.
2. D. E. N. Davies, "A fast electronically scanned radar receiving system", *J. Brit.I.R.E.*, **21**, p. 305, April 1961.
3. B. S. McCartney, "An improved electronic sector-scanning sonar receiver", *J. Brit.I.R.E.*, **22**, p. 481, June 1961.
4. D. G. Tucker, V. G. Welsby, L. Kay, M. J. Tucker, A. R. Stubbs and J. G. Henderson, "Underwater echo-ranging with electronic sector scanning: sea trials on R.R.S. *Discovery II*", *J. Brit.I.R.E.*, **19**, p. 681, November 1959.
5. D. G. Tucker, V. G. Welsby and R. Kendell, "Electronic sector scanning", *J. Brit.I.R.E.*, **18**, p. 465, August 1958.
6. D. G. Tucker and V. G. Welsby, "Sector-scanning sonar for fisheries purposes." (Presented at the World Fishing Gear Congress, London, May 1963.)

Manuscript received by the Institution on 21st June 1962. (Contribution No. 67/SS16).

POINTS FROM THE DISCUSSION

Professor M. Federici†: The equipment scans continuously for all bearings during one pulse duration. I would like to know if the signal for any given bearing is just sampled making input filtration necessary, or if it is utilized during all of its duration for any given bearing and if so, how?

† Polytecnico di Milano.

The authors (*in reply*)*:* The scanning process is equivalent to the sampling of the field in any particular direction. The scanning rate is directly related to the pulse duration τ, assuming that the receiving transducer circuits are restricted in bandwidth to $1/\tau$ c/s. The signals occupy a total bandwidth of n/τ c/s after the scanning process has taken place, where n is the number of channels.

U.D.C. 621.314.23

Square-loop Ferrites with Sinusoidal Magnetization

By

J. WOOD, B.Sc.†

Summary: The switching speed of square-loop ferrite cores is limited mainly by the restricted speed of domain wall movement, and not by eddy currents as in a metallic core. Their use in magnetic amplifiers and modulators allows the excitation frequency to be raised, giving greater bandwidth. This paper uses a simple theory of flux reversal in square-loop ferrites to predict switching speed with sinusoidal magnetizing force, and gives the results of experiments carried out to test this theory.

1. Introduction

The non-linear characteristics of metallic core materials have been used for many years in electrical engineering in applications such as magnetic modulators and magnetic amplifiers.[1, 2] The maximum useful excitation frequency which may be used with these materials is limited by eddy currents, which prevent rapid changes of flux in the core.

More recently, ferrite materials with square-loop *B-H* characteristics have been widely used in computer storage and logic. Cores made of this material may be switched rapidly between their magnetic saturation limits, since eddy currents in the core no longer impose a limit on the speed of switching. In magnetic modulators and amplifiers, large bandwidths are attainable only by using high excitation frequencies, thus where fast response and large bandwidth are essential, the use of a ferrite core material is preferable to a metal. In this paper the switching behaviour of square-loop ferrites with sinusoidal excitation is considered.

2. Operation with Low Carrier Frequencies

With low rates of change of magnetizing force, the behaviour of square-loop ferrites with a specified waveform of magnetizing force of sufficient amplitude to cause saturation may be predicted from the static *B-H* characteristics of the material. Figure 1 shows an idealized representation of these static characteristics and the ideal flux and voltage waveforms that are obtained with a triangular magnetizing force. (Because the peaks of this triangular waveform occur while the core is saturated, this wave-shape can often be used to represent, with little error, a sinusoidal m.m.f. having the same rate of change when crossing the zero axis). Rectangular voltage pulses are obtained as a consequence of the trapezoidal flux waveform. This simple analysis predicts that as the excitation frequency or amplitude is raised, there should be a proportional increase in amplitude and a decrease in the width of the pulses.

In practice, for frequencies below about 5 kc/s with peak magnetizing forces of about 1·5 H_C, and proportionately lower frequencies with higher magnetizing forces, the observed behaviour is in good agreement with this prediction. However, outside these limits, this is no longer true because the width of the pulse is then largely determined by the time taken for the magnetic domain walls to move in response to the excitation. The static *B-H* characteristics of the material and the magnetizing force waveform no longer provide enough information to enable the pulse width to be found.

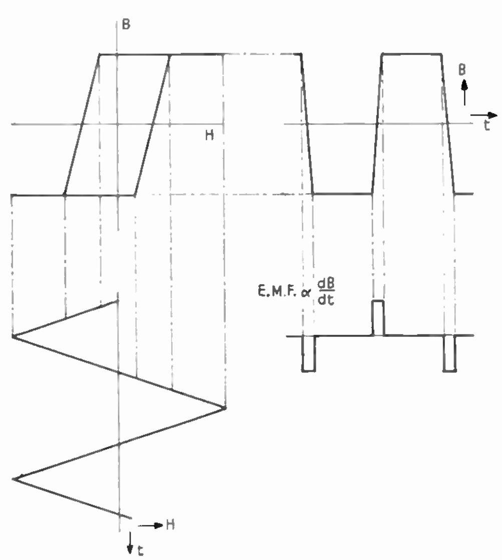


Fig. 1. Graphical determination of flux and e.m.f. waveforms from the static *B–H* loop.

3. Switching Time Theory

3.1. *General Theory*

The time taken for the flux in a square-loop core to reverse from positive to negative saturation or vice-

† Department of Electrical Engineering, University of Newcastle-upon-Tyne.

versa, called the switching time, τ, may be calculated using a simple theory based on three assumptions. These are that:

(1) Change of flux takes place entirely as a result of irreversible domain wall displacements. This neglects change of flux due to reversible rotation of the magnetization vectors, which in square-loop ferrites is a small proportion of the total flux change, provided that the magnetizing force is sufficiently large to switch the core completely.[3]

(2) The domain-wall velocity is proportional to the magnetizing force in excess of a threshold value H_0, which is approximately equal to the coercive force H_C. Strictly, this assumption is valid only for a single crystal.[5]

(3) The rate of change of flux, and hence the e.m.f. induced in a winding, is proportional to domain wall velocity. This assumes that the domain wall area remains constant during changes of flux.

From assumptions (2) and (3)

$$\frac{dB}{dt} = K(H - H_0)$$

where K is a constant depending on the material parameters. Hence τ may be found from

$$2B_m = \int_0^\tau K(H - H_0)\,dt \qquad \ldots\ldots(1)$$

where B_m is the saturation flux density.

For a step change in magnetization H_S, eqn. (1) gives

$$\tau = \frac{2B_m}{K} \cdot \frac{1}{H_S - H_0} \qquad \ldots\ldots(2)$$

This is the well-known switching equation for square-loop ferrites, which is usually written in the form

$$\frac{1}{\tau} = \frac{H_S - H_0}{S} \qquad \ldots\ldots(3)$$

where $S = 2B_m/K$ and is a constant for a particular material. Equation (1) may therefore be rewritten

$$S = \int_0^\tau (H - H_0)\,dt \qquad \ldots\ldots(4)$$

The simple theory outlined above indicates that a step change in magnetization should give a rectangular voltage pulse. In fact, the pulse observed in practice is more nearly triangular. It has been suggested that this is due to variations in the total domain wall area, which rises from zero to a maximum as regions of reverse magnetization form and grow, and falls to zero again as the regions of reverse magnetization meet and coalesce.[4] Nevertheless, if the reciprocal of switching time (defined now as the time interval between 10% points on the voltage waveform) is plotted against applied magnetizing force, a straight line is obtained over quite a wide range of applied magnetization, as predicted by eqn. (3). A typical result is shown in Fig. 2. The curve departs from a straight line at fields just above the threshold value H_0 due to the reversible component of flux change, which at low magnetizing forces switches much faster than the main irreversible component, giving a higher amplitude e.m.f. pulse.

The success of eqn. (1) in predicting the form of the relationship between switching time and magnetization for a step change of magnetization suggests that it (or eqn. (4)) may also be used to predict the performance of a core with sinusoidal magnetization.

3.2. *Theory with Sinusoidal Magnetization*

A half-cycle of sinusoidal magnetization, peak value H_p, is shown in Fig. 3. Core switching starts when H exceeds H_0, at $t = 0$. This corresponds to time T after $H = 0$, where

$$H_p \sin \omega T = H_0$$

Therefore

$$T = \frac{1}{\omega} \sin^{-1} \frac{H_0}{H_p} \qquad \ldots\ldots(5)$$

From eqn. (4)

$$S = \int_0^\tau [H_p \sin \omega(t + T) - H_0]\,dt$$

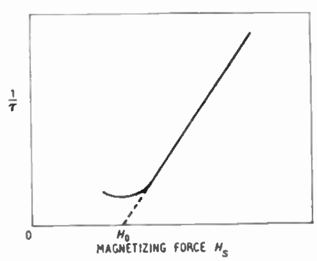


Fig. 2. Typical switching time graph for square-loop ferrite switched by a step magnetizing force.

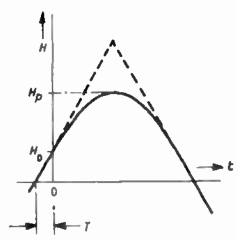


Fig. 3. Waveform of sinusoidal magnetizing force.

$$= \frac{H_p}{\omega}[\cos\omega T - \cos\omega(\tau+T)] - H_0\tau$$

Substituting from eqn. (5) and simplifying:

$$\omega = \frac{1}{S}[\sqrt{H_p^2 - H_0^2}(1-\cos\omega\tau) + H_0(\sin\omega\tau - \omega\tau)] \quad \ldots(6)$$

Unfortunately it is difficult to rearrange this expression to give an explicit value of τ in terms of ω. Numerical solutions for particular cases are of course possible, but their computation is laborious.

An approximate solution may be obtained from eqn. (6) for the case when the switching time is much less than the half-period. Then $\omega\tau \ll \pi$ and

$$\sin\omega\tau \simeq \omega\tau$$

$$1-\cos\omega\tau \simeq \frac{\omega^2\tau^2}{2}$$

From eqn. (6) this gives:

$$\omega \simeq \frac{\omega^2\tau^2}{2S}\sqrt{H_p^2 - H_0^2}$$

Therefore

$$\tau = \sqrt{\frac{2S}{\omega(H_p^2 - H_0^2)^{\frac{1}{2}}}} \quad \ldots\ldots(7)$$

For $H_p > 2H_0$, this may be further simplified to

$$\tau \simeq \sqrt{\frac{2S}{\omega H_p}} \quad \ldots\ldots(8)$$

This latter approximation introduces an error of less than 10%. Equation (8) may be shown to correspond to the case where H is of ramp, or triangular form such that ωH_p, the slope of the sine-wave at $H = 0$, is the same as the slope of the ramp. This ramp waveform is shown dotted on Fig. 3.

The expression given in eqn. (6) will only remain valid provided that the core switching is complete by the time the magnetization reverses through H_0, that is for

$$\tau < \frac{\pi}{\omega} - 2T$$

or

$$\omega\tau < \pi - 2\sin^{-1}\frac{H_0}{H_p} \quad \ldots\ldots(9)$$

In any case, before this limiting value is reached, the assumptions made to obtain the approximate expression, eqn. (8), cease to be justified.

4. Tests on Square-Loop Ferrite Cores

When a sinusoidal m.m.f. of sufficient amplitude to cause saturation is applied to a square-loop ferrite core, the e.m.f. induced in the excitation windings consists of pulses, shown in idealized form in Fig. 1.

It is important that these voltage pulses should not affect the current drive waveform. The winding should therefore be driven from a source having a high impedance to all frequency components of the voltage pulses. To avoid unnecessary power loss, an inductor may be used, as shown in Fig. 4. At the required drive current, the voltage drop across the inductor should be many times the peak voltage across the winding on the square-loop core. The known inductance L then swamps the unknown non-linear inductance of the winding, and the current waveform is closely sinusoidal. To reduce the current drawn from the voltage source, the circuit may be resonated at the drive frequency by a capacitor C.

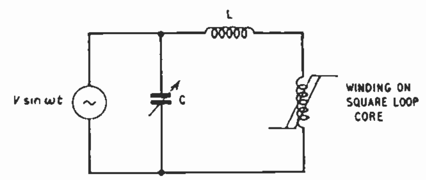


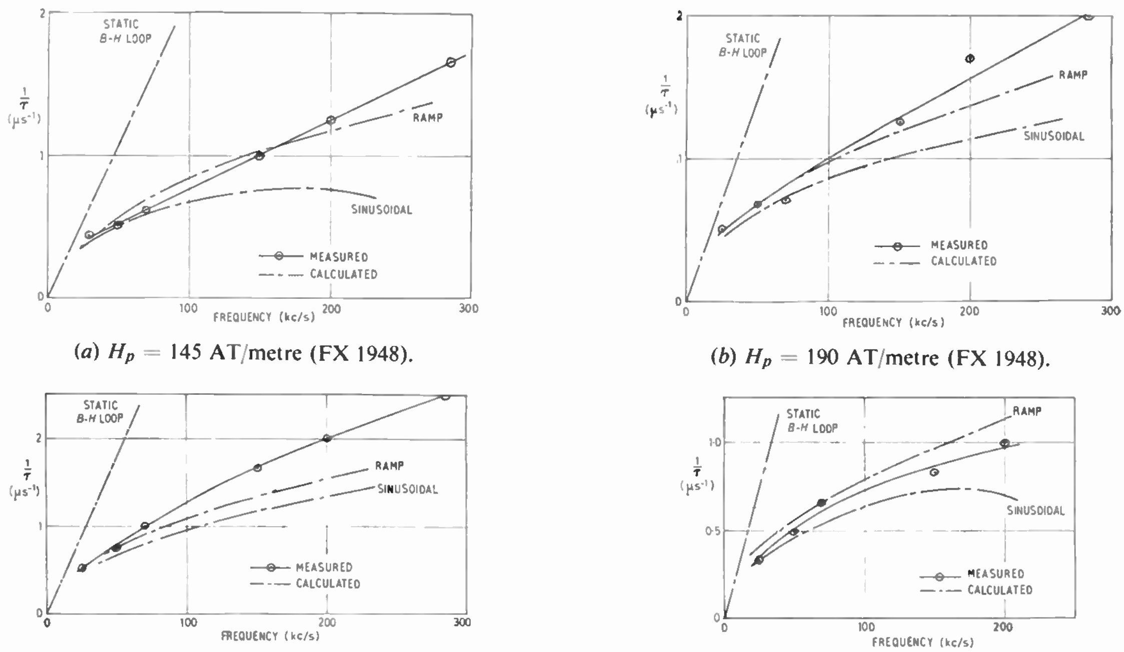
Fig. 4. Circuit for providing a sinusoidal current in a non-linear inductor.

The voltage waveforms appearing across the core windings were of approximately triangular shape, and the measured core switching time was taken between 10% points on these waveforms. Typical results are shown in Fig. 5. On each graph, measured switching times are compared with those predicted from the static *B-H* loop, from the theoretically derived eqn. (6), and from the approximate 'ramp' solution of eqn. (8). Figures 5(*a*), (*b*), (*c*), are for a 2-mm outside diameter core for various values of peak magnetization, and Fig. 5(*d*) is for a 4-mm o.d. core of different material. Details of the cores used are given in Table 1. At high frequencies, especially with the larger core, hysteresis heating gave rise to an appreciable increase in core temperature. Each result was therefore taken as rapidly as possible, starting with the core cold. It is known that an increase in temperature lowers the core switching constants S and the magnetization threshold H_0.[3]

Table 1

Details of Square-loop Ferrite Cores Tested

Core	Dimensions (mm.)			Magnetic Properties		
	Inside diameter	Outside diameter	Thickness	H_0 (AT/m)	S (ATs/m)	B_m (wb/m²)
FX 1948	1·3	2·0	0·6	77	$6{\cdot}4\times10^{-5}$	0·21
FX 1897	2·2	3·8	1·5	51	$5{\cdot}0\times10^{-5}$	0·26



(a) $H_p = 145$ AT/metre (FX 1948).

(b) $H_p = 190$ AT/metre (FX 1948).

(c) $H_p = 240$ AT/metre (FX 1948).

(d) $H_p = 100$ AT/metre (FX 1897).

Fig. 5. Graphs of switching time against excitation frequency for square-loop ferrite cores.

5. Conclusions

Figure 5 shows that the experimental results agree reasonably well with the theory derived in the paper at frequencies above about 10 kc/s. At lower frequencies, the response of the core depends also on the shape of the static B-H loop, and the theory does not apply.

As shown in Fig. 2, a theory based on the assumptions listed earlier breaks down for values of field near the threshold value H_0 when applied to the case of step changes of magnetization. With sinusoidal magnetization, the field in the core passes quite slowly through this region where the theory is invalid, and this may account for some of the discrepancy between calculated and measured switching times shown in the graphs.

A further source of error in the small cores used in the tests is that the ratio of inside to outside diameters is considerably less than unity, giving rise to variations of field strength with radius within the core.

To ease calculation of switching times, an approximate expression corresponding to a linear ramp of magnetizing force was derived (eqn. (8)). The graphs show that over most of the frequency range, this expression is a fair approximation to the more exact expression (eqn. (6)). Fortuitously, it happens to give a more accurate representation of the measured performance.

6. Acknowledgments

The work described in this paper was carried out with the financial support of the Department of Scientific and Industrial Research. The author wishes to thank Mr. F. J. U. Ritson and Mr. R. C. Foss for their advice and encouragement in the preparation of this paper, and Professor R. L. Russell for the use of the facilities of the Department of Electrical Engineering, University of Newcastle-upon-Tyne.

7. References

1. F. C. Williams and S. W. Noble, "The fundamental limitations of the second-harmonic type of magnetic modulator as applied to the amplification of small d.c. signals", *Proc. Instn Elect. Engrs*, **97**, Pt. II, pp. 445–59, 1950.
2. W. A. Geyger, "Low level magnetic amplifiers", *Proc. National Electronics Conf.*, pp. 558–93, 1961.
3. J. Smit and H. P. J. Wijn, "Ferrites", Section 63 (Philips Technical Library, Eindhoven, 1959).
4. N. Menyuk and J. B. Goodenough, "Magnetic materials for digital-computer components, 1A: Theory of flux reversal in polycrystalline ferromagnetics", *J. Appl. Phys.* **26**, pp. 8–18, 1955.
5. Ref. 3, Section 36.2.

Manuscript first received by the Institution on 19th December 1962 and in final form on 26th March 1963. (Contribution No. 68.)

U.D.C. 621.396.44: 621.398: 654.93: 622.698

A Telecommunication and Telecontrol System for a Crude Oil Pipeline

By

W. T. BROWN (Member)†

Presented at a meeting of the Institution in London on 10th April 1963.

Summary: The paper describes a comprehensive telecommunications, telemetry, telecontrol, and instrumentation system which was installed on a crude oil pipeline in North East India. The philosophies which led to the final design are discussed. Descriptions of the equipment used are given together with tables of functions controls and alarms.

1. Introduction

The crude oil pipeline described in this paper extends for a distance of 720 miles along the North East borders of India. It is in two parts; the first section consists of approximately 250 miles of 16-inch pipe and runs from the oilfields in Nahorkatiya to the main control centre at Nunmati (near Gauhati). A subsidiary oil field at Moran, some 30 miles from Nahorkatiya also supplies a quantity of oil to the pipeline. The second section of 14-inch pipe proceeds from Nunmati to Barauni, a distance of 470 miles.

There are two refineries concerned in the project, one being located at Nunmati and the other at Barauni. The oil in the 16-inch section divides at Nunmati and the refinery there takes an input of 0·75 million tons per year; the remainder of the oil (2 million tons per year) is pumped on to the refinery at Barauni. Figure 1 shows the general geographical layout of the pipeline.

The quality of the crude oil which is being pumped is such that special arrangements have to be made to condition it prior to its injection into the pipeline at Nahorkatiya and Moran. A conditioning plant at Nahorkatiya heats and cools the oil in a controlled manner, so that the maximum degree of fluidity (or pumpability) is obtained.

In the early design stages of the pipeline, it was decided that because of the length of the pipeline and the type of terrain which was traversed, automatic control and remote instrumentation would be employed. The obvious choice for the control centre was at Nunmati.

The advantages of applying remote and automatic control techniques to a trunk pipeline are manifold. Apart from the obvious benefits of the saving of man-power, accommodation, etc., three of the less obvious benefits which accrue are:

† Burmah Oil Co. Ltd., London.

(1) Valve and engine controls can be operated accurately and safely from a central point.

(2) Tele-control and data handling systems are unlikely to be affected by industrial disputes.

(3) Both the short and long term recording and/or display of measurements are accurate and reliable.

None of these benefits, however, is possible unless a reliable system of communication is available, and the satisfactory and economical operation of a pipeline depends entirely upon its communication system for security, continuity and operation. A trunk radio system capable of carrying a maximum of 48 carrier channels was chosen as being the most reliable communications system for the purpose. In passing, it should be noted that a coaxial cable system, whilst entirely adequate for the technical purpose, suffers from the disadvantage that were it to be buried with a pipeline (which is the only economical way of installing it) a catastrophic failure of the pipeline could automatically paralyse the communications network at a time when it was most required. Furthermore, a radio system is much more secure from civil disturbances, etc.

There are nine pump stations and seventeen radio repeater stations distributed throughout the pipeline and each pump station is provided with its discrete channels connecting that pump station to both H.Q. and its upstream neighbour (Fig. 2).

It was never envisaged that the pump stations would be entirely unattended since the amount of normal routine maintenance work which is required presupposes a small maintenance staff at each station. Each pump station is, therefore, provided with a local control desk from which the normal running of the station can be carried out. Oil throughput rate, engine speeds, alarms, etc., are located on the local control desks. Since the station alarms are located there and the station is normally manned during working hours, in the event of emergency the local engineer

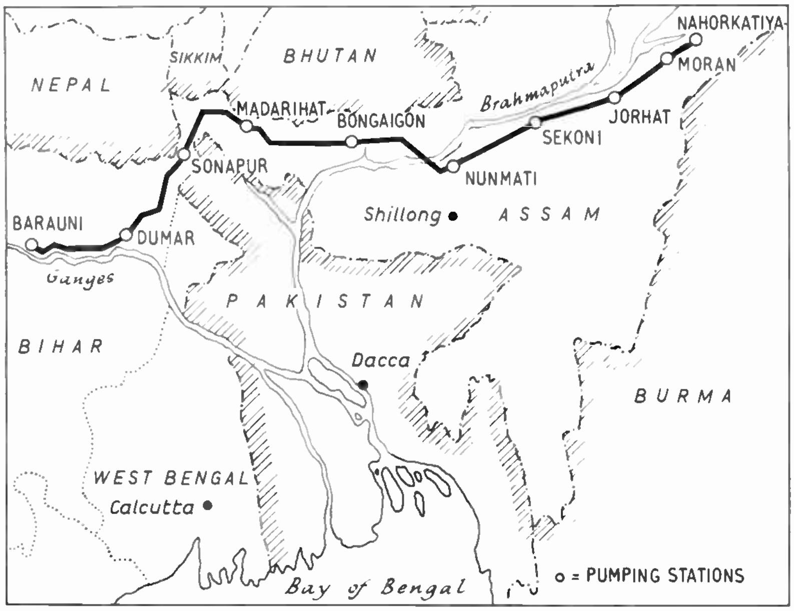


Fig. 1. General pipeline route.

can take the appropriate steps. Certain discrete alarms are forwarded to the despatcher at Pipeline HQ at Nunmati.

These are:

Pump inlet pressure.
Discharge pressure.
Balance tank 'out of band'.
Tank levels.
Fire.
Multiple (or combined alarms).

The basic philosophy for the control of the pipeline envisages that the despatcher at Pipeline H.Q. shall be in a position to receive all parameters associated with the oil flow, pressure, temperature, etc., and can regulate pressure/flow on the pipeline accordingly. He is not necessarily concerned with operational states of equipment in detail, and consequently the multiple alarm provides an indication that all is not well and voice and/or teleprinter communication from the despatcher to the engineer will enable him to determine the seriousness of any situation causing the multiple alarm to operate. Available also to the despatcher at Pipeline HQ are certain controls enabling him to increase or decrease the throughput of the pipeline, close down engines at any pump station, open and shut valves at pump and repeater stations, etc.

The telecommunications, instrumentation and control systems may be separated as follows:

(a) A main trunk radio comprising a v.h.f. system, capable of carrying up to 36 channels. In practice, however, the system is currently equipped to 17 channels, one of which is for the exclusive use of the telemetering system. The remaining channels are distributed between the various stations in

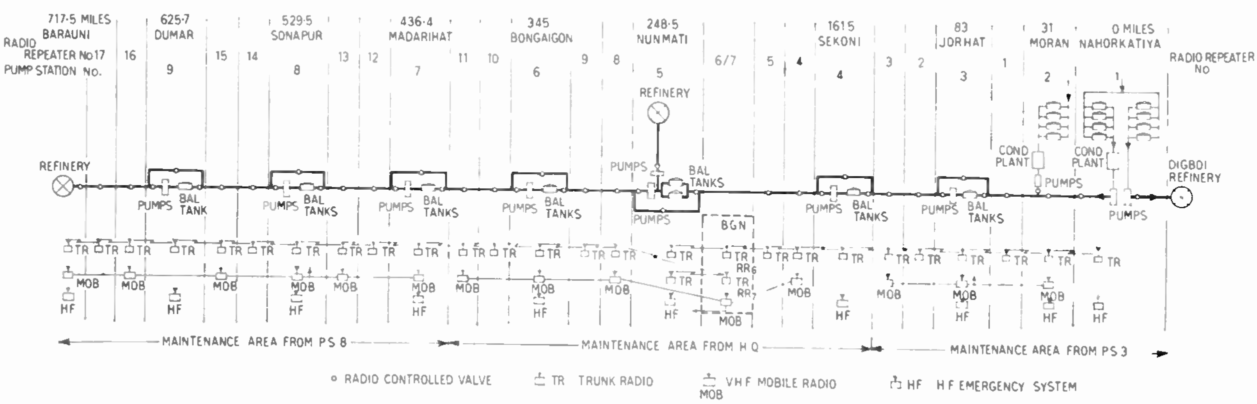


Fig. 2. General schematic of the radio system.

order to provide the best possible communication to these points. Provision has been made for through-channel connection if required. In order to provide maximum reliability for the trunk radio, duplicate equipment is installed at each station with an automatic change-over device, so that in the event of a failure, the trunk link is re-established with the minimum of delay.

(b) A v.h.f. mobile system to facilitate the patrol and maintenance of the pipeline. It is possible for mobile patrols and maintenance teams to communicate directly with their maintenance headquarters and, if necessary, the engineer in charge at three headquarters can patch through this channel to the despatcher.

(c) An h.f. emergency system connecting each pump station to the despatcher on a simplex basis, the receiver listening out at all times.

(d) An automatic/manual control system for pumping oil. With the exception of Nunmati pump station (PS.5) all the intermediate Pump Stations are self-controlling in so far as adjustment of through-put is concerned.

(e) Oil parameters which are of interest to the operational staff are measured at their functional point and are converted to a direct current in the range 4–20 mA in proportion to the value of the parameter. Examples of these are the temperatures and pressures of crude oil in the main pipeline.

(f) A telemetering system which converts the direct current mentioned in (e) into a signal suitable for transmission over a radio link. This signal is, in fact, a discrete series of pulses, directly related to the measured value of the parameter. The digital form of the pulse is converted to a frequency-modulated signal which, in turn, is transmitted by the radio link. At the receiving end the digital information is recovered, converted to a percentage of the maximum value being measured and displayed on a tele-type record. In order to assist the despatcher in quick appreciation of any operational situation, oil flow data are separated from mechanical data and are typed out on an Addo machine. For example, precise tank level, flow and summated throughput appear on the Addo, whilst line pressure, discharge pressure, etc., are displayed on the Teletype. Normal operational states, i.e. valves open or closed, are displayed on a mimic diagram of the panel, whilst certain alarms such as 'discharge pressure high', 'fire', etc., are indicated on the annunciator panel, situated immediately above the mimic diagrams.

(g) The tank levels at intermediate pump stations are held constant and, all other things being correct, engine speeds should therefore be constant. However, in the event of a leak of oil taking place in any one section of the pipeline, the input to the downstream station will fall and, therefore, the engines at this station will tend to slow down in order to maintain a constant tank level. Means are provided throughout the system for automatically checking and comparing engine speeds at all stations in a sequential manner. If any two stations show discrepancy in speeds, it could be indicative of a leak between those stations and suitable indications to this effect are given to the despatcher.

2. Instrumentation

The instrumentation for the pipeline has been designed on the principle that all stations will run automatically and operating staff are only required to start up major units or take corrective action on the receipt of non-urgent alarms. Each station has a control room, overlooking the pump and engine rooms, which houses the station control desk, the telemetering equipment, the telephone conference channel equipment and the teleprinter. This arrangement minimizes the cable runs between the engines, control desks, and telemetry. The physical arrangement of the desk is dictated by the shape of the room and the desirability of being able to view the pump and engine rooms while operating the controls.

The stations on the pipeline fall into two basic types, (1) pumping stations and (2) repeater stations. The pumping stations are best described by reference to the various parts of the control desks. They can be subdivided into two types:

(a) Those in which the rate of pumping is controlled manually at the station or by the dispatcher at H.Q.

(b) Those at which the rate of pumping is controlled automatically within the station.

At repeater stations, the only data handled are those showing pressure at each side of the mainline valve, and the running state of the generators. It is also possible to close/open the mainline valve remotely from H.Q. Since all data and control techniques at the repeater are identical to those at the pumping stations, they are not described separately.

The control desks can conveniently be divided into four sections (Fig. 3):

(1) The recording and display of essential continuously varying parameters.

(2) A mimic panel which displays the state of the mechanical remotely-controlled valves, machinery tanks and control systems.

(3) A control panel to operate the valves, engines and ancillary services.

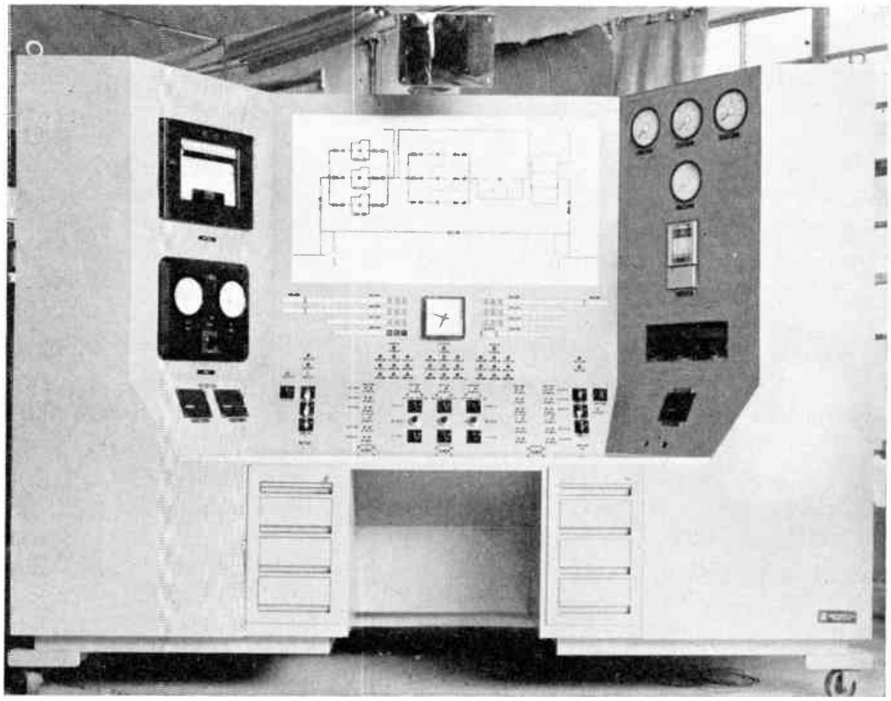


Fig. 3. Pump station control desk.

(4) An alarm display which automatically takes protective action to prevent damage to plant.

The measurement of the crude oil parameters is complicated by the fact that the oil is a non-Newtonian fluid with a high wax content (approx. 14%) and a high set point (approx. 85° F). Oil conditioning is to be provided but it is necessary to isolate all transducers from the crude oil by the provision of demountable diaphragms at the measurement point and by filling the transducers with a suitable fluid. In the case of tank level measurement it is also necessary to provide non-contacting probes. The measurement system throughout is based on a range of transducers, transmitters, indicators, recorders and controllers using a transmission or control signal of 4–20 mA d.c. There are also ancillary units such as millivolt to current converters, monitor switches for high and low limit detection and current limiter units for ensuring minimum and maximum output signal limiting.

The recorders/controllers are based on the force balance principle, force being applied to a pivoted beam by the process variable via a diaphragm, Bourdon tube etc. (Fig. 4). The resulting deflection of the beam varies the air-gap between a pair of ferrite cores, rigidly mounted on the beam which form part of the tuned circuit of an oscillator. The change of inductance causes the oscillator to draw more current from the power supply and hence gives a proportional indication of the input variable. The current is fed through a magnet unit which opposes the direction of motion and thus balances the beam. The recorders and indicators work in reverse with the magnet unit disturbing the beam which is restored by a spring coupled to the pen or pointer which in turn is driven by a torque motor.

Since a security check is carried out over the line it is necessary to measure tank levels to ± 1 mm in 10 metres. This is achieved by the use of a servo-driven probe containing a bridge which balances out the capacitance between itself and the surface of the oil. A transmitter is coupled to the drive tape take-up pulley between the probe and the servo

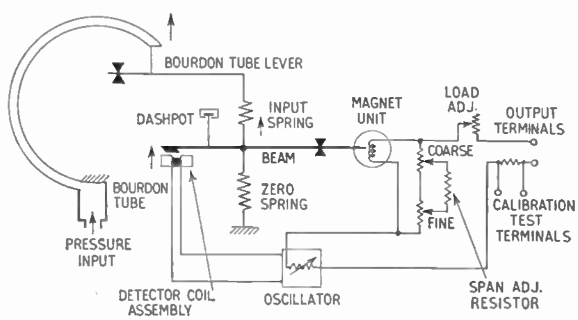


Fig. 4. Pressure transducer simplified circuit.

motor. For stations at which automatic control of the engines is required, the transmission consists of two analogue signals, one representing 0–10 m in 20 cm steps, the other 0–20 cm. At issuing stations where the tanks are of larger diameter a digital transmission system is used.

The mimic panel shows the state of the mechanical valves (open, shut, or in transit), which pumps, alternators and ancillary equipment have been selected and are running and which tanks are on flow. The control panel enables all pumps to be started and stopped, valves to be operated, the speed of the pumps to be varied and, where applicable, to be put under automatic control. A further facility is a switch with associated warning light to switch off the telecontrol thereby leaving the station completely under local control. Since a power switchboard is situated adjacent to the sets no controls are provided on the main control desk for the alternators apart from the automatic shut-down in the event of a major fault occurring.

By arranging for the normal state to be 'power on' all the alarm systems are made to 'fail safe'. If any damage occurs to an external circuit or there is a loss of site power, the appropriate alarm registers. The alarms are grouped for each major system (i.e. each engine or alternator) and are common to the station alarm system. In addition to the non-urgent alarms the occurrence of a low lubricating oil pressure, lubricating oil high sump level or high water temperature alarm causes the individual engine to shut down. The occurrence of a station low inlet pressure, high discharge pressure or of a fire alarm shuts down all pumping sets but leaves the alternators running.

Table 1 shows the alarm functions. The alarm system is arranged so that any alarm present prior to a shut-down remains displayed but any occurring during or after the shut-down (such as low lubricating oil pressure) will be inhibited. In the case of pumping sets, a permissive start circuit is coupled to the alarm system to ensure all services are normal prior to starting. A system of key switches ensures that alternators cannot be switched off accidentally and with the appropriate key removed, a pumping engine is safe for personnel to work on. The overall safety of the system is ensured by the provision of an emergency alternator which starts automatically on the failure of the last main alternator.

At intermediate pumping stations the engines are controlled by a signal derived from the station balance tank level gauge in order to ensure that the level remains within ± 1 cm of the mid-tank height. The signal (0–100 mV) is derived from the tank level receiver corresponding to $\pm$ 10 cm on mid-tank level, and is converted to the standard 4–20 mA

Table 1

Alarm and Permissive Starting Functions

Function	Alarm	Shut-down	Permissive Start
	ENGINES		
Temperature			
Water	77° C	80° C	50° C
Lubricating Oil (H)	71° C	—	—
Exhaust Differential	adjacent cylinder ± 28° C	—	—
Exhaust (H)	43° C	—	—
Fuel Oil (L)	—	—	35° C
Pressure			
Lub. Oil (L)	0·35 kg/cm²	—	0·35 kg/cm²
Lub. Oil (L)	1·2 kg/cm²	0·84 kg/cm²	1·3 kg/cm²
Gas (L)	0·7 kg/cm²	—	—
Air	18·0 kg/cm²	—	18·3 kg/cm²
Flow Water (L)	no flow	—	—
Level			
Lub. Oil (H)	yes	yes	—
Lub. Oil (L)	yes	—	—
Fuel Drain Tank (H)	yes	—	—
Load (H) & (L)	extreme rack positions	—	—
Lub. Oil Priming Pump	—	—	running
	PUMPS		
Pressure Lub. Oil (L)	yes	—	—
Temperature Lub. Oil (H)	yes	—	—
Level			
Lub. Oil (L)	yes	—	—
Drain Tank (H)	yes	—	—
	STATION GENERAL		
Fire Oil Area	yes	yes	—
Pressure			
Dis. (H)	85 kg/cm²	85 kg/cm²	—
Suction (L)	1·4 kg/cm²	1 kg/cm²	—
Strainer Diffl. (H)	0·35 kg/cm²	—	—
By-pass Line (H)	3·5 kg/cm²	—	—
Power Failure	yes	yes	—
Temp. Control Room (H)	yes	—	—
Viscosity	yes	—	—
Tank (H)	yes	—	—
Tank (L)	yes	—	—
Tank Out of Band	yes	—	—
Radio and Carrier	yes	—	—

signal and fed to a controller with adjustable set-point. If the level deviates from this point, a control voltage is fed to the electric governor on the engines via a magnetic amplifier to speed up or slow down the engines until the tank level is restored to mid-point. The magnetic amplifier gives the necessary power gain and enables two or more engines to be fed from one control signal without interaction. If the level variation exceeds $\pm$ 1 cm the engines are locked at the maximum or minimum revolutions (whichever is appropriate) and an alarm initiated.

All necessary measurements are presented to the telemetry in the standard 4–20 mA form and all alarms as change-over contacts. In addition to the specific telemetered alarms a further multiple alarm is presented at H.Q. This alarm combines various station functions which are not necessary of immediate interest to the despatcher, but alerts him to await an explanatory call from the particular station engineer.

3. Telemetry and Control

The telemetry of the Oil India Pipeline is a digital system and is used with continuously operating plant. This contrasts with much missile telemetry where the telemetry is a recording system and the information is played back later. The presentation of the telemetry is to an operator. It provides automatic detection of changes on the plant and the alarms, together with automatic numerical recording of data relating to the plant. Extensive manual remote control is also included.

3.1. *Outline of the Technical Approach*

The remote supervisory system was one of the first to employ solid state circuitry since it is only by the use of transistors that a time-multiplex system can achieve a fast enough performance. The circuit elements used are slow by computer standards but are at least two orders faster than electro-mechanical devices. The operation of the system is, therefore, quite fast enough to enable a signalling speed to make full use of the available bandwidth.

The master station is a true master in that all actions are initiated at the master, normally automatically but additionally manually, for 'on demand' controls. The out-stations are genuine slaves with no initiating power whatsoever. In operation the master station sends out in a predetermined sequence a series of questions to each out-station and receives, immediately after each question, a reply which normally contains information as to the state of a specific piece of plant. These interrogations are continuous throughout the life of the equipment. Because at any one instant only one question can be sent, there results a possible delay of a complete scanning cycle between the appearance of a change of state and the reply to the master station. The transmission rate, however, is adequate to enable the maximum delay to be well within the specified twenty seconds and, on average, will be half this maximum.

The requests for information are programmed within the master station equipment. This program is unalterable as to sequence but arrangements are made for breaking into the program when control action is required or when some other non-continuous function, such as the logging of information, is necessary. Under these circumstances a system of interlacing is used whereby every second signal is the special one but the normal interrogation sequence continues,

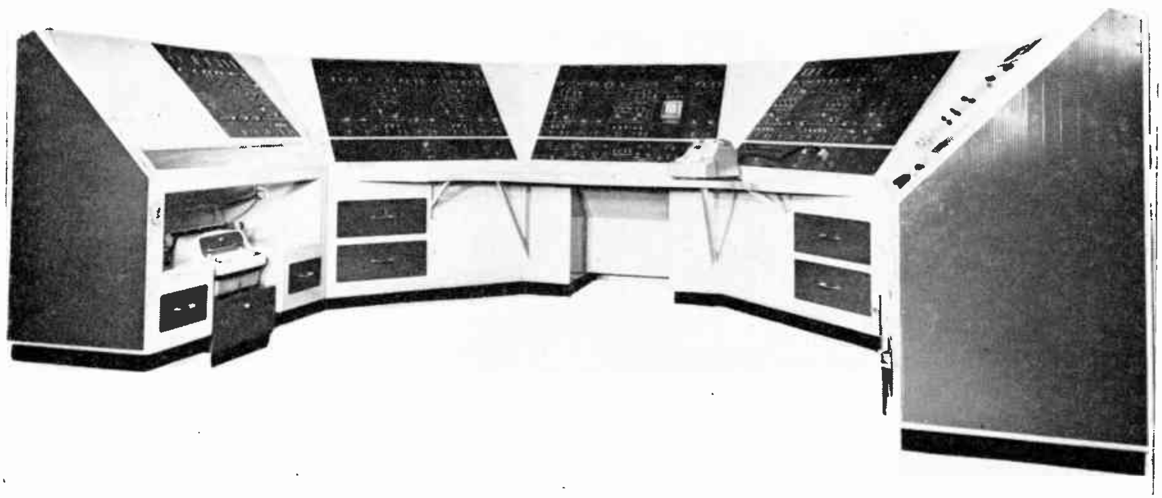


Fig. 5. Despatchers control desk.

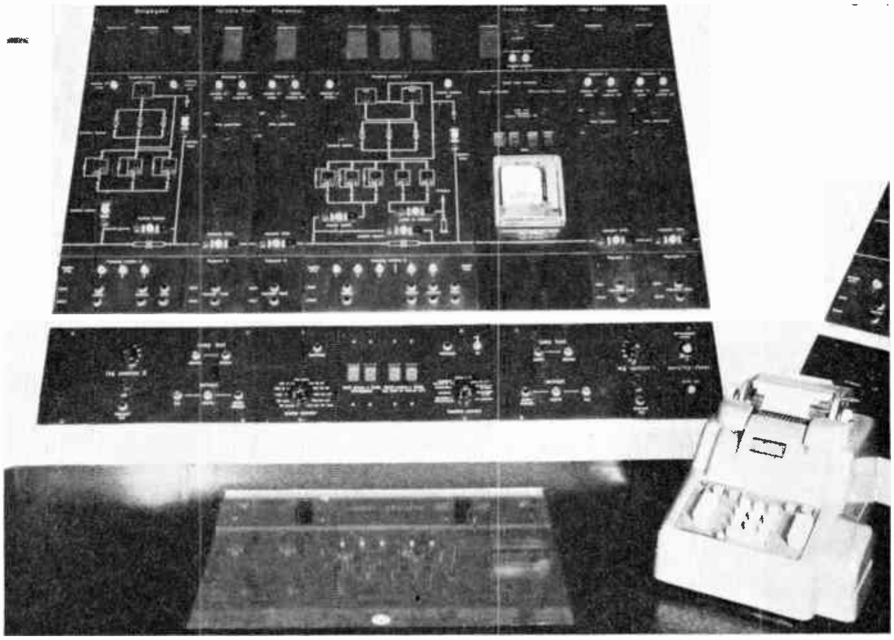


Fig. 6. Centre section of main control panel.

although the rate is reduced. Even under the worst conditions the rate of interrogation still enables alarms to be detected within the specified period.

The advantage of an active system is that the entire system is, in effect, being continually tested. The absence of reply information, which is readily noted from the display panel, indicates failure either of the telemetry data processing equipment or of the communications channel and this failure is signalled within a few seconds of its occurrence. The maximum chance is given of avoiding the situation whereby an alarm is not signalled due to failure of the transmission system as a whole. The digital system used provides a means for transferring numbers in both directions between the central control room and the 25 out-stations. These numbers are in a standard 3-digit decimal form for the whole scheme and are represented in various codes according to convenience of the equipment or the form of display required. The 3-digit numbers are sent from the central control room at a rate of 10/second; outgoing numbers are called addresses. Each address is recognized at one and only one out-station and this out-station replies with a 3 digit number, the value of the reply being the information of the item of plant called for by the address. Each address is followed by its reply and then by the next address. The addresses need not be sent in any given order, and since the system is a random access one consecutive addresses can call for information from different stations and they can also be used to operate controls on the plant at out-stations. There is, therefore, no difference between addresses for control action on the plant and addresses which call only for information from the plant.

3.2. *Displays*

The information displays and the control buttons for the operator are situated on the desk, which is separate from the logical equipment (Fig. 5). Both displays and controls are in terms of the plant and not the coded numbers, i.e. the controls are labelled according to the name of the item of plant they operate. Equipment in the control room converts the numbers used for transmission into the form required by the different types of display.

3.3. *Controls*

Each control function has its own labelled button on the desk and, when operated, the appropriate address is sent at intervals (Fig. 6). There are two numerical indicators, each of which displays the percentage value of one measurement on the plant, the items of plant being selected by the operator. The controls operate within half a second and the numerical indicator readings are revised every two seconds.

The controls which stop pumps, shut the main line valves and the flow counters are provided with a modifier at the out-station. This prevents any action taking place unless a particular store has been set at the station and the modifier store is set then by a different address. The important controls, therefore, require two different addresses to occur before the control action takes place. Extra redundancy is therefore provided on the important controls without lengthening the measurement addresses.

The equipment uses frequency division multiplexing to transpose 13 speech channels each of 300 c/s to 3400 c/s bandwidth and their associated signalling tones into the frequency band of 60 to 108 kc/s. The two modulation stages used firstly transpose the 12 speech channels into 4 sub-groups and secondly places these sub-groups into the final group frequency band of 60 to 108 kc/s. The carrier frequencies and 60 kc/s synchronizing pilot are generated within the equipment. The translating frequencies are derived from a 12 kc/s master oscillator whose frequency is stabilized by means of a crystal mounted in a thermostatically-controlled oven. The basic signalling frequencies, however, are derived from a 19·825 kc/s oscillator which again is crystal controlled but without the oven.

For this project the racks are only equipped with those channels necessary for each station, although each channel translation rack is wired to provide for a maximum of 36 channels complete with carrier supplies.

The group translation equipment, which occupies one rack side and assembled to the manufacturer's EC2 practice, has been provided in order to transpose the three basic 12 channel groups, A, B and C each in the frequency band of 60 to 108 kc/s, into three positions in the frequency band of 6–156 kc/s. At the terminal stations, where all 36 channels are terminated, a full complement of group translation equipment is provided. At intermediate stations the group translation equipment provides a facility which enables two groups, A and B, of 12 channels to be 'dropped' that is, extracted from and injected into the basic group band frequencies in both directions for transmission over the v.h.f. radio link.

4.1. *Frequency Modulated Voice Frequency Telegraph*

Normally the 24 channels of the f.m. v.f.t. are assembled by an amplitude modulation process from 4 identical sub-groups each of 6 basic frequency modulated channels having mid-frequencies of 1140, 1260, 1380, 1500, 1620 or 1740 c/s. Three of the four sub-groups are transposed by modulation, and are then combined with the unmodulated sub-group, to produce a group of 24 frequency modulated channels with a frequency bandwidth of 420–3180 c/s.

However, for this project only two channels, extracted from the unmodulated sub-group, are equipped at certain allocated stations in order to provide one teleprinter bearer channel and a standby. To enable selective and simultaneous calling over the teleprinter circuits special selective call relay sets, associated with each teleprinter, are fitted on the same rackside as the f.m. v.f.t. installed at the respective stations.

4.2. *Omnibus and Local Speaker Equipment*

This equipment has been provided for application to 4-wire circuits, and ensures communication with selective or simultaneous calling between two or more stations similarly equipped and connected to each other by only one 4-wire circuit. The bridging speaker panel ensures that full speech communication in both directions is obtained at all stations with the exception of certain terminal stations such as Nahorkatiya and Barauni, where one side of the panel only is terminated.

The primary purpose of these circuits is to permit a conference channel to be set up between stations for engineering discussions; at certain stations a further switching facility enables vehicles operating along the pipeline to communicate with their nearest maintenance base and Pipeline H.Q.

5. Radio Equipment

For operational purposes, the pipeline is divided into three maintenance sections. Each maintenance area is provided with an integrated mobile v.h.f. radio scheme to allow mobile patrols to be in touch with the area maintenance engineers, each of whom is resident at a pumping station near the centre of his section.

Although the main radio system is duplicated and power supplies at unattended stations are triplicated, the reliability of the communications for operational purposes must be beyond reproach, since complete trunk radio failure would entail cessation of pumping. To avoid this, an independent h.f. radio system has been supplied. This is remotely controlled from the pumping station control desks and is always in a standby state.

5.1. *Trunk Radio System*

The performance requirements were for a system in the 165–188 Mc/s band giving a total signal/noise ratio of better than 48 dB in the 36th telephone channel for 99% of the time. There are few really difficult paths on this project and careful planning of the station sitings has kept the path losses to a minimum. With the exception of one site, all sites have reasonably easy access and are located on the pipeline route. The exception is located at the top of a 1400 ft hill adjacent to Pipeline H.Q. at Nunmati and is used as a co-sited repeater station for both the East and West sections. The planning level of the median received signal is 700 μV e.m.f.; this level is required to give the desired fade margin whilst maintaining the thermal noise component of the signal/noise ratio of 16 links in tandem within the specification. The performance was achieved by using single, two- and four-stack 6-element Yagi aerials as required, mounted on self-supporting towers ranging between 75 ft and 250 ft in height.

A radio survey was carried out over most paths and confirmed the theoretical predictions. It also provided much valuable information on local terrain and conditions which was extremely useful during the planning of the installation.

The final frequency plan created many problems because the maximum allocation obtainable was fourteen discrete frequencies. Eight of these were selected for use at the Pipeline H.Q. area and care was necessary in the allocation of the remainder to ensure no interaction could occur. The frequency plan was simulated and proved by laboratory experiment.

Automatic changeover equipment is installed at all stations and the state of the equipment is continuously reported for display at Pipeline H.Q. Although this system is independent of the main telemetry scheme provided on the project, a remote radio failure alarm is displayed on the despatcher's desk.

The towers are of a straight-sided, tapered, self-supporting type, supplied and erected by an Indian Company. They are fitted with a platform at the top to facilitate aerial maintenance.

The aerials are of a rugged construction designed to stand up to the severe environment including winds up to 110 miles/h. A v.s.w.r. that is better than 1·12 : 1 at the centre frequency and better than 1·26 : 1 at ± 1 Mc/s is obtained. Single, double and four stack 6-element Yagi aerials are used. The multi-stack aerials are fitted with full-size rodded-back screens, and are fed by a pressurized air-spaced cable.

The radio rooms are air-conditioned, but in case of failure of the conditioning plant, a telemetered alarm occurs at 45° C. However, the equipment is shut down automatically if the temperature exceeds 50° C. This latter condition can occur with rooms that are almost sealed and where the outside ambient temperature may rise to 45° C. If the temperature falls again, the power is restored. A memory circuit in the automatic changeover ensures that the original working equipment is restored to traffic.

The v.h.f. trunk radio is a frequency-modulated broadband system capable of carrying a maximum of 48 telephone channels multiplexed in the band 12–204 kc/s. In this application, the system can be expanded to 36 channels, with a maximum transmitter output of 50 watts.

By the re-arrangement of various i.f. amplifier leads and certain coil taps identical equipment may be used as terminals of non-demodulating repeaters. This design feature simplifies servicing and allows full use to be made of spare units.

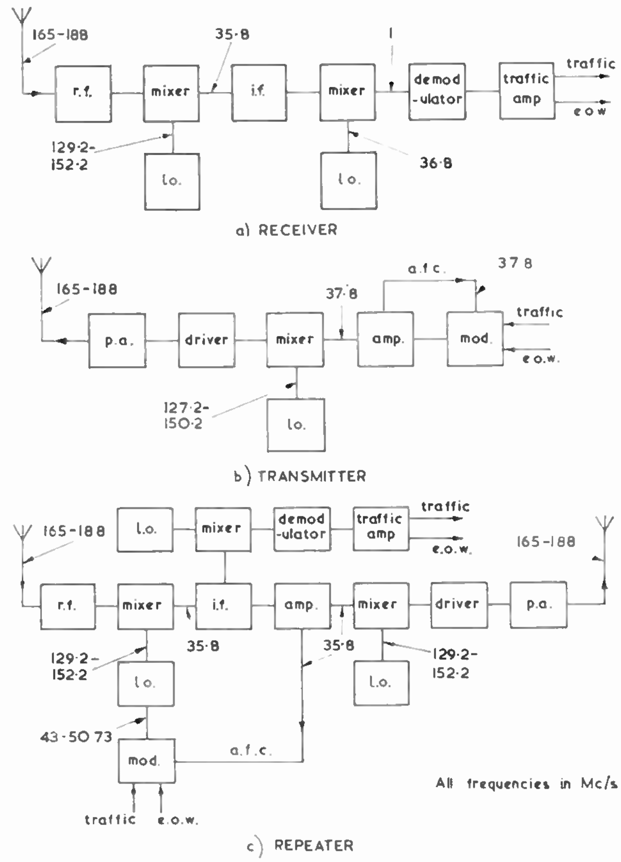


Fig. 7. Block schematic of trunk radio equipment.

The design of the transmitter and receiver follows established practice in these matters and is shown in block schematic form in Fig. 7.

The frequency modulated i.f. is 37·8 Mc/s and is produced by a free running oscillator with three reactance modulator valves across its tuned circuit. Two of these modulator valves are connected in push-pull and are driven by the multiplexed baseband signal. The third is used for automatic frequency control and engineer's order wire modulator combined. The modulated i.f. signal after amplification is mixed with the output of the crystal multiplier chain. The output from the mixer feeds the driver valve preceding the power amplifier.

The receiver r.f. input circuits comprise a neutralized cascode input stage followed by a mixer. The local oscillator chain is crystal-controlled and similar in design to the transmitter chain. The output of the mixer is fed to the i.f. amplifier at 35·8 Mc/s. The output of the i.f. amplifier is mixed with a local oscillator of 36·8 Mc/s. The resultant 1 Mc/s output is amplified and limited in a squarer circuit. The resultant frequency-modulated square-wave, which has been accurately controlled in amplitude and rise time is demodulated by a counter type discriminator. The derived baseband signal is suitably amplified for connection to the telephone carrier equipment.

In repeater applications the 35·8 Mc/s i.f. is fed directly to the transmitter (Fig. 7(*c*)). The master oscillator in the modulator is adjusted to the required frequency and in this application is used to replace the receiver local oscillator crystal. A demodulator is also incorporated which allows channels to be dropped off at repeater stations and for traffic to be inserted into free channels.

Fig. 8. General construction of trunk radio bays.

Built-in metering facilities allow both valve currents and important supply rails to be monitored. Transmitter deviation can also be monitored as well as tuning errors as shown by the a.f.c. circuit.

The engineer's order wire is incorporated in the direct voice band of 300–3000 c/s and out-of-band signalling at 3·8 kc/s is used. Access to the e.o.w. is available on all racks and with common apparatus housed in the top compartment. Calling is by lamp and buzzer, which may be extended if required.

The equipment is housed in a 7 ft high, 19 in rack, 7 in deep. Only front access is required for routine maintenance; all units may be operated on extension cords and rear doors are also fitted for convenience.

A general view of a suite of equipment including automatic changeover, is shown in Fig. 8. The design is generally to C.C.I.R. and DEF. 5000 specifications and the equipment will operate in the temperature range −20 to +55 deg C with relative humidities up to 95%.

5.2. *Mobile Scheme*

The pipeline is regularly patrolled and a mobile radio scheme has been engineered to allow communications between the mobile patrols and an area maintenance engineer.

By siting the mobile base station aerials at the top of the main radio towers, coverage is obtained over a radius of approximately 30 miles. In general base stations are sited at every other radio station.

The system is double frequency simplex f.m. in the 70–80 Mc/s band using Indian equipment. Two pairs of frequencies are used, designated 'A' and 'B'; alternate base stations use the same frequencies. This arrangement was adopted to avoid 'press to talk' at fixed stations.

All base stations in a maintenance area are connected on a party-line basis and appear at a jack panel located on the control desk at a maintenance area station. The base receivers are normally muted; on receipt of a mobile call the mute is removed and a signal appears on the jack panel; when the speak key is pressed, all transmitters in that maintenance area are switched on. The actual mobile stations operate as simplex transceivers operating on either the 'A' or 'B' frequencies. Since alternative base stations use the same frequency it is necessary for a mobile to try frequency selection to obtain the optimum result.

The despatcher can be patched through to a mobile party-line, and control of the base transmitters is then by the despatchers 'speak' key. This facility allows Pipeline H.Q. to communicate directly with maintenance crews in the field.

5.3. *H.F. Emergency Equipment*

H.f. equipment is installed and designed to provide reliable communication with maximum simplicity of operation. The transmitter/receivers are located in the telecommunications rooms of the pumping stations and are connected to horizontal half-wave dipole aerials.

The equipment is remotely controlled by a control unit embodied in a desk telephone positioned on the control desk at each pumping station. An alarm is given if trunk radio communication with Pipeline

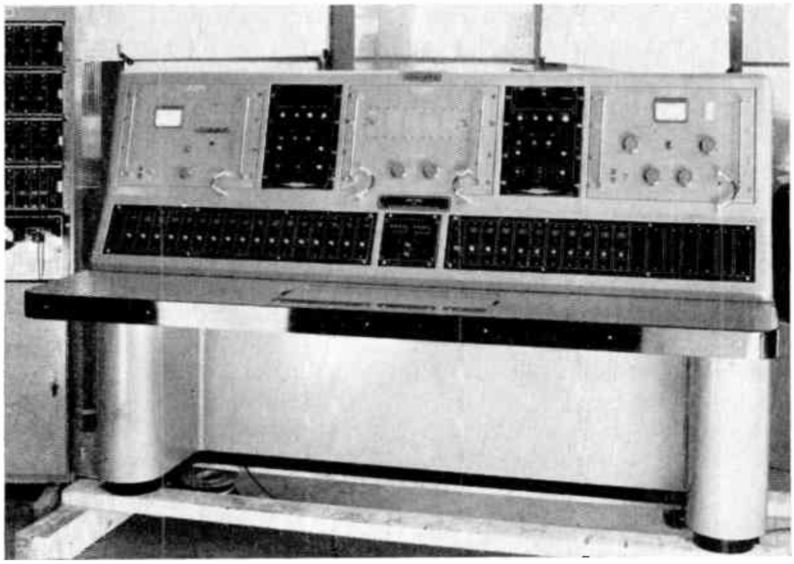


Fig. 9. Trunk radio control and monitoring desk.

H.Q. is lost, whereupon the pump station operator immediately establishes contact over the h.f. network. Calling is by a loudspeaker mounted on the pump station control panels, all stations being called together.

5.4. *Automatic Changeover System*

The automatic changeover system supplies comprehensive supervisory and control facilities for the main trunk radio. It is of unit construction for adaptability and is housed in a cabinet to match the radio equipment. Control and supervisory tones are in the band 4–12 kc/s. Single path operation with the standby equipment ready for immediate selection was adopted.

A control console (Fig. 9) is installed at Pipeline H.Q. A changeover taking place at any station is displayed on the console by the appropriate station lamp. The display may be reset by pressing the correct station acknowledge key; the supervisory system can then report a second changeover at that station should this occur, whilst circulating pilots continuously prove the 'go' and 'return' paths. Reverse operation of an acknowledge key and at the same time operating a master control causes the selected station to loop itself round at r.f. on the side farthest from the control station. All normal traffic is muted during loop round. Progressive signal/noise ratio and intermodulation measurements may be made along the system by means of a white noise test set integrated with the desk to localize any observed degradation in circuit quality.

By means of local switching either the 'A' or 'B' equipment may be selected as the duty equipment; the standby equipment runs with heaters on and reduced h.t. Aerials and feeders are not duplicated; a coaxial relay panel makes the necessary aerial circuit connections.

The facilities provided by the changeover system are as follows:

(a) By local switching, selection of either 'A' or 'B' radio equipment as the 'duty' equipment.

(b) Monitoring of the correct functioning of the following parameters of the 'duty' equipment:
    1. Transmitter power output.
    2. The a.f.c. control.
    3. The modulator stages.
    4. The receiver.

(c) Logical decision in the event of an indicated failure in any part of the equipment as shown in (b) above to initiate a changeover or, by different logic, to shut down a station.

(d) Local indication of the fault condition and the state of all equipment.

(e) Automatic transmission of a fault indication to Pipeline H.Q. station, with re-set and remote check facilities.

(f) Automatic transmission of a station tone, in lieu of the circulating pilots which disappear in the event of a failure of incoming signals to any station. This action maintains a normal condition at all subsequent stations. A crystal-controlled signal generator is switched on at a level of -15 dB on the normal received signal to quieten the receiver in the absence of the wanted signal. This condition is removed automatically when the circuit is restored to normal.

(g) An indication of 'path complete' from the Pipeline H.Q. is available at all manned stations.

(h) On command, a loop round condition at a selected station is established. This is accomplished by switching the 'go' path transmitter into a dummy load and feeding a part of this signal into a crystal-controlled frequency changer and from there by a coaxial relay into the 'return' path receiver. An indication of the satisfactory connection of the loop round condition is given on the control console. Stations beyond the looped round station operate normally because of the facility in (f) above. Under loop round conditions all external traffic is muted.

(i) In the event of a failure in the automatic change-over circuitry, the 'A' equipment is always selected as 'duty'.

The complete changeover equipment consists of a selection of standard plug-in sub-units and frames connected, by multi-way plugs and sockets, with fixed main units. The units are housed in a shallow 19-inch rack. Access to all units is available by front opening doors. As in the main radio equipment, comprehensive metering facilities are incorporated.

Maximum use is made of printed circuits, and except for the signal generator and frequency changer, transistors are employed. The printed circuit boards used are constructed from continuous film glass fibre bonded with epoxy resin.

Frequency sensitive units, e.g. detectors, tone generators etc., use the same unit design; only the actual frequency selector elements are varied.

6. Acknowledgments

The author wishes to thank the Directors of the Burmah Oil Co. Ltd. for permission to publish this paper and acknowledges the assistance given by: Rank-Bush Murphy Ltd., Mr. J. G. Ball, Associated Electrical Industries Ltd., Mr. D. Gibbon, Serck Controls Ltd., Mr. H. Cox, Honeywell Controls Ltd., Mr. D. Harbour, The Burmah Oil Co. Ltd., Mr. J. W. Smith and Bharat Electrics Ltd.

Manuscript received by the Institution on 12th September 1962. (Paper No. 845.)

U.D.C. 621.372.512.24

Direct Wide-Band Phase Shifters

By

P. V. INDIRESAN, Ph.D.

(Associate Member)†

Summary: Two functions which differ only in phase and that by 90 deg, are related to each other by the Hilbert transform and to derive one from the other it is necessary to know the values of the input function at all times, past as well as future, right up to infinity. However, the amplitude error will be less than ± 1 dB if the input values are considered up to a time duration equal to the period of the lowest frequency component in the signal. It has been explained that speech waves vary comparatively slowly and hence, their future form can be taken to be the same as that at the present instant. A direct wide-band phase shifting circuit which utilizes these ideas to give an arbitrarily small error in phase and an amplitude response constant to within ± 1 dB, over a 14 : 1 range of frequencies, is described.

1. Introduction

While several methods for obtaining a constant phase difference over a wide range of frequencies are known,[1-7] no method appears to have been described so far which gives a direct phase shift. That is, given a function

$$f_1(t) = \sum a_n \cos(\omega_n t + \theta_n) \qquad \ldots\ldots(1)$$

no method is known for obtaining the conjugate function

$$f_2(t) = \sum a_n \sin(\omega_n t + \theta_n) \qquad \ldots\ldots(2)$$

The difficulty is that this transformation requires a system which keeps both the attenuation and the non-zero phase shift constant. As Bode has shown,[8] the attenuation and phase shift characteristics of any network are not independent of each other. In fact, the present system, being of the non-minimum phase shift type, is more general than the one considered by Bode[9] and its input and output functions are related to each other by the Hilbert transform[10] and it is necessary to know all the values, past as well as future, of one function in order to derive the other.

Ideally, therefore, a direct phase shifter should be capable of predicting, right up to infinity, all the future values of the input function. However, in practice, it is sufficient if the future values are known for a limited period only. Fortunately, in communication engineering, the signals vary comparatively slowly and hence a reasonably accurate forecast can be made up to the requisite amount of time. Therefore, it is not impossible to obtain a constant, direct phase change over a wide range of frequencies. Such wideband phase shifters are described here. The amount of prediction possible for a typical speech wave and the error introduced by taking into account only limited future values have also been calculated.

† Electrical Engineering Department, University of Roorkee, India.

2. Principle of Direct Wide-band Phase Shifting

The conjugate functions $f_1(t)$ and $f_2(t)$ of equations (1) and (2) are related by the Hilbert transform

$$f_2(t) = \frac{1}{\pi}\int_0^{\infty} \frac{f_1(t-\tau) - f_1(t+\tau)}{\tau}\, d\tau \qquad \ldots\ldots(3)$$

While $f_1(t-\tau)$ defines past values of $f_1(t)$, $f_1(t+\tau)$ gives the future values. If $f_1(t+\tau)$ is known only up to, say, $f_1(t+T)$, an approximate solution can be obtained.

$$f_2'(t) = \frac{1}{\pi}\int_0^{T} \frac{f_1(t-\tau) - f_1(t+\tau)}{\tau}\, d\tau$$

$$= \frac{1}{\pi}\int_0^{T} \sum a_n \cos[\omega_n(t-\tau)+\theta_n] -$$

$$- \cos[\omega_n(t+\tau)+\theta_n]\frac{d\tau}{\tau}$$

$$= \frac{1}{\pi}\sum 2a_n \sin(\omega_n t + \theta_n)\int_0^{\omega_n T} \frac{\sin \omega_n \tau}{\omega_n \tau}\, d(\omega_n \tau) \qquad \ldots\ldots(4)$$

The integral term in eqn. (4) is given by the Si function and hence,

$$f_2'(t) = f_2(t) . 2/\pi . \mathrm{Si}(\omega_n T) \qquad \ldots\ldots(5)$$

Equation (5) explains why it is difficult to obtain a constant amplitude response in a wide band phase shifter. As the frequency varies, Si $(\omega_n T)$, and hence the amplitude response, vary. However, the Si function approaches a constant value of $\pi/2$ as $\omega_n t$ increases and will remain within ± 1 dB of this value when $\omega_n T$ exceeds 2π.[11] Therefore, if $\omega_n T$ is 2π or more for the lowest frequency component in the signal, Si $(\omega_n T)$ will remain a constant within ± 1 dB for the entire frequency range, and $f_2'(t)$ also will equal $f_2(t)$ to the same degree of accuracy. In telephony, the lowest frequency component is about 250 c/s and therefore, prediction up to 4 milliseconds will be necessary.

3. Persistence of Speech Waves

There is no problem in predicting the future values of a continuous signal and the difficulty arises only when the wave is building-up or when it is decaying. Fortunately, the rate of rise and decay of speech waves is quite slow and is further reduced by reverberation. This latter effect is disconcertingly obvious in a reverberent room. Even under ideal conditions, speech sounds do not rise more than 3% nor decay more than 2% per millisecond.[12] That is, for a 4 millisecond period considered above, the amplitude variation will once again be less than 1 dB. In fact, single sideband communication systems inherently assume speech waves to be invariant for periods of this order.[7, 13] Hence, a speech wave may be predicted a few milliseconds ahead simply by assuming it to be of the same form as that at the present instant. This fact has been utilized here to devise wide-band phase shifters.

4. Direct Phase Shifting Devices

The phase shift operation through equation (3) may be performed by direct analogy. An R-C network can be used as a delay line to give the function $f_1(t-\tau)$ and the function $f_1(t+\tau)$ can be obtained by using a similar network with negative, in place of, positive resistances.[14] Even if the problem of stabilizing the negative resistances is solved, this method will not be of practical interest because a prohibitive number of sections will be required. A second method, as shown in Fig. 1, divides the signal frequency range into a number of narrow-band sections.

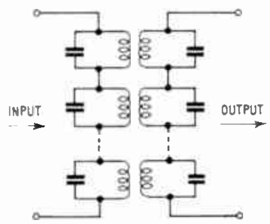


Fig. 1. Direct wideband phase shifter using tuned circuits.

The narrow-band components from each of these sections may be shifted in phase separately and then summed together to produce the desired output. If double tuned circuits are used as shown in Fig. 1, the two operations of selection and phase shifting will be performed simultaneously. It has been estimated that each tuned circuit should have a bandwidth of 50 c/s, that is, the arrangement needs over 50 sets of tuned circuits to cover the telephone band.

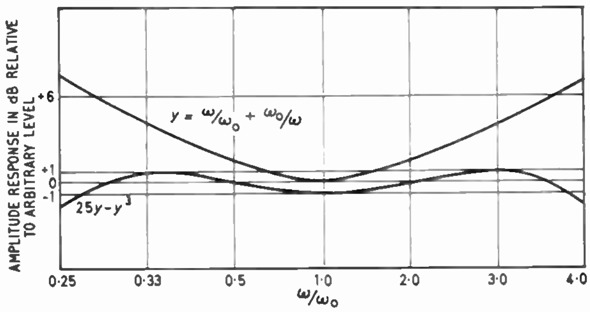


Fig. 2. Variation of amplitude response with frequency.

A simpler method uses integrating and differentiating circuits to give constant 90 deg phase delay and phase advance respectively. The difference between their outputs provides a transfer function of the form $j(\omega/\omega_0+\omega_0/\omega)$. This varies in magnitude comparatively slowly near the region $\omega = \omega_0$. The amplitude response can be made even more uniform, as shown in Fig. 2, by extending this principle by using the transfer function

$$Ky - y^3; \qquad y = \omega/\omega_0 + \omega_0/\omega \qquad \ldots\ldots(6)$$

When the constant K equals 25, the amplitude response is uniform to within $\pm$ 1 dB over a 14 : 1 range of frequencies. Where greater error is permissible a wider range of frequencies can be used.

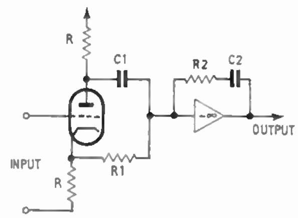


Fig. 3. Circuit to obtain a transfer function of the form $\omega/\omega_0 + \omega_0/\omega$.

The circuit of Fig. 3 is helpful in performing the above operation. The amplitude response of this circuit is

$$(1/R_1 - j\omega C_1).(R_2 + 1/j\omega C_2) = jR_2/R_1.(\omega/\omega_0+\omega_0/\omega) \qquad \ldots\ldots(7)$$

when $R_1C_1 = R_2C_2 = 1/\omega_0$.

The complete schematic circuit diagram for this direct wide-band phase shift system is shown in Fig. 4. The adjustments could be made without much difficulty, but noise due to the triple differentiation involved was a troublesome feature.

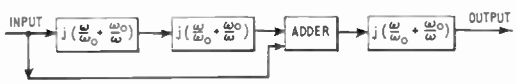


Fig. 4. Complete schematic diagram of direct wideband phase shifter.

5. Conclusion

It has been explained why, in a wide-band constant phase shift system, it is necessary to forecast the future values of the input function. It has been shown that, due to the slow rate at which speech wave forms change, it is permissible to assume that their future form up to several milliseconds is the same as that at the present instant. This permits the design of wide-band phase shifters to give an arbitrarily small error in phase, an error in amplitude response less than $\pm$ 1 dB, for about 14 : 1 range of frequencies. This is adequate for telephone transmission and these phase shifters can be used for the generation of single sideband signals or where the orthogonal conjugate of a given function is required.

6. Acknowledgment

The author thanks Professor C. S. Ghosh, Head of the Electrical Engineering Department, University of Roorkee, for the encouragement given in this project.

7. References

1. R. V. L. Hartley, U.S. Patent No. 1,666,206 (1928).
2. J. F. Byrne, "Polyphase broadcasting", *Trans. Amer. Inst. Elect. Engrs*, **58**, p. 347, 1939.
3. R. B. Dome, "Wideband phase shift networks", *Electronics*, **19**, p. 112, December 1946.
4. H. J. Orchard, "Synthesis of wideband two phase networks", *Wireless Engr*, **27**, p. 72, 1950 and **28**, p. 30, 1951.
5. S. Darlington, "Realization of a constant phase difference", *Bell Syst. Tech. J.*, **29**, p. 94, 1950.
6. A. A. Ahmed, "A wideband phase shifter," *Proc. Inst. Radio Engrs*, **48**, p. 945, 1960.
7. K. H. Powers, "The compatibility problem in single sideband transmission", *Proc. Inst. Radio Engrs*, **48**, p. 1431, 1960.
8. H. W. Bode, "Relationship between attenuation and phase in feedback amplifier design", *Bell Syst. Tech. J.*, **19**, p. 421, 1940.
9. J. C. West, "Servomechanisms", p. 101, (English University Press, London, 1953).
10. E. C. Titchmarsh, "Theory of Fourier integrals", p. 119, (Oxford University Press, 1937).
11. "B.A. Mathematical Tables", Vol. I, "Circular and Hyperbolic Functions", p. 34, (Cambridge University Press, 1951).
12. H. Fletcher, "Speech and Hearing in Communication", p. 62, (Van Nostrand, New York, 1953).
13. H. S. Black, "Modulation Theory", p. 174, (Van Nostrand, New York, 1953).
14. H. W. Bode, "Network Analysis and Feedback Amplifier Design", p. 196, (Van Nostrand, New York, 1945).

Manuscript first received by the Institution on 31st December 1962 and in final form on 28th March 1963. (*Contribution No. 69.*)

New Speaking Clock using Magnetic Recording

After 27 years of continuous service, during which nearly 800 million calls have been made by London telephone subscribers alone, the London speaking clocks (TIM) using optical recordings, have been replaced by a new installation using magnetic recording. The speaking clock service was started in 1936 with a pair of clocks at Holborn telephone exchange. It has since been extended to the rest of the country and a second pair of clocks was installed at Liverpool in 1942. These two installations are fed into a double 'ring main' round the country.

The constituent parts of the announcement were recorded as concentric photographic tracks on four glass discs. One disc carried the initial 'phrase' and the six different 'seconds' announcements, the next disc carried the 12 'hours', and the last two discs carried the 60 'minutes' and the 'pips'. The recordings were played by scanning them with six beams of light, the outputs from the associated photo-electric cells being combined in sequence to form the complete announcement. The speed of rotation, and hence its time keeping, was controlled by a pendulum swinging freely in a temperature-controlled cabinet. With hourly corrections from the Royal Observatory an accuracy of $\pm$ 0·1 second could be maintained.

The new clocks use magnetic recording but it was found that conventional magnetic tape is unsuitable for this purpose since both the tape and the replay heads would wear out too quickly. Within the last few years, however, a new magnetic recording material has become available that consists of a homogeneous mixture of a synthetic rubber (neoprene) and magnetic iron oxide. This material has already been used for other information service machines. It has a resilient non-abrasive surface and, so long as a thin film of silicone fluid is maintained on it, neither the material nor the reproducing head in contact with it suffer any appreciable wear over long periods.

The constituent parts of the speech announcement are recorded as circular tracks on a thick tube of magnetically loaded neoprene stretched on to a balanced brass cylindrical drum which rotates at constant speed. These phrases are assembled by cam-operated contacts driven by gears from the drum shaft. The drum is driven by a motor running in synchronism with a quartz-crystal-controlled oscillator so avoiding speed fluctuations due to variations in the mains voltage. This oscillator also provides a direct 1000 c/s signal which is broken up into three 100 ms pips at 1 second intervals by means of a pair of cam-operated contacts. The pulse shape is squarer than in the case of the optical recording and provides a more accurate means of automatic correction.

The new clocks automatically check and correct themselves against Greenwich Observatory time once a day and are not normally expected to be in error by more than five-thousandths of a second. This degree of accuracy cannot be passed on to subscribers for two main reasons: (a) delays in the distribution network and lines, (b) irregularities in the motion of the earth make the definition of time rather complex. The published claim is in fact "normally accurate to one-twentieth of a second". Because of their high inherent accuracy the clocks are capable of running for several days without any correction if this should be necessary.

If the two clocks in a pair, which are controlled by completely separate oscillators, become out of step by more than one-twentieth of a second, or should they start repeating announcements which differ in any way, both clocks are automatically taken out of service and the distant centre then supplies the entire country until the necessary correction is made.

Transistors have been used throughout for the new installation, with consequent saving in space and power. If the public mains supply fails the exchange battery takes over without a break.

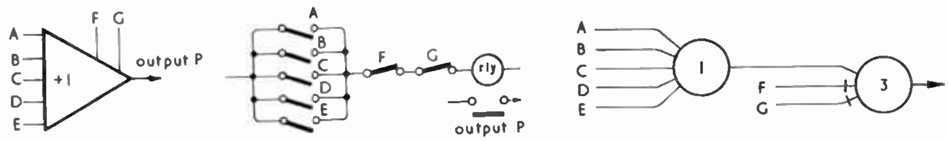
STANDARD FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

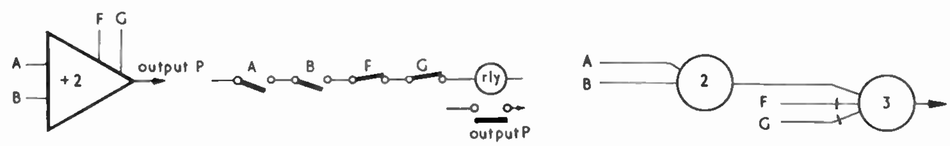
Deviations, in parts in 10^{10}, from nominal frequency for **August 1963**

1963 August	GBR 16 kc/s 24-hour mean centred on 0300 U.T.	MSF 60 kc/s 1430–1530 U.T.	Droitwich 200 kc/s 1000–1100 U.T.
1	− 129·7	− 130·4	+ 23
2	− 129·3	− 130·2	+ 23
3	− 130·6	− 130·1	+ 26
4	− 130·5	− 130·3	− 34
5	− 131·0	− 131·0	− 30
6	− 130·0	− 129·9	− 31
7	− 130·0	− 130·2	− 26
8	− 130·3	− 130·2	− 27
9	− 130·0	− 129·5	− 27
10	− 129·7	− 130·1	—
11	− 129·7	—	—
12	− 129·8	− 130·4	− 20
13	—	—	− 21
14	− 130·8	− 130·7	− 18
15	− 130·8	− 130·9	− 18
16	− 130·7	− 131·0	− 13
17	− 130·5	− 129·9	− 15
18	− 128·9	− 129·2	− 13
19	− 128·4	− 129·9	− 14
20	− 127·9	− 130·3	− 14
21	− 130·6	− 132·3	− 12
22	− 131·4	− 132·8	− 9
23	− 130·9	− 131·1	− 7
24	− 130·1	− 130·5	− 5
25	− 130·0	− 130·4	− 5
26	− 130·0	− 130·7	− 5
27	− 129·6	− 130·7	− 4
28	− 130·8	− 131·0	− 4
29	− 130·1	− 129·7	− 1
30	− 130·8	− 130·8	0
31	− 130·0	− 129·8	+ 1

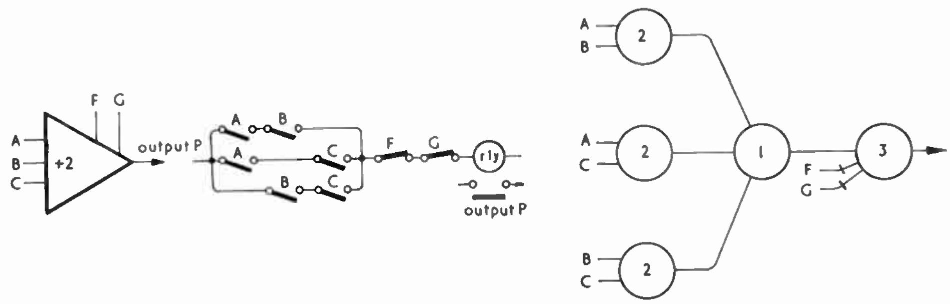
Nominal frequency corresponds to a value of 9 192 631 770 c/s for the caesium F,m_(4,0)–F,m (3,0) transition at zero magnetic field.



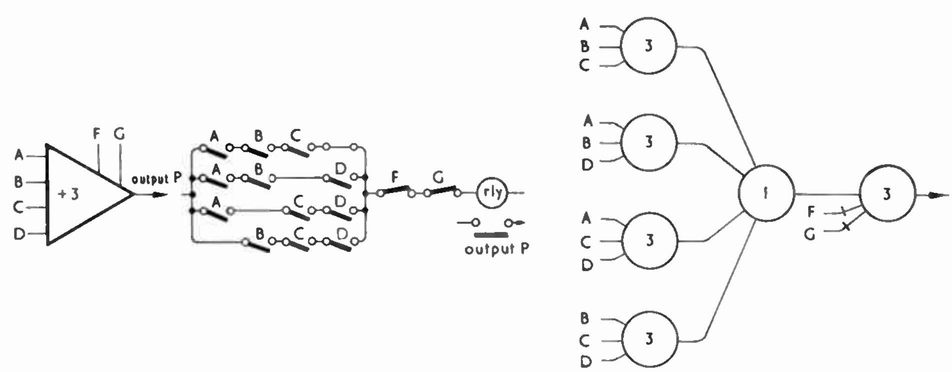
(*a*) PLUS 1 unit.



(*b*) PLUS 2 unit with two inputs.



(*c*) PLUS 2 unit with three inputs.



(*d*) PLUS 3 unit with four inputs.

Fig. 4. Equivalent circuits for combinations of PLUS unit inputs.

The output stage of the PLUS unit has a transistor VT3 connected in an emitter follower arrangement and this reduces the regulation of the unit.

When the unit is 'off' the base of the first transistor VT1 is kept slightly positive by the current through R8 and D1. VT1 is therefore non-conducting or cut off, the second transistor VT2 is fully conducting due to the base current through R9 and R10 and the unit gives no output. The application of −36 V to any one of the inputs is sufficient to overcome the positive bias, and the base becomes negative allowing VT1 to conduct. VT2 is now cut off, the emitter follower VT3 conducts and the unit gives an output.

If the base of VT1 is given additional positive bias by connecting at 56 kΩ selectivity resistor externally, shown as Rs, the single input signal is insufficient to overcome the bias. Two input signals however are sufficient, the unit gives an output and is then termed a PLUS 2 unit.

Similarly, lower values of the selectivity resistor will require three or four inputs for operation.

The second transistor VT2 can also be made to conduct by applying an input directly to its base through the blocking inputs, and this inhibits the output regardless of the input conditions.

The relay contact arrangements which are equivalent to the input conditions are useful for illustration and are informative for control engineers, although this kind of comparison is not the recommended approach for designing a control scheme.

Some of the simpler arrangements for the PLUS unit are shown in Fig. 4. The equivalent circuits using conventional logic symbols are also shown. (N.B. In the relay equivalents certain contact functions are duplicated and are put in for symmetry).

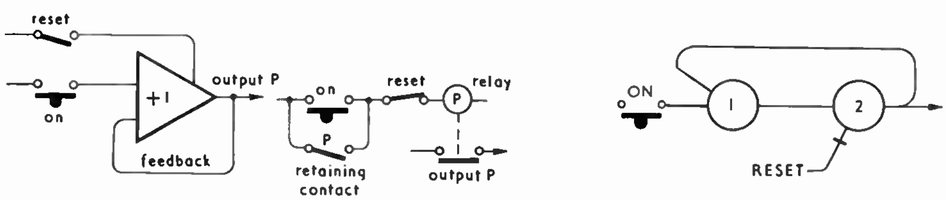
The two blocking inputs can be represented by two normally-closed contacts directly before the output—shown as F, G in all diagrams. Operation of either will prevent an output appearing. The PLUS 1 unit (Fig. 4(*a*)) is effectively an OR circuit. Operation of A or B or C or D or E will cause an output to appear. The PLUS 2 unit however (Fig. 4(*b*)) requires both A and B to be present together, and if we take three inputs to a PLUS 2 unit as in Fig. 4(*c*) the requirements are such that combinations of any two of the three will give an output, that is (A and B) or (A and C) or (B and C). If we extend this to a PLUS 3 unit with four inputs, we get the more elaborate arrangement shown in Fig. 4(*d*) and as there are five input terminals even more complex equivalent circuits can be drawn up.

The PLUS unit has further versatility as shown in Fig. 5. In control systems 'memory' functions are frequently required, in which a momentary event, e.g. a push button being operated, requires to be turned into a continuous signal. This is generally performed by a relay with a retaining contact.

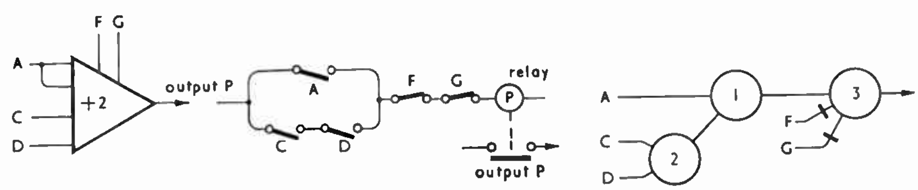
In the multi-function system a signal is fed back from the output of a PLUS unit to one of its inputs. When the push button is operated and released the unit still gives an output due to the feedback connection, and this output is maintained until an 'off' or re-set signal is applied to the blocking input. This is shown in Fig. 5(*a*).

By taking a particular signal to two inputs on the PLUS unit more emphasis is given to that signal. In Fig. 5(*b*) two inputs are required for the PLUS 2 unit and these can be satisfied by either A alone or by (D and C). In fact this input can be made over-riding as shown in Fig. 5(*c*). Here C, D and E combined are insufficient to operate the PLUS 4 unit and the presence of A is essential together with two of the remaining three input signals. The feedback circuit also can use this over-riding feature.

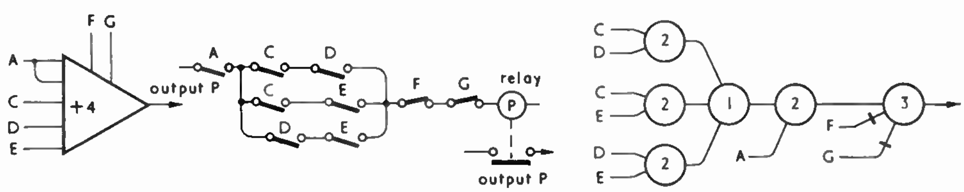
The PLUS unit is the principal unit with which most of the logic in a control scheme is performed, but the following two units are found useful.



(*a*) PLUS 1 unit with push button and retaining contact.

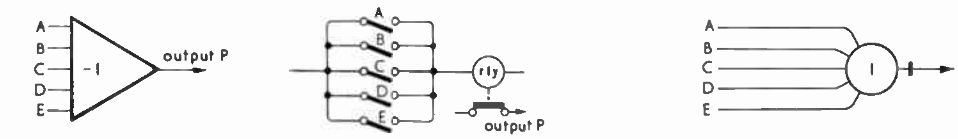


(*b*) PLUS 2 unit with two inputs.

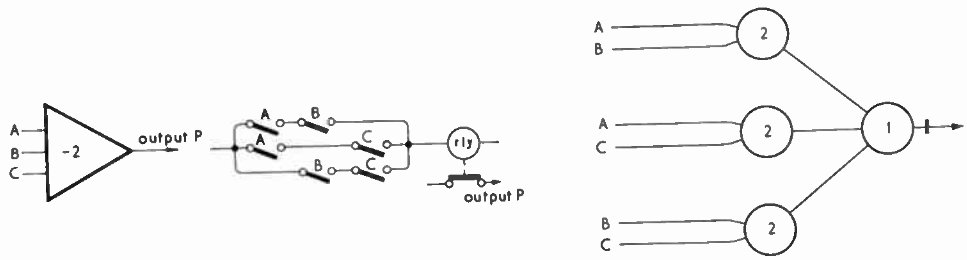


(*c*) PLUS 4 unit using three inputs.

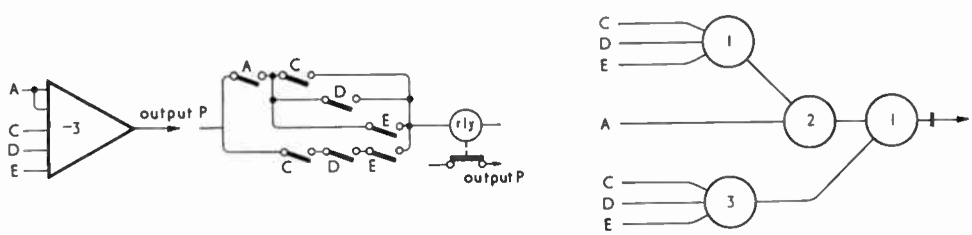
Fig. 5. Equivalent circuits for several particular arrangements of PLUS unit inputs.



(*a*) MINUS 1 unit.



(*b*) MINUS 2 unit with three inputs.



(*c*) MINUS 3 unit using three inputs.

Fig. 6. Equivalent circuits for typical MINUS unit input arrangements.

5. The MINUS Unit

This is a unit which is normally 'on' until certain input conditions are fulfilled for which the unit switches 'off'. As with a PLUS unit the selectivity feature is available so that one, two or three inputs are required to remove the output. This unit is used where inversion is required and can be compared to a relay with a normally closed contact. Several equivalent circuits are shown in Fig. 6.

6. The OR Unit

This unit consists of pairs of semiconductor diodes which can be connected as required to enable more than one signal to be fed into a particular input. The operation is similar to a PLUS 1 unit without the blocking or memory feature. The OR unit would generally be used where possible for economy purposes.

7. Other Units

The above three units will perform all logic required in a control system, but several units are included in the range to perform other necessary functions.

7.1. *Timing Units*

These also have the input selectivity feature and the output appears a fixed time after the input requirements have been satisfied. This time can be adjusted if required over a range of 20 : 1 by means of a potentiometer and the whole timing range covered by standard components is from 50 milliseconds to 40 seconds. By using additional capacitors time intervals of several minutes can be obtained.

7.2. *Power Output Units*

The control system as a whole is intended to operate large contactors, brakes and solenoids etc. which require higher power than is used by the static switches. A range of power output units energized by the static switches provides up to 200 W for this purpose.

7.3. *The Voltage Detection Unit*

This unit is a specialized device which operates from a varying voltage input signal. The pick-up and drop-off points can be independently adjusted, and the unit gives the standard output of -36 V which can be used directly by the logic units. The electro-mechanical equivalent is the voltage relay.

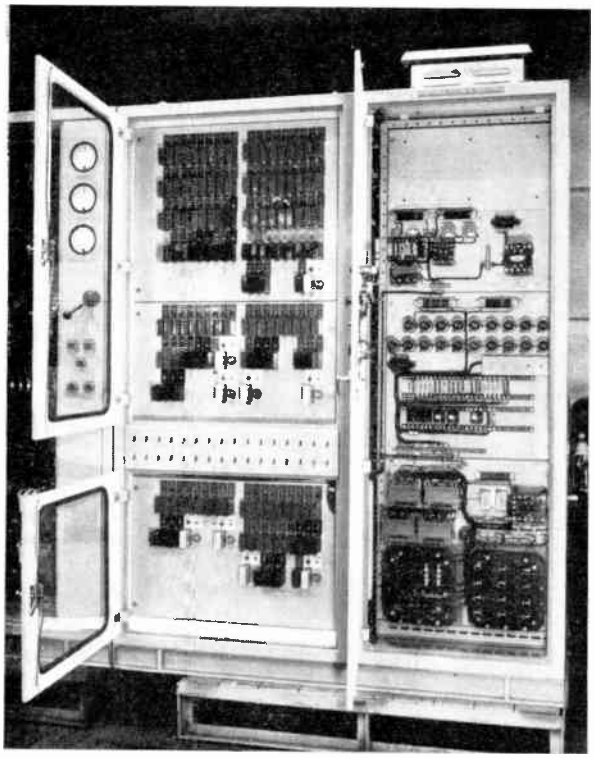


Fig. 7. Blast furnace hoist control incorporating static switching.

7.4. *The Voltage Protection Unit*

In a large scheme where power supplies are heavily loaded the voltage protection unit can be used to monitor the -36 V and $+50$ V supplies and if either deviates from specified limits the equipment is shut down and an alarm sounded.

8. Mechanical Construction of the Units

The units are available in two forms of construction.

8.1. *Encapsulated Units*

This is the original form of the unit. The circuits described above are built up as a network of components on the back of the Jones plug and the whole assembly is encapsulated, giving a very robust unit. The encapsulating medium is coloured for identification and each type of unit has its own characteristic colour: red for the PLUS unit, yellow for the timer, black for the power output units etc. The output transistor of the power output unit is bolted to an aluminium cooling fin which protrudes from the front of the unit.

The units are mounted in rows with a coloured label identifying each socket, and with the neon indicator, if required. Figure 7 shows a typical assembly.

For industries where the control room is liable to be penetrated by corrosive fumes or where rough handling may occur the encapsulated units are particularly suitable.

8.2. *Standard Units*

The appearance of highly reliable connectors for use with printed circuits has enabled the units to be redesigned for use in industries where the mechanical duties are not so rigorous. A 15-way gold-plated plug is used and components are flow soldered on to a printed circuit.

For protection against dust and the atmosphere the unit is given a coat of a specially developed resin. The component side of the printed board is enclosed by a black anodised aluminium shield for protection and handling. This shield is used as a heat sink for the large transistors of the power output units.

The units are identified by coloured labels on the shield.

8.3. *Mechanical Coding Facilities*

The units have been designed so that, as far as possible, insertion of a unit into a wrong socket will not damage the unit, although there may be an incorrect output from the control system as a whole. However, for applications where extreme safety is required a mechanical coding system is provided for the standard units. To each unit is bolted a flange which has an arrangement of holes that is unique for each type of unit. On the mounting racks adjacent to the sockets pegs are fitted which correspond to the holes and only the correct type of unit can be inserted into the socket.

9. Special Facilities of the System for Small Control Schemes

A large control scheme could consist of rotating machines, heavy current switchgear, cabling etc. of which the ancillary control gear is a small part. Consequently the extra expense due to the particular features of static switching tends to become lost in the total cost. Many control schemes in industry perform simple operations, however, and there are a number of economic problems to be overcome before a static switching system can cater for this type of application.

A small power press for instance may be controlled by half a dozen relays operated directly from the mains supply and mounted in a box on the nearest wall. To perform the same operation with static switches, the cost of mountings and power supplies for the units could be a significant proportion of the total cost.

9.1. *Mounting Arrangements*

For mounting static switching units an assortment of folded 'top hat' section steel sheeting has been designed. This can be assembled into a frame by an

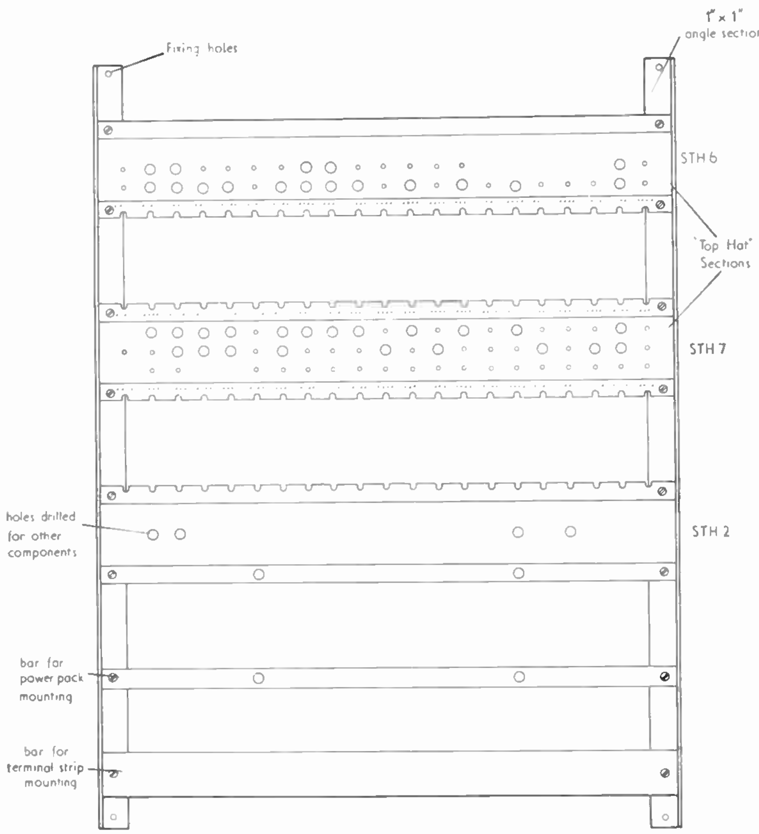


Fig. 8. Typical mounting rack for standard units.

unskilled fitter and Fig. 8 shows an assembled rack before sockets are installed. This frame may then be wired up on the bench and the whole transferred when complete to the cubicle for the insertion of units and testing.

9.2. *Power Supplies*

To reduce the cost of power supplies a small simple unit capable of supplying about 12 logic units has been introduced. An external alternating voltage is required from a small mains transformer or from additional windings on a transformer already required elsewhere in the system.

For the larger scheme standard power units, stabilized against input mains variations, can supply up to 120 units.

9.3. *Operating Output Devices*

Output devices such as clutches, contactors, solenoids etc. are generally operated directly from the factory mains supplies, either a.c. or d.c. Power output units, however, can switch only a maximum of about 40 V d.c. due to the limitation of existing transistors. Additional power supplies have to be used to provide power at 40 V for the output devices and the cost can be significant for a small scheme. If an existing control scheme is being converted for static switching, operating coils may require replacement by those operating at 40 V.

For control schemes operating only four or five output devices a new type of unit has been developed which can accept the low level output of the static switch and operate a mains energized device directly. This unit uses the silicon controlled rectifier semiconductor device and it is small enough to be fitted directly to the mounting racks.

9.4. *Designer's Manual*

An unexpected problem arises when the static switching equipment supplier is requested to design a control scheme. The ensuing analysis, may, in fact, cost more than the equipment is worth.

This is essentially due to the static switching designer being unfamiliar with the particular problems involved. Therefore the obvious person to design the control scheme is the factory engineer, who is in possession of all the facts, and who very probably has considerable experience in handling relay circuits.

Relays and associated equipment have been purchased from suppliers, and control equipments have been built by factory engineers for many years and it

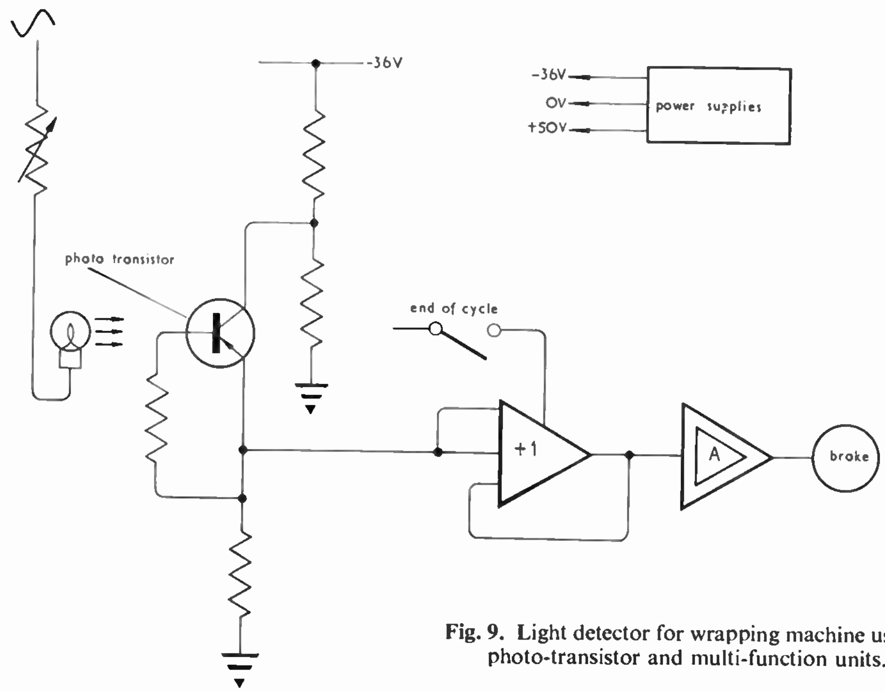


Fig. 9. Light detector for wrapping machine using photo-transistor and multi-function units.

is this attitude that should be encouraged in industry with regard to static switches.

Consequently a considerable amount of thought has been put into producing a designer's manual. In the applications section of it the designer is introduced to each unit, and its versatility is explained. Control scheme examples of increasing complexity are designed in detail and mistakes in approach to a problem are made and corrected, illustrating the desirable techniques to be acquired. An engineer who understands the fundamental operating requirements of his control system and knows the techniques of solving the problems with relays will in a short time become sufficiently familiar with the static switching units to design his own schemes.

Optimum design will obviously only come with experience and an engineer will probably at first require 20 units to perform a function which would reduce to 18 as he realizes the adaptability of the units. This is of course true of all industrial design.

The first part of the designer's manual, intended for quick reference, consists of data sheets giving condensed information on all the units of the system.

10. Application of the Multi-function System

Two particular applications will be described, the general problems only being discussed and not the detailed design.

10.1. *Wrapping Machine Control*

A preprinted reel of paper must be held in register with respect to the cutting and wrapping operations. Small holes are punched in the paper for location purposes and at the correct position the paper is stopped, guillotined and the cut-off portion wrapped around the customer's product.

Originally electromechanical sensing fingers detected the holes and relays applied the brake. This was not entirely reliable as the paper and holes were liable to wander.

The problem was solved by a photo-transistor feeding directly into a single PLUS unit (Fig. 9). The holes were detected by light passing through them and falling on to the photo-transistor and the PLUS unit is here used as a simple amplifier. The signal-to-noise ratio was minimized by masking the holes with a narrow slit (which was made longer than necessary to allow for paper movement). A 2 : 1 input voltage ratio was obtained and setting-up was accomplished by adjusting the brightness of the lamp with a variable resistance.

After the detection of the holes the PLUS unit locks on due to a feedback connection and the brake is applied to stop the paper (in fact the holes will have moved on due to the inertia of the mechanical system). On completion of the wrapping operation a brief signal is applied to the blocking input which turns the unit 'off', and the paper moves on. A similar circuit without feedback was also used for controlling the slack loop of paper.

D.c. power was already used in the machine and this was modified for use with the power output units.

10.2. *A Blast Furnace Hoist Control*

In operating a blast furnace absolute reliability is of prime importance. The furnace is expected to function continuously for ten years and shut downs of more than a few minutes can only be tolerated after careful production planning concerning the whole steelworks production line.

Towards the end of 1961 a blast furnace hoist control was commissioned in which the automatic sequence and program controls were performed by multi-function units. This includes operation of the Ward-Leonard speed control for the skip hoist, and control of the hopper doors, sealing bells and load distributor on the top of the furnace. A program for loads of coke and ore is set up with switches in the control room and the system ensures that the furnace is charged in accordance with this program.

It is essential for the proper control of the furnace that if the power supplies fail or are removed for maintenance, the control is set up at the same point in the sequence afterwards.

This permanent 'memory' feature was provided by magnetic toroidal inductors with a square hysteresis loop. The inductors were used in conjunction with binary units, a type of circuit which has two states, either 'on' or 'off'. The direction of the magnetic flux after removal of power determines the state that the binary unit takes up when the power is returned. These units perform the same function as the mechanically-latched relay.

The load program control of the furnace also used power-off-memory binary and other units arranged to simulate a uniselector.

The static switching equipment for the furnace control performed more complex operations than the relay control that was replaced yet required no more space. Consequently no enlargements to the control room were required which in this particular case was important.

11. Cost of Static Control

An engineer investigating static switching for a particular control is always concerned with the comparative costs of the equipment. Direct comparison between relay and static switching solutions for a particular problem are interesting but will not usually be informative about the general problem and, in fact, can often be misleading. Some applications may be particularly awkward and static switching could cost many times the price of the relay counterpart.

If a large conventional relay control scheme is compared in cost with an equivalent static control scheme, the price of the two schemes would probably be approximately the same. However, for a static control scheme that would be considered large as far as static switching is concerned, the control probably could not have been carried out with relays in any case.

For smaller controls the cost of power supplies and mounting becomes increasingly significant and it was once generally held that for applications that would require less than 40 relay contacts static switching would be less competitive. However, with the wider range of units and accessories now offered, simple applications requiring only two or three units can be considered.

11.1. *Maintenance and Shut-down Costs*

Of greater significance is the saving of routine maintenance and of shut-down time due to faults occurring. Production efficiency may also be improved due to increase in operating speed and a few actual examples may be informative.

A particular process line control included many interlocks performed by about a hundred relays using an average of about four contacts each. Shut-down due to control failure was amounting to about a dozen hours a year of which perhaps two or three could be tied down to relay contact trouble. The cost of shut-down is of the order of £1,000 an hour so that a reduction of shut-down time by a half amounts to a saving of thousands of pounds and the difference between relay and static switching initial costs becomes insignificant. In this particular factory every contact was cleaned monthly.

Correction of faults in the multi-function system of static switching consists of pulling out the units associated with the faulty part of the system and plugging in a spare until correct functioning occurs. If neon indicators are mounted with the units the effect of the new unit can be readily seen. In a few minutes, all units could be tested by substitution by an unskilled maintenance worker who did not understand the system. Test facilities can be built into a cubicle if required.

11.2. *Casting Machine Economies*

An example of another sort is given by a metal casting machine of a type in which the wall thickness is controllable. This was originally a manual operation and the castings were tested for wall thickness by weighing, the satisfactory castings lying between an acceptable minimum and an acceptable maximum weight. The range of weights actually produced by the operator during manual control generally lay towards the top of the range, medium to maximum acceptable weight. On replacing the operation by static switching it was possible to control the wall thickness more accurately and towards the bottom of the range, minimum to medium, resulting in a lower average weight for the castings.

The saving of raw material alone enabled the cost of the entire equipment to be recovered within three months and the production line was also speeded up.

12. Future Developments

When a fault occurs in a control incorporating static switching it is usually found to be due to the input devices, limit switches, cam switches etc., and these can be considered the weak link.

A range of input units is being produced which uses the dry reed switch. This is a highly reliable device capable of many millions of operations and, if suitably mounted, it can be made very robust. Operation of the switch is by a magnetic field, generally from a permanent magnet.

Although the multi-function static switching system has been specially adapted for use with small schemes, it is capable of being extended to the largest ones. A range of binary units is already being used for more complex controls, but detailed information has not yet been generally published. It is found necessary to have a fair amount of contact with typical control engineers in various industries in order to draw up the correct sort of information to include in an instruction manual.

The range of units is being extended to include analogue units such as amplifiers and other devices, for use in regulating and servo control applications.

13. Conclusions

The possibilities of static switching devices for large and complex applications in industry have been well known for some time. With the additional facilities described above it is considered practical to extend the use to all industrial applications. The increased speed of operation and reduced maintenance and shut-down time of static switching systems will increase the productivity of factories in which it is employed.

It is expected that eventually the control engineer will stock, experiment with and build up circuits containing static switches and will become as familiar with them as he is now with relays and associated devices.

14. Acknowledgment

The writer is indebted to the Directors of Associated Electrical Industries Limited for permission to publish this paper.

15. Bibliography

1. T. E. Barany, "Some recent advances in industrial electronic control techniques", *Trans. Soc. Instrum. Tech.*, **12**, p. 118 September 1960.
2. J. R. Barrat and T. E. Barany, "Logicon static switching: a shear control for Steel Peech and Tozer", *A.E.I. Engineering*, September 1961.
3. J. T. Gledhill, B. P. Jones and G. B. Hool, "A furnace charging system using Logicon static switching", *A.E.I. Engineering*, November/December 1962.
4. D. S. Simons, "Static switching—A new tool for the control engineer", *Rugby Engng Soc.*, **52**, p. 60, 1961/2.

Manuscript first received by the Institution on 19th March 1963. (Paper No. 846.)

DISCUSSION

Under the chairmanship of Professor A. D. Booth

Mr. J. R. Halsall: Mr. Cargill mentioned the high cost of the special low voltage d.c. supplies in relation to small scale applications of the 'Logicon' system. In this connection I would like to point out the widespread use of 24 V d.c. floating battery stand-by supplies for generating alarms and telecommunication equipment in modern chemical plants. I personally feel that equipment of the type described by the author would find more ready acceptance if it were designed to operate from the existing 24 V supply. At present all the half-dozen or so competing systems available on the market require different supply voltages so that it is not easy to combine the best elements of each in a single system without a high investment in special power supply units. Is there any good reason why this type of equipment should not be designed to operate from existing 0 and − 24 V supplies with or without a + 24 V rail?

The Author (*in reply*): A static switching system generally supplies output devices, such as contactors, clutches etc., and a relatively large power, low d.c. voltage power pack, is required for this purpose, usually supplied from the a.c. mains.

A 24-V battery supply would be quite suitable for the output devices providing the positive rail could be connected to the neutral rail of the static switching system.

The logic units would however, require separate battery power supplies for + 50 V and − 36 V, although these are of relatively low power, i.e. 1 or 2 A. The 'Logicon' system is intended for industry in general in which standard d.c. power supplies range up to 110 V. It was found that no standard supplies were sufficiently universal to be adopted and the particular voltages used were selected for the reasons given in the paper. A static inverter can be used to provide the required operating voltage from existing d.c. supplies.

The 'Logicon' system could be adapted to a − 24 V, + 24 V system if the supplies were sufficiently stable but, as the units would be to some extent of special design, the cost may be higher. The facility for neon indication of unit operation would be lost and the possibility of input switches failing may be slightly increased due to the lower operating voltage.

Mr. S. L. H. Clarke: The question of possible economies of relays over static switching elements has been raised. We have done a considerable amount of work analysing the number of elements needed to mechanize a logical mesh using relays, NOR units, AND/OR/NOT, etc. I am quite sure that the result is equally applicable to PLUS units. This result is that the difference in a reasonable sized mesh is insignificant between one 'sufficient set' and another. In a mesh multi-input with the single output is no better and no worse than the other way round. In fact I believe that we spend much too much time arguing this very point.

The Author *(in reply)*: We have arrived at the same conclusion concerning the minimum number of elements to solve a sufficiently large logical problem, with semi-conductors. In such a problem the ideal solution would be a tailor-made circuit.

When designing a building block system for general industry the circuit design must be a compromise between sufficient versatility and the minimum redundancy of components, which is compatible with the mechanical and overall economic consideration.

The 'Logicon' PLUS unit has three logical elements associated with it, combinations of which make it quite versatile. When the system was first proposed some indication of redundancy was desirable and several differing schemes were designed using 'Logicon' and an established NOR system. After equating the mechanical facilities of the two systems it was found that the overall cost was significantly the same.

An elementary building block system requires a certain amount of inter wiring whereas the logic elements of a PLUS unit are already interconnected and the comparison shows that the cost of additional redundancy in the 'Logicon' units is offset by the reduction in wiring.

Although the number of logic elements to perform a sufficiently large problem in 'Logicon' or another system would be the same, the number of actual building blocks would be less for 'Logicon'.

I agree with the multi-input—single-output etc. comments. Theoretically the logic is no more involved either way round. However, most control schemes are many-input—few-output problems and a layman would expect a multi-input—single-output logical unit to be easier to apply than the vice versa device. Relays, which came first, are vice versa and the established control engineer has no difficulty in single-input—many output reasoning.

Mr. A. Acres: In the post commissioning period of a Logicon control panel how do the many plug and socket connections affect the reliability?

The Author *(in reply)*: Encapsulated units with the Jones type plug were first installed in 1959 in a steelworks. Many other applications in various atmospheres have subsequently been commissioned and at no time during or after commissioning has there been any reason whatever to suspect the performance of the plug and socket. The Jones plug has been used on military equipment for many years.

The standard units with the 15-way gold-plated plug and socket have been in service for a year, during which time there has been no failure but this is not considered sufficient time to make firm claims on reliability. The standard units are recommended for general industry at present but when our investigations are complete we may be able to extend the recommendation. We are aware of one deficiency in the plug when used in corrosive atmospheres. We have taken steps to overcome it.

Television Standards Conversion

AN ELECTRONIC SWITCHING SYSTEM DESIGNED BY THE B.B.C.

The first television line standards converter, used by the B.B.C. since 1952, converted 819-line pictures originating in France for viewing on 405-line receivers in Britain.† The 'image-transfer', or 'optical' converter as it is called, consists of a television camera, operating on the required standard, viewing a display on a high-quality picture tube of the programme on the original standard. This is a relatively simple process which is still being used frequently for the exchange of programmes between the United Kingdom and other European countries.‡ It has also been adapted for conversion between American and European standards and is used whenever live transatlantic television programmes are transmitted via a communications satellite, or for broadcasting American and Canadian programmes in the U.K. from video-tape recordings.

One of the serious limitations of the optical converter is the resulting poor definition; focus and contrast of the original picture must be continuously controlled to maintain a reasonable standard of the converted picture. In addition there are apparent reflections between the optical surfaces, giving ghost images; the noise level is increased; rapid movement in the scene and panning produce 'smearing'.

These limitations, coupled with the need for constant skilled attention in the setting up and operation are fundamental to the system. Although the results of using an optical converter have shown to be of an acceptable, if not desirable quality, it has long been apparent that a simpler and more stable converter could be built if the optical-to-electronic transfer process could be eliminated.

When B.B.C.-2 starts in April 1964, in the London area, it will be on 625 lines whereas B.B.C.-1 will continue for some years on the present standard of 405 lines. It will therefore be important to have a reliable means of converting pictures from one standard to another without appreciable loss of quality, so that programme material recorded on video tape on the 625-line standard can be reproduced in either B.B.C.-1 or B.B.C.-2 as required. Standards conversion will become even more important when the time comes to transmit B.B.C.-1 on 625 lines also, because there will be an interim period, perhaps lasting some years, during which B.B.C.-1 will have to be broadcast on both standards. It is therefore essential that converters should be able to operate for long periods without attention and without appreciable loss of picture quality. This has been one of the most difficult problems in planning the change-over, because the converters hitherto available introduce some loss of picture quality and require continuous attention and adjustment. The availability of reliable converters at appropriate points in the network would make it possible for B.B.C.-1 to be originated on 625 lines and transmitted on that standard from some of the transmitting stations while still being transmitted on 405 lines from other stations. There is also a continuing need for standards conversion for Eurovision programmes, most of which originate on 625 lines and have to be converted to 405 lines for B.B.C.-1.

Electronic Conversion

The new converter, which has been built and tested on a broadcast programme, operates directly on the television signals of a 625-line transmission.§ The process used is perhaps best understood by imagining that the picture is

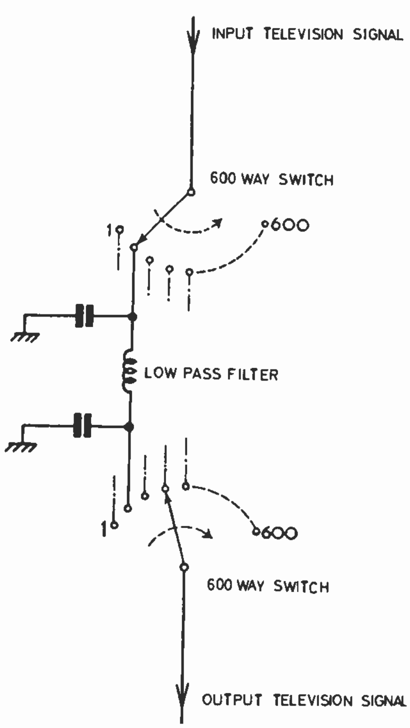


Fig. 1. The principle of the standards converter.

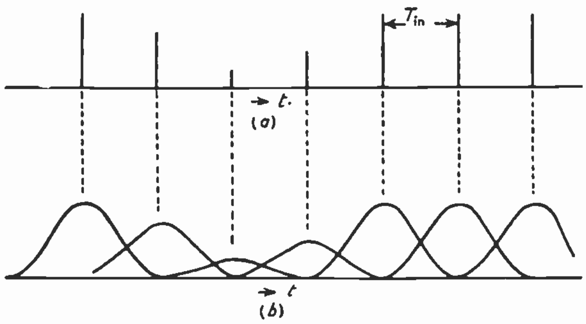
cut up into a large number of narrow vertical strips. Each line of the incoming signal may be considered as a number of picture elements defining the horizontal definition, say 600. A switch, actually electronic but shown diagrammati-

† T. Worswick, "The B.B.C. television standards converter", *J. Soc. Motion Pict. & Television Engrs*, **68**, pp. 130–5, March 1959.

‡ E. R. Rout and R. F. Vigurs, "A wide range standards converter", *J. Television Soc.*, **9**, No. 12, pp. 494–503, October–December 1961.

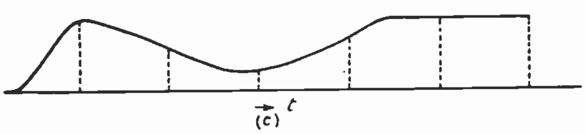
§ P. Rainger, "A new system of standards conversion", "Television Engineering", Proceedings of the International Television Conference 1962, pp. 170–2 (The Institution of Electrical Engineers, London, 1963).

T_{in} = incoming line duration.



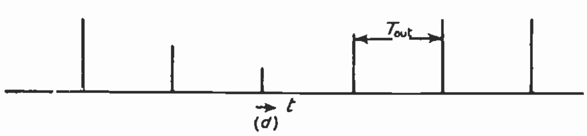
(*a*) Input train of pulses.

(*b*) Network response to each pulse.



(*c*) Total output of network in response to (*a*).

T_{out} = outgoing line duration.



(*d*) Pulse train resulting from sampling at output rate.

Fig. 2. Input and output signals of a network.

cally as a uniselector in Fig. 1, having a single wiper and 600 contacts, connects the picture signal in turn to each of the contacts, at the same rate as, and in perfect synchronism with, the line scanning of the 625-line input signal. At the outputs of the switch 600 samples of the picture signal are obtained spaced at equal intervals of time along any one line. As the switch continues to operate in synchronism with the scanning, all lines of the picture are sampled successively and any one contact of the switch will therefore carry samples of the waveform at line-period intervals; its output will be a series of pulses as shown in Fig. 2(*a*) spaced at the incoming line interval, in this case 60 μs. Thus, any one contact of the switch can be identified with a vertical strip of the picture, one picture element in width.

The components of the line repetition rate are eliminated by constructing the envelope of the pulses in the storage filter network shown connected to the switch in Fig. 1. The resulting smoothed pulses are shown in Fig. 2(*b*) where it is assumed that the response of the filter to a given pulse has died away by the time the response of the succeeding pulse has reached its maximum value. By the usual addition properties of the filter, the individual responses will combine to produce the output shown in Fig. 2(*c*).

The signals at the outputs of the 600 storage filters are sampled by a second high-speed switch which rotates at the line frequency of a 405-line signal and which is also shown in Fig. 1. This process results in a series of pulses spaced, one from the next, by a time corresponding to one picture element duration; in this case 1/600th of the duration of a line scan in the 405-line system. Thus, the picture is built up element by element and line by line by the series of pulses, and the required waveform is obtained by the insertion of a filter limiting the bandwidth to that normally occupied by a 405-line picture. That portion of the converted picture which is obtained at the output of the second high-speed switch from any one storage filter is illustrated in Fig. 2(*d*).

It should be noticed that the filter shown in Fig. 1 has stored the individual pulses and spread them out over two line periods; the characteristics of the network determine the individual contribution of two adjacent pulses to the waveform of Fig. 2(*c*). This is described as the interpolation characteristic of the network, and on it partly depends the

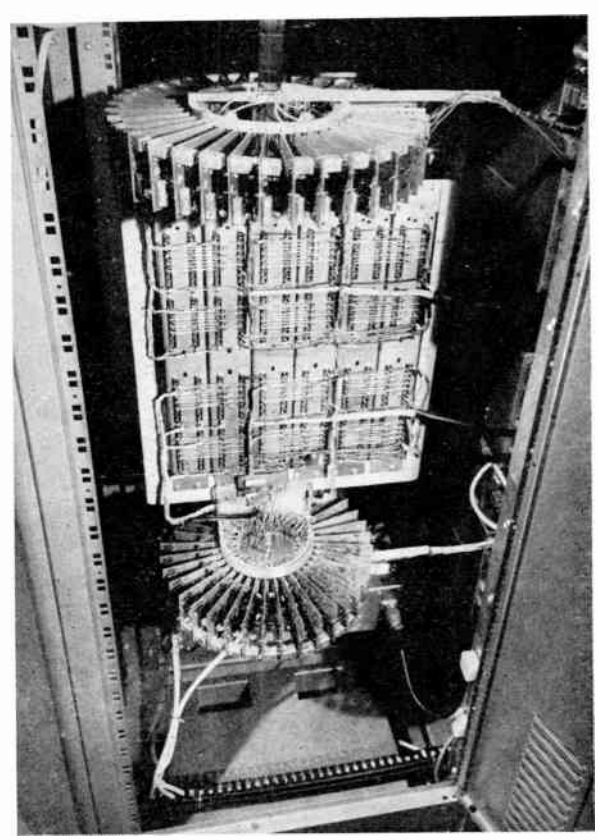


Fig. 3. Rear view of video circuitry showing the input (625-line) switch at the top, storage and interpolation units in the centre and the output (405-line) switch at the bottom.

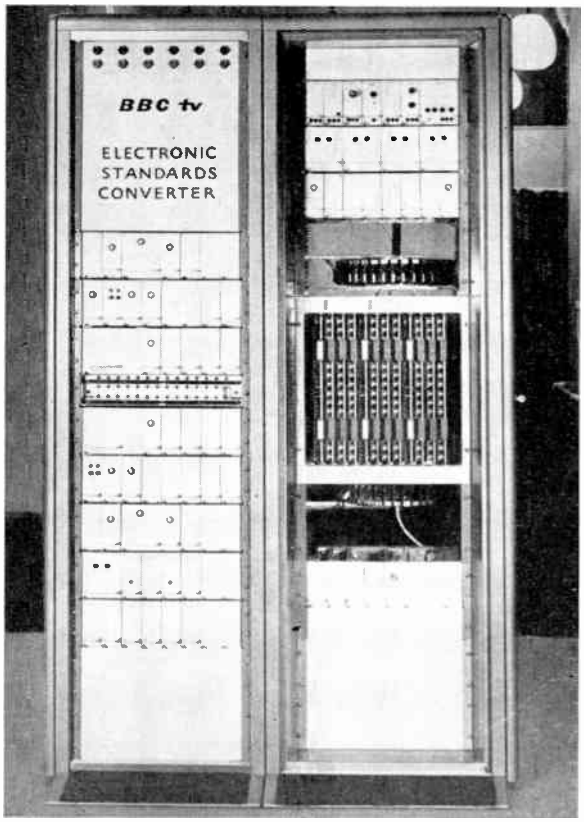


Fig. 4. Front view with doors open. The left-hand bay contains power supplies and pulse generators. The right-hand bay contains all video circuitry.

quality of the converted picture. If the pulses were spread out too much in passing from 2(*a*) to 2(*b*), vertical definition would be lost and 2(*c*) would not faithfully express the envelope of 2(*a*). On the other hand, if the waveform in Fig. 2(*c*) contains any substantial components of the original line frequency, spurious patterns will be generated when this is re-sampled at the new line frequency.

This explanation has been simplified by treating the switching as if it were performed by uniselectors. In fact, because of the high speed of operation required, each contact in each of the two banks shown in Fig. 1 is made by a high-speed on-off electronic switch. To reduce the number of such switches required, the switching is actually performed in stages, there being 36 high-speed switches and 576 slower-speed switches. Some of the switches operate at a rate corresponding to 18 million rev/min, and the apparatus includes over 2000 transistors. The constructional techniques employed in the converter are shown in Figs. 3 and 4.

The techniques used in the converter are in principle applicable to a change between any two line standards, but they cannot at present be used when it is necessary to change the field frequency.

Conclusions

There is no doubt that there is a notable improvement in the results of the electronic converter as opposed to those of the optical converter. It is difficult to detect on viewing that the converted signal picture is very much inferior to a 'live' 405-line picture. At a demonstration of the system (given to the press recently) it was sufficient to show that, apart from the expected differences in quality between 405-line and 625-line pictures, the strobing effect, noise, contrast and vertical resolution were considerably better than has been achieved with the optical system. There was one instance during fast camera panning that revealed a 'break-up' of the vertical resolution, but this does not appear as bad as it would have done by optical conversion It would have been interesting to a see a 'live' 405-line picture as well as the 625-line converted signal pictures: the demonstration was in fact given with a 625-line video tape recording as the subject.

U.D.C. 534.24: 550.81

Adaptation of Sonar Techniques for Exploring the Sediments and Crust of the Earth Beneath the Ocean

By
J. B. HERSEY,†
S. T. KNOTT,†
D. D. CAULFIELD,†
H. E. EDGERTON†
AND
E. E. HAYS†

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

Summary: The 'continuous seismic profiler' is an instrument system for studying the structure of sediments and rock beneath the sea floor by the seismic reflection method. Pulses of sound reflected at near-normal incidence are automatically timed and correlated by the recording method used in echo sounding. The sound source is chosen to give a repeatable pulse having a broad spectrum rich in low frequencies. One sound source called the 'sparker' has been developed into a unit with a power supply delivering 25 000 joules. A second source called the 'boomer' generates a sound pulse by means of the repulsive force induced between a flat spiral coil and a contiguous plate of aluminium when a large pulse of current passes through the coil. The latest 'boomer' is designed to accept 13 000 joules of electrical energy. Recordings of good quality can be made in 3000 fathoms of water from a ship proceeding at 6 knots and echoes have been recorded well over one second after the initial bottom reflection.

These sound sources and recording techniques have also been applied to studying layered structures in the sediments by the refraction method and by studying reflection of sound from these layers as a function of angle of incidence.

1. Introduction

The 'continuous seismic profiler' has been developed at Woods Hole to measure the shape and acoustic properties of geological structures in water-covered areas by the reflection and refraction methods. This system is similar to others developed in the past.[1, 2] Its operation at sea and the recording techniques used are similar to those of echo sounding. A pulsed sound source is employed having a broad spectrum somewhat like that of a high order explosion, but it is electrically timed to repeat on a precise, predetermined schedule. The receiving equipment is designed to accept frequencies in the range 20 c/s to 10 kc/s and has an almost flat frequency response. The schedule of the source is determined by the sweep repetition rate of a graphic recorder similar to those used in echo sounding.[3] This recorder is used to correlate successive echoes from the same reflecting surface, as in echo sounding. The received echo train is fed to a magnetic tape recorder on which the time-base of the graphic recorder is also imprinted, making it possible to re-examine observations in the laboratory after they have been made. In addition to the correlation made by eye with the aid of the graphic recorder, various filters have been employed to suppress noise and unwanted reverberation, to discriminate between echoes from deep structures and shallower side echoes, and by emphasizing high frequencies in broad bands to resolve events having slightly differing travel times. Various data processing devices have proved to be effective, such as varying the sweep speed of the graphic recorder to provide the best possible resolution and information rate, half-wave rectification to emphasize phase relations, and gating of the graphic recorder to improve signal/noise ratios.

2. The Sound Sources

Two sources, the 'sparker' and the 'boomer', have been used. Each is energized by discharging a large capacitor bank through a suitable switch. The 'sparker' has a pair of electrodes, one small, and one large and remote from the first.[4] Energy is released mostly from the smaller electrode as expanding gas formed in the discharge. The initial sound pulse is followed by at least two smaller pulses resulting from expansion and compression of the gases (Fig. 1). The 'boomer', developed by Edgerton,[5] is a flat, spiral coil embedded in plastic against which one or two aluminium plates are held firmly by springs in compression. The pulse of current from the discharging capacitor generates eddy currents in the aluminium plates repelling them suddenly from the coil. The sudden outward motion of these plates

† Woods Hole Oceanographic Institution, Massachusetts, U.S.A.

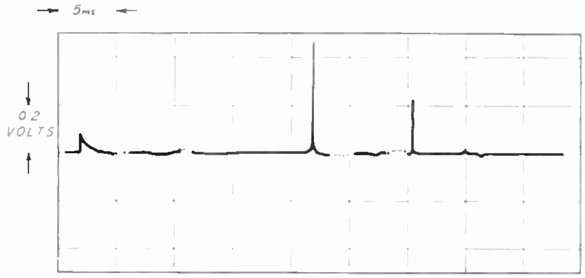


Fig. 1. The 'sparker'. Pressure *vs* time measured at 16 metres by an Atlantic research line BC-10 pressure gauge (sensitivity—117 dBV rel. 1 microbar). The surface-reflected pulse train has been removed.

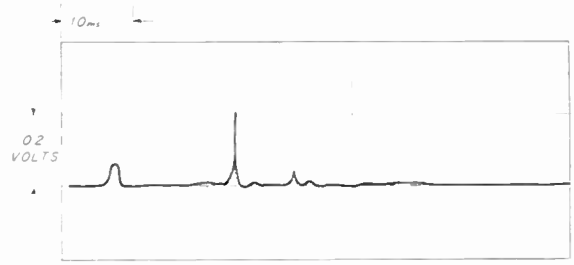


Fig. 2. The 'boomer'. Pressure *vs* time measured at 8 metres by an Atlantic research line BC-10 pressure gauge (sensitivity—117 dBV rel. 1 microbar). The surface-reflected pulse train has been removed.

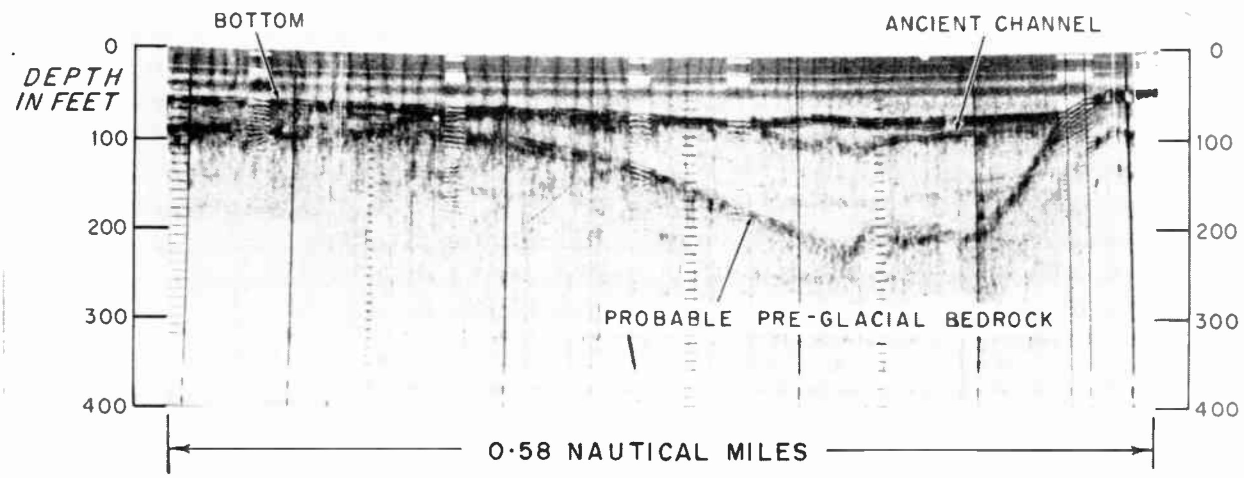


Fig. 3. A buried channel in Narragansett Bay, R.I., U.S.A. Source: a 4-joule 'sparker'.

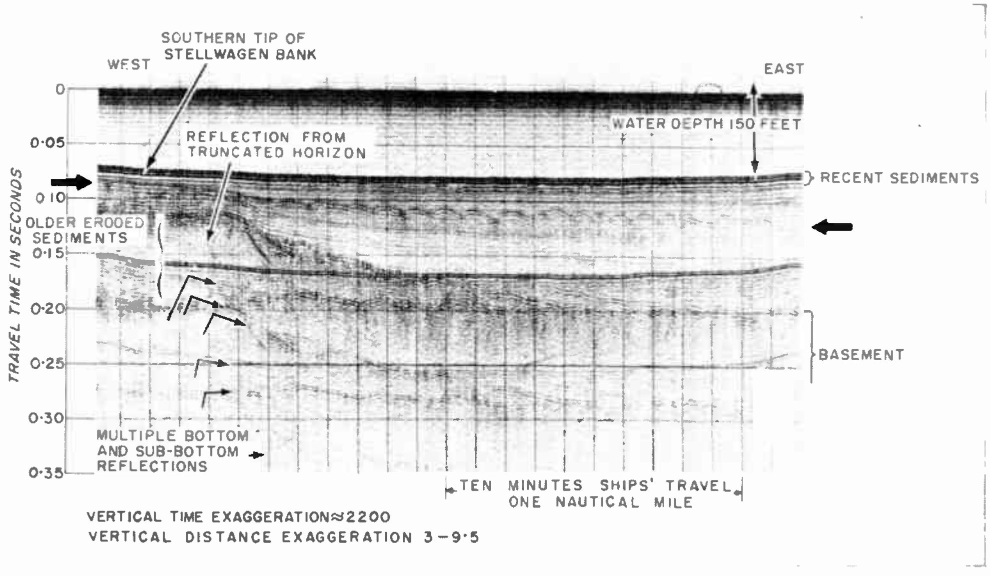


Fig. 4. Reflection profile in Cape Cod Bay, Massachusetts. (*Photograph by Hoskins and Knott.*)

creates a mass of cavitation bubbles between the coil and the plates which acts as a spring causing the plates to oscillate toward their rest position against the coil, radiating pulses somewhat similar to the 'sparker' (Fig. 2).

The source first used was the 'sparker'. Starting with 4 joules, the electrical energy stored for a single discharge has been increased to that of the present 'sparker' (25 000 joules, 500 μF). Meanwhile, the 'boomer' (formerly called the 'thumper') has been developed from an initial 1000 joules to the present model of 13 000 joules and 2000 μF.

3. Receiving Techniques

During early studies in Narragansett and Buzzards Bays, and later in Cape Cod Bay, the problems of identifying and allowing for the second and higher order reflections from the same surface became evident. Examples of the early recordings with the 4-joule 'sparker' are given in Figs. 3 and 5. In Fig. 3 most of the recording was done with a full-wave rectifier feeding the recording stylus. This causes the stylus to mark the recording paper on both positive and negative excursions of the electrical signal. When either high frequencies, low stylus speeds, or both are employed the signal is continuously recorded on the paper. Variations in average intensity appear as a variable density of the recording. Five times during the recording, the rectifier was changed to half-wave working thus displaying only the positive excursions of the wave. This mode of recording was found to bring out the correlation of successive sweeps clearly. The half-wave rectification is commonly used now; the fine detail which can be achieved with it is clear from Fig. 4, taken from the study by Hoskins and Knott in Cape Cod Bay, Massachusetts.[6] The second and third bottom reflections can be seen in Fig. 4. In places many multiple bottom-surface reflections were recorded from the early Narragansett Bay studies (Fig. 5).

As the stored energy was increased in successive models of 'sparker' and 'boomer', it became possible to obtain sub-bottom information in 400–500 fathoms with 1000-joule sources. In 1961, deep-sea recordings were successfully made with a 5000-joule Boomer.[7] Figure 6 is an example taken from this work.

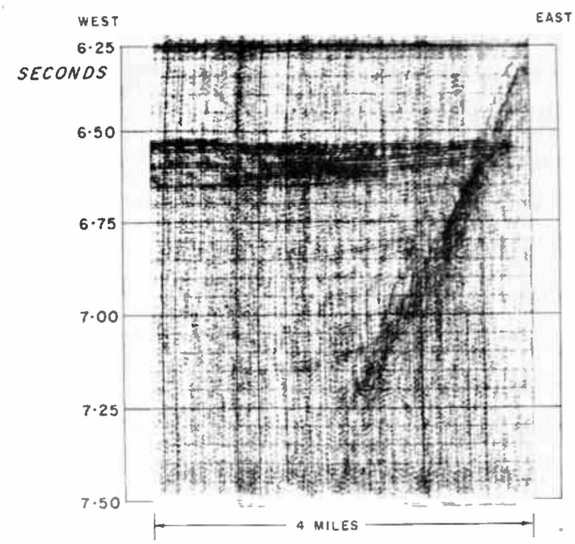


Fig. 6. Reflections interpreted as coming from layers of sediment laid down around Caryn Peak. 36° 35′ N—67° 55′ W.

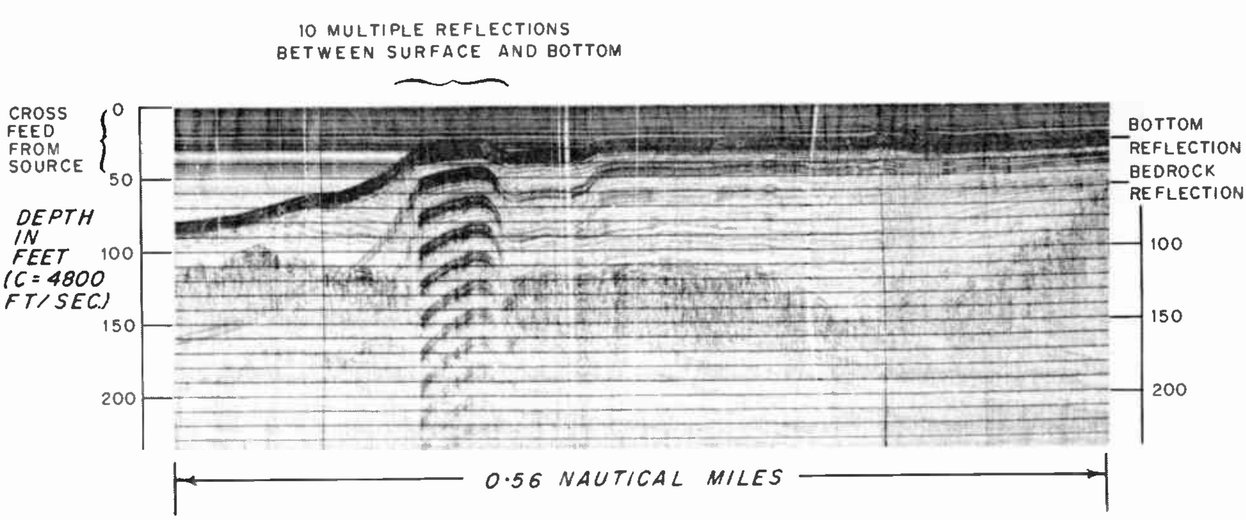


Fig. 5. An example of a highly reflective sea floor. Note that sub-bottom reflections cannot be traced through the part of the profile while multiple bottom echoes are strongest.

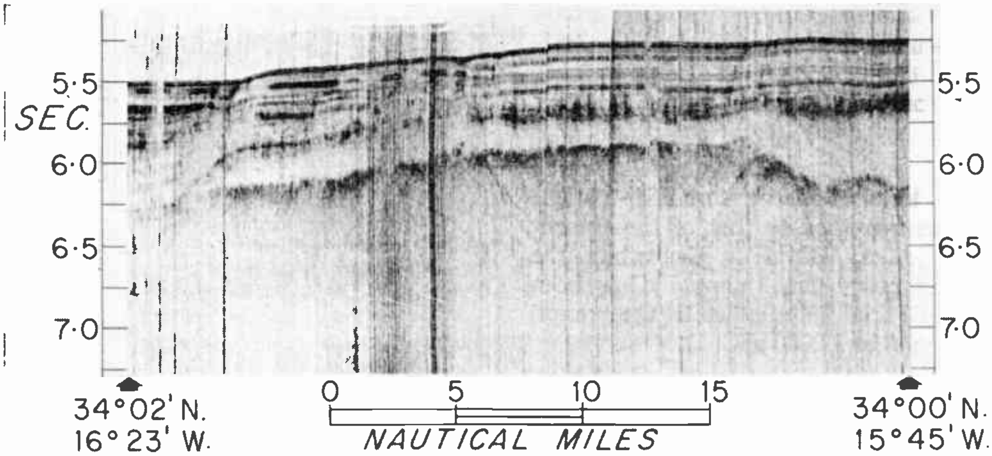


Fig. 7. A reflection profile showing sediments probably burying an ancient series of volcanic flows north of Madeira.

This increase in stored energy greatly improved signal/noise ratios, as illustrated by Fig. 7, a record taken from magnetic tape recordings of reflection profiling north of Madeira Island during the *Chain* Cruise 21. Subsequent increases in source strength noted above have made corresponding further improvements in signal/noise ratio particularly in the part of the spectrum between 40 and 200 c/s. Since the first presentation of this paper, line arrays of hydrophones have been used by several American groups to increase signal/noise ratios. In the continuation of the Woods Hole programme the line array enabled a group on the *Chain* to make good recordings at speeds up to 8 knots. At the time of writing 6 knots is the highest speed at which good recordings can be made in 3000 fathoms of water.

4. Results of Soundings

The system is useful not only for soundings on buried structures but also for oblique reflections and refraction studies. Because of the high energies required in the refraction method the relatively weak electrically powered sources do not compete with explosives except in shallow water for studies of shallow structures. But in oblique reflection studies, which are a powerful means of studying the sound velocity variations in sediments, the energy in the present 'sparker' and 'boomer' are adequate for many lines of investigation in deep water, including studies of multiple reflections over horizontal distances of many miles. Recently a radio-acoustic buoy, similar to that used by Hill and others[8] at Cambridge, has been built and used successfully with a 'boomer' towed by *Chain* to record oblique reflections up to ten to twelve miles away.

Figure 8 is a record of a single reflection from the bottom followed by a complex series of reflections from sub-bottom layers. This record was made while the ship, towing the sound source, approached the buoy at a distance of a few yards (see 'normal incidence sound source passing buoy' in the figure) and then proceeded away from the buoy on a straight course. Sounds were received at the buoy which were transmitted directly through the water and also by two or more reflections from the bottom. These are not shown on Fig. 8 but were recorded on magnetic tape for future analysis.

Up to the time of writing the principal use of this method of sounding has been to record what appears to be deposits of unconsolidated sediments. Since only a few sound velocity determinations have been made and even fewer cores are available for identifying reflections with geological structures, speculations must be based largely on the shape and attitude of the observed reflecting surfaces. Flat sea floors have been found to be underlain by several flat reflecting surfaces. These layered sequences are found to be interrupted at depth by a surface having an uneven profile. In some instances this surface can be identified with a high speed layer, 4 to 5·5 km/s,[9] from measurements by the refraction method; in other cases, mostly on the flanks of sea mounts, it has been traced to places where it crops out to become the bottom. In one instance, on the outer ridge north of the Puerto Rico Trench, echo sequences have been recorded which suggest flat layering below the uneven surface. All speculations about the nature of the material forming the deep, uneven surface are made difficult by the complicated train of side echoes which follow the beginning of this reflection. They tend to obscure evidences of structure which will be needed to determine whether the crustal layer (approx. 6·5 km/s) can be mapped by a reflection method.

5. Acknowledgments

The development and testing of the sound sources, 'sparker' and 'boomer', and the associated receiving and recording apparatus of the 'continuous seismic profiler' are the work of many individual scientists, engineers, and professional seamen on ships of the Woods Hole Oceanographic Institution. We gratefully acknowledge their help and their interest. This work was supported mainly under Contract Nonr-1367 with the Office of Naval Research. Early experiments with the 'continuous seismic profiler' were supported under contracts with the Bureau of Ships, the District of Public Works, 1st Naval District, and the U.S. Army Corps of Engineers. Administrators of all these federal agencies have been especially helpful in facilitating progress of this development and the accompanying scientific research. We gratefully acknowledge their assistance.

6. References

1. J. B. Hersey, "Continuous Seismic Profiling". In "The Sea: Ideas and Observations", M. N. Hill (Editor), (*Interscience*, New York). (To be published.)
2. R. Bowers, "A high-power, low-frequency sonar for sub-bottom profiling", *The Radio and Electronic Engineer*, **25**, No. 5, p. 457, May 1963.
3. S. T. Knott and J. B. Hersey, "High-resolution echo-sounding techniques and their use in bathymetry, marine geophysics, and biology", *Deep-Sea Research*, **4**, pp. 36–44, 1956.
4. D. D. Caulfield, "Predicting sonic pulse shapes of underwater spark discharges", *Deep-Sea Research*, **9**, pp. 339–48. 1962.
5. J. B. Hersey. H. E. Edgerton, S. O. Raymond and G. Hayward, "Pingers and thumpers advance deep-sea exploration", *J. Instrum. Soc. Amer.*, **8**, No. 1, pp. 72–7, 1961.
6. H. Hoskins and S. T. Knott, "Geophysical investigation of Cape Cod Bay, Mass., using the continuous seismic profiler", *J. Geol.*, **69**, No. 3, pp. 330–40, 1961.
7. J. B. Hersey, "Findings made during the June 1961 cruise of *Chain* to the Puerto Rico Trench and Caryn Sea Mount" *J. Geophys. Res.*, **67**, No. 3, pp. 1109–16, 1962.
8. M. N. Hill, "Seismic refraction shooting in an area of the Eastern Atlantic", *Roy. Soc. Phil. Trans. (G.B.)* Series A, **224**, No. 890, pp. 561–94, 17th June 1962.
9. J. Ewing and M. Ewing, "Reflection profiling in and around the Puerto Rico Trench", *J. Geophys. Res.*, **67**, No. 12, pp. 4729–39, 1962.

Manuscript first received by the Institution on 3rd May 1963, and in final form on 27th May 1963. (Paper No. 847/SS9).

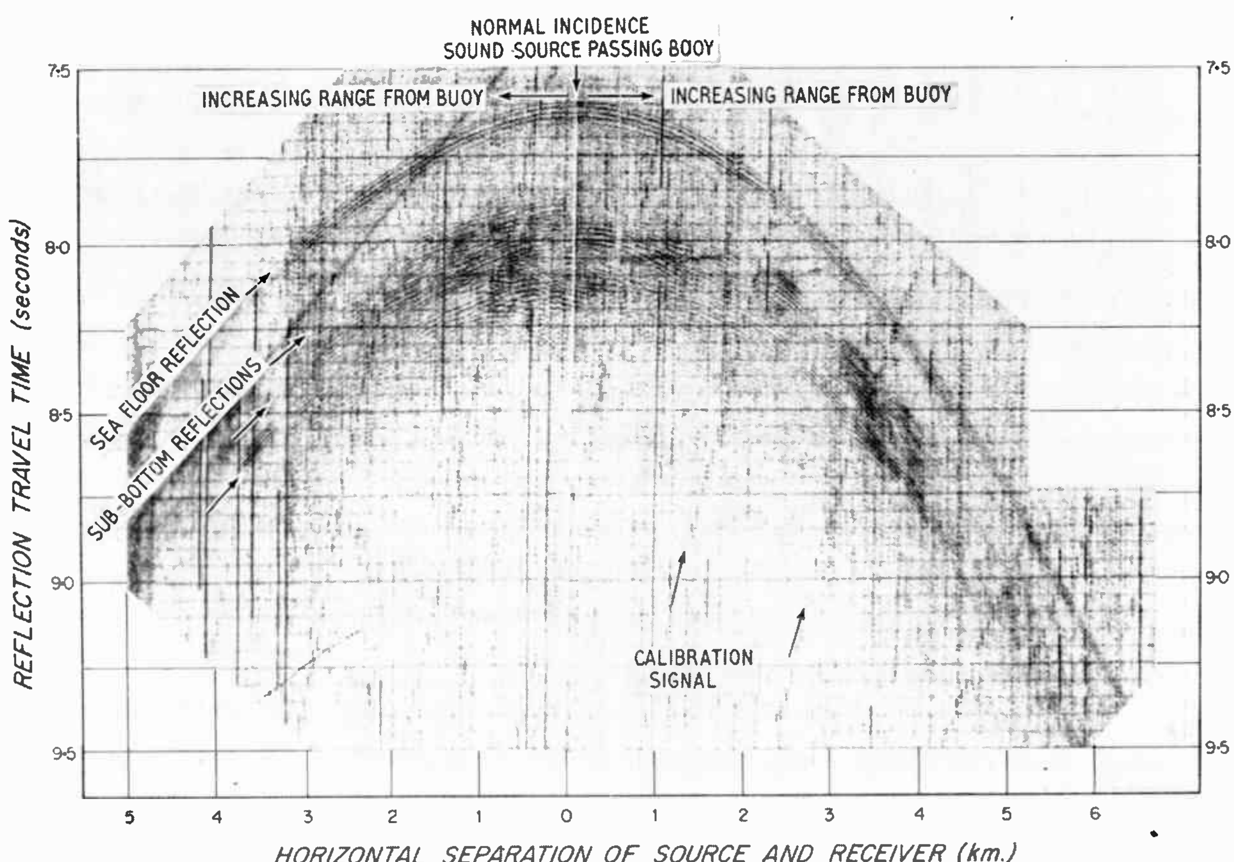


Fig. 8. A profile of sound reflections from the sea floor and underlying sediment and rock layers obtained while horizontal distance between sound source and receiver is varied.

POINTS FROM THE DISCUSSION

Mr. D. Davies†: Firstly I would like to ask the authors if the unusual pulse shape is explained by chemical recombination rather than by any bubble effect?

Secondly, to what extent have these instruments been used for refraction work?

Thirdly, have long monochromatic wave trains been observed in the records, and if so, is it possible to correlate them with observed horizons and sedimentary wave velocities?

The Authors (*in reply*): We would agree in part to Mr. Davies' first question although the shape of the pressure-time curve is not fully understood. A few experimental profiles have been detected in water a few hundred feet deep only. The wave trains you mention have been observed but up to now these records have not been fully analysed.

Dr. S. S. Srivastava‡: How far is this technique applicable in exploring the structure of the hard bottom granite surfaces of the ocean?

The Authors (*In reply*): It appears to be applicable to measuring the shape and depth of the top surface of hard rock overlain by sediment or water. There are some hints, as yet not substantiated, that reflections within or below masses of crystalline rock can be mapped by this technique.

† Department of Geodesy and Geophysics, University of Cambridge.

‡ Joint Services Staff College.

"SIGNAL PROCESSING IN RADAR AND SONAR DIRECTIONAL SYSTEMS"

(With special reference to systems common to Radar, Sonar, Radio Astronomy and Ultrasonics.)

A symposium on the above subject, sponsored jointly by the Institution and the Electrical Engineering Department of the University of Birmingham, will be held at the University of Birmingham from 6th to 9th July 1964§. Its purpose is to bring together those working in the fields of Radar, Sonar, Radio Astronomy and Ultrasonics to enable them to discuss common aspects of the signal processing techniques as applied in these different fields.

The development of signal processing techniques in these various applications has continued independently despite the common work involved, and it is clear that a useful purpose would be served in bringing such experts together to exchange new ideas and techniques. The scope of the symposium will include the following topics in so far as they are concerned with directional systems:

- Non-linear signal processing and correlation systems
- Directional pattern synthesis including aperture synthesis
- Electronic scanning systems
- 'Within-pulse' scanning
- Multiple beam-forming techniques
- Digital processing applied to directional systems
- Directional systems for automatic tracking
- Wide-bandwidth directional systems
- The effect of phase-coherence in directional systems
- Transient effects in directional systems
- Pulse compression

Contributions are invited on any of the above, or related topics, for consideration for inclusion in the programme of the symposium and offers of papers from overseas will be especially welcomed. Summaries should be submitted to the Institution as soon as possible. It is intended to preprint all papers; an announcement about subsequent publication will be made later. Details relating to registration and accommodation in a University Hall of Residence will be circulated shortly.

§ Amended date.

U.D.C. 539.16: 669.18

Activation Analysis Applied to Steel Production

By

G. I. CRAWFORD, Ph.D.,†

A. L. GRAY, B.Sc.,†

G. D. SMITH, C.G.I.A. (Member)†

Presented at the Convention on "Electronics and Productivity" in Southampton on 17th April 1963.

Summary: The techniques of activation analysis are discussed and equipment suitable for application to steel production described which uses a neutron generator based on the D–T reaction. These techniques offer the possibility of direct process control, since the analyses can be performed automatically and the results are presented in a form which could easily be applied to a controlling mechanism.

1. Introduction

Although the great interest at present shown in activation analysis is of comparatively recent growth, the technique has, by the standards of modern physics, an almost venerable history, stretching back as it does for 30 years. Activation analysis can be said to date from the discovery of induced radioactivity by Curie and Joliot in 1933. They found that an aluminium foil which had been exposed to alpha-particle bombardment continued to emit positrons for several minutes after the end of the irradiation. We now know that what they had observed was the reaction Al^{27} (α, n) P^{30}. The first use of the method for analysis purposes came in 1936 when von Hevesy and Levi determined the amount of dysprosium in impure yttrium oxide by thermal neutron irradiation. Some of the Dy^{164} (which amounts to 28% of natural dysprosium) in the sample was converted to active Dy^{165}, and the disintegration rate of this activity was then a measure of the amount of dysprosium in the sample.

Activation analysis consists, therefore, of three stages:

(1) A sample of the material to be analysed is exposed to a suitable source of radiation, which induces radioactivity in the constituent which is to be measured.

(2) The activity of the sample is measured.

(3) The percentage of constituent under study in the sample is deduced.

In principle gamma-rays, neutrons, protons, deuterons, alpha-particles or even heavier nuclei can be used as bombarding particles and, in fact, all of these have been used successfully. The use of charged particles is only practical when the analysis is to be confined to a very thin surface layer of the material, since their range in solid materials is only a fraction of a millimetre. The use of gamma-rays has few advantages over the use of neutrons and since it is more expensive to produce a given activity using gamma radiation than the same activity by neutron bombardment, most work in this field has been performed using neutrons. Various types of neutron source are discussed in more detail in Section 3.

The activity induced in the sample will nearly always be beta activity. It is possible, but unlikely, that alpha-activity will result, and the formation of isomeric states which de-excite only by gamma-emission is not very common. Beta activity can be measured by direct counting of the electrons or positrons which are emitted, or by detecting any subsequent gamma-emission or the gamma radiation emitted following annihilation of the positrons. In general (n, γ), (n, p) and (n, α) reactions produce electron-emitters while most positron-emitters are formed by (n, 2n) reactions.

Direct counting of electrons or positrons is not a practical technique, unless a very thin sample is of interest, because of the limited range of the particles. Because of the continuous nature of the energy spectrum of beta particles, it is impossible to use energy selection to distinguish between the particles omitted by different isotopes. Hence it is in general necessary to use radiochemical techniques to measure beta activities. Here conventional chemical methods are used to separate the active element which is to be measured from the other elements in the sample before the activity is determined. This technique, though powerful, is time-consuming and not suited to process control.

Measurement of the associated gamma-radiation has the great advantage that the gamma-rays have discrete energies and that a substantial proportion of those incident on a sodium iodide detector deposit all of their energy in the detector, thus giving a distinct peak (known as the 'photo-peak') in the detector

† Plessey Nucleonics Ltd., Northampton. (Dr. Crawford is now at the University of Glasgow.)

output spectrum. It is therefore possible, by counting only pulses in the photo-peak, to separate out to a considerable extent the contribution from different isotopes to a combined spectrum. It is possible therefore, in many cases, to measure the activity of a sample without recourse to radiochemical separation. This has the further advantage that the measurement is consequently non-destructive.

It will normally be necessary to place the detector remote from the neutron source to avoid activation of the detector.

2. Scope of Activation Analysis

It is possible, in principle, to perform activation analysis for almost every element, by using either fast or thermal neutrons. Even where activation analysis is impossible since all of the reactions which occur lead to stable end-products, it may be possible to perform an analysis by detection of the prompt radiations emitted in the reaction.

The number of counts recorded in a given analysis is given by

$$C = \frac{mN\sigma\phi(1-e^{-\lambda t_1})e^{-\lambda t_2}(1-e^{-\lambda t_3})E}{A} \quad \ldots\ldots(1)$$

where C = number of counts recorded in time t_3.

m = mass of isotope to be measured in the sample.

A = atomic weight of isotope.

N = Avogadro's number.

σ = cross-section for the reaction.

ϕ = neutron flux at the sample.

t_1 = irradiation time.

t_2 = delay between end of irradiation and start of counting.

λ = decay constant $\left(= \dfrac{0{\cdot}693}{\text{half-life}}\right)$ of activity.

E = detection efficiency.

t_3 = counting time.

Thus the sensitivity of the activation method depends upon the following factors:

(a) the neutron flux

(b) the cross-section for the chosen reaction

(c) the half-life of the activity produced

(d) the energy of the emitted radiation

(e) interference from other activities produced in the sample

(f) the general background level.

The influences of factors (a), (b) and (f) are obvious but some elaboration on the others is desirable.

If the half-life is short—less than a second, say—then a large proportion of the activity will decay in the time taken to transfer the sample from source to detector, and the sensitivity will be correspondingly reduced. If, on the other hand, the activity is longer than a few hours, only a small percentage of saturation activity will be reached in a reasonable irradiation period and also only a similar percentage of the total activity will be measured.

Factors (d) and (e) are best considered together. If little activity is induced in the other elements present compared to that in the element which is to be measured, then the energy of the emitted radiation is relatively unimportant. When, as is usually the case, the unwanted activities are stronger than the one to be measured, the measurement may only be possible if the wanted radiation is of higher energy than the interference.

Because of the interdependence of all these factors it is impossible to generalize on the possibility of performing an analysis for any particular element. Each individual problem must be considered on its merits. In typical applications, using a neutron generator, sensitivities of a few parts per million may be achieved.

This means, of course, that activation analysis with a neutron generator is considerably less sensitive than normal 'wet' chemical analysis. (It should be noted that when a reactor is used for activation analysis the technique is normally much more sensitive than chemical methods.) The main advantage of activation analysis lies in the speed of the operation; analysis times are typically reduced from 'hours' to 'minutes'.

In the particular case of steel production, the standard analytical procedures are already largely physical rather than chemical. Analysis for oxygen is most frequently carried out by the vacuum fusion process, where the steel sample is heated under vacuum in the presence of carbon and the oxygen content is deduced from the amount of carbon monoxide which is collected. The minimum analysis time is about 30 minutes, compared to the one and a half minutes taken by the equipment which is described in Section 7, and only a very small sample can be analysed. For other elements in steel, the analysis is usually by spark spectrograph which is faster than the corresponding activation measurement would be, but has the severe disadvantage of studying an even smaller effective sample than vacuum fusion does in the case of oxygen determination. This method is therefore open to errors caused by local inhomogeneity and segregation. In the case of elements which remain in solution during the solidification process these are not usually serious but this is not the case for oxygen in particular.

3. Neutron Sources

The sensitivity of useful reactions is usually such that neutron fluxes of at least 10^7 n/cm^2/s are required over the volume of the specimen. Even with small specimens total neutron outputs of upwards of 10^9 n/s from point sources are thus necessary. The most common laboratory neutron sources are of the type in which alpha particle or photon emission from a radioactive material is used to excite a neutron-producing reaction in a suitable substance surrounding or mixed with it. This technique is convenient for sources with outputs up to some 10^6 n/s, but such sources become very costly for higher outputs. They have the added disadvantages of emitting unwanted gamma-radiation and of requiring shielding even when the neutron output is not needed. Radioactive sources are in general of little value for activation analysis.

At the other end of the range of available sources lie the large heavy ion accelerators yielding neutron outputs of the order of 10^{12} n/s and small reactors which can produce neutron fluxes of 10^{12} n/cm^2/s in relatively large volumes of materials. A great deal of the pioneer work in activation analysis has been performed with such sources. Both types of installation are too costly to merit sole employment as analytical tools in any but a large research establishment and involve considerable commitments in staff and buildings.

While little cost reduction can be expected in small reactors, the use of the D–T reaction between deuterium and tritium ions with its release of high energy neutrons does mean that small heavy ion accelerator sources are possible; in the last few years very great strides have been made in their development. This reaction has a large peak in the yield versus particle energy curve at about 100 keV and, for an accelerator in which deuterium ions are projected at this energy against a tritium loaded target, neutron yields of the order of 10^{10} n/s per milliampere of beam current are obtainable. An additional advantage of this method of neutron production lies in the high energy, about 14 MeV, of the neutrons obtained. This enables a number of very useful activation reactions to be used which have high threshold energies. Beam currents of this order can be drawn from quite simple and compact ion sources; the necessary accelerating potentials can be obtained from many types of e.h.t. set such as those used for x-ray work.

Many neutron generators of this basic type have been developed and marketed in the last few years, although most of them have been essentially laboratory instruments not suitable for use by unskilled operators in process control work. In this application it is essential that a generator will run reliably with simple remote controls and will achieve a high degree of stability with a minimum amount of adjustment while running. The attainment of these ends depends mainly on good ion source design, good beam geometry and a high standard of vacuum practice. The performance of tritium-loaded targets is very dependent on this last requirement; the presence of the merest trace of organic vapour from even a well-baffled oil diffusion pump can adversely affect target performance. The most reliable operation is obtained in sealed tubes in which the target is not formed until after pumping is complete when no further contamination is possible. The life of such tubes is generally limited by that of the tritium loaded target, and neutron outputs of more than 10^9 n/s have not been easily obtainable.

Demountable tubes, which are continuously pumped and in which the target is easily replaceable, have recently been developed with reliability as high as sealed-off tubes by using an ion pump to maintain vacuum conditions better than those obtained in

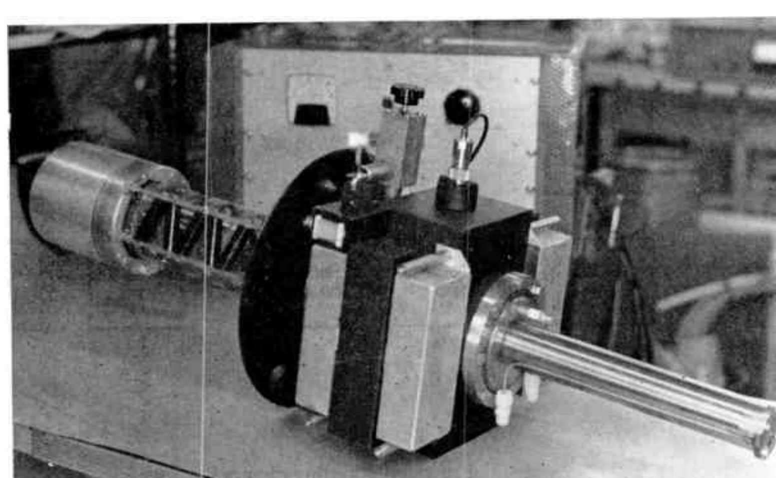


Fig. 1. Neutron generator tube.

most sealed tubes. Such generators are now readily available, with outputs up to more than 10^{11} n/s. At very high outputs target life, normally of the order of 10^{14} neutrons, becomes inconveniently short. In the most recent designs the target can be reloaded from an external source of tritium without opening the tube, so that such generators are virtually sealed ones; the only consumable items, deuterium and tritium, are fed in from external reservoirs. A generator of this type is shown in Fig. 1. This particular tube, shown with its electrical supply and control units, has an output of about 5×10^9 n/s.

The operation of such a generator as part of a process control analysis system is essentially very simple. The tube is supplied with an accelerator e.h.t. of 100 kV d.c. The ion source anode requires 8 kV d.c. and an adjustable low voltage a.c. supply heats the nickel leak admitting deuterium to the tube. The ion pump requires only a single direct supply of a few kilovolts. When the tube is not being used it is normal to leave the ion pump running, although this is not strictly necessary. The tube is brought to a standby condition by switching on the supply to the nickel leak which introduces gas to the tube. The gas pressure takes a few minutes to stabilize and fine adjustments can be made, if required, by varying the voltage supplied to the nickel leak. Neutrons are then obtained by switching on the accelerator and ion source e.h.t. supplies. The neutron output is adjusted by varying the accelerator e.h.t. and may be switched off by breaking either supply.

4. Shielding

Fast neutrons present a serious health hazard and if a neutron generator is to be employed safely then it is necessary to shield people working near it from the neutrons. Various materials are suitable. If a permanent structure is required, concrete is probably the best choice on the grounds of efficiency and cost, being the cheapest solution. A more flexible system, convenient in some cases, can be built up from steel tanks filled with water, since these are fairly easy to move when empty.

Further shielding of the gamma-ray detectors is normally necessary to maintain a sufficiently low background count-rate. Background counts can arise from activation of the detectors as a result of neutron interactions and also from gamma-rays emitted following activation of some of the constituents of the shield. For the first of these reasons it is desirable to site the gamma detectors as far from the generator as possible. (A compromise with the resulting transit time of the sample may be necessary for activities with short half-lives.) A few inches of lead around the detectors is normally sufficient to exclude gamma-rays originating in the shielding.

For outputs in the region of 10^{10} neutrons per second, approximately five feet of concrete or water are required to reduce the dose rate at the outside of the shield to below the normally accepted level of 0·75 mr/h. For the most powerful generators with outputs of 10^{11} per second an additional foot of shielding is necessary.

5. Neutron Output Monitoring

The generator output falls gradually during the lifetime of the target, as the tritium in it is consumed. Since the activity produced in a given sample is directly proportional to neutron flux, it is necessary to monitor the generator output during each irradiation. It is not normally necessary to know the absolute value of neutron output and hence the most convenient form of output monitor is a thermal neutron detector placed at a point in the shielding where the flux at maximum neutron output gives a counting rate from the detector of a few thousand pulses per second. This monitor can be calibrated in terms of absolute neutron output if desired. The schematic diagram of a complete neutron counting channel is shown in Fig. 2. This system, and the gamma

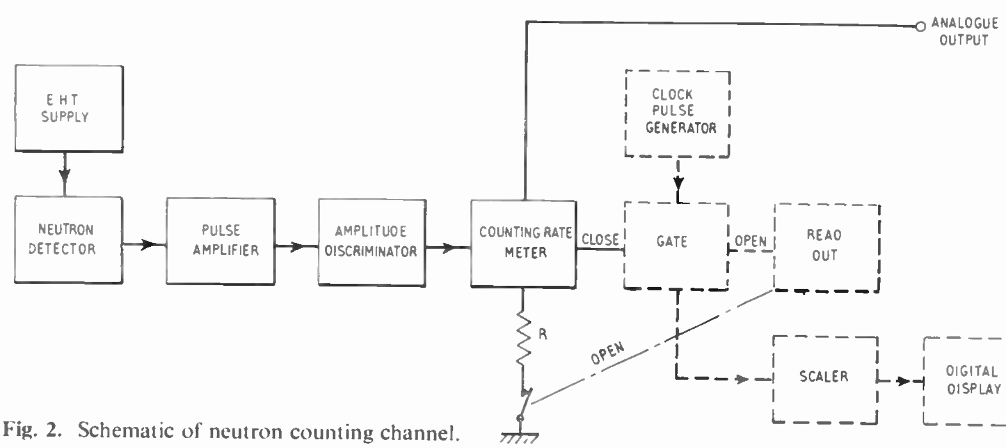


Fig. 2. Schematic of neutron counting channel.

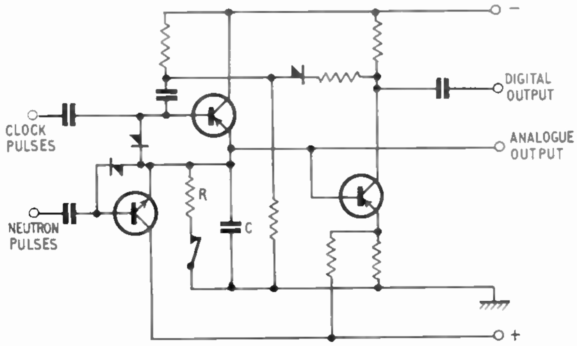


Fig. 3. Ratemeter and analogue/digital converter.

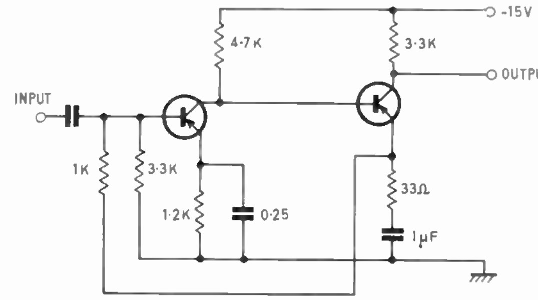


Fig. 4. Pulse amplifier.

channel shown in Fig. 8, have been developed using transistor electronic circuits throughout.

Either a fission counter or a boron trifluoride proportional counter can be used as the neutron detector; the BF_3 counter has the advantages of being a less expensive component and giving larger output pulses, but the fission counter requires a lower e.h.t. supply voltage and has a better count-rate/e.h.t. plateau, giving the prospect of better long term stability.

For a single element analysis, an analogue of the quantity $(1-e^{-\lambda t_1})e^{-\lambda t_2}$ (the activation constant) in equation (1), may be directly obtained by measuring the detector output with a counting rate circuit of the type shown in Fig. 3, whose C–R integrating time-constant is equal to the reciprocal of decay constant λ of the activity. At the end of the transfer time t_2, resistance R is disconnected and the activation constant remains stored in the capacitance.

Counter e.h.t. is derived from a push-pull transistor converter run from a stabilized low voltage d.c. line, and followed by a voltage multiplying rectifier. About 1000 volts is required for a fission counter (Plessey PNI 1049), or 2000 to 4000 volts for a BF_3 counter (actual voltage dependent on type used).

The pulse amplifier consists of three cascaded ring-of-two feedback circuits of the type illustrated in Fig. 4. A low input impedance matches the coaxial cable through which detector pulses are fed to a control console. The amplitude of the pulses developed across this impedance by the detector output current is approximately 200 μV from a fission counter, or 2 mV from a BF_3 counter. Voltage amplification factors are 60 dB and 80 dB for the two cases.

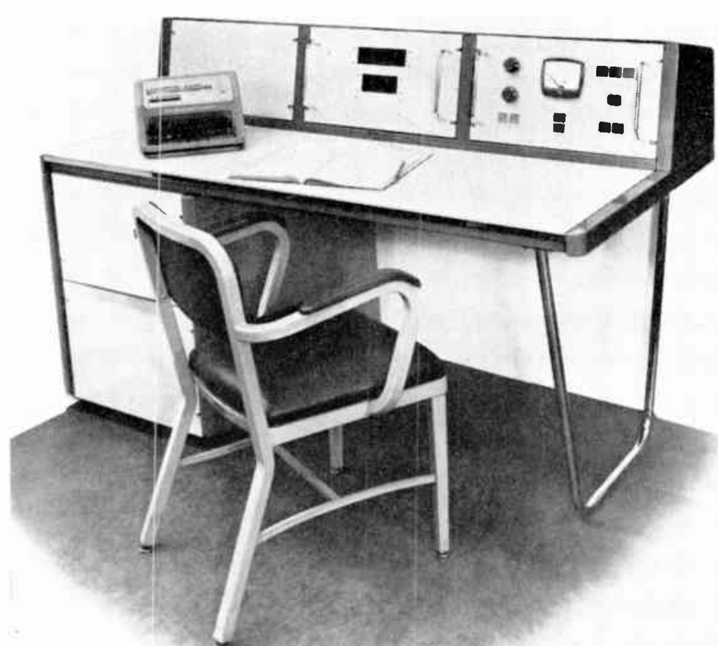


Fig. 5. Analox Mk II control desk.

The amplitude discriminator is a binary circuit, responding only to those pulses exceeding a predetermined amplitude. It also makes the duration of pulses fed to the ratemeter equal to the time interval between detector pulses; this normally allows adequate time for transferring charge to the ratemeter integrating capacitor C, whilst those random pulses which are spaced at intervals, which are short relative to the charge time, still contribute to the measurement and are not ignored as they would be if the driving circuit produced a constant duration output.

Output voltage could be displayed on a meter or fed to an analogue computer. In the equipment shown in Fig. 5, an analogue/digital conversion system provides a direct digital display. The integrating capacitor C in Fig. 3 is discharged by clock pulses of opposite polarity to the neutron pulses, and the clock pulses are counted by a three decade scaler. These pulses are fed to the scaler via a gating circuit which closes when the voltage in capacitor C falls to zero. Each decade of the scaler consists of four transistor binaries, with feedback to add in six pulses after the eighth input pulse. Each decade is connected to its appropriate digit in the display via a transistor gating system whose logic is shown in Fig. 6.

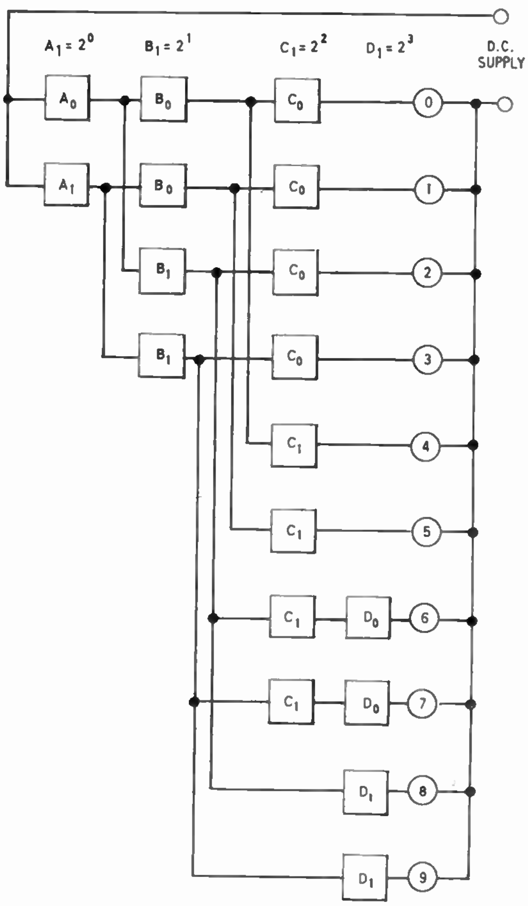


Fig. 6. Gating system for digital display

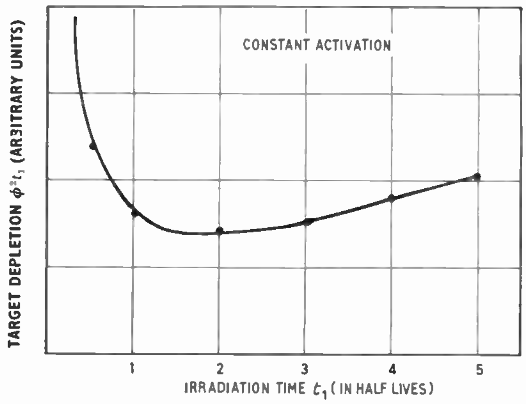


Fig. 7. Target depletion v. Irradiation time.

On the empirical observation that generator target life is inversely proportional to the square of beam current, the relationship between target depletion per irradiation ($\phi^2 t_1$) and irradiation time t_1 for a defined activation constant is as shown in Fig. 7, and is minimum when $\lambda t_1 = 1{\cdot}25$, i.e. $t_1 = 1{\cdot}8$ half-lives. Activation times between 1 and 4 half-lives are reasonably economic. When the half-life is very long however it may be necessary to use relatively short irradiation times.

If the irradiation period is terminated when a predetermined activation constant has been reached, and the beam current is initially adjusted to achieve this level in one half-life, then a 40% reduction in output flux is permissible before re-adjustment is necessary. Termination of the irradiate period is therefore initiated by an amplitude sensitive trigger circuit inspecting the voltage across the integrating capacitor C. A timer overrides this trigger should the desired activation constant not have been reached in four half-lives.

6. Gamma Activity Measurement

Using a scintillation counter with single-channel pulse amplitude analysis as depicted in Fig. 8, the gamma counting rate obtained will depend on sample activity, sample/detector geometry, detection efficiency, pulse amplitude distribution and channel width.

The energy resolution of a sodium iodide scintillator for a single gamma line is typically about 10% (except at low energies). The channel width will therefore be of the same order for many applications. Where the signal/background ratio is favourable, or where

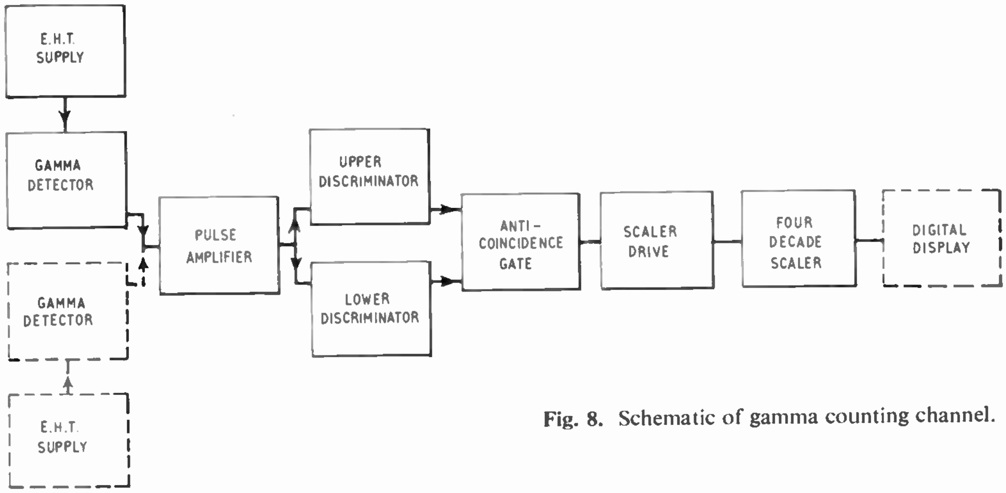


Fig. 8. Schematic of gamma counting channel.

there is more than one gamma line from a single isotope, count-rate may be increased by broadening the channel.

The sample will normally be brought into close proximity with the detector to secure the best practicable geometrical efficiency (approaching 50%). The absolute detection efficiency and the photo-fraction are both related to scintillator size, and at high gamma energies a scintillator large enough to make either approach unity is uneconomic.

For gamma energies of a few MeV, a scintillator 3 inches in diameter by 3 inches thick is a reasonable compromise; a duplicate detector assembly can be added on the opposite side of the sample if improvement in geometrical efficiency by a factor of two is essential. At 5 MeV this arrangement results in an overall counting efficiency of 20%. A rather higher figure might be obtained with a single 5 in × 5 in scintillator, particularly if of the re-entrant well type, but the additional cost would be several hundred pounds.

A count-rate measurement would be unsuitable where the half-life of the activity was short. The best statistical accuracy is obtained by counting detector output pulses on a scaler. For short lived substances, the count can conveniently extend for several half-lives.

The amplitude of photo-multiplier output pulses corresponding to a selected energy band can be of the order of a few hundred millivolts, and only a small amplifier gain is needed. The appropriate pulses are selected by a conventional analyser circuit comprising upper and lower level amplitude discriminators, followed by an anti-coincidence gate to pass only those pulses which exceed the lower level and do not exceed the upper level.

Output pulses are fed to a four-decade scaler which contains counting and display circuits similar to those described for the neutron channel.

Both long term and thermal stability of the pulse amplifier and analyser circuits are good, but photo-multiplier gain is very dependent on temperature and e.h.t. voltage. It is therefore desirable to provide a quick overall check on the gamma counting circuits. Many of the short half-life isotopes which are of interest in nuclear activation analysis emit high energy gamma rays which cannot be obtained from any long lived test sources. It is therefore convenient to switch to a higher amplifier gain for the test position, and a counting ratemeter is provided in the test facilities so that e.h.t. voltage may be adjusted if necessary to give a peak reading in the channel.

7. Oxygen Analysis Equipment as an Example

Although the equipment described in the preceding sections is of general application to activation analysis, its development was stimulated by the requirement for a rapid automatic measurement of the oxygen content of steel samples taken at various stages in the conversion process. Other methods of oxygen analysis in general use, although considered reasonably accurate, were slow and the use of activation analysis has been proposed by a number of workers. The n-p reaction of fast neutrons with oxygen 16 yields the isotope nitrogen 16 which has the very convenient half-life of 7·4 seconds decaying with the emission of gamma rays of 6·13 and 7·13 MeV.

An irradiation period of 10–20 seconds is thus convenient (see Fig. 7) and after irradiation a counting period of 30 seconds is sufficient to count most of the significant activity. Because of the short half-life rapid transit between source and counter is necessary and thus in principle the whole measurement can be completed within a minute.

Initial experiments were conducted with a laboratory type generator and after some initial difficulties, due equally to the oxygen content of the sample transport mechanisms and to the severe non-uniformity of the samples on which the check analyses had been made using small pieces, sufficiently encouraging results were obtained to design a prototype automatic equipment (subsequently named the 'Analox' †). This was built using a Kaman NT–60–9 neutron generator as shown in Fig. 1 and suitably modified conventional valve electronic circuits. The generator was mounted vertically in a central space inside the 4-ft thick octagonal shield wall of water enclosed in steel tanks, with its target about 12 inches off the ground. A simple mechanical transport mechanism working through a narrow slot in the shielding conveyed the sample from the target position to the gamma counter in about one second.

The operating cycle of the equipment is as follows. The normal stand-by condition of the equipment before an analysis maintains the neutron generator with the ion source working, but no accelerating potential applied. The operator loads the sample, weighing from 20–100 grams, into the carrier and presses the 'analyse' button. The carrier mechanism takes the sample through the shield and positions it below the target and at the same time the generator e.h.t. is switched on. Full output is obtained in less than 1 second and maintained for a period decided either by the attainment of a pre-set activation constant, or by the lapse of a pre-set time. At the end of this period the generator is switched off and the sample is moved rapidly back to the loading position. Here a latch releases the sample from the carrier and it drops down a chute into a space between the counters, two 3 in × 3 in sodium iodide crystals. Its arrival in this position starts the counting equipment, which operates for 30 seconds. Releasing the sample from the carrier avoids errors due to oxygen activity in the carrier itself. In the prototype equipment the oxygen count is displayed on a conventional dekatron scaler and the monitor output on a moving coil meter. This equipment has been operated for about a year and has successfully analysed many hundreds of samples. It has proved capable of consistent performance over long periods and even in prototype form has been simple to operate.

Once the prototype performance had been established the development of an engineered system (the Mk. II) was undertaken and the transistor circuits described in this paper for control and measurement functions developed. These have been built in printed card form and are all contained, together with setting-up equipment, in the desk shown in Fig. 5. The centre panel of this desk carries the display tubes. The right-hand panel carries the system controls and the left-hand panel, blank on this illustration, the neutron generator setting-up controls. Setting-up controls and a small ratemeter are contained in the drawer.

The generator, transport mechanism and counting systems of the Mk. II are essentially the same as in the Mk. I equipment but the generator is mounted horizontally to facilitate its removal for servicing. This entails the removal of the wooden shield blocks behind it, but avoids moving one water tank as was necessary in the Mk. I equipment.

8. Extension to Other Problems in Steel Making

Activation analysis is potentially of value in two main fields in steel making:

(a) in analysing the raw materials from which the steel is made;

(b) in analysing samples of the steel at any stage of the processing.

Two problems in raw material analysis have so far been considered. The first is the analysis of iron ore as it is delivered to the works, the second the analysis of the material which is fed to a sinter strand. These are obviously closely similar, and will be discussed here as a single problem.

The main constituents present in the raw material are iron, silicon, calcium and oxygen, and measurements of the first three of these may be required. The oxygen will inevitably be activated, as in the Analox determinations, but because its half-life is shorter than those of interest in this application, the N^{16} activity can be eliminated by increasing the delay time between activation and counting. The reactions which produce useful activities in this analysis are shown in Table 1.

Table 1

Element	Reaction	Half-life	Gamma Energy
Iron	$Fe^{56}(n, p)Mn^{56}$	2·56 h	0·84, 1·81, 2·13 MeV
Silicon	$Si^{28}(n, p)Al^{28}$	2·27 min	1·78 MeV
Calcium	$Ca^{44}(n, p)K^{44}$	22 min	1·16, 2·13, 2·7 MeV

In the case of very complex analyses, the use of a multi-channel analyser which effectively divides the pulse height spectrum from the gamma detector into a hundred or more channels and counts simultaneously in each, may well be justified. In such analyses a computer can be of great help in separating out the contributions from different isotopes.

† Registered Trade Mark.

For samples of finished steel, or for samples taken at different stages of the steel making process, analysis for a great many elements may be required. In almost all cases this is possible, at least in principle, by activation analysis. Table 2 shows many of the elements which are believed by the authors to be of interest divided into 3 broad groups labelled 'likely', 'difficult', 'unlikely'. This division is somewhat arbitrary and depends obviously on the lower limit of concentration which may be required and also on the amount of time and money which is available for the analysis.

Table 2

Likely	O, F, Fe
Difficult	N, B, Al, Si, P, S, Ca, Mn, Cu
Unlikely	H, C

9. Acknowledgments

The authors wish to thank the Directors of Plessey Nucleonics Ltd. for permission to publish this paper.

They would also like to record their appreciation of the assistance of the many members of the steel industry who have supplied samples and offered the benefit of their advice and experience on analytical problems in general which have materially assisted the development of the equipment described above.

10. References

1. R. F. Coleman, "The determination of oxygen in beryllium by activation analysis," "The Determination of Gases in Metals". A symposium report published by the Iron and Steel Institute 1960.
2. J. Harte, F. Bouten and F. Adams, "Minor-constituent analysis with neutron activation", *Nucleonics*, **19**, No. 3, p. 118, 1961.
3. E. L. Steele and W. Wayne Meinke, "Determination of oxygen by activation analysis with fast neutrons using a low-cost portable neutron generator", *Anal. Chem.*, **34**, No. 2, p. 185, 1962.
4. W. Wayne Meinke and R. W. Shideler, "Activation analysis: new generators and techniques make it routine", *Nucleonics*, **20**, No. 3, p. 60, 1962.
5. D. F. C. Morris, "Neutron activation analysis", *Metallurgical Reviews*, 7, No. 26, 1962.

Manuscript received by the Institution on 13th March 1963. (Paper No. 848).

Professor S. K. Mitra, F.R.S.

The Council of the Institution has learned with deep regret of the death of Professor Sisir Kumar Mitra, D.Sc., F.N.I., F.R.S. (Member) on 13th August 1963 at the age of 73 years. He was Professor Emeritus of the University of Calcutta and National Research Professor in Physics of the Government of India.

Professor Mitra was born in Calcutta and educated at the T.N.J. College, Bhagalpur, and at Presidency College, Calcutta. He received the D.Sc. degree from Calcutta University in 1919 and then went to Paris where he worked at the Sorbonne with Professor Ch. Fabry. As a result of research on the determination of spectroscopic standards in the near ultra-violet region (2100 to 2370 Å), he was awarded a Doctorate in 1922.

After leaving the University of Paris, Professor Mitra spent some time at the Institute of Physics, University of Nancy, with Professor Gutton where work on the three-electrode valve was being undertaken. On his return to India in 1923 he was appointed Khaira Professor of Physics in the University of Calcutta and organized teaching of "Wireless" at post-graduate level and also started research on wireless and allied subjects in his laboratory. A Post-Graduate Department of Radio Physics and Electronics was formed under his leadership at the University of Calcutta in 1949—now known as the Institute of Radio Physics and Electronics.

One of Professor Mitra's greatest contributions to Indian science was the part he played in establishing the Radio Research Committee of the Council of Scientific and Industrial Research, Government of India. He was the first Chairman of the Committee for five years from 1943 to 1948. Professor Mitra may thus be regarded a pioneer in the field of radio research in India. He built up an active school of research on ionospheric investigation at Calcutta and an Ionosphere Field Station, the first one of its kind in India, was established through his efforts at Haringhata, near Calcutta.

In 1947 Professor Mitra published his well-known treatise on "The Upper Atmosphere". The book, the first to be devoted solely to this subject, contains much original work and it received immediate world-wide recognition including translation of a subsequent edition into Russian.

In 1958 Professor Mitra was elected to Fellowship of the Royal Society of London for his contributions to the studies of upper atmospheric phenomena.

A man of wide interests, Professor Mitra was President of the Asiatic Society during the years 1951–2. He was General President of the Indian Science Congress in 1955, and the President of the National Institute of Sciences of India during 1959–60. He was also President of the Indian Science News Association (publishers of the journal "Science and Culture") from 1956–8. Professor Mitra was a member of the Indian National Committee for the International Geophysical Year and was on the Editorial Board of a number of Indian and foreign scientific journals.

Recognition of his scientific and other contributions was widespread and included the King George V Silver Jubilee Medal (1935), Joy Kissen Mookerjee Gold Medal of the Indian Association for the Cultivation of Science (1943), Science Congress (Calcutta) Medal of the Asiatic Society (1956) and the Sir Devaprasad Sarvadhikary Gold Medal of Calcutta University (1961). He received the Presidential Award *Padmabhushan* in 1962.

Professor Mitra held the Khaira Professorship of Physics in the Calcutta University up to the year 1935, when he was appointed Sir Rashbehary Ghose Professor of Physics. This post he held till his retirement in 1955. In September 1956, he took up the Administratorship of the Board of Secondary Education, West Bengal, relinquishing this post in April 1962, when he was appointed National Research Professor of Physics by the Government of India.

Elected a Full Member of the Institution in 1952, Professor Mitra was closely concerned with the setting up of the original Indian Advisory Committee (forerunner of the Indian Divisional Council) and he was chairman of the Calcutta Section for the first three years of its existence. His only personal contribution to the *Journal* was an appreciation of the career of his distinguished predecessor at Calcutta University, Sir Jagadish Chandra Bose, F.R.S., on the occasion of the centenary of the latter's birth in 1958.† He took a very active interest, however, in the Institution's scientific work through his encouragement of the submission of papers to the *Journal* by members of the Institute of Radio Physics and Electronics.

By his death, Indian science has lost a radio physicist of world-wide reputation.

† *J. Brit.I.R.E.*, **18**, page 661, November 1958.

U.D.C. 621.395.616

The Condenser Microphone and Some of its Uses in Laboratory Investigations

By

P. V. BRÜEL, D.Sc.†

AND

W. J. PARKER, B.A.†

Presented at a meeting of the Electro-Acoustics Group in London on 16th January 1963.

Summary: The development of techniques for measuring sound pressures are discussed, with special reference to the accuracy of various methods. After comparing the advantages and disadvantages of different kinds of laboratory microphone in use today, the characteristics of the condenser microphone are discussed more fully. The theory brings to light the factors which govern the frequency response, sensitivity, acoustic impedance, and other important qualities. Then, following some comments on the principles of designing miniature microphones, methods of calibration are discussed.

Taking the loudspeaker as an example of electro-acoustic equipment which is to be investigated, most of the possible acoustic test arrangements are considered, partly with a view to deciding which, if any, give the best objective criterion of fidelity. Other uses of the condenser microphone are also discussed briefly.

1. Introduction

Acoustic measurements before 1800 were surprisingly sparse and fundamental—cannons and stop-watches were typical methods of instrumentation—but during the nineteenth century enormous advances were made by a very few physicists, such as Lord Rayleigh and Helmholtz. These men showed a remarkable insight into the basic acoustic problems, but their understanding, which in many aspects can hardly be bettered today, was to languish untapped for several more decades. In recent years acoustics has been able to advance tremendously due to twentieth-century developments in electrical circuits and radio techniques. The microphone, for instance, was developed first as a communications transducer, but it has gradually evolved into a precision device capable of application to a great many measurement problems: sound surveys, architectural acoustics, noise control, gas dynamics, and of course electro-acoustics. Electro-acoustics covers many important aspects of radio engineering so it is worthwhile to discuss these laboratory microphones.

It was the Bell telephone receiver which was first used as a successful measuring microphone. No carbon microphone has ever been sufficiently consistent. Subsequently (as late as 1913), it was, however, reported[1] that the Rayleigh disc was found to be still the most reliable and sensitive sound measuring instrument. A real break-through did not come until 1917 when, at the Western Electric laboratories, the first condenser microphone was developed, together with its associated electronic circuitry.

Sound level meters, and then audio-frequency analysers, were evolved alongside developments in room-acoustics measurements. Techniques were further improved by wartime studies on sound direction finding and submarine activities, and more recently considerable interest has been shown in industrial and environmental noise problems. There has been a progressive improvement in accuracy (Table 1) and in the scope of measurements, many of which can now be made automatically.

Table 1

Overall Accuracy in the Measurement of Sound Pressure Level

Date	Accuracy	Comments about the sound level meter	Calibration reference
1930	± 15 dB	Barkhausen	Rayleigh disc—free
1935	± 15 dB	Objective s.l. meter	,, ,, ,,
1940	± 15 dB	Better version	,, ,, in tube
1950	± 3 dB		,, ,, ,,
1955	± 1 dB	Precision s.l. meter	Reciprocity
1960	± 0·5 dB	I.E.C. Precision	Pistonphone

From an analysis of the trend of these figures, the overall accuracy in 1965 may be expected to be ± 0·2 dB.

† A/S Brüel & Kjær, Copenhagen.

2. Laboratory Microphones

The increase in accuracy shown in Table 1 has been brought about solely by design improvements, and no new basic principles are involved. The choice is still between dynamic, piezo-electric or condenser microphones, although in the second category there is scope for finding materials with better properties. Lead zirconium titanate appears to be the most promising of these substances, but in the present state of the art, the condenser microphone is preferred for its excellent all-round performance.

Table 2

Laboratory Microphones Compared

(1 = very favourable, 2 = satisfactory, 3 = not favourable)

Factors to be considered	Dynamic micro-phones	Piezo-electric microphones		Con-denser micro-phone
		Barium titanate	Lead zirconium titanate	
Dynamic range	2	1	1	1
Sensitivity	1	3	3	2
Low frequency response	3	1	1	1
Frequency linearity	3	3	2	1
High frequency response	3	3	2	1
Dimensions	2	1	1	1
Need for associated supplies	1	1	1	3
Long-term stability	2	2	2	1
Range of working temperatures	3	3	2	1
Influence of temperature	2	3	2	1
,, ,, vibration	2	2	2	1
,, ,, moisture	1	2	2	3
,, ,, magnetic field	3	1	1	1
Fragility	2	1	1	3
Price	2	1	1	3

Microphones respond to the difference in pressure between the front and back of their diaphragm. This difference is obtained either by a difference in phase between the sound pressure at the two sides, as in pressure gradient microphones, or by keeping the pressure on one side of the diaphragm constant by mounting it as one of the walls of a closed cavity, as in pressure sensitive microphones. For measurement purposes the latter type is usual and much of this paper will therefore be confined to a discussion of the pressure sensitive condenser microphone.

3. Condenser Microphones

3.1. *Principle of Operation*

The condenser microphone cartridge itself (Fig. 1) is a high impedance device and so, to avoid cable loading problems, it is usually necessary to connect it directly to a cathode follower, the two parts forming one cylindrical body (Fig. 2).

Fig. 1. A condenser microphone cartridge (*left*) with its protection grid removed (*right*).

Figure 3 shows a microphone cartridge which consists essentially of a thin metallic diaphragm in close proximity to a rigid back plate. These two elements are electrically insulated from each other and constitute the electrodes of a capacitor. The space behind the diaphragm is only in communication with the outside via a hole having a pneumatic impedance which is very large at audio frequencies. Variations in pressure due to sound waves will therefore move the diaphragm, i.e. alter the capacitance between the two plates at a frequency equal to that of the original sound waves. A so-called polarization voltage, applied across the capacitor induces a charge which remains constant, since the charging time-constant is

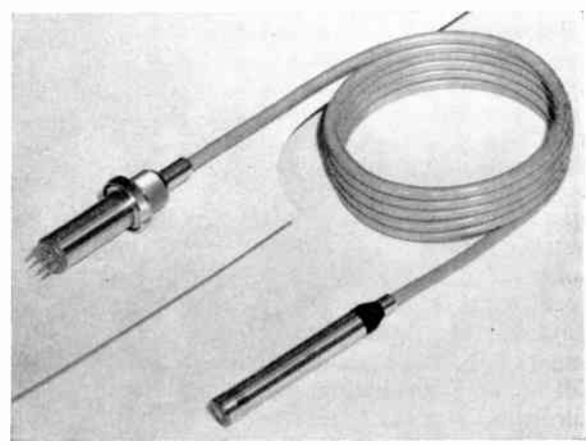


Fig. 2. Typical condenser microphone. This example has an outside diameter of $\frac{1}{2}$ in.

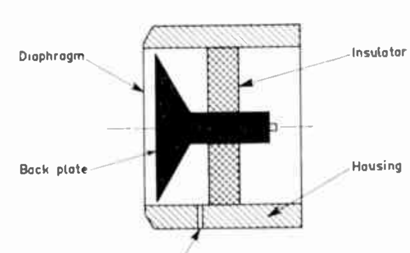


Fig. 3. Basic construction of a condenser microphone cartridge.

long compared with the period of sound pressure variation and consequently an alternating voltage component is generated.

By careful design it is possible to maintain proportionality between a.c. output voltages and sound pressures over a wide frequency range and a large dynamic range.

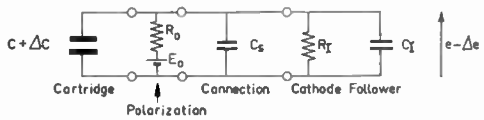


Fig. 4. Simplified diagram of a condenser microphone cartridge connected to a cathode follower and polarization voltage source.

C = capacitance of microphone cartridge.
C_s = stray capacitance of connection to cathode follower.
R_0 = series resistance in polarization circuit.
E_0 = polarization voltage.
R_I, C_I = cathode follower input resistance and capacitance.
e = output voltage.

The condenser microphone circuit diagram is shown in Fig. 4. Reducing this to the equivalent circuit of Fig. 5, the output voltage fluctuations Δe can be related to ΔC, which is proportional to the magnitude of the pressure fluctuations

$$\Delta e = \frac{\Delta C}{C_T} E_0 \frac{j\omega RC}{1+j\omega RC_T} \qquad \ldots\ldots(1)$$

It is seen from eqn. (1) that sensitivity is inversely proportional to total capacitance, i.e. C_T should not exceed C more than is absolutely necessary. (Hence the need for integral construction of the cartridge and the cathode follower.)

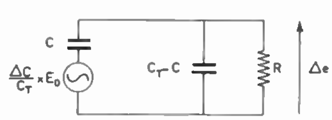


Fig. 5. Equivalent circuit of complete condenser microphone.

C = capacitance of microphone cartridge.
C_T = total capacitance in circuit.
R = shunt combination of R_0 and R_i.
E_0 = polarization voltage.

The output appears to be proportional to polarization voltage. This is very nearly so, but the critical gap behind the diaphragm is slightly dependent on electrostatic attraction forces, thus making the variation with voltage somewhat greater. In practice a particular stabilized voltage (say 200 V) is used for polarization so these variations are of no significant importance.

3.2. *Low Frequency Response*

Two factors influence the low frequency response:

(a) Electrical loading when the condenser impedance becomes comparable with the cathode follower's input impedance, and

(b) Acoustic cut-off where pneumatic leakage prevents full pressure variation over the diaphragm.

The electrical effect (a) is inferred by the last term of eqn. (1), the cut-off frequency being equal to

$$\frac{1}{2\pi RC_T} \qquad \ldots\ldots(2)$$

R is made as large as possible; the cathode follower input impedance is often 700 MΩ or more. The cut-off frequency (−3 dB point) is usually at about 10 c/s. This governs the microphone's low frequency response since under normal circumstances the pneumatic cut-off ((b) above) is at a lower frequency than this.

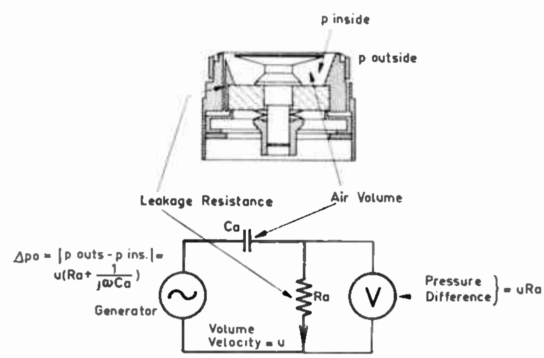


Fig. 6. Cross-section of a condenser microphone cartridge and equivalent circuit of the static pressure equalization system.

An equivalent circuit can be drawn up to show how the static pressure equalization system causes l.f. cut-off (Fig. 6). Any volume, such as that enclosed by the diaphragm, can be said to have capacitance given by the change in volume which results in a unit change of pressure, a close analogy with the electrical case. Furthermore, the pneumatic resistance presented by the leak can be portrayed by an electrical resistance R_a. Acoustic pressure is analogous to voltage, since the generator signal represents sound pressure on the microphone diaphragm, and current is analogous to volume velocity u. The transfer function is seen to be

$$\frac{\text{effective pressure difference}}{\text{outside pressure on diaphragm}} = \frac{j\omega R_a C_a}{1+j\omega R_a C_a} \quad \ldots(3)$$

As expected, the pressure difference, which is effective in operating the microphone, is the same as the outside sound pressure at higher frequencies, but there is no microphone output at zero frequency.

This is necessary if sensitivity is not to be severely affected by changes in ambient pressure. C_a, or the volume behind the diaphragm is strictly limited by other design considerations but R_a can be adjusted, usually by partially filling a capillary tube with fine wires, so that the cut-off frequency is optimized, usually at about 1 c/s.

It should be noted that the leak resistance thus arrived at is adequate to allow pressure equalization during rapid climb by flight vehicles carrying microphones.[2] This factor is also dependent on microphone sensitivity but as a guide, a low frequency cut-off below 1 c/s infers satisfactory operation for rates of climb up to about 100 000 ft/min.

Another very important reason for providing a path from the inside of a microphone to the outside is that condensed water vapour, which can cause serious noise problems, must be allowed to escape.

3.3. *Acoustic Impedance*

If the working air-volume outside the microphone diaphragm is small, then the change in this volume due to diaphragm indentation will be significant. The microphone, which works like an acoustic capacitance—a certain pressure change resulting in a certain volume change—is loading the source capacitance when it is of comparable magnitude. Acoustic measurements on earphones are often made in enclosed couplers where this condition applies so it is convenient to express the microphone's acoustic input impedance[3] in terms of the acoustic impedance of an equivalent volume V_e. V_e is then the volume by which the coupler volume is effectively increased and its acoustic impedance Z is derived from

$$Z = \frac{1}{j\omega \dfrac{V_e}{\gamma P_a}} \qquad \ldots\ldots(4)$$

where γ = ratio of specific heats, P_a = ambient pressure, and ω = angular frequency.

An interesting method, first reported by Embleton and Dagg[12] of the National Research Council of Canada, is used to determine the equivalent volume. The microphone diaphragm forms part of a hard-walled Helmholtz resonator which is excited by a frequency adjusted to give exact resonance. The microphone is then replaced by an externally identical solid 'stopper' and the shift in resonant frequency is noted. The small difference in volume of the resonant system is easily calculated. Measuring the acoustic impedance directly is more complicated but yields the real and imaginary parts of the equivalent volume (Fig. 7).

3.4. *Influence of Reflections*

At higher frequencies, where the wavelength shortens and becomes comparable with the dimensions of the microphone, reflection and diffraction occur near the diaphragm. Consequently the pressure P_1 in this region (see Fig. 8) is not necessarily the same as the undisturbed sound pressure P_0. This raises the distinction between free field response—indicating the sound pressure with no microphone present—and pressure response following whatever value P_1 happens to have. At low frequencies there are no differences between the two responses since P_1 is the same as P_0, but at higher frequencies the two responses differ, depending on the angle of incidence (defined in Fig. 9).

The pressure increase due to reflection is plotted for a simple case in Fig. 10, D/λ being taken

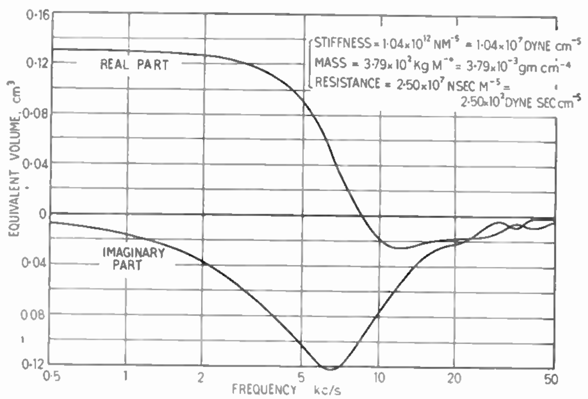


Fig. 7. Real and imaginary parts of the equivalent volume of a typical 1 in microphone often used in closed coupler work.

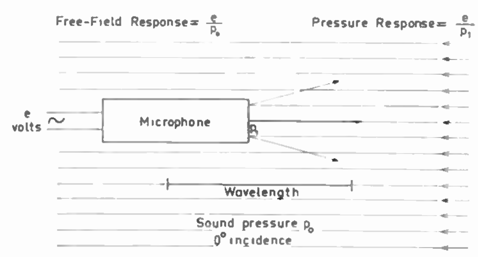


Fig. 8. Sketch showing the reflections from a cylindrically shaped microphone in a free sound field. The definitions of pressure and free field response are also given.

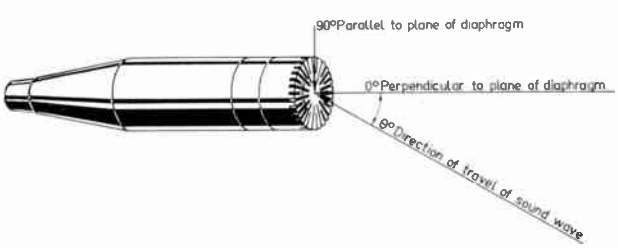


Fig. 9. Definition of angle of incidence used throughout this paper.

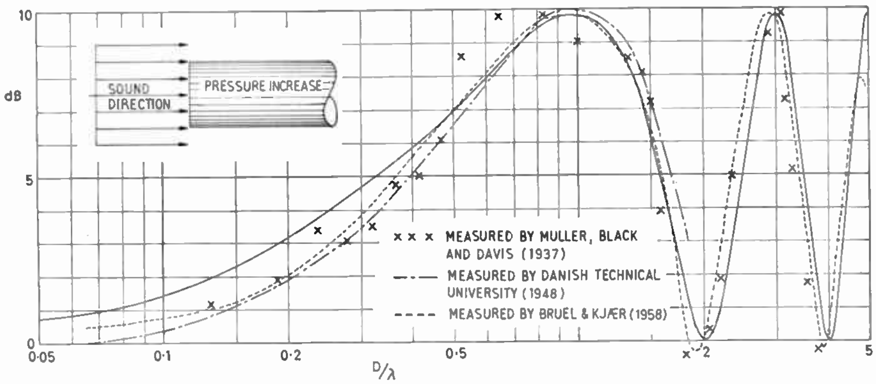


Fig. 10. Pressure increase on the axis of a cylinder placed in a free sound field.

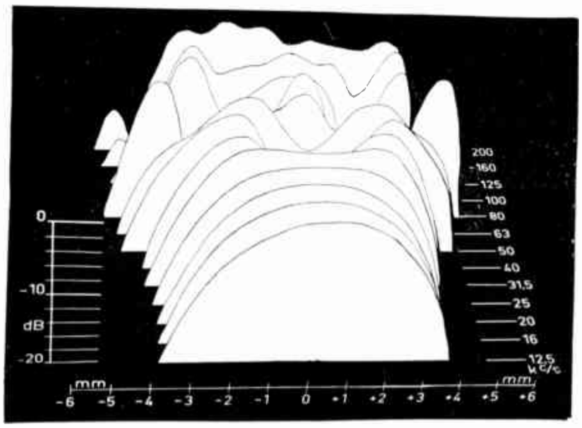


Fig. 11. Overall view of the sensitivity distribution of a ½ in o.d. condenser microphone.

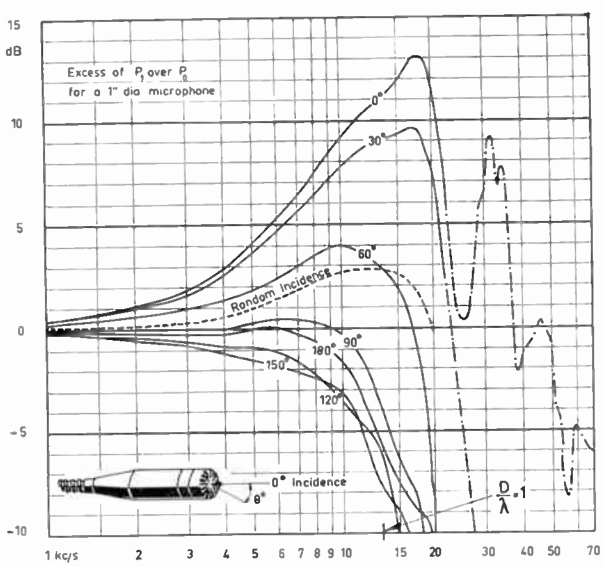


Fig. 12. Pressure increase due to reflections, shown as a function of frequency. Microphone diameter: nominal 1 in, 0·936 in exactly.

as abscissa where λ = wavelength and D = effective diameter of cylinder.

It is not, however, sufficient to know the pressure increase at the centre of the cylinder because it is the average pressure across the diaphragm which is of interest here. It is a complete study[4] in itself to find this, but briefly, the different pressure distribution curves have been determined experimentally for several frequencies (Fig. 11), and then at each frequency the pressure distribution has been simulated by an electrostatic 'pressure' field of the same shape. Curves such as those in Fig. 12 could be drawn up when the average intensity of the electrostatic field is known.

The frequency at which the effective diameter D is equal to the wavelength λ is marked on Fig. 12. D which from the reflection point of view should obviously be as small as possible, is only equal to the physical diameter if the top of the microphone is planar (Fig. 13(*a*)). A cavity such as is shown in Fig. 13(*b*) can make the effective diameter 50% greater than the physical diameter, thus shifting the onset of disturbances downwards in frequency as compared with a plane-ended microphone of the same dimensions.

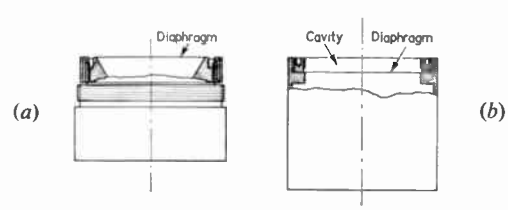


Fig. 13. Section through two microphone cartridges (*left*) with plane end, and (*right*) with a cavity above the diaphragm.

3.5. *Sensitivity and High Frequency Response*

The mechanical elements which determine the high frequency characteristics, and also the sensitivity of the microphone will now be considered.[2] Figure 14

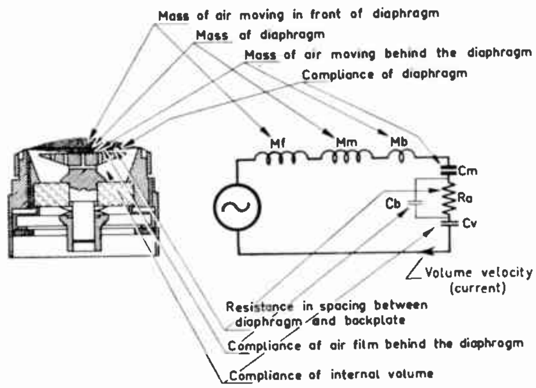


Fig. 14. Equivalent circuit of a condenser microphone (pressure type) around and below the first resonance.

shows a simplified mechanical model of a plane-ended microphone and also an electrical analogue valid below and around the principal resonance. Inductance represents mass and capacitance represents compliance. Compliance is reciprocal stiffness, i.e. it gives the volume change resulting from a certain pressure change.

The sound pressure (generator) has to move the mass of the diaphragm itself and also the air 'cushions' on both sides of the diaphragm. The compliance allowing movement of the diaphragm is made up from the compliance of the diaphragm itself, plus that of the air film between the diaphragm and the back electrode, plus that of the main air volume inside the cartridge. The motion of the diaphragm is only transmitted to the main volume via some holes in the back electrode which offer resistance R_a to the passage of air particles. The film is so thin (0·001 in) that its capacitance C_b is small enough to be neglected.

Inserting a proportionality constant K to relate pressure to diaphragm velocity $\dot{x}$, the following equation is obtained:

$$\frac{\dot{x}}{P_1} = \frac{K}{j\omega(M_f + M_m + M_b) + \dfrac{1}{j\omega}\left(\dfrac{1}{C_m} + \dfrac{1}{C_v}\right) + R_a} \qquad \ldots(5)$$

where ω = angular frequency.

Output voltage V is dependent on diaphragm deflection x, so in terms of another constant K_1

$$\frac{V}{P_1} = \frac{K_1}{j\omega\left[j\omega(M_f + M_m + M_b) + \dfrac{1}{j\omega}\left(\dfrac{1}{C_m} + \dfrac{1}{C_v}\right) + R_a\right]} \qquad \ldots\ldots(6)$$

This denominator reduces to a second-order factor:

$$\frac{V}{P_1} = \frac{K_1}{\left(\dfrac{1}{C_m} + \dfrac{1}{C_v}\right) + j\omega R_a - (M_f + M_m + M_b)\omega^2} \qquad \ldots(7)$$

At frequencies well below resonance (ω comparatively low) the sensitivity is

$$\frac{K_1}{\left(\dfrac{1}{C_m} + \dfrac{1}{C_v}\right)} \qquad \ldots\ldots(8)$$

providing the resistance and the masses are designed to be numerically much smaller than the stiffness. Now $1/C_v$ is an air stiffness and this will decrease with falling pressure, i.e. the capacitance C_v in the electrical analogy becomes larger. In a vacuum the sensitivity will be determined by the stiffness of the diaphragm alone, so to avoid too much sensitivity variation with pressure, the diaphragm stiffness is usually made six to ten times as great as this air stiffness.

At high frequencies, where the whole of equation (7) must be considered, the behaviour of R_a, M_f, and M_b with changing environment has to be taken into account. The resistance R_a does not depend on pressure until the mean free path of the air molecules is comparable with the spacing between diaphragm and back plate. Its variation with temperature is small enough to be neglected. M_f and M_b, however, become markedly smaller with decreasing air density and increasing temperature.

The denominator of equation (7) can be compared with the standard second-order factor

$$p^2 + 2\mu\omega_n p + \omega_n^2 \qquad \ldots\ldots(9)$$

where ω_n = resonant frequency and μ = relative damping ratio (near resonance $Q = 1/2\mu$).

It is seen that

$$\omega_n^2 = \frac{\dfrac{1}{C_m} + \dfrac{1}{C_v}}{(M_f + M_m + M_b)} \qquad \ldots\ldots(10)$$

and

$$\mu = \frac{R_a}{2\omega_n(M_f + M_m + M_b)} \qquad \ldots\ldots(11)$$

With medium pressure decreases, ω_n becomes larger, i.e. the resonance shifts to higher frequencies, on account of diminished air mass. However, when R_a starts to grow less, the high frequency value of C_v increases because the divided air masses within the microphone behave as one. As a result, ω_n starts to decrease.

It is strikingly shown in Fig. 15 that the damping diminishes rapidly as soon as the mean-free-path effect starts to influence R_a, whereas there is hardly any change in damping as the pressure falls from atmospheric to about one-third of atmospheric.

This analysis is based on only a simple model, but one extremely important factor is made clear. This is that the value of R_a is of great importance in determining damping, in other words the shape of the

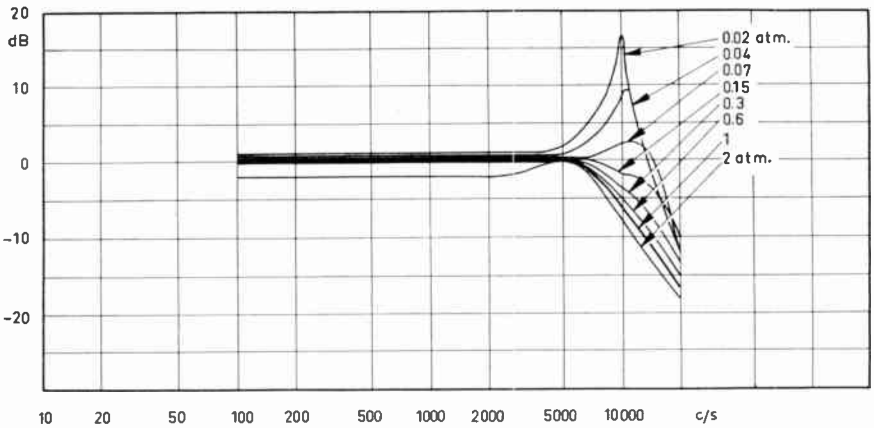


Fig. 15. Typical frequency characteristics for a 1 in diameter microphone at various ambient pressures.

high frequency characteristic can be governed by adjusting R_a, or by adjusting the diaphragm mass, which equation (11) shows to give the same effect.

Returning to the definitions of free-field and pressure response (Fig. 8), microphones are needed which have an optimum flat response for both these cases. For the best high frequency results the damping at standard pressure can be individually adjusted during production to give correct pressure response (Fig. 15), or exactly the right amount of overdamping can be introduced so that one or more of the P_1–P_0 pressure-excess curves (Fig. 12) can be compensated for. This principle is explained in Fig. 16. Normally the response is optimized for 0 deg incidence, the curves for other angles of incidence can be made nearly the same as this by screwing a specially-shaped grid on to the end of the microphone.

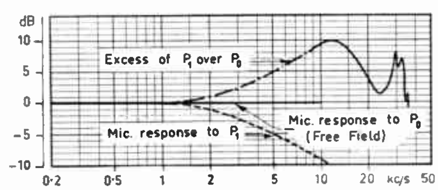


Fig. 16. The pressure measured at the microphone diaphragm is P_1, so by suitably damping the characteristic, a flat response to the free sound pressure P_0 can be obtained.

3.6. *Transient Response*

It can be inferred from the extensive frequency ranges shown in Fig. 17 that the transient response of the smaller microphones must be excellent. Figure 18 shows some typical step responses.

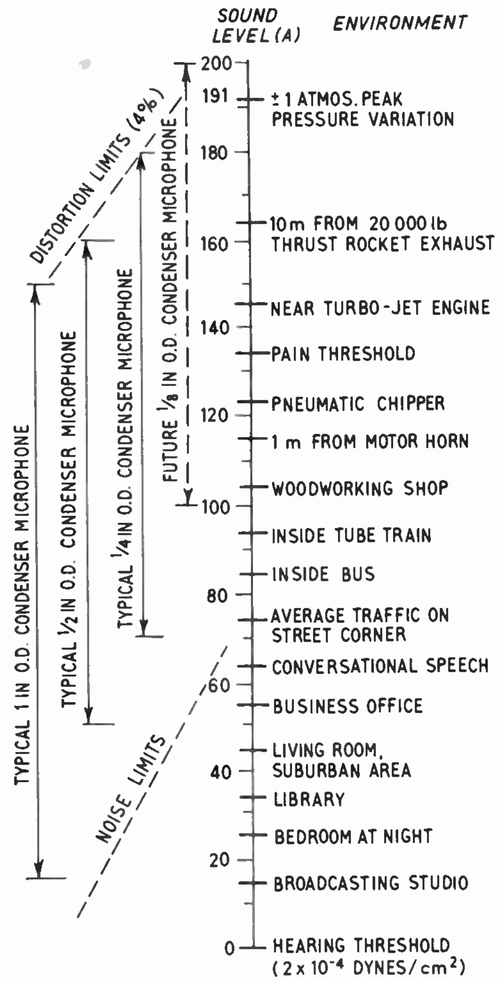
3.7. *Construction*

In addition to all the points considered so far, the designer must make the microphone as rugged as possible: for instance it must not be greatly influenced

Table 3

Useful Dynamic Ranges for Various Microphones

(The remarks to the right give some impression of what the various levels sound like)



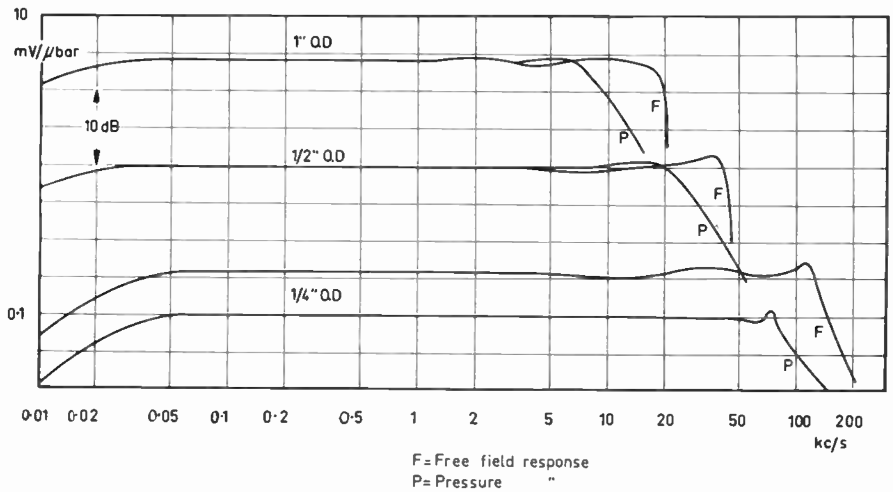


Fig. 17. Frequency characteristics of some typical condenser microphones. The useful dynamic ranges are shown in Table 3.

by vibration and temperature. The materials used therefore should have similar coefficients of thermal expansion and all parts must be mounted in a special way.

The diaphragm has long been a problem in the latter respect: differential expansion and vibration have allowed it to slip slightly from under its clamping ring. One solution to this is to 'electro-deposit' the diaphragm to the supporting ring so that the crystal structure is continuous between the two. Electro-deposition makes it possible to construct microphones of small diameter, yet maintaining a large percentage of the diaphragm area as effective. The advantages in high-frequency measurement offered by reducing the size are obvious; in the space of not many years the diameter has been reduced from a few inches down to a quarter of an inch and still further reductions are possible.[5]

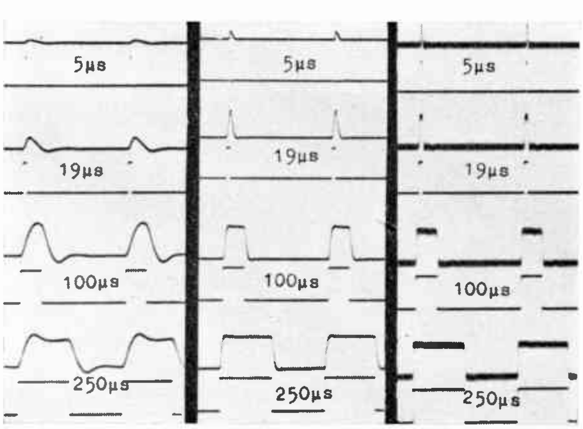


Fig. 18. Transient responses of typical microphones, 1 in o.d. (*left*), ½ in o.d. (*centre*), and ¼ in o.d. (*right*).

3.8. *Calibration*

3.8.1. Reciprocity method

The reciprocity method is an absolute calibration method in which two microphones are coupled together acoustically. One acts as a transmitter and the other as a receiver. Treating the condenser microphone as a linear, reversible transducer the product of the sensitivities of the two microphones are found. Then by using a third microphone as transmitter and the two microphones in turn as receivers, the ratio of the two sensitivities can be calculated.

Brief Theory. Consider Fig. 19 where two microphones are coupled by a close rigid cavity of known physical dimensions. The cavity is filled with a gas having a known ratio of specific heats.

The following symbols are used in the description:

P_a = ambient air pressure.

γ = ratio of specific heats of the gas in the cavity, (1·402 for dry air).

V = volume of cavity + equivalent volume of both microphone cartridges.

p = r.m.s. sound pressure.

u = r.m.s. volume velocity.

Z_a = acoustic impedance of the cavity $= \dfrac{\gamma P_a}{j\omega V}$

ω = angular frequency $= 2\pi f$.

e_c = r.m.s. voltage across capacitor C.

E_1 = r.m.s. value of the voltage of frequency f applied to microphone 1 used as a transmitter ($E_1 \gg e_c$).

e_{10}, e_{20} = r.m.s. voltage of the cartridge when loaded with a cathode follower.

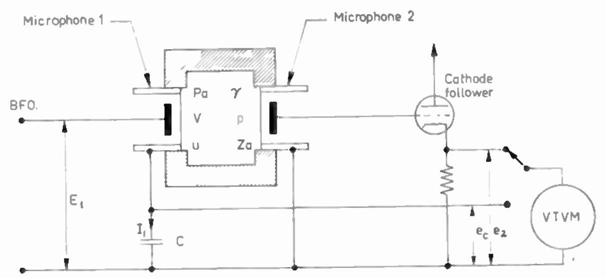


Fig. 19. Simplified diagram of reciprocity calibration method.

e_1, e_2 = r.m.s. voltage of the cathode follower output when looking into an infinite impedance.

I_1 = current through C, representing the current through microphone 1.

$\beta = e_{20}/e_2 = e_{10}/e_1$ = attenuation of the cathode follower.

The sound pressure produced in the cavity is

$$p = uZ_a = \frac{\gamma P_a u}{j\omega V} \qquad \text{......(12)}$$

The pressure sensitivity of microphone 2 is $T_2 = e_2/p$ so that

$$e_2 = T_2 \frac{\gamma P_a u}{j\omega V} \qquad \text{......(13)}$$

The reciprocity theorem can be expressed as follows:

In a passive, linear, reversible microphone the ratio of input current I_1 to the open circuit volume velocity u of the diaphragm of the microphone used as a transmitter, is equal to the ratio of the open-circuit pressure p to the open circuit voltage e_{10} developed across the electrical terminals of the microphone when used as a receiver.

This means that

$$\frac{I_1}{u} = \frac{p}{e_{10}} = \frac{p}{\beta e_1} = \frac{1}{\beta T_1} \qquad \text{......(14)}$$

where e_1 is the output voltage from the cathode follower. The factor β is due to the attenuation of the cathode follower used in conjunction with the microphone. T_1 is the sensitivity of the microphone. (Units depend upon the system adopted, e.g. volts/μbar.)

$$T_1 = \frac{e_1}{p}$$

From eqns. (13) and (14):

$$e_2 = T_1 T_2 \frac{\beta \gamma P_a I_1}{j\omega V} \qquad \text{......(15)}$$

The difficulty here is the precise measurement of I_1.

By connecting the transmitter microphone as shown in Fig. 19 in series with a known capacitance C which is large compared with the capacitance of the condenser microphone, and measuring the voltage drop e_c across C with the same input voltage E_1 as used for the reciprocity measurements, I_1 can be calculated from: $I_1 = j\omega C e_c$

When this is substituted in eqn. (15) we obtain

$$\frac{e_2}{e_c} = T_1 T_2 \frac{\beta \gamma P_a C}{V} \qquad \text{......(16)}$$

so that the product of the sensitivities is then found simply by using a third microphone as a transmitter and measuring the output from the two microphones in turn for the same input voltage. If we measure e_3 volts from microphone 1, and e_4 volts from microphone 2, we have

$$\frac{T_1}{T_2} = \frac{e_3}{e_4} \qquad \text{......(17)}$$

and from eqns. (16) and (17) the two sensitivities can easily be calculated.

It should be noted that this method of calibration gives the absolute sensitivity of the microphone + cathode follower, so that if a cathode follower having a different attenuation is used, this must be taken into account when the new combination is employed for absolute measurements.

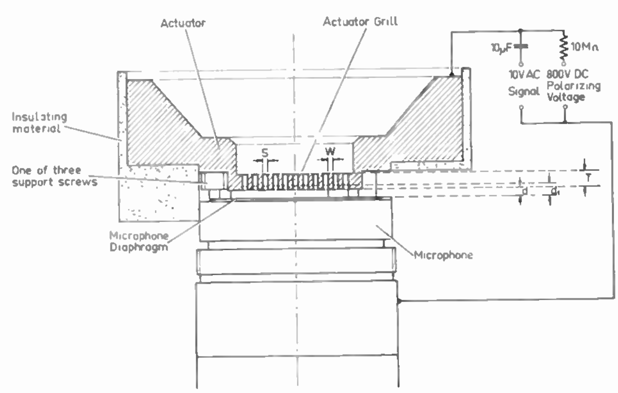


Fig. 20. An electrostatic actuator simulating sound pressures on a microphone diaphragm.

3.8.2. Electrostatic actuators

The sound pressure on a microphone diaphragm can be simulated by a known electrostatic attraction between the diaphragm and a special grill placed above it (Fig. 20). This method does not suffer from the same high frequency limitation as the reciprocity method, where wave motions can be set up in the coupler, although at very low frequencies the analogue is not valid because the electrostatic and acoustic forces are not equivalent. However, throughout the useful frequency range of a microphone the actuator is excellent for determining the shape of the characteristic. Its absolute accuracy is not so high as that obtainable by reciprocity calibration or by using a pistonphone.

3.8.3. The pistonphone

A very convenient device for rapid calibration of the condenser microphone is the pistonphone. A sectional drawing of the pistonphone is given in Fig. 21, and the principle of operation is shown in Fig. 22.

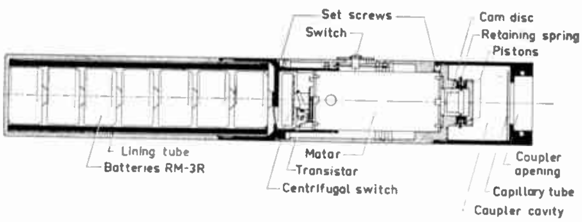


Fig. 21. Assembly drawing of the pistonphone.

The two pistons are driven symmetrically by means of a cam disc mounted on the shaft of a miniature electric motor. When rotating the cam will give the pistons a sinusoidal movement at a frequency four times the speed of rotation. Consequently the cavity volume is varied sinusoidally and the r.m.s. sound pressure will be:

$$p = \gamma P_0 \frac{2A_p S}{V\sqrt{2}} \qquad \ldots\ldots(18)$$

where γ = ratio of specific heats for the gas in the cavity.

P_0 = ambient pressure.

A_p = area of one piston.

S = peak amplitude of motion of the pistons from the mean position.

V = volume of the cavity with pistons in the mean position + equivalent volume of the microphone.

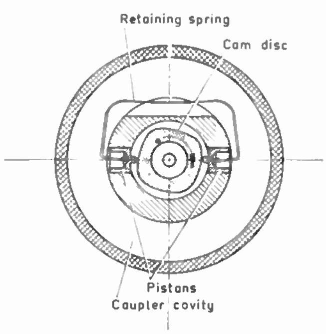


Fig. 22. Cross-section showing the principle of operation of the pistonphone.

The sound pressure level in dB is then s.p.l. = 20 log p/p_T where $p_T = 2\times 10^{-4}$ dynes/cm² which is the reference pressure usually adopted for acoustical measurements in air. It is worth noting that the speed of rotation does not enter into the equation so that the speed of the motor is immaterial as long as the sound pressure frequency is higher than about 5 c/s, below which the compressions and expansions are not adiabatic and eqn. (18) is not valid.

4. Applications of the Condenser Microphone in Investigating Electro-Accoustic Equipment

The foregoing discussion has shown that a properly chosen condenser microphone is a nearly perfect acoustic transducer. That is, its electrical signals can normally be taken for granted, without having to account for its characteristics during investigations. This reduces the electro-acoustic measurement problem to that of deciding on the electrical measurements necessary for determining the performance of some piece of equipment. The loudspeaker[6] will be taken as an example, so that not only the basic characteristics of interest to the systems engineer but also the more subtle and intangible 'fidelity' factors enter the picture. In the sections which follow, the more important performance criteria will be considered in turn.

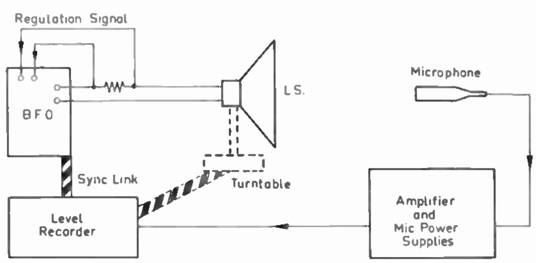


Fig. 23. Arrangement for recording the frequency response of a loudspeaker in an anechoic chamber or in the open air.

4.1. *Frequency Response*

To avoid disturbing reflections, acoustic tests on loudspeakers should be carried out either in an anechoic chamber or in the open air.

It is usually logical to drive the loudspeaker with constant current throughout the frequency range and so an oscillator with automatic output control should be used if an automatic frequency sweep is required (Fig. 23). For convenience the loudspeaker characteristic can be directly recorded on frequency calibrated chart paper. This runs through a logarithmic level recorder which at the same time drives the frequency scan on the beat frequency oscillator. Note that a microphone with optimum free-field response is required here, whereas, if measurements were being made in closed couplers on telephone receivers etc., an optimum pressure response microphone would be needed.

4.2. *Directional Characteristics*

Concentrating on one frequency at a time, the directional properties of the loudspeaker should be found. This is easily done when the level recorder used can take polar paper and can also control the rotation of a turntable which is substantial enough to carry any loudspeaker. Such a turntable, and the link which synchronizes its orientation with the angles marked on the polar paper, are shown by broken lines on Fig. 21.

4.3. *Acoustic Output Level*

Here it is necessary to make an absolute measurement of sound pressure level in some specified acoustic environment around the loudspeaker. The instrumentation following the microphone need not be so complicated as in Fig. 23—a sound level meter is adequate. Usually this will be calibrated but if necessary a pistonphone can be used to find the sensitivity. The measurements of sound pressure which result are usually[6] normalized by some dimensional and electrical characteristics of the system so that a 'characteristic sensitivity' is obtained. This figure, which gives the sound pressure at a specified position when one electrical watt is dissipated in the speech coil, is much more practical than a measurement of total radiated sound power.

4.4. *Nonlinear Distortion*

Nonlinear distortion, which is due to amplitude non-linearities in the loudspeaker, is usually treated separately from frequency distortion (due to a poor frequency response characteristic), and from phase distortion (caused by a non-uniform phase characteristic). If a pure sine wave is applied to a non-linear system, harmonics will be produced, and if two or more sine waves are applied simultaneously, intermodulation results. A measure of nonlinear distortion can therefore be obtained by investigating either harmonics or intermodulation.

Harmonics are easier to measure than intermodulation products, but tests on the latter are probably the better gauge of quality because harmonics, which are frequently in the original sound, are not always judged to be unpleasant, whereas intermodulation products are not heard naturally. Even the intermodulation test relies on only two pure tones for excitation and this may not be very realistic. It has therefore been proposed that white noise be used as an input signal, and a method of discriminating the resulting distortion is mentioned below under section 4.7.

4.5. *Harmonic Distortion*

The harmonic distortion in the output from a good beat frequency oscillator is so small compared with that expected in a loudspeaker that the driving voltage can be taken as being purely sinusoidal. Any sounds detected at frequencies which are multiples of that generated by the oscillator are therefore harmonics generated by the loudspeaker. Instead of using a linear microphone amplifier as in Fig. 23, a selective amplifier system must follow the microphone. Broadly speaking, there are two possible arrangements:

(a) For a single-reading measurement which does not show the importance of any particular harmonic, a rejection filter is set to reject the fundamental frequency and to give equal weight to all the significant harmonics. The ratio

$$\frac{\text{r.m.s. value of total harmonics}}{\text{r.m.s. value of complete signal}}$$

gives the distortion factor.

(b) For individual measurement of each harmonic filters having narrow pass bands coinciding with the various harmonic frequencies of interest are required. The sets of $\frac{1}{3}$ octave filters used for sound analyses are selective enough up to the fifth or sixth harmonics. These have the advantage that they can be automatically switched successively into circuit to follow the sweep of the b.f.o. An off-set can be introduced between the b.f.o. frequency (which is the same as the chart paper frequency graduation) and the filter centre frequencies. In this way automatic plots of any low harmonic can be made for all frequencies considered. The discrimination of the filters is limited, so if harmonics at widely different levels are to be recorded, the fundamental or powerful harmonics may have to be rejected by a special filter.

4.6. *Intermodulation Distortion*

Basically, an input I_{in} which is the sum of two different sinusoidal signals is applied to a nonlinear device. A Maclaurin expansion gives the output I_{out} from this device in the form

$$I_{out} = A_0 + A_1 I_{in} + A_2 I_{in}^2 + A_3 I_{in}^3 \ldots A_n I_{in}^n \ldots \quad \ldots(19)$$

where A_n is the coefficient dependent on the shape of the characteristic.

Putting $I_{in} = P \sin pt + Q \sin qt$, the output can be shown to contain frequencies which are:

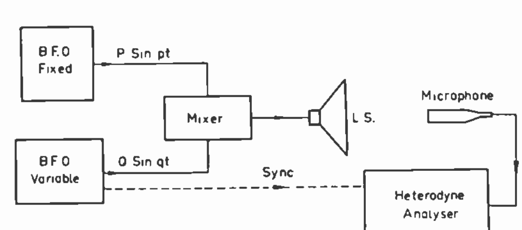


Fig. 24. Arrangement for measuring intermodulation distortion.

(a) multiples of p and of q,

(b) sums and differences of multiples of p and of q, (e.g. $p+q$, $2p+q$, $p+2q$, etc. $p-q$, $2p-q$, $p-2q$, etc.).

The lower orders of the intermodulation product can be selected by a heterodyne analyser while the input frequencies, p and q, are varied relative to one another (Fig. 24). Various schemes and more details about these rather complicated measurements are given in references 7 and 8.

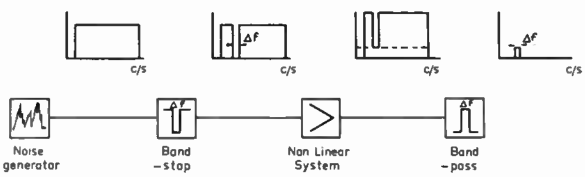


Fig. 25. Principle of measuring nonlinear distortion by wide-band noise.

4.7. *Noise Test*

White noise can be looked upon as a signal containing an infinite number of sine waves of all frequencies within its band. If a narrow band of frequencies is removed before applying the noise to a perfectly linear device, no frequencies within the blocked band would be detected at the output of the device. However, if nonlinearities are present, harmonics and intermodulation products from the other frequencies will appear within the critical bandwidth. The noise level in this band is therefore some criterion of distortion (Fig. 25). This method, which is discussed more fully in ref. 9 is not fully developed since, to the authors' knowledge, no equipment offering ideal characteristics is available commercially.

4.8. *Transient Response*

A loudspeaker having an excellent frequency characteristic may lack fidelity because certain

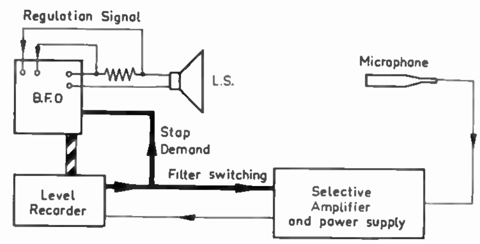


Fig. 26. Arrangement for automatically recording several decay curves on one chart to see how the transient response of the loudspeaker varies with frequency.

mechanical resonances in the speech coil or the diaphragm are not adequately damped and oscillations persist after the electrical signal has ceased. Figure 26 shows a measuring arrangement which is an adaptation of the B.B.C. 'impulse glide method'.[10, 11] An endless belt of recording paper runs through the high speed level recorder and each cycle of the belt takes slightly less time than the cycle of associated events so that a series of spaced out curves appear, as in Fig. 27. The associated events, which are all controlled by the level recorder, are as follows:

The oscillator signal is cut dead so that the acoustic decay can be plotted.

The b.f.o. moves to a new frequency and its signal level is restored before another curve is recorded. The selective amplifier follows the b.f.o. frequency so that a good signal/noise ratio is obtainable.

Subjective impressions of the quality of the loudspeaker dealt with in Fig. 27 agree very well with the impulse glide analysis, namely that the treble response is very clean whereas the bass response is 'woolly'. Another useful feature of such a family of curves is that the top envelope is actually the frequency response characteristic.

4.9. *Production Tests*

The test systems which have been mentioned can yield a great deal of information but do not give a

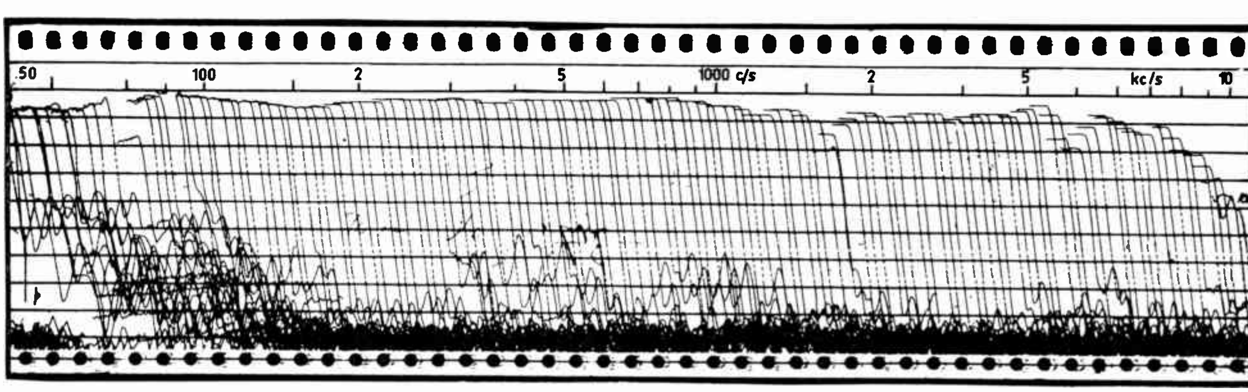


Fig. 27. Family of transient response curves, showing the decay of sound after input signals of several different frequencies have been muted. Note the poor damping at low frequencies. (Paper speed 100 mm/s, writing speed 1000 dB/s.)

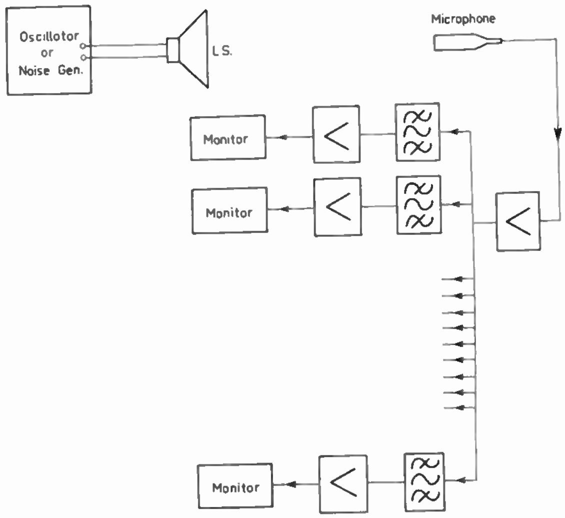


Fig. 28. Principle of rapid test system monitoring in several frequency bands simultaneously.

sufficiently rapid and concise check for production inspection of loudspeakers. A frequency response tracer is a useful instrument for this purpose. It is essentially a cathode-ray oscilloscope in which the X-scan is dependent on the frequency of an associated b.f.o. feeding the loudspeaker. The loudspeaker output, picked up by the microphone, gives rise to a Y-deflection so that the complete frequency response curve appears on a long-persistence screen. It is then a simple matter to draw in tolerance limits on a graticule covering the screen.

An alternative check is to use several selective amplifiers which are all fed from the same microphone but tuned to different frequencies (Fig. 28). The outputs from all the amplifiers can be monitored simultaneously to give the frequency response of a loudspeaker fed with white noise, (i.e. the speaker is driven at all relevant frequencies concurrently). If the loudspeaker is excited by discrete sine-wave frequencies, instead of by white noise, distortion can be assessed immediately from the output levels in channels which are tuned to the harmonic or intermodulation frequencies.

5. Application of the Condenser Microphone to Probe Techniques

When it is necessary to investigate sound pressure levels in places of small dimensions, the condenser microphone may be large enough to disturb the sound field considerably and thus invalidate any measurements carried out, or the variations in the sound field may take place over distances comparable to or smaller than the microphone diameter and so would not show up in the results. In such cases the probe technique is of great value. An illustration of the condenser microphone with probe is given in Fig. 29.

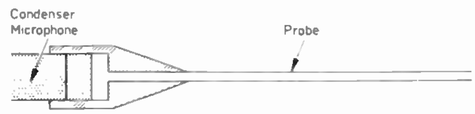


Fig. 29. Condenser microphone with probe.

The frequency response of such a probe will vary with length and diameter of the probe tube, and to obtain the best possible frequency characteristic it is advisable to use a tube as short and wide as possible. For low frequencies the response characteristic will be flat, (see Fig. 30) but at higher frequencies resonances in the tube will cause large variations in sensitivity. It is however possible to damp out these resonances by inserting a tuft of wire wool inside the probe tube. The flat part of the characteristic is thereby extended considerably (Fig. 31). An example of the application of probe microphones has already been given in Fig. 11, where the pressure distribution in front of a cylinder placed in a sound field was investigated.

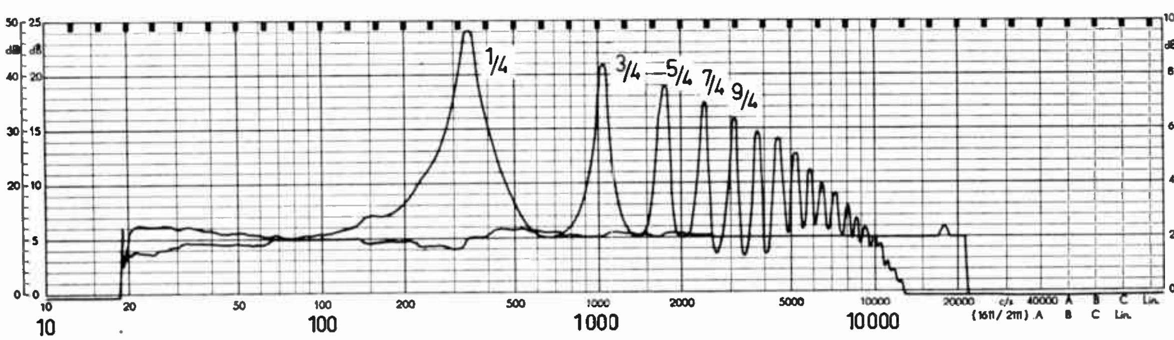


Fig. 30. Frequency response of probe microphone undamped.
The peaks are marked $\frac{1}{4}$, $\frac{3}{4}$ etc. indicating wavelength resonances in the probe tube.

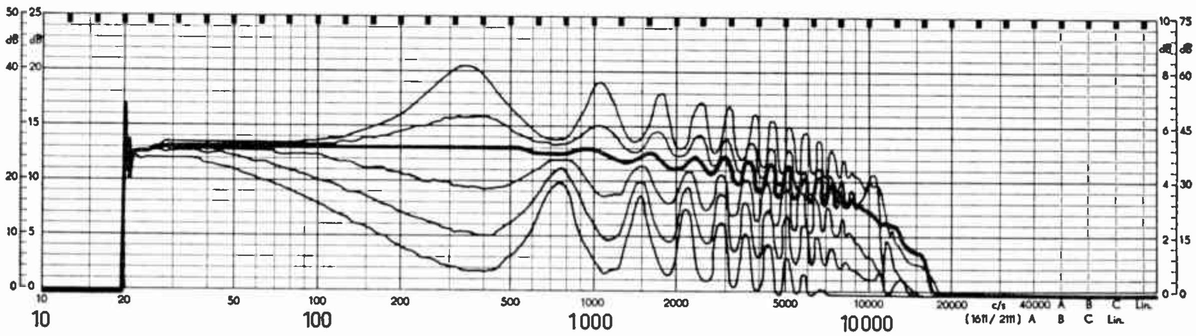


Fig. 31. Probe tube with various degrees of damping.

6. Application of the Condenser Microphone to Sound Level Measurements

A very important application of the condenser microphone is in the absolute measurement of sound pressure levels. Used in conjunction with properly designed amplifiers and indicating devices, the condenser microphone can be calibrated to give exact readings of sound pressure levels in the audible frequency range. This is of particular value in noise control, building acoustics, etc., but also in laboratory measurements where absolute values of sound pressure levels are required, e.g. in psycho-acoustical investigations.

7. Conclusions

For laboratory measurement purposes, microphones are usually used which respond to pressures and not to pressure gradients. On balance a microphone working on the principle of a constant-charge variable-capacitance is preferred, although piezo-electric microphones show great promise.

Condenser microphones are extremely rugged and are able to withstand great changes in environmental conditions. Above all, they are perfect enough as acoustic-electrical transducers for their characteristics to be neglected during measurements if properly chosen and set up. Size is an important factor; a typical 1 in o.d. microphone has a useful frequency range of 20 c/s to 18 kc/s and a dynamic range of 15 to 150 dB. The corresponding figures for a typical $\frac{1}{4}$ in o.d. microphone are 30 c/s to 100 kc/s and 70 to 180 dB. Once calibrated, the accuracy of a complete sound measuring system can be within $\pm$ 0·2 dB.

The condenser microphone can be used with advantage in the great majority of investigations of an acoustical nature, such as the determination of performance of electro-acoustic equipment, the accurate measurement of sound pressure levels etc.

8. References

1. H. O. Taylor, *Phys. Rev.*, **2**, p. 270, 1913.
2. G. Rasmussen, "Pressure equalization of condenser microphones and performance at varying altitudes", *Brüel & Kjaer Tech. Rev.*, No. 1, 1960.
3. American Standard Specification for Laboratory Pressure Microphones, ZS 24.8–1949.
4. P. V. Brüel and G. Rasmussen, "Free field response of condenser microphones", *Brüel & Kjaer Tech. Rev.*, Nos. 1 and 2, 1959.
5. G. Rasmussen, "Miniature pressure microphones", paper presented at the 4th International Congress on Acoustics, 1962.
6. "Recommended Methods of Measurement for loudspeakers", I.E.C. Draft Recommendation, December 1961.
7. F. Ingerslev, "Måling af lineær og ulineær forvrængning i elektro-dynamiske højttalere", (Teknisk Forlag, Copenhagen 1953).
8. W. J. Warren and W. R. Hewlett, "An analysis of the intermodulation method of distortion measurement", *Proc. Inst. Radio Engrs*, **36**, p. 457, April 1948.
9. B. E. Bang, "A.f. non-linear distortion measurement by wide band noise", paper presented at the 4th International Congress on Acoustics, 1962.
10. D. E. L. Shorter, "Loudspeaker transient response", *B.B.C. Quarterly*, **1**, No. 3, p. 121, October 1946.
11. T. Sommerville, "Acoustics in Broadcasting", *Building Research Congress 1951*, Division 3, Part 1, p. 53.
12. T. F. W. Embleton and I. R. Dagg, "Accurate coupler pressure calibration of condenser microphones at middle frequencies", *J. Acoust. Soc. Amer.* **32**, No. 3, p. 320, March 1960.

Manuscript first received by the Institution on 10th January, 1963, and in final form on 5th May, 1963 (Paper No. 849/EA10).

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".

TIME-SHARING COMPUTERS

Following the development of two transistor computers in Japan, investigations were made of the problems related to time-sharing systems. The waiting at input-output devices and the waiting in computations are very important problems; the probabilities of the waiting at input-output and their mathematical expectations are given by input-output time intervals and the probability distribution of the computing time. The probability distribution of both input-output time intervals and computing times can be regarded as normal distributions.

By way of example, the value and the variance of a program execution time in a serial computer are given by determining the mean value and the variance of each instruction execution time, from which it is found that the waiting at input-output devices can be made small enough by dividing each computation into many blocks. Furthermore, the improvement factor of the information transaction capabilities in time-sharing systems can be defined. The dependency of the improvement factor upon the number of input-output devices is analysed for a typical computation program.

"Waiting and improvement factor in time-sharing systems", Y. Ishi. *N.E.C. Research and Development*, No. 4/5, pp. 119–30, April 1963. (In English.)

PARAMETRIC FREQUENCY MULTIPLIERS

An analogue computer can be applied successfully to yield useful solutions for the 'large-signal' analysis of parametric frequency multipliers. The quantitative results, obtained in graphical form, give a vivid picture of the behaviour of the non-linear circuit. The main advantage of this approach is the relative ease with which the parameters can be changed and adjusted to bring about optimum solutions. A whole range of non-linear, time-varying systems, such as parametric amplifiers, phase-distortionless limiters etc., can be investigated in this way. All these problems have the common factor that the otherwise linear circuit, which contains one or more non-linear capacitor diode, can be described by a set of ordinary non-linear differential equations. If the non-linear charge-voltage relationship of the capacitor diode is different from the square-root law, as in the case of an 'hyper-abrupt' junction, it can be mechanized in general by employing an electronic function generator.

"Analysis and simulation of parametric frequency multipliers", W. M. Rupp. *Archiv der Elektrischen Übertragung*, **17**, No. 7, pp. 338–44, July 1963. (In English.)

DOUBLE FREQUENCY CONVERSION TELEVISION RECEIVER

Since many television receivers now incorporate a u.h.f. channel selector, the possibility arises of making such receivers simpler and cheaper by taking advantage of this unit for reception on v.h.f. A v.h.f. converter described in a German paper does not contain any tuning devices for channel selection, but only a selector switch for bands I and III. Both v.h.f. bands are transposed in their entire width into the u.h.f. range. The selection of the required v.h.f. channel then takes place in the u.h.f. channel selector.

The problems arising from the need for wide-band amplification and from the double frequency conversion are discussed; the results of measurements enable a comparison to be made with the circuits in normal use up to the present.

"Double frequency conversion as a novel method for the reception of television transmissions on v.h.f.", H. Hein. *Nachrichtentechnische Zeitschrift*, 16, pp. 317–20, June 1963.

DISTORTION IN COLOUR TELEVISION

Investigations by a Czech engineer into the signal quality of N.T.S.C. colour television have shown that, parasitic phase modulation and asymmetrical transmission, causes differential gain and differential phase distortion in the circuits following the detector and this in turn distorts the colour rendering on the receiver picture tube. Two cases of detection are considered, namely synchronous detection and envelope detection. The relations for calculating the distortion are stated and distortion values of the deepest base and complementary colours are given for the case of the envelope detector.

"The influence of parasitic phase modulation in the video transmitter on the signal quality of N.T.S.C. colour television", Z. Hajoš. *Slaboproudý Obzor*, **24**, No. 4, pp. 209–13, April 1963.

BROADBAND TUNED CIRCUITS FOR FREQUENCY-SHIFT KEYING

A recent paper, published in Germany describing work carried out in America, shows how high signalling rates or large transmission bandwidths are achieved over high-Q circuits and antennae by modifying the response of the circuit in such a way that the steady state response is obtained instantaneously.

In f.s.k. transmissions the frequency may be changed and an additional reactive element, with the appropriate amount of stored energy, is either inserted or removed at those instants when the stored energy of this element is maximum. Inaccuracies in establishing the initial conditions for the switched elements, timing errors of the switches and incorrect circuit tuning will cause residual transients, which have been calculated.

"Electronic broadbanding of high-Q tuned circuits or antennae", J. Galejs. *Archiv der Elektrischen Übertragung*, **17**, No. 8, pp. 375–80, August 1963. (In English.)

SUB-MILLIMETRE WAVE GENERATION

The need for sub-millimetre wavelength radiation has become important in recent years with growing activity in the field of plasma physics. For plasma densities of thermonuclear interest (i.e. greater than $10^{15}/\text{cm}^3$) only wavelengths below one millimetre can propagate; for these waves the plasma behaves as a dielectric producing a phase shift which is a measure of charge density, a parameter of considerable importance.

An Australian paper surveys some of the techniques which are being employed for generating these wavelengths. These include incandescent body radiators which operate at 5000° K but give very low power, decreasing inversely with the fourth power of the wavelength. Frequency multiplication of a 10 kMc/s fundamental can give a seventh harmonic level of 10 mW. Backward wave oscillators have been used recently in France to produce 10 mW power at 0·7 mm. Solid-state devices, for example the tunnel diode and the mixer diode, show intriguing possibilities. A point contact tunnel diode has been developed which can produce an oscillation range tunable from 93 kMc/s to 103 kMc/s with an output of a few tenths of a microwatt, when mounted in a reduced-height rectangular waveguide.

"Sub-millimetre wave generation", L. C. Robinson. *Proceedings of the Institution of Radio Engineers, Australia*, **24**, pp. 507–13, June 1963.

WAVE PROPAGATION

Using the well-known sheath model a German paper calculates the propagation characteristics of the helical line for the case of a homogeneous dielectric in the inner space of the helix or in the space between the helix and the outer conductor. It is then shown that at frequencies below 150 Mc/s the behaviour of this line can no longer be explained on the basis of the sheath model. A different mode of wave propagation makes its appearance which leads to far higher values of delay and characteristic impedance. The propagation characteristics are approximately calculated with the aid of a ladder network model consisting of a standard 'fourpole'. For both cases some measurements have been carried out whose results agree with the calculations to a good approximation.

"The propagation of microwaves and long waves on a helical line with dielectric and coaxial outer conductors", H. Heynisch. *Archiv der Elektrischen Übertragung*, **17**, No. 5, pp. 254–60, May 1963.

INFRA-RED RADIATION

A group of papers on infra-red radiation have been published in the French journal *Acta Electronica.*

In the paper "Transmission of infra-red radiation in the atmosphere" by A. Arnulf and J. Bricard, two phenomena exerting an influence on the transmission of radiation by the atmosphere are studied, namely the absorption by gases (involving the emission of a spurious radiation) and the absorption resulting from the diffusion due to the particles constituting fogs and mists. A method of computing the average values of the absorption co-efficients of the components of the atmosphere is given and, as an example, the results for the absorption by vapour. The spectrographic devices used for the measurement of the absorption by diffusion are described and the results obtained for different kinds of fogs and mists are compared with the results given by the granulometric measurements of the atmosphere. It is shown that the use of infra-red radiation detectors improves considerably the limits of perception of an object in a highly absorbing diffusing medium (pp. 409–25).

The second paper "Noise phenomena in infra-red ray detection" by H. Dormont and M. Auphan makes a survey of noise phenomena and the limitations they set to the precision of physical measurements. The use of correlation functions makes it possible to compute the noise generated by a signal passing through a linear system. The statistical properties of photons in thermodynamical equilibrium with a black body emitter are summarized. The noise generated by various types of detectors is analysed using the notion of a perfect detector which is shown to be an energy detector, e.g. the radiation bolometer. The noise properties of a perfect quantum detector are then computed in comparison with those of the true perfect detector. Photo-electric cells are studied as concrete examples of quantum detectors (pp. 427–58).

Another paper by J.-J. Brissot and A. Dauguet, entitled "Infra-red radiation detectors" covers specially the 'punctual' detectors, i.e. cells by which the received energy can be measured in a unit of time and at a given point. The principal infra-red detectors are: thermal detectors (bolometers, thermocouples, pneumatic cells) and selective detectors (photo-emissive, photo-conductive, photo-voltaic and photo-electromagnetic cells) (pp. 459–516).

The fourth and final paper, by F. Desvignes on "Analysis of the operation of the optical radiation detectors" is a comprehensive study of physical phenomena involved in optical radiation detectors and leads to the establishment of the relations between geometrical and chemical structure, main ambient parameters and the optical and electrical properties of these detectors. The limits of detection of the 'perfect detector' are examined, taking into account the influence of problems such as matching of detector area and beam structure, optical coupling, spectral range, ambient temperature, stray light; these factors are related to the actual detectors, giving the main values of active and reactive components of the parameters of the equivalent circuit (pp. 517–92).

Acta Electronica, 5, No. 4. (Dated October 1961 but published 16th March 1963.)