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Commentary

AN event of far reaching importance in the world of communications took place in London last month when an agreement was signed between the General Post Office, the American Telephone and Telegraph Company, the Canadian Overseas Telecommunication Corporation and the Eastern Telephone and Telegraph Company of Canada for the provision of a telephone cable to be laid in 1955 across the Atlantic Ocean between Scotland and Newfoundland.

Only preliminary details have so far been announced of this great project which will provide some 29 telephone circuits between the United States and this country with six to Canada as well as a number of telegraph circuits.

The existing coaxial cable from London to Scotland will be extended to Oban which is the eastern terminal of the new cable and a microwave radio link will be provided between Nova Scotia and New York.

The cost—about £12 500 000—will be shared between the countries concerned and the major British contribution will consist of the manufacture and laying of some 2 250 miles of cable on the transatlantic section, together with some 360 miles of cable between Newfoundland and Nova Scotia.

The submerged repeaters on the transatlantic section will be of American manufacture for the reason that more experience has been gained in America in the performance and reliability of repeaters at great ocean depths.

This is an important factor, of course, for the replacement of a faulty valve or component can be a most expensive undertaking.

Work on submerged repeaters has reached an advanced stage in this country as well and recently one of the longest submarine telephone cables with submerged repeaters was laid in the North Sea between Denmark and Holland, although not at such great depths. Submerged repeaters of this type will be laid on the short section between Newfoundland and Nova Scotia where the depth is less.

It is barely one hundred years ago that the small tug *Goliah* steamed out of Dover Harbour on August 28, 1850, carrying on her decks a submarine telegraph cable to be laid across the English Channel between Dover and Cap Gris Nez. This was, in fact, the first submarine cable to be laid, and although it was very short lived it provoked tremendous feelings of wonder and amazement throughout the world. A second and more successful cable was laid in the following year and there then followed a period of great activity of cable laying between England and her continental neighbours. There were many failures, of course, due to inexperience both in design and manufacture and in the techniques of cable laying, but the early pioneers, notably the Brett brothers, were men of resource and courage, and it is primarily due to such men that Britain became the centre of the world's system of communications.

On the other side of the Atlantic there had not been any great need for a network of submarine cables due principally to geographical considerations, but the overland telegraph line system had by this time extended to the Eastern States and Newfoundland. It was therefore a logical development that the next step should be an attempt to link the New World with the Old with a telegraph cable spanning the Atlantic.

This proved to be a most difficult undertaking and the first transatlantic cable was a disastrous and costly failure, involving a loss of about £500 000. After several set-backs in 1857 the laying of the first cable was finally completed on August 5, 1858, with the exchange of congratulatory messages between Queen Victoria and the President of the United States. But almost immediately the insulation of the cable began to fail and within a few months the cable had gone completely dead.

Great as were these disappointments, they did not cause complete despondency and plans were prepared for an improved design of cable and for better methods of cable laying.

On July 13, 1865, the *Great Eastern* left London with nearly 2 300 nautical miles of the new cable in her holds, but no sooner had the cable laying commenced than several mysterious faults made their appearance and this, together with the difficulties encountered in the methods of recovery at great depths led the expedition to be abandoned for the time being. However, in the following year the *Great Eastern* made a second voyage across the Atlantic which this time was successful, and on July 27, 1866, the two continents were firmly joined together.

From then onwards a considerable expansion took place and by the turn of the century practically all the corners of the world were linked together by a vast network of submarine cables. But this virtual monopoly of long distance communication was not to remain unchallenged, for the early experiments of Marconi had shown that other methods were possible, and in 1907 the first transatlantic radio telegraph circuit was opened for public use.

A fierce competition grew up between the cable and wireless companies for the world's traffic, and with the opening of the direct beam-radio service between this country and Australia, Canada, South Africa and India in 1927-8, the advantage appeared to be with the wireless companies. There was a tendency at the time to assume that the submarine cable was obsolete and that the beam service with its impressive performance would eventually replace all other methods of long distance communication. Experience had shown that the radio system could operate at considerably higher speeds and both initial costs and maintenance of radio links were very much lower. But experience also showed that radio links suffered bad fading periods during which reception was impossible.

Until 1927, however, the world's long distance traffic, whether by wireless or by cable, was still in telegraph form and it was not until January 7 of that year that the first transatlantic radio telephone circuit was opened between London and New York.

With the subsequent unification of the rival companies it soon proved that the two services were, in fact, complementary rather than competitive. The cable could provide a highly reliable system for the transmission of telegrams all the year round, while the radio promised the only long distance method of communication by the spoken word.

A High-speed Precision Tachometer

By W. R. Bland*, A.M.Inst.E., and B. J. Cooper*, B.Sc.

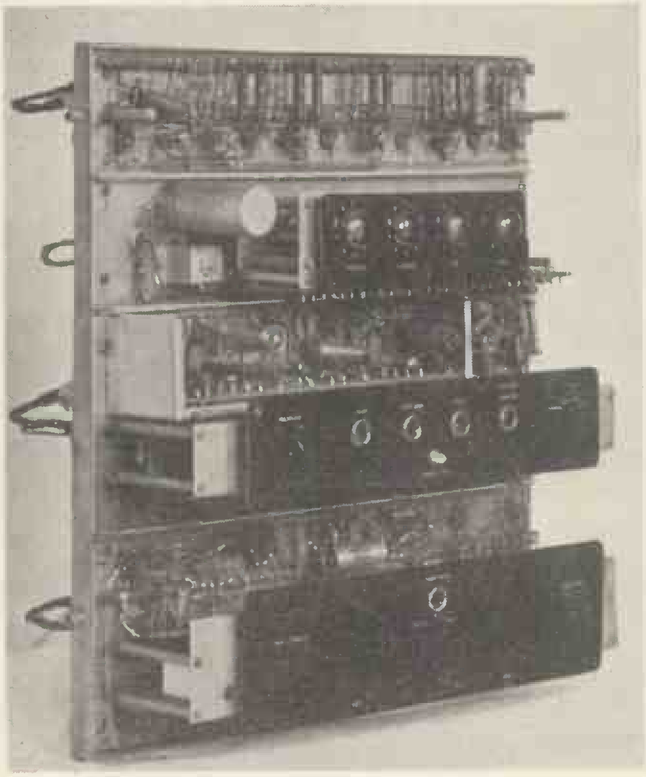
The availability of cold-cathode counter tubes prompted the development of an electronic precision tachometer of the integrating type having ranges of 0—8 000 R.P.M. *and* 10—80 000 R.P.M. *with a reading accuracy of* 0·01 *per cent. The instrument provides either single or repetitive indication and may be used with most types of pick-up device. The equipment may also be used as a frequency meter.*

IN the investigation of vibration problems on rotating machines it is often necessary to measure angular velocity with considerable accuracy. Since tachometers commercially available did not seem to fulfil the existing requirements it was decided to investigate an alternative method of making precise measurements.

A method using a cathode-ray oscilloscope in conjunction with a highly stable calibrated oscillator and a small alternator driven from the machine shaft was tried and found to be unsatisfactory due to difficulty in making rapid measurements of changes in speed. Another difficulty encountered was that this arrangement did not provide a direct numerical indication of speed and was therefore unsuitable for use by semi-skilled operators. It was eventually decided to devise an all electronic method making use of the recently developed Ericsson cold-cathode glow discharge tubes[1] to give a visual display. The basic system devised was as follows.

A pulse signal having its repetition rate proportional to angular velocity was obtained from the shaft pick-up and the total number of pulses occurring during a known interval of time was to be counted.

Owing to the wide speed range to be covered, i.e., from zero to 8 000 R.P.M. it was decided initially to make use of a shaft pick-up consisting of a toothed wheel rotating in close proximity to an electro magnetic pick-up. This provides an output pulse each time a tooth passes the pick-up. In the interest of accuracy it is desirable to have a maximum number of teeth on the wheel and in this case it was decided that 30 teeth would be sufficient. Thus 30 pulses per revolution were obtained from the pick-up. In order to record a direct indication of speed in revolutions per minute it was necessary to count the number of pulses generated in a two-second period. At a later stage in the development of the equipment it was decided to incorporate an additional refinement in the form of a one-second period which would be useful in some applications when measuring rotational speeds, and also enabling the equipment to be used for direct measurement of frequencies up to 4 000c/s, this being the maximum counting rate of the cold-cathode glow tubes used.



The tachometer with covers removed.

In the application envisaged it would be required to continuously monitor the speed and in order to do this it was necessary to design the equipment to be fully automatic in operation. Thus after the two-second counting period the count should remain visible on the display unit for a time sufficient to enable the reading to be recorded, the display then being cancelled immediately before recounting. To ascertain the optimum value of this reading time a series of tests were made and it was found that the minimum time required for the average operator to interpret and record the reading was 5 seconds. On the finalized equipment it was thought desirable to make this period variable to allow for different operators and operating conditions. In addition to the automatic operation of the system, provision was made for manual operation whereby a single reading would be obtained by the operation of a push-button, the count remaining visible until the button is again operated.

Development

TIMING PERIOD.

Since the ultimate accuracy of the instrument is absolutely dependent upon the accuracy of the period in which the pulses are counted, the duration of this period must be accurately defined. It was decided that the most suitable frequency standard to employ was that utilizing a 100kc/s quartz crystal having an accuracy of ±0·005 per cent. The crystal was to be operated in an oscillator circuit giving a pulse output which would be fed to a frequency divider chain giving an output of one pulse per second or one pulse per two seconds corresponding to counting times of one second and two seconds respectively. It was found that the maximum reliable division ratio of each stage of the frequency divider was 10 so that the most convenient arrangement was as shown in Fig. 1.

* *The Brush Electrical Engineering Co. Ltd.*

This consists of two decade dividers followed by a binary divider (which may be switched out) followed by three further decade counters. One of the advantages of this system is that it is possible to obtain standard frequency outputs at 100, 10 and 1kc/s for testing and calibration purposes. Various types of frequency dividing circuit were investigated including blocking oscillators, phantastron type dividers, energy storage counters and several types of multivibrator dividers and it was concluded that the most satisfactory type for this application was the type of assymetrical multivibrator divider described by Attew[2] (Fig. 6), each having a division ratio of 10. Two of these multivibrators were tried and found to give reliable long term stability in dividing from 100kc/s to 1kc/s. For the binary divider the obvious choice was an Eccles-Jordan type circuit which was driven by a pulse forming stage in order to ensure reliable operation. This stage can be switched out in order to give a one-second timing interval.

For the final three decade dividers it was decided to use Dekatron counter tubes because of their inherent reliability. These tubes are preceded by a pulse shaping circuit. The three tubes are coupled by amplifiers and the final output pulse is used to trigger the signal gating circuits.

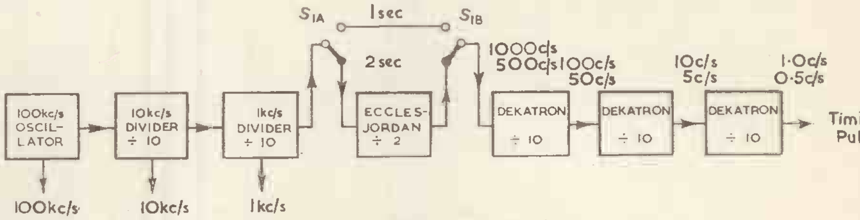


Fig. 1. The dividing stages

Gating Circuit,

The output pulse from the divider chain must be made to actuate a gating circuit, such that the gate is open for the required timing period during which the signal is passed to the counter circuit. Since each counting period must be followed by a reading period during which the gating circuit must be inoperative it is evident that some form of self-locking circuit must be used in conjunction with the gating circuit. There are several forms which this may take and it was found that the most reliable was that based upon the Eccles-Jordan circuit. This is shown in Fig. 2.

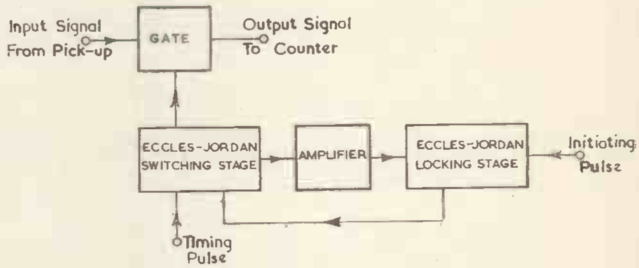


Fig. 2. The gating circuit

The gate valve is controlled by an Eccles-Jordan circuit which is itself locked by a second Eccles-Jordan circuit. The timing pulses are applied to the first stage and are ineffective until an initiating pulse has been applied to the locking stage. In order to render the operation of the system automatic the initiating pulses must be applied at the correct instant. These pulses are obtained from a pulse generating circuit as shown in Fig. 3.

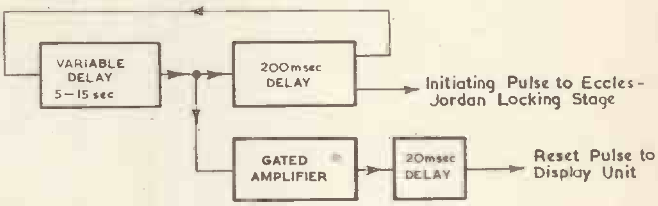


Fig. 3. The pulse delay circuits

The delay circuits are connected as shown to form a closed loop and run independently of the timing circuit. Assuming that the 5-15 second delay flip-flop is pulsed, then after the requisite delay, the 200msec and 20msec flip-flops will be initiated, the latter via a gated amplifier. Thus after 20msec the counters are cleared and after 200msec the locking stage is reset and the variable delay flip-flop reinitiated.

The gated amplifier is fed from the stage in phase opposition to the signal gate so that while counting is in progress no stray pulses can trigger the 20msec flip-flop and cause the counter to be falsely reset to zero.

The output from the gating circuit is fed to a pulse generator driving the first counter tube. The output from this tube is fed through an amplifier to another pulse forming stage driving the second counter tube. The output of which is fed to the third and fourth counter tubes as shown in Fig. 4. During the development of the display unit cold-cathode trigger tubes were originally used as coupling valves but were found to be rather sensitive to change in H.T. voltage and were replaced by thermionic valves.

Resetting.

In order to reset the display counters to zero the glow discharges must be transferred to the output cathodes. The conventional method of carrying out this operation is to apply a negative pulse to the output cathodes. This was tried, and although the glow discharges were transferred to the output cathodes the trailing edge of the resetting pulse was transmitted to the next counter tube in each case and caused it to register. Several modifications were tried including the use of rectifiers to prevent this but none of these were entirely satisfactory. It was. therefore, necessary to design a new system for resetting in which a positive pulse was applied to the ordinary cathodes and guide electrodes, thus causing the discharge

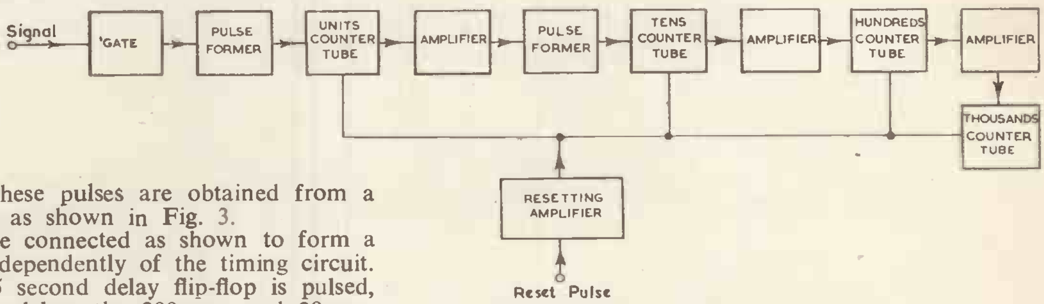


Fig. 4. The display unit

to transfer to the output cathodes. This arrangement was found to be completely reliable under all conditions of operation.

While work was in progress it was decided that it would be advantageous to incorporate a decade frequency divider in the input circuit so that the instrument could be used to measure speeds up to 80 000 R.P.M. Since the accuracy required was the same as at 8 000 R.P.M. no modification to the display unit would be required. The input signal is fed to an amplifier biased to cut-off (Fig. 5), the output of which is fed through a diode clipper to a second limiting amplifier, this in turn feeds a Schmitt trigger circuit, the output of which can be switched to the decade divider when required. This divider is a scale-of-ten unit of the Eccles-Jordan type, the output from this circuit being fed to a flip-flop circuit driving the gate valve.

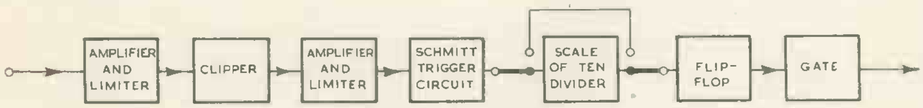


Fig. 5. Input stage

Circuit Description

TIMING PANEL. (Fig. 6.)

The crystal oscillator V_1 is a modified Colpitt's type over-compensated in the anode circuit to provide a large positive over-shoot which is differentiated and shaped by cathode-follower V_2. The output from V_2 feeds a synchronizing pulse to a 10kc/s asymmetrical multivibrator V_3 whose output is differentiated and fed to a second multivibrator V_5 operating at 1 000c/s. The output from this multivibrator is differentiated and the negative pulse is used to trigger a flip-flop circuit which generates a square wave having a short wavefront suitable for triggering the following Eccles-Jordan circuit formed by V_8.

This stage may be switched out by switch S_2. Following this stage a second flip-flop is used as a driver stage for the first Dekatron divider V_{10} which gives an output frequency of 100c/s or 50c/s depending on the position of S_2. The two halves of a double triode valve V_{11} are connected as cut-off amplifiers to couple between successive Dekatrons.

Thus the output pulse from the last Dekatron consists of a positive pulse occurring at one second or two second intervals as required. This pulse of approximately 30 volts amplitude is squared by trigger circuit V_{14}, the output of which is differentiated and fed to a pulse forming flip-flop V_{15} which supplies negative going output pulses of approximately 200 volts amplitude. It was found necessary to include V_{14} since the rise time of the output pulse from V_{13} was too long to ensure reliable triggering of V_{15}. Cathode-followers V_4 and V_6 were incorporated to provide calibration signals of 10kc/s and 1kc/s respectively. A neon indicator N_1 connected to V_{14a} anode is used to monitor the timing pulses.

INPUT CIRCUIT AND RELAY CIRCUIT (Fig. 7).

The input circuit is designed to work from positive pulses or sine wave input, the minimum input signal being 5 volts. The first valve V_{16} is a conventional pentode amplifier. The output being fed through a germanium diode clipper which eliminates spurious signals, to a phase invertor and limiting amplifier V_{17}, and hence to a Schmitt trigger circuit V_{18} which provides a constant amplitude square wave signal. By means of switch S_3 the signal from V_{18} can be fed direct to V_{19}, the following pulse shaper or to V_{19} via the decade scaler. This stage was included to give a pulse of constant amplitude and duration to drive the gate circuit V_{20}. The gate circuit V_{20} consists of a short suppressor base pentode gated on its suppressor grid, the signal being applied to its control grid. The gating pulse selector circuit consists of V_{21} and V_{22} arranged in an Eccles-Jordan circuit, V_{23} the isolating amplifier and V_{24a} and V_{24b} the locking stage.

In the quiescent state V_{24b} is conducting and V_{24a} is thus cut-off, its grid being at approximately 60 volts negative. Hence V_{22} which has its suppressor direct connected to the grid of V_{24a} is cut-off. Thus the input timing pulses which are applied to V_{21} suppressor and V_{22} control grid as shown cannot alter the state of the circuit. Hence the gating valve V_{20} which has its suppressor tied to the control grid of V_{22} is cut-off. If a negative initiating pulse is now applied to the grid of V_{24b} the stage is changed over so that V_{24a} conducts. This removes the negative potential from the suppressor of V_{22} and hence the next input timing pulse changes over V_{21} and V_{22} removing the negative potential from the suppressor grid of the gating valve V_{20}, which allows it to pass the signal to the display unit.

The next timing pulse changes over V_{21} and V_{22} again and cuts off V_{20}. This also causes a pulse to be passed through amplifier V_{23} to change over V_{24a} and V_{24b} so that the stage is again locked until another negative pulse is applied to the grid of V_{24b}.

AUTOMATIC OPERATING CIRCUIT

This consists of V_{25-29}. V_{28} is a flip-flop having switched time-constants of 5, 10 and 15 seconds, the output of which is applied directly to V_{29}, a flip-flop having a time-constant of 200msec. The output from V_{28} is also fed via a gated amplifier V_{26} to a third flip-flop V_{27} having a time-constant of 20msec. The output from the 200msec flip-flop V_{29} is fed back to the input of V_{28} so that V_{28} and V_{29} form a closed loop and run continuously independently of the remainder of the circuit.

The positive pulse at the grid of V_{27b} is fed to the display unit resetting valve V_{52} which is used to zero the Dekatron tubes. This pulse causes V_{52} to conduct and so discharge C_{105} through VR_4. The negative output from V_{29} is applied to the grid of V_{24b} to reset the locking stage and initiate the next counting period.

For manual operation the chain is broken by disconnecting the H.T. from V_{28} by means of S_1 and the 200msec and 20msec flip-flops are actuated by V_{25} and V_{26} respectively by means of a push-button in the grid circuits. On depressing the push-button a positive pulse is applied to the grids causing both valves to conduct, and hence apply negative pulses to the two flip-flops.

Neon indicator lamps N_{2-5} are connected at various points in the circuit to indicate the sequence of operations.

DEKATRON DISPLAY PANEL (Fig. 8.)

The input signal from the gate circuit is fed to the pulse shaping stage V_{41}. In order to achieve the maximum possible counting speed, i.e., 4 000c/s of which the Ericsson Dekatron counter tube type GC10B/S (CV2271) V_{43} is capable, the negative pulses applied to the guides must be specially shaped and sequentially applied. These conditions are fulfilled by use of the flip-flop stage V_{41}. The output from V_{41b} anode consists of a negative going square wave of 50μsec duration and approximately 200 volts amplitude.

This pulse is attenuated by R_{219} and R_{220} and applied to the first guide of the tube. The positive going square wave at the anode of V_{41a} is differentiated and applied to the second guide. The negative over-shoot producing the requisite glow transfer in the tube. (Fig. 10.) The diode V_{42} is connected to guide 1 to ensure that the guide bias level is unaffected by frequency. Guide 2 is self biased. The output from V_{43} is fed via an amplifier biased to cut-off, V_{44}, to a second driving flip-flop V_{45}. The negative going output pulse from the anode of V_{45b} is fed to both guides of V_{46}. The signal applied to the second guide being integrated by R_{237} and C_{98} to provide the required pulse waveforms.

It was found essential to reform the pulse obtained from the output cathode of V_{43}, since at the highest frequency

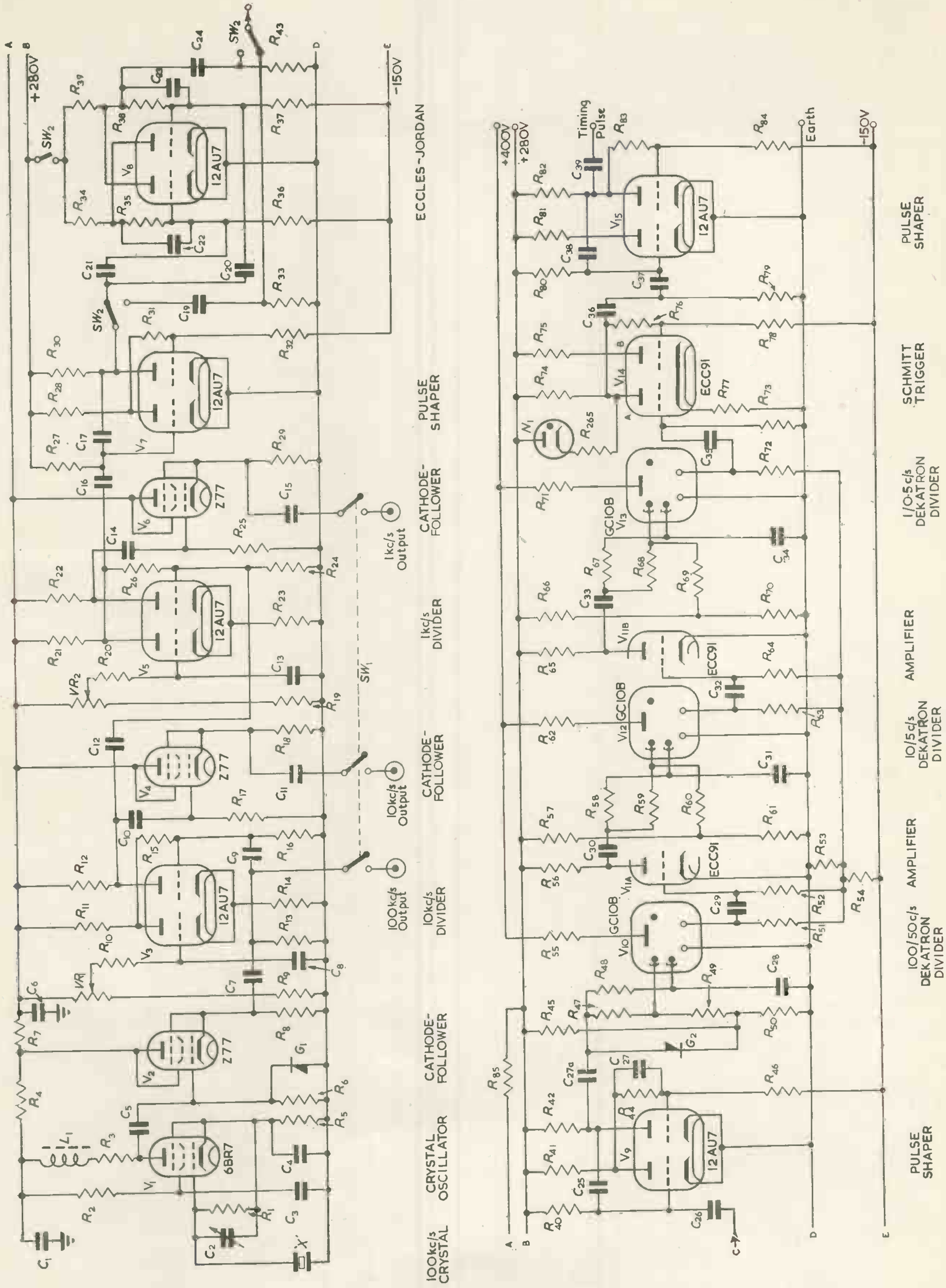


Fig. 6. Timing unit

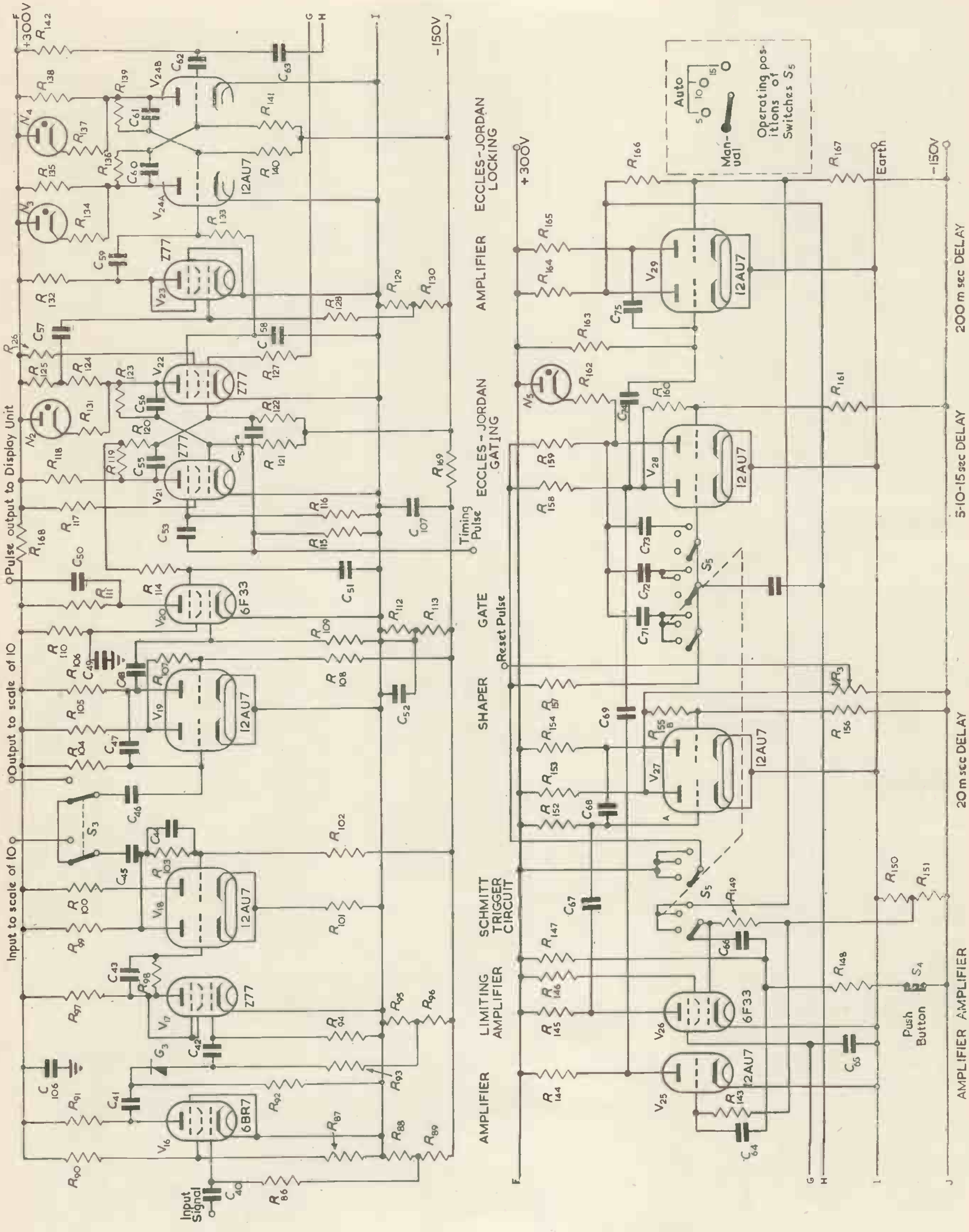


Fig. 7. Gating unit

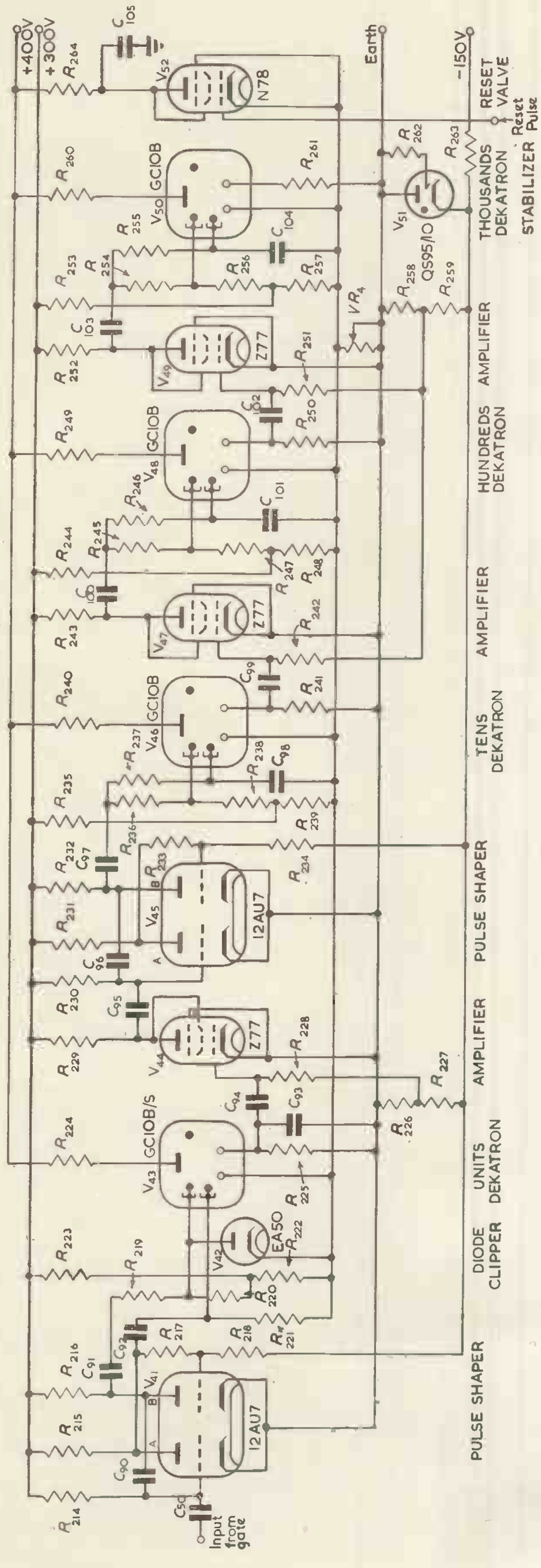


Fig. 8. Display unit

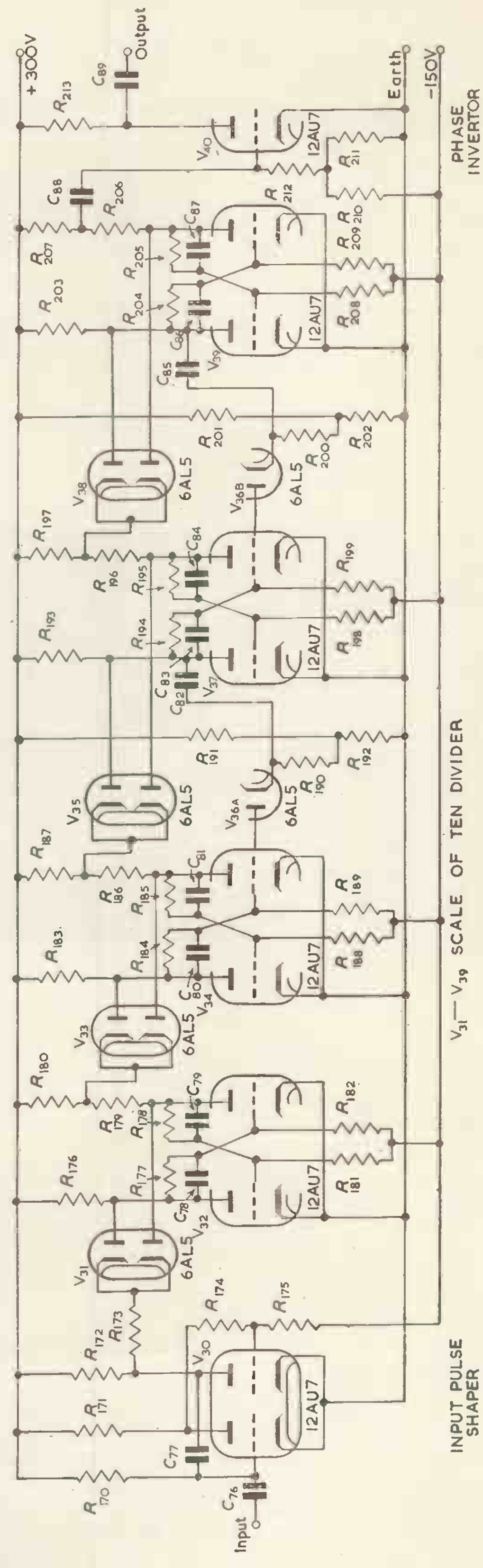


Fig. 9. Scale-of-ten divider unit

of operation the pulse duration at the output cathode of V_{43} was insufficient to trigger the following Dekatron tube reliably. Succeeding stages are fed via amplifiers V_{47} and V_{49}.

The reset valve V_{52} is a low impedance pentode having its anode fed from a 400 volt supply via a 100kΩ resistor R_{264}. The valve is normally cut off so that C_{105} connected between V_{52}, anode and earth is charged to approximately 400 volts through R_{264}. The grid is D.C. connected to the anode of V_{27a} (the 20msec flip-flop) so that when V_{27} is triggered the grid of V_{52} is driven up to earth potential and C_{105} is discharged through V_{52} and resistor VR_4, thus applying a positive pulse to the Dekatron guides and cathodes and causing the glow to transfer to the output cathodes, i.e. zero set the tubes.

Decade Scaler Panel (Fig. 9.)

This consists of a conventional type of Eccles-Jordan decade divider[3-6]. The input is fed to a pulse shaping stage V_{30} the output of which is fed to the scaler via a coupling diode V_{31}.

The scaler consists of $V_{32,34,37,39}$, forming the Eccles-Jordan stages coupled by means of diodes $V_{33,35,38}$. The feedback circuits are coupled via diodes $V_{36a,36b}$. The output from the unit is taken from the anode of V_{39b} through a differentiating circuit through the amplifier V_{40}.

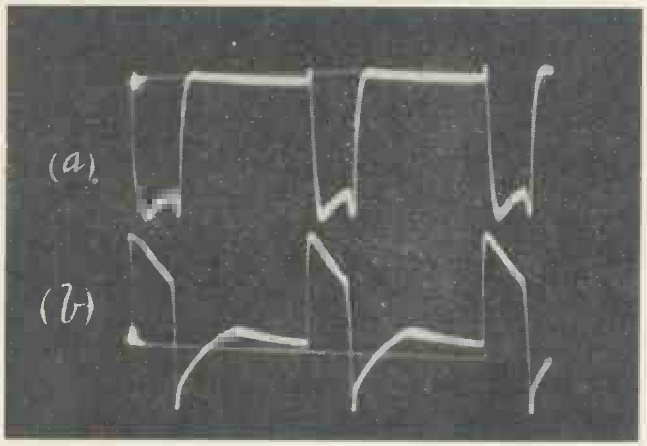


Fig. 10. Circuit waveform

a) V_{41b} anode waveform applied to guide 1 of V_{43}

(b) Differentiated waveform applied to guide 2 of V_{43}

This scaler unit employing diode couplings will operate at frequencies in excess of 200kc/s and so for the frequency range involved in this equipment completely reliable operation is obtained.

Power Supplies.

Power is supplied to the equipment at 400, 300 and 280 volts positive and at 150 volts negative. These supplies are all obtained from a conventional series valve stabilized power unit. Fuse protection is incorporated in all supplies and a no-volt relay circuit on the negative supply guards against bias failure.

Conclusions

While it was originally intended to feed the counter in conjunction with an electro-magnetic pick-up, the addition of a decade scaler to extend the range to 80 000 R.P.M. made the use of this type of pick-up impossible. Accordingly, a capacitance type pick-up was designed which could be used in conjunction with a frequency modulated carrier system to provide suitable input signals over the entire speed range. It is anticipated that a photo multiplier type of pick-up would also be suitable although this has not yet been tried. The main limitation to this type of counter for general use is the relatively long counting period required when using a shaft pick-up having 30 teeth. In the application for which the counter was designed, however, i.e., for measuring the speed of high inertia rotating machinery, this period was not considered to be unduly long. The limitation being that instantaneous speed is not recorded.

For other applications it is obviously possible to reduce the two-second counting period, with a consequent reduction in the frequency response of the instrument. The limit in this respect is at present set by the maximum operating frequency of the counter tubes used. With the future promise of higher frequency counter tubes becoming available it will be possible to greatly reduce the counting period if required. Future modifications to this equipment will be the simplification of the timing pulse generator by the substitution of a low frequency 1kc/s crystal for the present 100kc/s unit and subsequent elimination of the first two divider stages.

Acknowledgment

The authors wish to express their thanks and appreciation to the Brush Electrical Engineering Co., Ltd for permission to publish the article.

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DEVELOPMENT IN DIVERS' INTERCOMMUNICATION EQUIPMENT

Considerable improvements have been made in the efficiency of divers' intercommunication equipment as a result of collaboration during the past two years between C. E. Heinke and Co., Ltd., submarine engineers, Kenton Laboratories Ltd., sound recording specialists, and The General Electric Co., Ltd.

The new equipment, which enables divers to keep in touch with each other and with the diving party on the surface, is a development of the Diverphone equipments which have been widely employed since the war. It differs from them in that the H.T. supply, previously obtained from a motor generator in the back of the surface unit, is now provided by a vibrator. The circuits have been overhauled, miniature valves have been introduced into the amplifier circuit to help render it more compact, and the construction is now on the unit principle to simplify servicing. The results of all this work are louder surface signals, less service troubles and a smaller drain on batteries. A further innovation has been the substitution of a head-operated transceiver (a speak-listen transducer and switch unit in the diver's helmet) for transceivers of the earlier chin-operated type.

So great is the gain obtainable from the amplifier that the merest whisper in the divers' helmets is audible to all on deck, while divers have even reported being able to hear the calling of gulls over the water. Naturally there are serious snags attaching to so high a performance. Not only must mechanical vibrator hum not be transmitted through the chassis and metal case of the set, but air-vibration affecting the relatively large speaker-cone must be eliminated unless the divers are to hear a perpetual background hum.

The most troublesome problem was that of hum in the vibrator pack. It was solved by the production of a special low-loss transformer, and by replacing as much of the steel in the chassis, frame and case as possible by copper and brass. Special attention was devoted to component layout, and the resilient mounting of the vibrator unit in a sponge-rubber tube, held away from the chassis by a moderately flexible metal stay.

The possibility of damage due to inadvertent reversal of polarity at the accumulator is obviated by a circuit design which embodies a cut-out incorporating a selenium rectifier and relay.

Because the air pressure under which a diver works affects the pitch of his voice, which rises as the pressure rises, it becomes progressively more difficult for his signal to overcome the perpetual air delivery noises into his helmet. Frequency response has therefore been so controlled in the new equipment that interference noises are reduced to the absolute minimum.

Valve Reliability in Digital Calculating Machines

By L. Knight*, A.M.I.E.E.

This article discusses measures which can be taken to reduce the number of valve failures and to minimize the inconvenience caused by the remaining unprevented failures. Although these measures are considered primarily with regard to digital calculating machines many are applicable to other electronic equipment.

ONE of the most important problems associated with the design of digital electronic calculating machines is that of reliability. This is especially true now that such equipment is beginning to be used for accountancy where work must often be carried out to a rigid timetable and frequent interruptions for servicing cannot be tolerated.

The most common cause of breakdowns in electronic equipment is the valve and it is therefore this component which must receive most of the designer's attention. He must not concentrate solely on reducing the number of breakdowns due to valve failures but must remember that this is only one aspect of a wider problem, that of minimizing the inconvenience which valve failures cause to the user.

The procedure for tackling the problem would appear to be divisible into the following distinct steps:

(1) Choice of inherently reliable valve types.
(2) Measures to minimize valve deterioration.
(3) Designing the circuits to be as tolerant as possible to changes in valve characteristics.
(4) Testing valves before insertion into the machine.
(5) Comprehensive testing of the complete machine.
(6) Preventive maintenance in the field.
(7) Provision of facilities for dealing rapidly with any unprevented failures.

Choice of Valve Types

Considerable research into valve reliability is now being done by many valve manufacturers. When the improved types become readily available for industrial purposes they should prove valuable, even though most of the research at the moment does not appear to be directed at the same kind of reliability that is required in calculating machines. The main emphasis at present seems to be on valves which will operate reliably for 1 000 to 2 000 hours under extremes of ambient temperature, in equipment which may be subjected to rough mechanical treatment[1]. In a calculating machine the main requirement is for valves which will work for 20 000 hours or more in a stationary equipment, usually in a temperate climate.

Valves often vary very considerably from one manufacturing run to another and it is not very helpful to use large scale life-tests to determine which types and makes are the best. A life-test takes several years to conduct and at the end of that time the results cannot be taken as applicable to current production valves. Consequently it would seem that the designer of electronic apparatus must rely more on an understanding of the causes of valve failures and combine this with the experience of those who have used large quantities of valves in the past to predict which contemporary valves have the greatest potentialities for reliability and longevity.

One factor that must be borne in mind when choosing types is that valves are most reliable when manufactured in continuous and smooth-flowing production runs. It is wise, therefore, to avoid types which are only produced in small quantities or which are obsolescent and this suggests that the choice should fall on types from the popular range of miniature glass-based valves.

These valves, however, have some disadvantages. The contact between the pins and the valve holder is less reliable than with the octal base and sometimes gives rise to noise or intermittent failures. There is also the danger that, when the valve is inserted into the holder, stresses may be set up in the glass. These may not be evident immediately but may cause cracks to develop at a later date. In spite of the development of special techniques, which include the avoidance of stiff wiring to the valve holders and the use of wiring jigs, the situation does not appear to be completely satisfactory. Rowe[1], for example, claims that damage due to valve insertion alone jeopardizes the chance of obtaining a failure rate of less than 1 per cent per 1 000 hours.

The safest course seems to be the elimination of conventional valve holders altogether. The G.P.O. have found it advantageous in repeaters to spot-weld short flexible leads to the valve pins and then solder these wires into a special valve holder[2].

In other respects the reliability of miniature valves appears to be comparable with that of larger ones. The scaling down of the electrode assemblies has produced certain manufacturing difficulties and resulted in higher bulb temperatures, but these set-backs seem to have been successfully overcome. As regards invulnerability to physical shock and vibration, the miniature is generally better than the larger types.

Since one of the greatest causes of valve breakdown will be heater failures it is worth while to consider which valves can be expected to have the lowest heater mortality. It would seem that thick heaters tend to last longer than thin ones and that, because of this, a low voltage heater is preferable to a higher voltage one of the same power rating. On the ENIAC it was found that the failure rate of valves with folded heaters was considerably higher than that of those with twisted heaters[4]. Thus, for example, one would expect the ECC33, which has two 3·15V 0·4A spiralled heaters in series, to be better than most makes of 6SN7, which have two 6·3V 0·3A folded heaters in parallel.

Minimizing Valve Deterioration

Having selected valve types which seem to have the greatest potentialities for long life it is essential to ensure that these can be realized. It is advisable to operate all valves at very conservative ratings, to adhere to the standard code of practice[3] and particularly to any special recommendations of the manufacturer.

A lower heater voltage than the nominal one specified by the manufacturer will lengthen the life-expectation of

* *The British Tabulating Machine Co. Ltd.*

the heater, but any cathode deterioration which may occur will then be far more pronounced. However, provided that the anode voltage is low, there may be some overall advantage in slightly under-running the heater. In the ENIAC, where anodes and screens are operated at about half their maximum continuous service voltage ratings and about a quarter their maximum current ratings, experiments tended to show that the optimum heater voltage for 6·3V valves lay somewhere between 6·0V and 6·3V[5].

It seems likely that heater failures will be reduced if it can be arranged to avoid the heavy heater current surge which normally occurs on switching on. This is admittedly only conjecture and tests on the ENIAC failed to prove that it gave any measurable advantage[4]. Nevertheless, it seems worth while to err on the cautious side and incorporate this feature. If voltage stabilization is used on the heater supply it will only require a minor modification to give a gradual initial voltage rise.

Another unconfirmed supposition is that it may be mildly injurious to the cathode to apply any anode or screen potentials before the heater has fully warmed up. Thus it may be beneficial to arrange for a delay of about two minutes on the H.T. supplies.

Wherever possible it should be arranged that if any fault occurs in the machine it does not have any disastrous repercussions in any other part of the machine. Obviously it is impossible to cover every contingency but it is essential, at least, to ensure that no widespread valve destruction can ensue from a failure in a master circuit or in a bias voltage supply.

Special attention should be paid to valve cooling. With the large number of valves usually found in calculating machines a forced air draught is desirable. To avoid complications with dust filters and to enable the internal temperature to be independent of the room temperature, it may even be advisable to have a completely enclosed air circulation system with its own refrigeration plant[6].

Tolerant Circuits

A large proportion of the valves will not have sudden catastrophic failures. Their life span will be the time taken for their characteristics to deteriorate to a level which prevents reliable operation in a given circuit. Thus in order to obtain a life of, say, 10 000 hours, circuits must be used which will accept the changes in valve characteristics occurring during that time. Before this can be done, some estimate must be made of what these changes will be.

With valves made for repeaters, Eaglesfield[7] has found that the deterioration is due mainly to the growth of resistance of the interface between the cathode core and the oxide coating. After rising to a certain value within the first few thousand hours this resistance has been found to remain essentially constant at a value given by the following formula:—

$$R \simeq \frac{40}{A_k} \simeq \frac{120}{W_h}$$

where R is the interface resistance in ohms,

A_k is the area of the cathode coating in sq.cm,

W_h is the rated heater power in watts.

Results of a valve life test conducted by the author have shown that with standard commercial valves a greater deterioration can sometimes occur. 80 type ECC33 double triodes, consisting of samples from several different deliveries, were run for 3 250 hours under conditions which simulated a typical circuit application. The majority showed little change in anode current during the period but with 14 of the 160 triode units there was a marked fall. Fig. 1 shows the histories of several of the worst of these. The curves show a tendency for the rate of deterioration to fall after several thousand hours and it seems reasonable to suppose that over a period of 5 000 hours at least, and possibly over 10 000 hours, few of the valves would have fallen much below the level of d. Subject to any future evidence to the contrary, this level has been taken as representative of the worst deterioration to be expected with ECC33 valves over a period of 10 000 hours.

The static characteristics of d after 3 250 hours were found to be similar to those of an average new triode unit with a resistance, R_k in series with the cathode and another R_a in series with the anode. It is presumed that R_k is a measure of the internal cathode resistance and that R_a, since it limits the maximum possible anode current, is indicative of a drop in cathode emission. The values of R_k and R_a were approximately 1 000 and 7 000Ω respectively.

In the Hollerith type 541 multiplier* it has been made a criterion for any circuit using an ECC33 that it must operate satisfactorily both with a valve artificially aged to this level and with a top-limit new one. Moreover, any circuit element, such as an Eccles-Jordan trigger, which uses a plurality of triode units must be capable of working satisfactorily with any combination of good and bad triodes.

Under dynamic conditions at high frequencies a genuine aged valve may not behave so badly as a synthesized one,

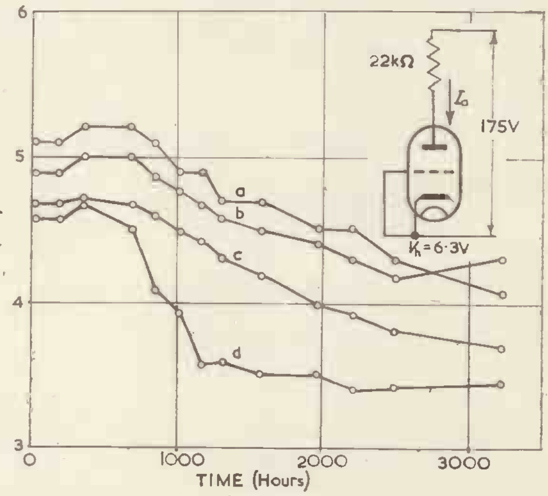

Fig. 1. Characteristics of several of the worst valves tested

The anode load was raised to 47kΩ between measurements to make the average anode current during the soak period approximate to that under equal mark-space ratio pulsed conditions.

due to the fact that the cathode interface resistance of the former will be shunted by a capacitance. This gives an added safety factor.

No comparable evidence has yet been obtained on what values of R_k and R_a to expect with any other valve types and tentatively it has been supposed that they will be related to the cathode area, and therefore to the heater power. A single ECC33 heater is rated at 1·26 watts and thus equivalent values for R_k and R_a for other valve types are estimated as below:—

$$R_k \simeq 1\,000 \times \frac{1{\cdot}26}{W_h} \simeq \frac{1\,300}{W_h} \text{ ohms}$$

$$R_k \simeq 7\,000 \times \frac{1{\cdot}26}{W_h} \simeq \frac{9\,000}{W_h} \text{ ohms}$$

In a pentode the screen current will also be restricted when the cathode emission fails. This can be simulated in an artificially aged pentode by a resistance in series with the screen, its value being such that it will limit the screen current by the same proportion as R_a limits the anode current.

* *An electronic multiplier, capable of working in decimal or £.s.d. notation, which is one of the range of Hollerith punched card accountancy machines manufactured by the British Tabulating Machine Company Ltd.*

The above methods of determining the characteristics of aged valves must be taken as very approximate. They are based on rather meagre evidence and make certain questionable assumptions. Nevertheless, they give a limit which, while providing a generous latitude for deterioration, does not make circuit design unduly difficult.

The actual procedure for designing equipment to accept the required range of valve characteristics will depend on the circuits involved but the principles will be broadly as follows:

(a) To make each circuit element accept the widest possible variations of amplitude, shape and timing of the input pulses.

(b) Then to ensure that the output waveforms of each circuit element will be always within the limits required by the succeeding stages.

Normally in a digital machine a valve will only be used as a two state device. It will either be cut off or conducting, the grid in the latter state normally being clamped at the cathode potential by grid current. The input waveform, therefore, must traverse sufficiently negative to ensure that the grid is beyond the cut-off point and sufficiently positive to override the voltage drop across R_k. If two valves have their grids commoned, allowance must be made for the fact that the valves may have different values of R_k. Each grid should be connected through a resistance to prevent them both limiting at the same potential.

To ensure the minimum variation in the amplitude of the anode waveform the anode load should be as high as possible so as to swamp the effects of R_a and R_k. If necessary, the sharpness of the waveforms can be improved by the use of limiting diodes[6].

In certain instances, such as the "read" amplifiers in a magnetic drum store, valves may be used in class-A. Here the effects of deterioration can be minimized by the use of negative feedback. Even though this may necessitate the use of extra valves to obtain the required gain, it will probably result in fewer valve replacements.

In the circuit design, account must be taken of the fact that component values will deviate from the nominal figure and that any valve must give satisfactory results in spite of this. The fact that a component has a certain tolerance printed on it means that it was within those limits when tested at the time of manufacture. It does not necessarily mean that it will remain within those limits for a number of years, and an allowance must be made for the maximum anticipated drift. Even if cheaper components appear to be adequate it seems wise to err on the cautious side and use stable and close tolerance resistors and capacitors if they will give a greater margin for valve variations. In fact, it seems a sound principle that everything possible should be controlled to tight limits to allow the greatest safety margin for the most uncontrollable item—the valve. In line with this policy, the power supplies should all be stabilized, and the D.C. ones free of ripple.

The use of simple circuits is usually preferable. It is unwise to rely on characteristics which are not utilized in the normal applications of a valve. The manufacturer may not be controlling that characteristic and there may not be evidence available on what variations to expect. It is probably advisable even to avoid the use of pentodes. The variations of all the different electrode characteristics within their permitted tolerances, together with changes due to cathode deterioration, make it difficult to assess what variations in performance to expect. With a little ingenuity a pentode can often be replaced by a double triode and it would seem preferable to do this whenever possible.

One factor which should not be overlooked is the possibility of parasitic oscillations. If one side of an Eccles-Jordan trigger oscillates at a very high frequency the trigger may become monostable. A pulsed valve which oscillates can radiate to another, apparently unconnected, circuit where it may be rectified and produce a spurious pulse. Faults due to parasitic oscillations may occur sporadically as stray capacitances and inductances change and can be very baffling. It is infinitely preferable to avoid the possibility of such faults by making a strict practice of including stopper resistors in all control grid and suppressor grid leads. Long earth leads from the valve holders should also be avoided, especially when they are common to several electrodes.

Valve Testing

The only satisfactory type of valve reliability test seems to be a batch test, every valve in the batch being rejected if the performance of the batch as a whole is not satisfactory. Samples from the batch can then be tested to destruction. For such a test to be most useful the valves should be kept in the same batches as those in which they were manufactured. This, coupled with the economic consideration that vast numbers of valves may be rejected to effect the elimination of a few faulty ones, makes it rather impracticable as far as the user of valves is concerned.

Even though no quantitive measurement of reliability is possible it is still worth while to test each valve individually before insertion in the equipment.

The anode current can be measured under conditions which simulate the conducting state of the valve in the circuit in which it is to be used. Those valves with the lowest anode current can be rejected on the assumption that they are likely to reach the lowest acceptable limit sooner than the others. The choice of a test limit will be to some extent an economic one, dictated by what proportion of rejects can be tolerated. A reasonable test figure seems to be one which rejects about 5 per cent. This eliminates any which are exceptionally low without incurring a large wastage.

Another safeguard is to reject those valves with high reverse grid current, this being indicative that the gas pressure is high and that, in consequence, there is more danger of cathode poisoning. Here again the test limit will have to be determined on economic grounds.

Checks can also be made of other characteristics, such as the grid cut-off voltage, to ensure that they are within the required limit. There can be a mechanical inspection for any visible defects, such as loose electrode assemblies or hot spots on the heaters, which may impair reliability.

Testing of Complete Machine

Considerable attention must be paid to the test procedure of the complete machine. The tests must ensure that everything is working exactly as the designer intended and that there are not any faulty components enjoying the tolerances which he has meant as a safeguard for valve deterioration. An extensive check should be made to verify that the output waveforms of every stage are within the limits to be expected with new valves. It is also extremely desirable to check that each stage will tolerate the required variations in input waveforms. This latter test will only be practicable if the machine is constructed of small units which can be tested individually prior to their incorporation in the machine.

It is highly advisable to test every component. Even with the highest quality components there are occasional ones outside limits or wrongly marked and, although these should show up in a comprehensive functional test, it is safer, and often quicker in the long run, to have a 100 per cent test of all components. Often it is possible to do this when they are wired into sub-assemblies, in which case it will check that no damage has resulted from handling or soldering.

Care should be taken before switching on for the first time that there is no fault which may endanger the valves. Immediately after switching on it is a good plan to check that all the grid potentials are safe. Although any faults in this respect would be discovered eventually, this action prevents the possibility of any valves passing excessive current for any length of time.

In electronic equipment generally it has been found that a large proportion of the valve failures, especially catastrophic ones, occur during the first 250 hours. It is prudent, therefore, to make the testing include a soak run which is long enough to ensure that the valves have been running for this period before the machine is passed to the user.

Preventive Maintenance

Even if the utmost care has been taken in the design, production and testing of an electronic calculating machine there will inevitably be some valve failures. Precautions must be taken to prevent these causing any serious inconvenience to the user. It is extremely valuable, for example, to institute periodic checks to detect any valves whose characteristics are approaching the danger level. These valves can then be replaced before they actually cause any errors. The most obvious way to perform this check would be to remove each valve individually to test it. This would be extremely tedious and it is very likely, especially with glass-based valves, that it would introduce more faults than it prevented.

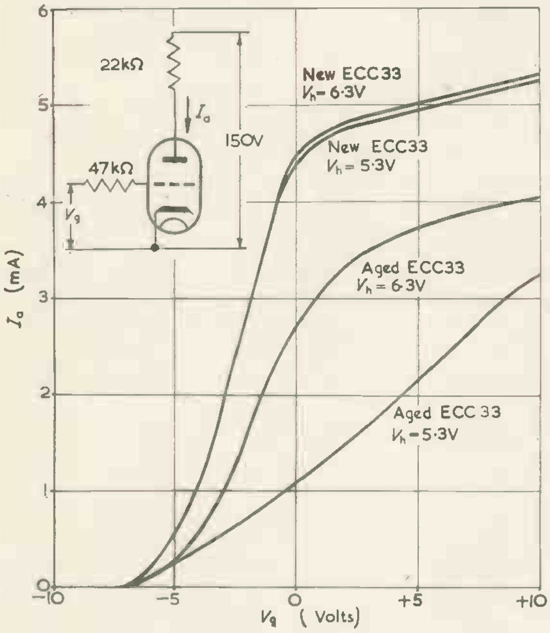

Fig. 2. Effect of heater voltage

The procedure which has been adopted with a number of computors is that of marginal checking. This usually involves putting the machine through a test run with one or more of the power supply voltages offset from their normal values and comparing its results with predetermined answers. Any valve which has deteriorated badly and has only been operating with a small safety margin will then cause an error which can be traced. On the ENIAC the pulse repetition frequency is raised for marginal checking[8].

If the marginal checking is made semi-automatic, as on the Whirlwind[9], it can be performed by the normal operator. The serviceman then need not be called in unless the result of the check is unsatisfactory.

An interesting point with regard to marginal checking is the effect of the heater voltage on the performance of a valve. If the valve has suffered no cathode deterioration there will be negligible change in its characteristics (with low anode voltages) if the heater voltage is lowered by 15 per cent. But if there has been deterioration this will become much more apparent as the heater voltage is lowered. Fig. 2 shows this effect in a typical application. It is supposed that an ECC33 is to be used with a 22kΩ anode load and an effective grid circuit resistance of 47kΩ. Assuming that in the conducting state V_g will be +5V, a typical new valve will pass 5mA and a badly aged one (d of Fig. 1) will only pass 3·7mA. If this latter figure is near the danger point the valve will fail when the heater voltage is reduced to 5·3V and the anode current falls to 2·1mA.

Facilities for Rapid Servicing

Not all valve failures will result from progressive deterioration and measures must also be taken to deal with unpredictable failures. The first requirement is that any fault should be discovered as soon as possible and the best way of ensuring this is to incorporate some form of self-checking. Often this can be done by arranging that the machine automatically performs every calculation by two different methods and compares the two results. With complex computors this is not always possible and some other system must be devised. In the S.D.C. Raytheon computor[10], for example, each number has associated with it a check number which is the weighted sum of the binary digits. These check numbers are used to verify arithmetic operations and also to check that numbers do not suffer any mutilation in transfers or in storage. In the UNIVAC[11] the binary code for each character (which may be a decimal digit) contains a check digit which is chosen to make the sum of that group of digits odd. During all transfers each character is examined to check that the sum of the digits is still odd. Periodically every character in the acoustic delay line storage is also examined. Arithmetic operations are checked by performing them simultaneously in twin arithmetic units which check against each other.

There must also be a service engineer within easy reach and facilities for him to diagnose the fault easily. Neon indicators can be included in the machine to give a visual display of the contents of valve registers. They can even be used to indicate short-circuit capacitors and open-circuit resistors. It may prove worthwhile to build an oscilloscope into the machine. If the machine does not use a dynamic storage medium it is a great advantage to have facilities for pulse by pulse operation at low speed.

An engineer who is thoroughly acquainted with the functioning of the machine can often diagnose a fault very quickly from the nature of the errors it causes and it is extremely helpful to have a series of test problems especially prepared for this purpose. This method of fault-finding is invaluable for intermittent faults.

There must also be provision for the engineer to rectify faults quickly. The machine should be made of easily replaceable units of as few different types as possible. The engineer then need not isolate the precise fault but only narrow it down to the faulty unit and replace this.

In many cases it will be preferable to have several small identical calculating machines sharing a job rather than one large machine doing it unassisted. A fault then will not immobilize the whole installation, but only one unit, and it will still be possible to continue at reduced speed.

Valve Failure Figures

Most of the measures outlined above were adopted with the Hollerith type 541 multiplier. Miniature valves were not used, however, because at the time of the conception of this machine their suitability for calculating equipment was not felt to be sufficiently well proved. The two main types employed were the ECC33 double triode and the 6F32 short suppressor base pentode, both octal based valves. The original valve testing and maintenance techniques were not as thorough as outlined above.

None of the multipliers has been in service long enough yet for any extensive evidence on valve failures to be obtained but the failure rate to date, excluding the 250 hour soak period, has been less than 1 per cent per 1 000 hours. With only a small number of failures it is impossible to draw any definite conclusions, but there is an inference that the ECC33 is more reliable than the 6F32. On later versions of the multiplier where valve failure prevention is being tackled more strictly in accord with this article it appears probable that the failure rate will be less than ½ per cent per 1 000 hours.

In the Manchester University computor the valve failure rate has been approximately 1 per cent per 1 000 hours over a period of more than 5 000 hours[6]. It is interesting to note that 180 out of the 239 valve failures during this time were due either to low emission or faulty valve bases. This suggests that could these two faults have been eliminated the valve failure rate on this machine would also have been less than ½ per cent per 1 000 hours.

It would appear from the evidence of these two machines that, even with standard commercial valves, it should be possible to obtain a valve failure rate of ½ per cent per 1 000 hours for the first few thousand hours. If failures continue at this rate indefinitely the average valve life would be 200 000 hours, that is to say, about 100 years if run for 40 hours a week. It seems presumptious to expect this to be achieved and it must be concluded, therefore, that after a time the failure rate will increase. Experience may prove it expedient to withdraw a machine from service when the failure rate exceeds a certain level and replace all the valves.

Conclusion

Summarizing, it would seem that the problem of valve failure involves the consideration of a variety of widely different factors. Although the greatest onus for valve reliability falls on the valve manufacturer, it would appear possible for the designer of electronic calculating equipment to achieve, for the first few thousand hours at least, a valve failure rate of less than ½ per cent per 1 000 hours.

Having reduced the number of failures to a minimum there are still other important measures which can be taken to reduce the inconvenience caused by the remaining unprevented failures. Although this article has been confined to valves it is worth noting that some of these measures apply equally well to failures due to other causes.

Acknowledgments

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AN AMPLIFIED-TORQUE CLUTCH

By R. Voles*

The device described gives an approximately linear control of large torques at slow speeds. The torque is provided in the first place by a magnetically-actuated plate clutch. This torque is then amplified by means of a torque amplifier of the tape type.

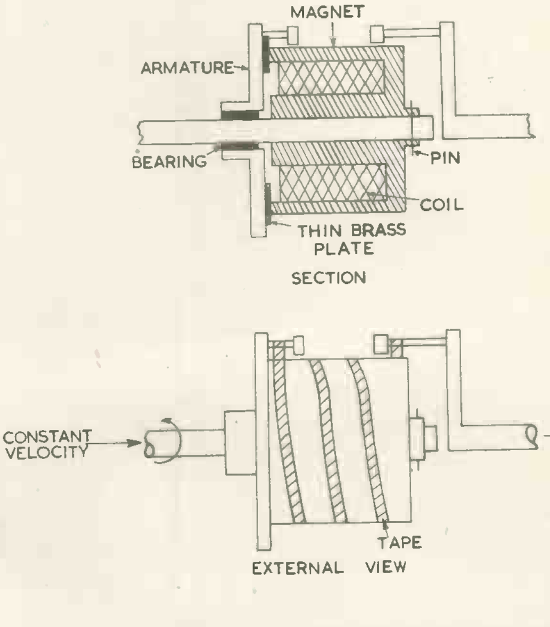
The primary torque produced is that required to cause slip between a flat-faced armature and a "pot" electromagnet which is rotated at constant velocity. The magnet is mounted on the driving shaft, and the armature is mechanically free to move on a bearing along the same shaft when the magnet is energized. The magnet current is conveyed to the coil through slip rings and control of the current gives an approximately linear control of the torque, which is therefore virtually independent of slip speed. A thin non-magnetic sheet is fixed to the face of the armature to reduce the effects of remanance.

In the torque amplifier the tape is wound around the cylindrical outer surface of the magnet. The tape is pinned to a point on the circumference of the armature, passed around the surface of the magnet the requisite number of times and fixed to a crank on the load shaft which is coaxial with the magnet and the driving shaft.

The primary torque and the torque amplifier are both unidirectional; for servo purposes, therefore, it is necessary to use two clutches driven in opposite senses. This arrangement has the added advantage that small constant currents may be maintained in the magnet coils to ensure that the tapes are permanently taut.

The chief advantage claimed for this method over one in which the tape amplifier is controlled by a torque motor is a reduction in size and complexity. Also, unless two torque motors are used, the unit cannot be split into two halves and situated conveniently in the equipment. Another advantage is that the device automatically takes up any movement which might be caused by a variation in tape length.

Cross section and external view of the clutch.



* *E.M.I. Engineering Development Ltd.*

D.C. Amplifiers

Methods of Amplifying and Measuring Small Direct Currents and Potentials

(Part 1)

By J. Yarwood*, M.Sc., F.Inst.P., and D. H. Le Croissette†, M.Sc.

The principles of the single-valve amplifier for the amplification and measurement of small unidirectional currents and potentials are established. Grid current and electrical leakage and methods of measuring and reducing them are discussed. Data on high value resistors and on electrometer valves are tabulated with critical notes. Triode and tetrode electrometer valves in single- and double-valve circuits are reviewed in relation to the current and voltage sensitivities attainable. The attenuation of the effects of supply variations leads to a consideration of the stabilization of power supplies for D.C. amplifiers and the use of parallel-balance circuits, differential input circuits, cathode-coupled balanced amplifiers and the use of a valve as a cathode load. There is included accounts of receiver valves suitable for electrometer use, methods of inter-stage coupling, and relevant information on noises and drifts.

THERE are numerous applications for amplifiers of direct currents and potentials: in photo-electric measurements, for *p*H meters, to the so-called biological amplifiers, for thermocouple thermometry, in mass spectrometry, using ionization chambers in recording fundamental particle phenomena and in X-ray and γ-ray techniques, to vacuum gauges of the thermionic and radioactive ionization types, in cathode-ray oscillography, and so on. In general, for currents greater than 10^{-7} to 10^{-6}A it is preferable to dispense with an amplifier and make direct use of a suitable moving-coil meter unless it is essential that a robust or panel-mounted meter be used or that considerable power be developed. The chief concern here, therefore, is the amplification of direct currents of less than 10^{-6}A (1μA). For amplifying very small currents, between 10^{-15}A and 10^{-8}A, perhaps the best solution in many cases is to use a vibrating reed electrometer[1]. This instrument converts the D.C. to A.C. by the use of a vibrating reed capacitor and employs then a high gain tuned A.C. amplifier. Such instruments are commercially available with selectable input resistances of 10^{8}, 10^{10} and $10^{12}\Omega$. The P.D. developed by the applied direct current can then be read to an accuracy of 2 per cent or less with a maximum reading of 1 volt on a multi-range, panel mounted moving-coil meter. The advantage of such apparatus is the high stability attained. There is no directly-coupled amplifier on the market with the high performance of the vibrating reed electrometer. However, the expense of such apparatus is often unjustifiable, it is very difficult to design and build in the laboratory, it is fairly bulky, and the range of D.C. inputs that can be used may be inconvenient. The concern in these articles, therefore, is the principles of the directly-coupled electron tube amplifiers which can be readily constructed for the expenditure of a few pounds and which, provided various precautions are taken, give satisfactory performances in a variety of applications. The galvanometer amplifier[2] and the magnetic amplifier[3] will not be considered.

The Simple, Single-stage D.C. Amplifier

Though subject to many disadvantages and incapable of satisfactorily amplifying currents of less than 10^{-8}A, the study of the use of commercial receiving valves at normal operating potentials forms a useful preliminary.

A source of direct current of high impedance such as a photo-electric cell (Fig. 1(b)) or ionization chamber at a saturation potential, is placed across a resistor R which is in series with the negative operating bias, V_g (Fig. 1). This combination is made the input to the valve, which has a moving-coil meter in its anode circuit if the output current is to be measured, or where an anode load R_L is used if an output voltage—usually to be fed to a subsequent stage—is needed. The negative bias V_g is to the linear part of the mutual characteristic of the valve at the operating anode potential.

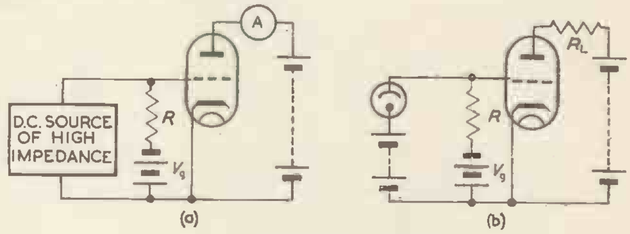


Fig. 1. Simple D.C. Amplifier
(a) *Output current measured.*
(b) *Voltage across anode load R_L measured.*

Suppose the D.C. input is changed by ΔI, or a D.C. input of ΔI is established. It is assumed that I changes to $I \pm \Delta I$ and that the current then remains at its new value for a considerable length of time compared with the time-constants of the circuit. The corresponding change of P.D. across R is $R \,.\, \Delta I = e$. If the static mutual conductance of the valve in Fig. 1(*a*) is g_m then the consequent anode current change is $\Delta I_a = g_m \,.\, e = g_m \,.\, R \,.\, \Delta I$.

The current gain m_i is therefore given by:

$$m_i = \Delta I_a / \Delta I = \frac{g_m \,.\, R \,.\, \Delta I}{\Delta I} = g_m \,.\, R \quad \ldots\ldots (1)$$

This is often expressed in terms of the current sensitivity S_i of the amplifier where $S_i = d/\Delta I$, d being the deflexion in millimetres of a moving-coil galvanometer used as the anode current meter A (Fig. 1) where this meter is a lamp and scale instrument with a working distance of one metre, and ΔI is the D.C. input which produces the deflexion d. It is readily seen that:

$$S_i = \Delta I_a / \Delta I \,.\, \sigma = g_m \,.\, R \,.\, \sigma \quad \ldots\ldots\ldots (2)$$

where σ is the current sensitivity of the galvanometer in mm/A, and S_i is also in the units of mm/A. It is noteworthy that high values of R, g_m and σ are necessary to obtain an amplifier of high current sensitivity.

The voltage sensitivity will likewise be $S_v = d/e$.

$$\therefore S_v = d/e = \Delta I_a / e \,.\, \sigma = g_m \,.\, \sigma \quad \ldots\ldots (3)$$

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† *The Physics Department, The University of Southampton.*

In this case a high g_m and σ are required. Since the voltage sensitivity is independent of R, so the input source impedance can now be low, as in the case of a thermocouple. It is, however, often difficult to achieve high values of S_v whereas S_i, dependent on R, can be made very large.

If an anode load R_L is used (Fig. 1(b)) then equation (1) becomes:

$$m_i = g_m' . R \qquad (4)$$

where g_m' is the dynamical mutual conductance which is less than g_m because of the presence of R_L.

It is preferable, however, to use the normal stage-gain formula derived from the equivalent anode circuit theorem where the output voltage e_L across the anode load R_L is related to the input voltage change $e = \Delta I . R$ by:

$$\text{voltage gain} = e_L/e = \frac{\Delta I_a . R_T}{\Delta I . R} = \frac{\mu R_L}{R_L + r_a} \qquad (5)$$

and

$$\text{current gain} = \Delta I_a/\Delta I = \frac{\mu R_L}{R_L + r_a} . R/R_L \qquad (6)$$

where r_a is the anode slope resistance of the valve.

Instead of reading the anode current change brought about by the D.C. input a simple null method is to balance the input voltage $e = \Delta I . R$ against an opposing change of the bias potential using a potentiometer R_1 across the bias battery (Fig. 2) so as to return the anode current to its initial value after establishing the input. Then e equals the measured change of bias. This method has two advantages over the simpler procedure: (a) it is independent of the linearity of the mutual characteristic of the valve; (b) the grid current is the same at both times when a reading is taken.

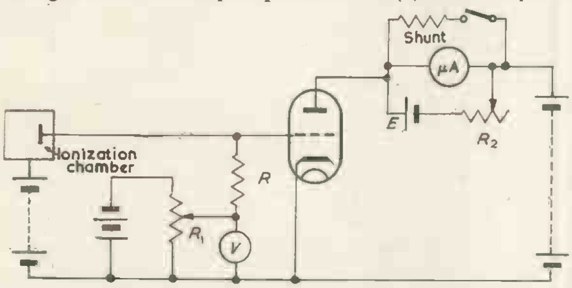

Fig. 2. D.C. amplifier with provision for "backing" out steady, no input anode current.
Null method of measurement used.

It is often desirable to balance out the initial steady anode current prevailing before the D.C. input to be measured is applied. Then zero reading of the anode current meter will conveniently correspond to zero D.C. input. This is readily achieved by the adjustment of the variable resistor R_2 in the "backing" circuit of Fig. 2. Here R_2 must be considerably greater than the resistance of the galvanometer if the sensitivity of the latter is not to be reduced by shunting. This is usually the case in practice. An extra 1·5V cell E is involved. A further outstanding advantage of such a provision is that the anode current meter can be a sensitive microammeter which, for the purposes of preliminary adjustment, is shunted. The null method described in the preceding paragraph is much more sensitive using such an arrangement (Fig. 2).

The most satisfactory method of achieving a sensitive balance method is, however, to use the two-valve bridge circuit described later.

Grid Current and Electrical Leakage

Referring to equations (1) and (4) it is seen that the current gain is directly proportional to the resistor R used across the D.C. source. It is therefore desirable to make R as large as possible. There are, however, two severe restrictions to increasing R inordinately: (1) the impedance of the source of current; (2) the input impedance of the valve due to (a) grid current, (b) electrical leakage. These all reduce the effective gain. Using a directly coupled input of the type so far discussed, a high current gain demands a high impedance source. There still remains, nevertheless, the necessity of reducing grid current and electrical leakage. Indeed, using commercial receiving types of triode and pentode valves at their usual operating potentials, the grid current restriction is such that R cannot exceed 10^7 or possibly $10^8\Omega$ in practice. Here grid current is much more significant than electrical leakage so it will be considered first.

Sources of grid current when the grid potential is negative are[4,5,6]:

(1) Initial thermal velocities of electrons from the heated cathode.
(2) Positive ions formed in the residual gas by the accelerated electrons passing through it.
(3) Positive ions emitted from the cathode.
(4) Thermionic emission from the grid which is heated mostly by radiation from the cathode.
(5) Photo-electric emission from the grid.
(6) Ionization of the gas by the soft X-rays produced at the anode by the arriving electrons.

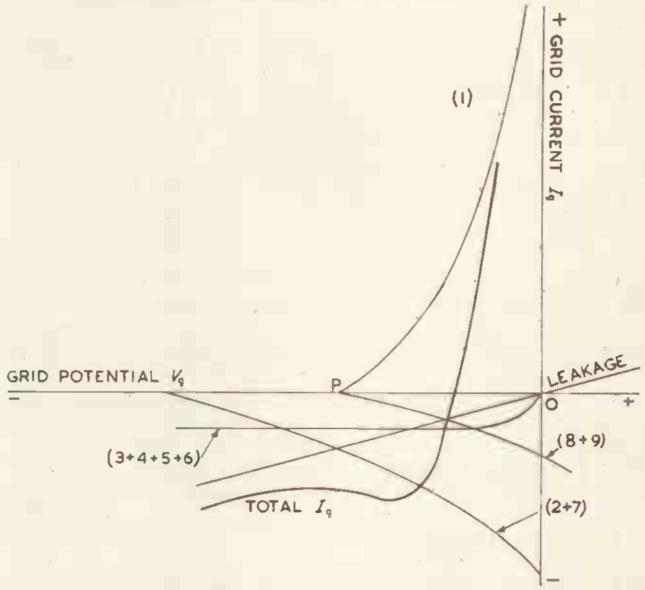

Fig. 3. Components of grid current and the resultant grid characteristic of a triode valve

(7) Positive ions released from the anode because it is heated and bombarded by the arriving electrons.
(8) Secondary emission from the grid due to positive ion bombardment.
(9) Secondary electrons from the cathode due to positive ion bombardment.

The manner in which these components contribute to the total negative grid current is shown in Fig. 3.

By simple methods it is possible to reduce all these contributions to grid current:

(i) Operate the valve at a sufficiently large negative bias. This will reduce effect (1). By far the majority of electrons emitted from an oxide-coated cathode at normal operating temperatures (1 050°K) have thermal velocities corresponding to less than one volt. Hence if the negative bias is at least one volt this effect (1) will be small. In practice the existence of contact potentials between cathode and grid leads to a preference for a bias of −2V or more. However, the grid current characteristic (Fig. 3) is best studied before deciding on the bias. If the negative bias is greater than the value corresponding to the point P on Fig. 3 then this effect (1) will be eliminated.

(ii) Operate the valve at reduced anode and screen potentials. This will reduce the effects (2), (6), (7), (8) and (9).

F

If the anode potential is less than the minimum ionization potential of the residual gas then the effect (2) will disappear completely, and this production of positive ions in the residual gas is the largest source of grid current, if the negative bias is sufficient to eliminate the effect (1). Ionization by collision usually begins at $V_a = 10V$ and is considerable at 24V. In practice V_a is set at about 12V. This reduces the gain seriously, hence a higher V_a up to 24V may be used if high gain is required rather than very low grid current. The residual pressure in the evacuated valve is also important: it should be as low as possible. It might be possible—nobody has tried it apparently—to permit higher anode potentials by ensuring that the residual gas is primarily one like helium with a high minimum ionization potential (24V). This might be easier than ensuring a very low pressure in a sealed off valve envelope.

(iii) Run the valve at lower cathode temperature than normal, e.g., heater voltage of 4·5 instead of 6, say. This will reduce, in particular, effects (1), (3) and (4), as well as reducing the effects enunciated in (ii) because the anode current will be less. The effect (4)—thermionic emission from the grid—is usually negligible with small cathode powers[6].

(iv) To avoid much of the photo-electric emission from the grid the valve is best in a light tight box. Nielsen[7] states that a ¼in diameter hole in such a box near the valve will, because of room light, give a photo-electric current of 10^{-12}A in some cases. There will remain photo-electric emission due to light from the cathode. This is, of course, reduced by lowering the cathode temperature. Photo-electric currents of 10^{-14} to 10^{-15}A are probably produced by light from the cathode at normal operating temperatures[6].

With the valve at reduced operating anode and screen potentials the chief source of grid current remaining—apart from electrical leakage—will be positive ions emitted from the cathode. This is important when endeavouring to amplify currents of less than 10^{-13}A. It is reduced by the use of a space-charge grid as in the electrometer valve described later. Warren[5], however, asserts that if the cathode is underrun at a temperature of less than 950°K then positive ion emission in an electrometer valve with a filamentary cathode power of 0·1 watt is only 10^{-17}A and there is hardly need for a space-charge grid. On the other hand, Metcalfe and Thompson[4] indicate that positive ion emission currents as great as 10^{-13}A are obtained with a filamentary cathode power of 0·25 watt.

Electrical leakage (see Fig. 3) is effectively added in parallel with grid current, the two together reducing the input impedance of the valve. Electrical leakage may be due to:

(a) getter on the pinch and micas of the valve,
(b) imperfect insulation of the valve base and socket,
(c) leakage between leads in the input circuit,
(d) moisture films on the valve, associated leads and the resistor R.

The effect (a) can be reduced by direction of the gettered metal away from the pinch and micas during valve manufacture and by using pinch and micas with roughened surfaces produced by sand-blasting or spraying with alumina.

To avoid (b) as far as possible it is preferable to use valve sockets of ceramic or alkathene or, for very small current measurements, remove the valve base and connect directly to the pinch wires. These precautions are relatively unimportant if a valve with a top-cap grid connexion is used.

In connexion with (c) leads in the input circuit should be screened and spaced by air or good quality insulators such as polystyrene.

Moisture films (d) are reduced by enclosing the valve and resistor R in a light-tight box of earthed metal to provide screening, which box contains a silica-gel drier. In extreme cases, using acorn valves, an evacuated box is recommended[7].

Further data on avoiding electrical leakage, especially by using water-repellant silicone laquer, is given later.

The Measurement of Grid Current

If it is required to amplify currents of less than 10^{-10}A it is best to measure the grid current characteristic of the valve to be used before embarking on a particular design. These characteristics are, of course, supplied by the manufacturers for electrometer valves, but such figures can—especially under reduced cathode temperature conditions—only be the average figures for a batch of valves. Any particular valve may depart considerably from this average as regards grid current.

The method of measuring grid current[4,6,7,8,9] depends on its order of magnitude. For currents in excess of 10^{-12}A a simple method is available provided a series of high resistors, say 10^{10}, 10^{11} and $10^{12}\Omega$, are at hand. A similar method, using a vibrating reed electrometer, enables grid currents down to 10^{-14}A to be measured, but for these smaller currents a rate of drift of grid potential procedure has been commonly adopted in the past.

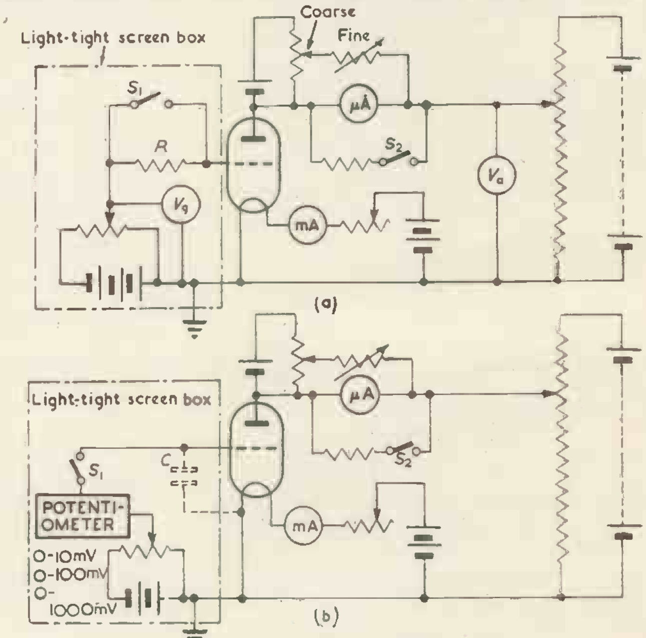

Fig. 4. The measurement of grid current
(a) *Using a known resistor R.*
(b) *Measurement of rate of drift of grid potential.*

With refinements described by Hay[6] a suitable circuit for grid currents greater than 10^{-12}A is shown in Fig. 4(a).

The resistor R, say $10^{12}\Omega$, is inserted in series with the grid and a variable bias supply. A high insulation switch S_1 is provided to short-circuit R when required. The whole of the input circuit arrangements are in a light-tight metal shield with screened cable leads to the input of the valve. To obtain a grid characteristic, I_g/V_g, where I_g is the grid current, the anode and filament voltages are first set at the desired values. With the negative bias at V_g the anode current is recorded with R in the circuit. This anode current can be brought to a convenient low value, sensitively indicated by a microammeter, by using the "coarse" and "fine" "backing" controls in conjunction with a shunt switched in and out by S_2. The resistor R is then short-circuited by closing S_1 and the alteration of V_g necessary to restore the anode current to its previous value is found. This alteration ΔV_g of V_g must be equal to I_gR where I_g is the grid current at grid bias $V_g - I_gR$. Since ΔV_g is recorded by the meter, V_g

and R are known, so I_g is found. Hence an I_g/V_g characteristic is obtainable. Using a vibrating reed electrometer, R will be 10^8, 10^{10} or $10^{12}\Omega$, the input resistance of the electrometer, and the P.D. across R will be directly recorded to an accuracy of 1mV.

To measure grid currents of less than 10^{-12}A the usual method is to measure, in effect, the rate of drift of potential of the "floating" grid. In this case (Fig. 4(b)) the static input capacitance of the valve including leads needs to be known. Let this be C. With switch S_1 closed, measure the change in I_a for a change of V_g of, say, 10mV. A potentiometer of the Cambridge Instrument Co. pattern is useful for this purpose in series with the main grid bias supply. If the switch S_1 is now opened then the negative grid current will charge C in such a direction that the anode current I_a decreases. Find the time t in seconds for the change of V_g due to this charge of C to be 10mV, as determined by the decrease of I_a.

Then $Q = C \,.\, \Delta V_g = 10^{-2}C$, for $\Delta V_g = 10\text{mV} = 10^{-2}\text{V}$, where Q is the charge accumulated on the input capacitance. Since C is known and t is measured, so I_g is found. In practice C would be about $5\text{pF} = 5 \times 10^{-12}\text{F}$.

$$\therefore I_g = \frac{5 \times 10^{-14}}{t}$$

This method is capable of measuring 10^{-15}A and indicating 10^{-16}A. The limit in sensitivity is due to the drift of grid bias battery voltage which would take place if t were of the order of 500sec or more.

The switch S_1 is necessarily arranged so that the difference of its capacitance between opened and closed positions is very small. A highly insulated switch of thin platinum wires is preferable.

High Value Resistors

It is obvious from a consideration of equations (1) and (4) that an essential requirement in using valves for the measurement of D.C. after amplification is an accurately known, stable, low temperature coefficient, non-polarizable high resistor R. This resistor is also best shielded or mounted in a vacuum jacket so as to be independent of changes in the relative humidity of the surrounding atmosphere.

Since the war, resistors of values of up to $10^{13}\Omega$ have been commercially available. Most experimenters are well advised to use these for D.C. amplifier work. However, several ways of constructing high value resistors in the laboratory have been described. Brewer[10] discusses various types and considers a lead pencil line on a polished amber rod mounted in a chamber with drier to be satisfactory for values up to $10^{12}\Omega$. Curtiss[11] describes a less bulky resistor for values of 10^8 to $10^{12}\Omega$ fabricated from pyrex rod with lead pencil rubbed into ground grooves arranged along its surface and then coated with lacquer. Preston[12] uses ordinary 10MΩ resistors mounted zig-zag fashion in a paraffin wax block to give a total resistor of $10^8\Omega$, Van Atta[13] discusses resistors of up to $10^{12}\Omega$ made from thin, sputtered films on pyrex glass, White and Van Atta[14] give a design for electrolytic resistors of values 10^8 to $3 \times 10^{10}\Omega$, while Gemant's book[15] discusses liquid resistors in detail and Craggs[16] gives further data on their stability. Practical circuits for the measurement of high resistances are described by Rose[17], Turner[18] and Higgs[19].

Since most users would nowadays prefer to buy resistors rather than indulge in the uncertainty of making their own, the following is a summary of the commercially available high value resistors, the data being provided by the courtesy of the manufacturers concerned.

DUBILIER ULTRA HIGH RANGE RESISTORS

The resistance coating is applied to a small diameter glass rod, thereafter electrically aged, and assembled in a ceramic or glass tube with metal alloy end caps. The resistance tolerance is normally ±30 per cent with a minimum of ±20 per cent.

TYPE	MAX. RESISTANCE (Ohms)	MAX. WORKING VOLTAGE (Volts)	TUBE MATERIAL
FH1	5×10^8	500	Ceramic
MG3	10^{10}	1,000	Glass
MG6	2×10^{10}	2,000	Glass
MG12	5×10^{10}	4,000	Glass

MORGAN CRUCIBLE CO. LTD

The resistance element is sealed into an evacuated glass envelope of which the outside is coated with water repellent, silicone lacquer. These resistors are suitable in the working range from 0 to 300V. The change of resistance with applied voltage over the range 1 to 100V is less than $-0{\cdot}25$ per cent per volt for all types. These Megistors, as they are called, will work satisfactorily in the temperature range 0 to +60°C and probably down to −60°C. Lying idle, they are subject to less than 3 per cent change in 12 months. Polarization is extremely small. If touched by hand or otherwise contaminated it is recommended that they be cleaned in alcohol.

RESISTANCE (Ohms)	TEMPERATURE COEFFICIENT
10^7 to 10^9	Less than $+0{\cdot}2$ per cent per °C.
10^9 to 10^{11}	" " $+0{\cdot}3$ per cent per °C.
10^{11} to 10^{13}	" " $+0{\cdot}4$ per cent per °C.

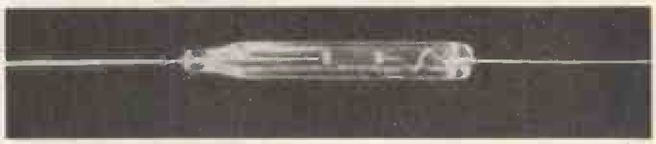

Fig. 5. Megistors. High value resistors in range 10^7 to 10^{13} ohms.
(Courtesy of Morgan Crucible Co. Ltd.)

VICTOREEN INSTRUMENT CO.

A series of so-called Hi-meg resistors, vacuum sealed in a glass envelope coated with silicone varnish. Aged electrically. They are satisfactory over the temperature range from −40° to +120°C, over the voltage range ·0001 to 1 000V, and over a range of relative humidity from 0 to 98 per cent.

	MIN.	NOM.	MAX.
Resistance range ..	$10^8\Omega$	—	$10^{12}\Omega$
Tolerance from specified resistance	−10%	—	10%
Tolerance from labelled resistance	−1%	—	1%
Temperature coefficient	−0·06%	−0·1%	−0·15%
Voltage coefficient ..	—	−0·02%/V	−0·03%/V

WELWYN ELECTRICAL LABS. LTD.

Welwyn Pyromatic resistors are available as follows:

TYPE	MAX. VOLTAGE (Volts)	TOLERANCE (per cent)	RESISTANCE (Ohms) MIN.	MAX.
PY3611	500	10	$2{\cdot}5\times10^6$	2×10^8
		20	$2{\cdot}5\times10^6$	2×10^8
PY3623	1,000	10	$2{\cdot}5\times10^7$	2×10^9
		20	$2{\cdot}5\times10^7$	$2{\cdot}5\times10^{10}$
PY3635	2,000	10	10^8	5×10^9
		20	10^8	10^{11}

Electrometer Valves

The design of a special electrometer valve with very low grid current and electrical leakage suitable for use in D.C. amplifier practice with a very high input resistance was undertaken by both Metcalfe and Thompson[4] and by

Nelson[20] in 1930. An input resistance of $10^{16}\Omega$ was achieved in both designs. Nelson used an oxide-coated filament in a triode design operated with the anode potential less than the minimum ionization potential of the residual gas. Metcalfe and Thompson used a thoriated tungsten filament in a tetrode design using an extra space-charge grid to repel positive ions emitted from the hot filament. This valve was the first well-known General Electric Co. (America) FP54. " Pliotron " with which a great deal of the work in subsequent years on electrometer valve circuits was carried out. Metcalfe and Thompson asserted that even with the highest vacuum obtainable and with potentials of less than 8V, ion currents of 10^{-13}A were obtainable in a triode design. These ions, mostly due to positive ion emission, were therefore repelled from the control grid by an intermediate space-charge grid at a positive potential of 4 to 6V. This grid also increased the mutual conductance of the tube. Low filament power and open grid structures were employed to avoid grid heating, the control grid was mounted on quartz beads shielded to prevent surface contamination, and had a top-cap connexion to reduce electrical leakage. The thoriated tungsten filament consisted of two legs in parallel so as to reduce the filament voltage required, and was run at reduced temperature. The grid current was then about 10^{-15}A minimum, this residual current being attributed mostly to positive ions produced by X-rays from the anode with $V_a = 6$V. The valve was mounted on a UX base.

The principal characteristics of this early FP54 were: $V_f = 2{\cdot}5$V; $I_f = 0{\cdot}1$A; $V_a = 6V$; $V_g = -4$V; $V_s = 4$V; $I_a = 40\mu$A; $g_m = 25\mu$A/V; $\mu = 1$; $R_a = 40$kΩ; $I_g \approx 10^{-15}$A; input capacitance = 2·5pF.

Since 1930 electrometer triodes and tetrodes have been developed on somewhat similar lines to the early work of Nelson and of Metcalfe and Thompson. Innovations have been miniature electrometer valves with low filament currents suitable for dry-cell working in portable apparatus and with very low inter-electrode capacitances, connexions being straight to the electrode leads; double electrometer valves have also appeared, in particular the Ferranti double tetrode, described by Darbyshire[21], for use in a balanced bridge circuit. Of recent years the use of indirectly heated cathodes for electrometer valves[22], particularly double valves like the Ferranti DBM6A, has come to the fore. Such cathodes have the advantages that they are (a) much more easily located symmetrically with respect to the electrode structure than are filaments; (b) the indirectly-heated cathode has an equi-potential surface which is more efficient when used with low potential electrodes; (c) they are less prone to microphony, (d) the larger space-charge developed around the cathode leads to increased stability. A further advantage of double electrometer valves, both of the directly heated and indirectly heated types is that, compared with the use of two separate valves as used

Fig. 6. Electrometer valve types ME1400 and sub-miniature models ME1401 and ME1402

(Courtesy of Mullard Ltd.)

in a balanced circuit, they do not give rise to fluctuations and drift due to the emission of one cathode varying or falling off with respect to the other. A disadvantage attending the use of indirectly heated cathodes—alleged not to be serious[22]—is that the grid current will be greater due to increased thermal and photo-electric emission.

A summary of the characteristics of some of the electrometer valves commercially available is given by courtesy of the manufacturers concerned.

FERRANTI LTD.

Indirectly-heated Single Tetrode Electrometer Valves

TYPE	V_h* (Volts)	I_h* (Amps.)	V_a (Volts)	V_s (Volts)	V_g (Volts)	I_a (μA)	I_s* (μA)	g_m (μA/V)	I_g (Amps.)
BM3A	2·0	0·345	6	4	−3	70 to 200	90 to 220	45 to 110	$<2\times10^{-14}$
BM5A	2·0	0·345	4·6	6	−3	90 to 250	150 to 450	60 to 160	2×10^{-14} to 30×10^{-14}
BM4A	4·0	0·24	6	4	−3	100 to 350	140 to 300	70 to 140	$<2\times10^{-14}$
BM6A	4·0	0·24	4·6	6	−3	150 to 550	250 to 500	120 to 220	2×10^{-14} to 30×10^{-14}
BM8A	8·0	0·125	6	4	−3	100 to 350	140 to 300	70 to 140	$<2\times10^{-14}$
BM12A	8·0	0·125	4·6	6	−3	150 to 550	250 to 500	120 to 220	2×10^{-14} to 30×10^{-14}

* V_h=heater volts, I_h=heat current, I_s=space-charge grid current.
N.B.—The range of values I_a, I_s and g_m encountered in a batch of valves is quoted.

The cathodes of types BM3A and BM5A should be maintained at a potential of 1 to 2 volts *negative* relative to heater negative. For types BM4A and BM6A this potential should be 6 to 10 volts *negative*. For types BM8A and BM12A it should be 8 to 10 volts *positive*. The guard rings provided are operated at the mean operating potential of the control grid.

Indirectly-heated Double Tetrode Electrometer Valves

TYPE	V_h (Volts)	I_h (Amps.)	V_a (each anode) (Volts)	V_s (Volts)	V_g (Volts)	I_a (μA)	I_s (μA)	g_m (μA/V)	I_g (Amps.)
DBM4A	4	0·24	7	5·5	−3	50 to 275	300 to 550	30 to 100	$<3\times10^{-14}$
DBM6A	4	0·24	8	6	−3	75 to 350	300 to 650	30 to 110	3×10^{-14} to 30×10^{-14}
DBM8A	8	0·125	7	5·5	−3	50 to 275	300 to 550	30 to 100	$<3\times10^{-14}$
DBM12A	8	0·125	8	6	−3	75 to 350	300 to 650	30 to 110	3×10^{-14} to 30×10^{-14}

The cathodes of types DBM4A and BDM6A should be maintained at a potential 6 to 10 volts *negative* relative to heater negative. For types DBM8A and DBM12A this potential should be 8 to 10 volts *positive*. The potential of the compensating grid is usually maintained at a potential equal to the mean operating potential of the signal control grid. The signal grid has the lower grid current. It is distinguished for connexion purposes from the compensating control grid since the latter is marked with a white band.

GENERAL ELECTRIC CO. LTD.

Directly-heated Single Triode Electrometer Valves

TYPE	V_f (Volts)	I_f (Amps.)	V_a (Volts)	V_g (Volts)	I_a (μA)	g_m (μA/V)	I_g (Amps.)	C (pF)
ET1 ..	1	0·1	4 6	−2 −2	100 325	50 100	5×10^{-16} at V_a=4V	2
ET3 ..	1·25	0·025	6	−2	110	70	8×10^{-15} at V_a=4V	1·7

C is the capacitance between the control grid and all other electrodes strapped together. In both valves the grid current $I_g = 0$ at $V_a = 6$, $V_g = -2$. The ET1 valve has the control grid terminal mounted at the end of a special, high resistance glass stem. The valve is capped. Internal and external guard rings and an electrostatic shield round the electrode system are brought to a pin in the valve base: this is usually connected to the negative end of the filament. The ET3 miniature triode has a glass envelope coated with a water repellant substance which should not be handled. Connexions are directly to the electrode leads with the control grid lead isolated at one end.

MULLARD LTD. (see Fig. 6).

An ME1400 pentode valve of comparatively orthodox design with indirectly heated cathode is available, which has a grid current of 10^{-11}A approximately, where a standard octal base is used and the control grid is to top cap. The screen grid is used in the conventional manner where both screen and anode potentials are 45 volts. With $V_g = -2$V, $I_a = 80\mu$A and $g_m = 240\mu$A/V. $\mu = 1200$ approximately. Connected as a triode, i.e. with the screen and anode joined, $I_a = 100\mu$A., $g_m = 300\mu$A/V, but μ is only 20. The input capacitance is 5·5pF. Two other sub-miniature electrometer valves have principal characteristics as follows:

TYPE	V_f (Volts)	V_a (Volts)	V_s (Volts)	g_m (μA/V)	μ	I_g (Amps.)
ME1401 .. .. sub-miniature triode	1·25	9	—	80	1·7	$<1{\cdot}25\times10^{-13}$
ME1402 .. .. sub-miniature tetrode	1·25	4·5	4·5	100	1·0	$<5\times10^{-15}$

RAYTHEON MFG CO.

TYPE	V_f (Volts)	I_f (Amps.)	V_a (Volts)	V_s (Volts)	V_g (Volts)	I_a (μA)	g_m (μA/V)	μ	I_g (Amps.)
CK5697 .. triode	0·625	0·02	12	—	−3	220	135	2·1	5×10^{-13} max.
CK5884 .. double tetrode	1·25	0·01	4·5	4·5	−3	20	15	0·75	10^{-14} nom.
CK5885 .. double tetrode	1·25	0·02	13·5 triode	13·5 connected	−3	185	160	2·4	10^{-13} nom.
CK5886 pentode	1·25	0·01	10·5	10·5	−3	200	160	2	2×10^{-13} max.
CK5889 .. pentode	1·25	0·075	12	4·5	−2	14	14	—	3×10^{-15} max.

All types are sub-miniature tubes with directly-heated filaments of very low current consumption.

THE VICTOREEN INSTRUMENT CO.

TYPE	V_f (Volts)	I_f (Amps.)	V_a (Volts)	V_s (Volts)	V_g (Volts)	I_a (μA)	g_m (μA/V)	I_g (Amps.)
5803 .. .. sub-miniature triode	1·25	0·01	7·5 10·0	— —	−1·7 −2	100 250	150 200	2×10^{-14} 3×10^{-13}
5800 .. .. sub-miniature tetrode ..	1·25	0·01	4·5	3·4	−3	12	15	10^{-15}

(To be continued)

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Some Recent Developments in Electronic Engineering

IN setting forth this survey of the more important developments in electronic engineering which have taken place during the past year, some omissions must inevitably occur, due primarily to lack of space.

Many of the more interesting developments are those sponsored by the Ministry of Supply and are concerned mainly with defence measures. For very good reasons, much of this work is governed by security restrictions and it is therefore only occasionally that a brief glimpse can be given of these activities which absorb so much of the national effort.

Telemetering in Guided Missiles

Considerable importance has been attached to a recent statement by Mr. Duncan Sandys, the Minister of Supply, in which he announced the present state of progress in the guided weapons programme. With the military implications of these new missiles we are not primarily concerned, but this survey must obviously include more than a brief reference to the important subject of the telemetering equipment.

As is well known, much information on the behaviour of missiles in flight can only be obtained by the firing of test rockets carrying equipment which transmits data by radio to ground stations.

Failure of such telemetering equipment during a flight means that the test missile has been wasted, with the consequent loss of a good deal of money and productive effort, and delay in the compilation of important information.

Apart from a high standard of efficiency and reliability, the main requirements for airborne telemeter transmitters are high power output, good frequency stability against power supply run-down and mechanical stress, low weight and small volume and, since the life of the equipment is short, economy in cost.

Two main types of equipment are needed. For the measurement of aerodynamic properties such as control surface positions, strains, pressures, and torques, the time sharing multiplex system is adequate. This provides a large number of separate channels for the transmission of data, each with a low frequency response.

For measuring the waveforms occurring in electronic equipment carried by a missile, however, channels with a high frequency are required, for which the P.P.M. (pulse position modulation) and time division systems are usually employed.

Two systems fulfilling these requirements have been evolved by the Ministry of Supply and are at present in quantity production. Each consists of one type of ground receiver with several varieties of airborne transmitter equipment.

These two sets provide adequate and largely complementary coverage for frequency responses up to 200-230c/s. The first system provides 23 channels for the transmission of data. a further channel being reserved for synchronization. The second set gives up to 20 higher frequency channels any one or any number of which can be subcommutated to give lower frequency channels.

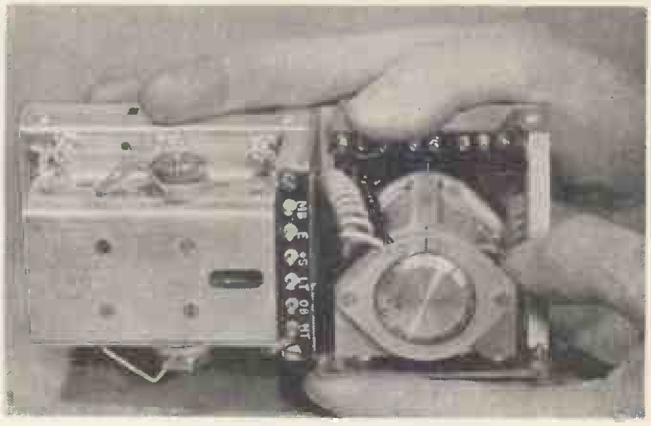
If necessary, and if sufficient space is available, both sets can be installed in a missile, or two of the first systems and one of the second can be fitted to provide 46 lower frequency and 20 higher frequency channels.

TIME MULTIPLEX SYSTEM

The basis of the airborne transmitting section of the system is a rotating sampling switch, the contacts of which are connected to pick-ups which convert the physical data required into electrical quantities.

The switch samples the pick-ups in turn and allows each to send its particular message to the ground for a small percentage only of the total time. This, coupled with the mechanized switch speed, accounts for the low frequency response of the system.

The output from the pick-ups passes, via the switch, to a modulator. This generates a sub-carrier whose frequency is directly related to the voltage, or inductance fed into it by the switch wiper. The resultant time multiplexed frequency modulation is used to amplitude-modulate an R.F. oscillator.



A typical time multiplex transmitter.

(Crown copyright reserved)

Transmitting Equipment

Since the set must be fitted to projectiles of varying sizes, from 3in diameter upwards, flexibility is an important factor in its design. The components of the transmitting set have therefore been designed as standard units which can be arranged on various sizes of chassis.

The dimensions of the smallest chassis yet fired were $5\frac{1}{4}$in × $3\frac{1}{4}$in × $2\frac{1}{4}$in (telemetry chassis) together with a power-supply chassis measuring $4\frac{1}{2}$in × 3in × $2\frac{1}{4}$in.

Pick-ups

Both the voltage and inductance types of pick-ups used in the system are produced in a variety of types and ranges.

Inductance pick-ups available include linear and angular accelerometers, pressure pick-ups and linear and angular movement pick-ups.

Voltage pick-ups usually take the form of a resistance potentiometer device. A number of such designs in the form of pressure pick-ups, general movement pick-ups from control surfaces or temperature pick-ups have been designed.

The Switch

On the speed of the switch unit depends the channel frequency response of the system.

Three models of the switch, with speeds of 120, 80 and 40c/s respectively have been designed.

The switch itself consists of a miniature motor, working from 6·3V D.C., driving a four finger nickel silver brush assembly in a 2-way commutator, moulded in "Araldite".

Modulator Circuits

Three types of modulator are in use with the system, one accepting inductance inputs only, one voltage inputs only and one responding to both types of input.

In the case of the inductance responsive modulator, a normal triode oscillator is used, the pick-up providing the variable inductance of the tuning circuit. The output of the oscillator is then amplified to give the power needed to modulate the R.F. oscillator.

In the second modulator, responding to voltage inputs only, the frequency of oscillation of a tuned grid triode oscillator is controlled by a conventional reactance valve which forms part of its tuned circuit, the output frequency being again amplified to modulate the R.F. oscillator.

Dual-purpose Modulator

To provide a combined circuit responding to both voltage and inductance inputs, these two circuits are combined in a four-valve version. The first valve acts as an oscillator for inductance inputs, the second as an amplifier which locks the frequency of the third valve to that of the first. The fourth valve is a straight amplifier.

In the case of voltage inputs, the first valve is rendered passive by the damping of its tuned circuit and the voltage applied to the input to the second valve, which, acting as a reactance valve, determines the frequency of oscillation of the third valve. The fourth valve is, again, a straight amplifier.

In all cases the range of frequencies generated is 130-160kc/s which in the case of voltage inputs is covered by a change of 6V.

Oscillators

To provide adequate power for all ranges, while keeping the direct current supplies needed as small as possible, three oscillators were designed for use in the system, with outputs, measured on a C.W. basis, of 0·2, 1·0 and 8 watts respectively. The earliest version was the 0·2W oscillator, built in the form of a push-pull tuned-anode, tuned-grid oscillator, using a CV858 (6J6). This is now obsolescent and is being replaced by the 1·0W version which has a similar circuit arrangement but is built around a specially-developed sub-miniature valve.

The larger 8W unit, used for longer ranges, consists of tuned-anode, tuned-grid oscillator built around the CV397a disk sealed triode.

As the modulator, unaided, cannot control this oscillator, a driver-unit, supplying the necessary power, is included in this model.

Ground Station

The ground equipment of the system provides two records, a histogram (i.e. high-speed record) and the main record.

For the latter the signal received from the airborne transmitter, after demodulation, is changed to a series of direct current levels and the step function thus produced is applied simultaneously to the Y plates of 15 cathode-ray tubes. All the 15 tubes are photographed side by side on a moving film twenty-four inches wide.

Timing pips at each tenth of a second are recorded on each tube as also is the firing pulse. On the completion of each firing, short calibration lines from a crystal controlled frequency generator are added to the record.

The histogram recorder is designed as a self-contained unit for producing records from single channel senders or histogram traces from multichannel senders. It contains a receiver and discriminator similar to those used in the main equipment and is fed from a separate helical aerial. The display consists of three cathode-ray tubes, one large and two small, which are photographed simultaneously by a continuous 35mm camera.

The large tube displays the single channel record or the series of 24 direct current levels corresponding to the channels in the sender. One of the small tubes is used to display a time scale, the lateral position of which is calibrated in terms of field strength, and the other small tube indicates any radio frequency deviation throughout flight.

THE P.P.M. SYSTEM

In this system, channel information is given by the time interval between the trailing edge of a reference pulse (about 50μsec long) and the incidence of a 1μsec channel pulse.

The carrier frequency is pulse modulated by the reference pulse and the channel pulse, the recurrence frequency of the complete pulse train being between 2 000 and 5 000 per second, according to the number of channels in the sender.

Up to 20 channel pulses, each of which can vary over a range of 400μsec, can be accommodated between successive reference pulses. Normally, however, groups of pulses are confined to consecutive intervals of 80μsec, the range of time displayed on each of the five cathode-ray tubes in the ground equipment.

Each channel is arranged to produce, either directly or via a transducer, a variation of voltage or a variation of capacitance which is used to control the time interval between the trailing edge of a reference square wave and the generation of the respective channel pulse.

A P.P.M. transmitting set for use in a test rocket.
(Crown copyright reserved)

Airborne Transmitter

The airborne transmitting equipment of the set, the main parts of which are a modulator and an oscillator, is made in the form of a cylinder with a diameter of 4½ inches and an overall length of 6¼ inches for a 12-channel set.

The waveform generated by the modulator consists of a wide pulse of approximately 40μsec width, the trailing edge of which is the time reference for the system, followed by a number of narrow pulses of 1μsec width.

The oscillator normally used is a 2W (rated on C.W.) model, but for longer ranges an 8W oscillator similar to that used in the time multiplex system has been produced. With varying direct current supplies, this meets all requirements from 2 to 8W.

A greater number of information channels can be obtained by introducing a time division switch on any one or more channels with the corresponding sacrifice of frequency response. Standard switch wafers at present in use have eleven contacts which can be arranged in a number of different ways, four capacitance and four voltage channels or eleven voltage channels with the same voltage excursion being typical assemblies.

As these channels may be arranged to overlap and up to four may be interlaced, some method of identification is needed. This is provided by an identification switch which runs at a low speed connecting each input to a fixed voltage, usually earth, in sequence so that once one channel has been identified all the rest follow in order.

Ground Equipment

The ground equipment of the P.P.M. system produces a maximum of five displays depending on the number of channels in use.

The time scales are used as the Y time bases for five separate cathode-ray tubes, the channel pulses being used as brightening pulses for these traces. Each tube is photographed by a continuous 35mm camera, the speed of which can be varied from 2 to 120 inches per second.

To achieve this the signal is received on the ground and, after demodulation by a sensitive receiver, the resultant train of pulses is fed into a voltage circuit box unit. Here the pulses are limited and the reference square wave is separated from the channel pulses.

A trigger pulse is then generated from the trailing edge of this square wave, and goes to each of the five time-base generators, where two phantastron delay circuits are used to develop the appropriate Y axis time-base for each display tube. At the same time, the channel pulses, after further amplification and limiting in the voltage circuit box, pass through a pulse forming channel, and are used as brightening strobes for the cathode-ray tubes. Initially, the time-base is adjusted to scan the whole face of the cathode-ray tube, and to check that no variation of the 80μsec needed to do this occurs during flight, brightening strobes from the time-base circuit are used to mark the two ends of each scan.

Before each firing the time-bases are set up accurately and checked by means of a built-in crystal calibrator, which is fed into the voltage circuit box, and the scan on each tube adjusted accordingly.

A timing trace, obtained from the range master timing system, is displayed on a small cathode-ray tube by the side of each of the channel recording tubes so that this is photographed simultaneously on the 35mm film. On firing, a pulse from the firing switch is arranged to displace this trace and indicate the instant of firing.

Radar and Radio Apparatus

In the spring of last year, an operational display of military electronic equipment was held at the Royal Aircraft Establishment, Farnborough, before representatives of N.A.T.O. This display, which was organized by the Radio Communication and Electronic Engineering Association at the instigation of the Ministry of Supply, was claimed to be the largest and most comprehensive of its kind and was to demonstrate the wide variety of equipment required to meet the operational needs of the armed forces, including fixed and mobile radio communications equipment, radio navigational aids and radar. Service aircraft fitted with various types of electronic equipment carried out demonstration flights, landing visitors at Service aerodromes for inspection of the ground counterparts of these airborne equipments. Army equipment included tanks with signals and gun control equipment, a modern tank turret with all its electronic equipment, vehicles of all kinds and electronic equipment required for anti-aircraft purposes.

Much of this radar communication equipment finds application in civilian airline operation and was later on view at the S.B.A.C. air display.

Among the Farnborough exhibits were the following.

RADIO AND NAVIGATIONAL AIDS

Among the radio communication equipment exhibited was an S.H.F. radio link manufactured by Standard Telephones and Cables Ltd. This link comprises a light weight equipment forming a complete transmitter and receiver operating in the frequency range 3 600-4 750Mc/s and is intended for providing main or standby circuits in multichannel carrier telephone networks. The equipment can also be used for television purposes.

The transmitter which is shown opposite has a velocity-modulated coaxial line valve delivering a power of 250mW into a 4ft diameter parabaloid through a waveguide and electromagnetic horn. By limiting the power delivered, the directly modulated coaxial line oscillator can feed the antenna direct, simplifying the apparatus and making it more compact.

The receiver unit, which is similar in appearance, employs an antenna which is identical with that of the transmitter. The received signal is fed through the waveguide to the receiver head unit in the rear of the parabaloid. Here the modulated energy is fed direct to a crystal to which is added the output from a coaxial line valve oscillator.

The resulting frequency-modulated intermediate frequency of 30Mc/s is then amplified in a four stage amplifier using grounded-grid triodes.

The receiver output from one link unit may be connected to the transmitter unit of another link so as to form a repeater.

Equipment for an S.H.F. radio link.

The whole equipment, which is weatherproof, is designed for ease in mobility and no individual piece of equipment forming the link is more than 60lb in weight.

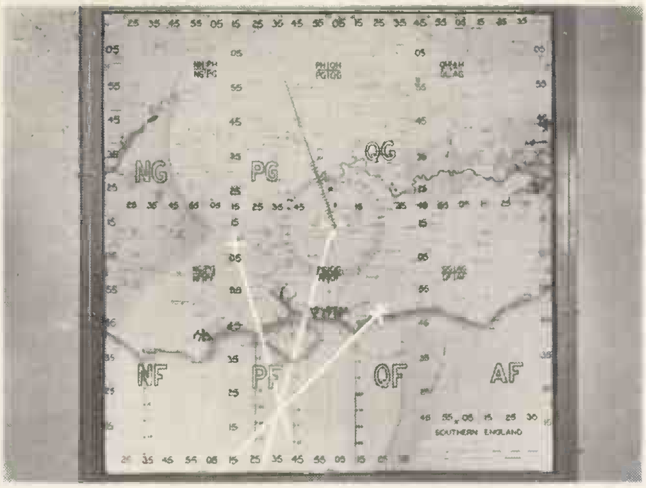
An interesting navigational aid produced by the same firm is an automatic triangulation equipment which will give automatically and almost instantaneously the position of an aircraft on a screen about 5ft square.

The screen shown overleaf has a specially prepared map printed on it covering the area of the network and illustrated with useful topographical features such as towns and airfields. Grid reference lines and other similar data can also be included.

The equipment is designed for use primarily with the type PV1-B V.H.F. Automatic Direction Finder. A network of up to ten such automatic D.F. equipments is required, each feeding its directional information to the Triangulation Centre where the hearings are displayed automatically on the illuminated screen as a line of light. The root of each line of light or bearing trace is centred on the map position of each D.F. station, and a "fix" is indicated by the intersection of two or more traces.

Between each D.F. station and the Triangulation Centre only one conventional two-way speech channel, either by land line or radio link, is required to carry the D.F. bearings, speech received from aircraft, and the essential control facilities.

Bearings are derived from the normal V.H.F. air-to-ground radio equipment of the aircraft and no additional



The viewing screen used with the automatic triangulation equipment.

equipment is required. Identification of aircraft is achieved simultaneously with a bearing "fix" by aural monitoring of the aircraft's signals received at the centre from which the visual presentation is derived.

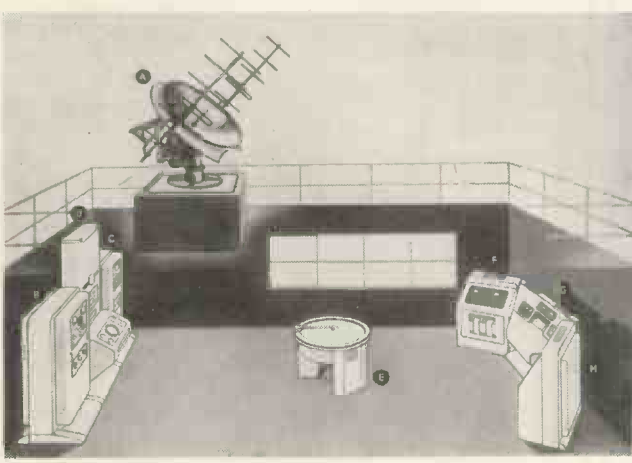
The equipment at the D.F. stations generates four co-ordinated D.C. potentials to produce a radial trace or bearing information on its own cathode-ray tube. These voltages are converted into a form suitable for transmission to the Triangulation Centre where they are restored to operate a cathode-ray tube in a projection unit mounted behind the screen. This cathode-ray tube in the projector unit is mounted so as to correspond as near as possible with the geographical position on the map screen of its associated D.F. station and in operation a small cross is projected on to the screen to correspond with the map position of the D.F. station. The bearing trace from the D.F. station is centred on its associated cross on the screen.

The network of D.F. stations can be subdivided into two groups so that two aircraft calling on different frequencies can be dealt with simultaneously.

Meteorological Equipment

Considerable attention has been paid in the post-war years to meteorological conditions in the upper atmosphere due principally to the demands made by modern aviation. With aircraft flying at higher altitudes and at higher speeds, accurate forecasting of weather conditions over the

Layout of radar-sonde station
A—*aerial unit,* B—*transmitter,* C—*receiver,* D—*display unit,* E—*wind computor,* F—*wind recorder,* G—*telemetering unit,* H—*sonde recorder.*



principal air routes has become an essential factor.

To this end, therefore, equipment has recently been installed at a new meteorological station at Crawley, Sussex, and it is claimed to be the world's first fully automatic station for accurate observations of wind speed, temperature pressure and humidity at a range of at least 100 nautical miles.

The new meteorological equipment has been developed and built for the British Meteorological Office by the Research Laboratories of Mullard Ltd. in collaboration with other laboratories in the organization. It is based on initial research work undertaken some years ago by the Telecommunications Research Establishment of the Ministry of Supply, and the project has been carried through with their collaboration.

The system consists basically of an automatic ground interrogator and computing station, working in conjunction with airborne transponders. Two types of transponder are available: one is confined to wind measurements only; the other, in addition, provides sonde information. During the operation of the system, one of these units, depending on the extent of the observations to be made, is suspended from a hydrogen-filled balloon. The maximum slant range of the system is about 100 nautical miles (185km). In order to reach this extreme range under normal conditions, the balloon would have to ascend to a very great height. A standard balloon, as at present used, would burst before this height is reached. Thus, the height to which soundings and wind measurements can be made is, at present, largely limited by the height performance of the balloon. It is important to note, however, that the equipment is capable of operating at a ceiling height of at least 100 000ft (30·5km).

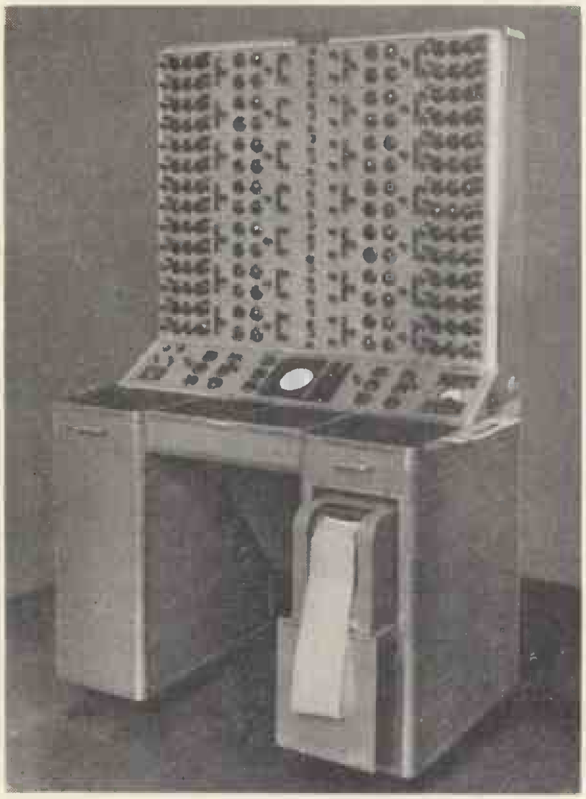
The balloons are released from the ground station at certain internationally agreed times. As the balloon ascends into the upper air, its airborne transponder is interrogated from the ground station by radar pulses of 2μsec duration. These are transmitted from a 50kW (peak) transmitter, operating on a primary frequency of 152·5Mc/s. The interrogating pulses are received by the balloon-borne unit. Here they are used to make the airborne transmitter send back pulses to the ground station on a secondary frequency of 2 850Mc/s (10cm band). These pulses, in turn, are received at the ground station by an automatic following aerial.

The slant range of the airborne unit is readily determined by measuring the transit time of the pulse to and from the transponder. From this, and from the values of the azimuth and elevation angles of the aerial unit, the drift of the airborne unit in space is computed. Wind speed and direction are computed automatically and continuously from the rates of change of the measured parameters. The height of the balloon is also computed and a correction for the curvature of the earth is applied automatically.

Telemetering of temperature, pressure and humidity information, registered by the meteorological elements in the sonde unit, is effected by causing the sonde to transmit twin pulses each time it is interrogated from the ground station. The degree of separation between the two pulses serves as a measure of the meteorological parameter in circuit at any particular time. The three elements for measuring pressure, temperature, and humidity are switched into the telemetering circuit in sequence by a motor-driven switch. In this way, a complete cycle of sonde observations is telemetered to the ground station every 15 seconds during the flight of the balloon. A fourth element is included in the cycle providing a constant reference reading. Any change in the constants of the telemetering circuits affecting the accuracy of the meteorological readings will be indicated by a change in the reference reading. When the balloon bursts on reaching its maximum altitude, a parachute opens and the transponder unit slowly descends. During the descent, the airborne unit continues to be interrogated by the ground station provided, of course, that it remains within the 100 nautical mile range of the equipment.

Computors

The past year has seen considerable development in the field of computors and the main effort has been largely concentrated on increased reliability and simplification of design. Progress has to some extent been halted by the introduction of the transistor, particularly in regard to its application to computor techniques and it may be some time yet before the all-transistor computor makes its appearance.



A general purpose analogue computor.

An interesting general purpose analogue computor has recently been designed by the Precision Engineering Division of Short Brothers and Harland Ltd.

In this instrument three basic types of computing element are provided, as follows:—

1. *Scaling Units*—which are essentially three decade variable resistors.
2. *Function Units*—comprising resistance-capacitance networks which may be selected by switches.
3. *Amplifiers.*—These are high gain 5 valve units which may be used in combination with the scaling and function units for addition, integration and all other linear operations in computation and simulation.

The appearance of the prototype instrument is shown above. It is built in the form of a double-pedestal desk surmounted by a shallow rack, and is mounted on castors for ease of movement. The overall dimensions are only 5ft 9in high, 3ft 6in wide and 2ft 2in deep. All the sub-assemblies (such as the computing units) are designed to slide into the framework and be secured with quick-release fasteners, electrical contact being made automatically with suitable connectors.

The computor derives its power (approximately 1kW) from the normal mains supply via a number of power units which are fitted into the left-hand pedestal of the desk. The pen recorder, with its associated power unit and amplifier, is accommodated in the right-hand pedestal.

Set-up Panel

The computing units are interconnected for each problem by inserting small plugs into a Set-up Panel which is recessed into the top of the desk and covered by hinged flaps when not in use. This feature avoids the usual tangle of telephone plugs and cords which disfigures most computors. As a further simplification, blank charts are provided which can be marked with the positions of the plugs required for any problem. The chart can then be used as a guide when setting up the computor, and subsequently filed as a record.

Display

With a computor of this type two methods of working are possible. One is to choose a slow time scale, of the order of seconds, initiate the forcing function (usually a voltage step) manually, and obtain the solution as a graph from a recorder of some kind. The other is to operate repetitively with a fast time scale, reducing the time constants of the system if necessary, and observe the solution with a cathode-ray oscilloscope.

The second method is generally more convenient, since it gives a steady picture on a screen, and enables the operator to observe the result of changing the constants of a system immediately. However, it may not be satisfactory for simulation, because if actual components of a system are being used in a set-up they will generally have to work on a true time scale.

Facilities have therefore been provided for both kinds of display by building a cathode-ray oscilloscope into the computor for normal use, and also arranging for a pen recorder to be fitted optionally. This enables an operator to set up a problem, determine the optimum constants by trial and error visually on a fast time scale, and then turn to a slow time scale to obtain a permanent record from the recorder.

Electro-Medical Applications

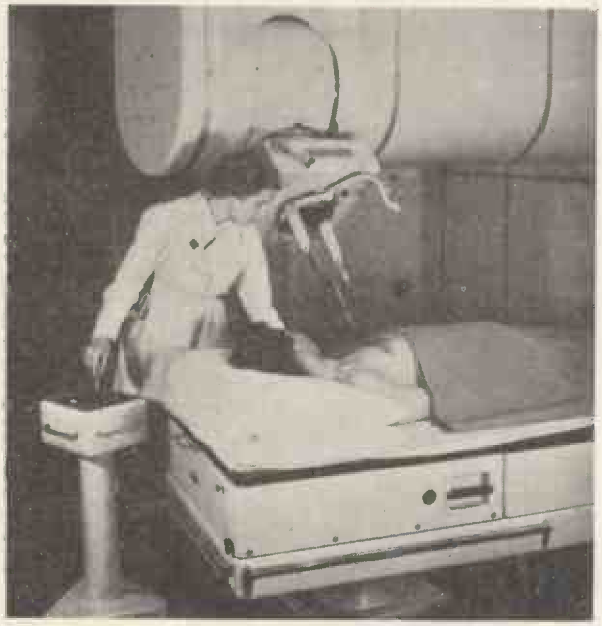
Two applications of electronics to medical research which have taken place during the past year merit inclusion in this survey. They are the 8MeV linear accelerator installed at the Hammersmith Hospital in London and the artificial heart-lung apparatus.

The linear accelerator was built by the Metropolitan-Vickers Company and was developed in conjunction with the Atomic Energy Research Establishment and the Radiotherapeutic Research Unit of the Medical Research Council. The machine, which is the first of its kind to be built specifically for X-ray therapy, has completed its performance tests and is now in regular use.

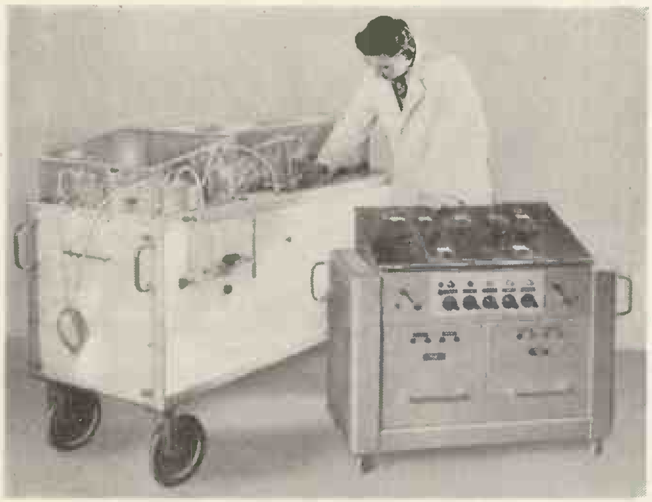
Electrons are injected into the accelerator tube from a gun operating at 45-50kV and power flux of some 3·6MW peak power is required. On emerging from the accelerator tube the high energy electron beam is made to strike a gold target and X-rays are produced which can be directed accurately into the patient at any desired angle. The patient is positioned in the X-ray beam by moving the floor of the treatment room up or down and by the use of a specially designed moving couch. The movements of the X-ray beam, the floor and the couch are controlled automatically by electronic methods.

For the first time there are incorporated in this one machine all the facilities needed to permit the full exploitation of the technical advantages to be gained by using very high energy radiations. It makes possible the delivery of an adequate dose to a deep-seated tumour of highly penetrating 8 million volt X-rays, collimated in a well-defined, high intensity beam, which can be directed into the patient at any angle. The high intensity shortens the time of each treatment to about two minutes and the extremely penetrating nature of the X-rays allows the delivery of a high dose to the tumour without any danger of damage to the skin.

Undergoing construction in the same hospital is a 45in cyclotron. This machine will be used for the production



Automatic positioning of the couch prior to use of the 8MeV X-ray unit.



The extra-corporeal heart and lung equipment.

of very short life radio-active isotopes. It weighs some 180 tons and is housed in a large room with concrete walls and ceilings over 6ft thick and is sealed in by concrete sliding doors weighing over 200 tons. The whole installation is operated from a remote control room.

Extra-Corporeal Heart and Lung Equipment

The extra-corporeal heart and lung equipment, produced by New Electronic Products Ltd. for the University of London Post Graduate Medical School, is designed to duplicate the functions of the human heart in pumping blood through the system and the lungs in oxygenating the blood. These aims are met by using a large rotating plastic cylinder in which the blood can be oxygenated by spreading it over the surfaces of internal blades, and two pumps, which draw blood from the body to the oxygenator cylinder and then return it to the arteries.

It is essential to provide close control of both the suction (venous) and pumping (arterial) pressures as well as the quantity of blood in the oxygenator.

In practice the pumps and oxygenator, with its associated cylinders of oxygen and carbon dioxide, are built into one mobile cabinet and the control gear into another, smaller one.

The pumps are driven by variable speed electric motors which can be run from a standstill to full-speed by transductor controlled power supply units. The blood pressures in and out, as well as the blood level in the oxygenator, are measured by special gauges incorporating differential transformers, the output voltages of which are used in combination to supply the control current to the transductors.

Provision is made for manual or fully automatic control. Speeds, pressures and level are displayed on meters on the control panel, danger limits being indicated by warning lights.

The oxygenator cylinder is thermostatically controlled normally at blood heat. Extra blood can be drawn from a transfusion bottle into the oxygenator by pressing a button on the control trolley when necessary to make up patient's blood-loss.

The equipment has wide applications in resuscitation and in special treatments where large-scale blood transfusion is involved, as well as opening the way to new surgical technique especially in heart and lung operations.

The machine illustrated is the first prototype equipment and has been used in animal trials over the past year. Careful analysis of its performance has proved its fitness for human service and further development is aimed towards smaller simpler and utterly reliable equipment for routine operating theatre use.

Industrial Applications

There have been many applications of electronics to industrial uses during the year under review, but space permits mention of only one of these.

This is the installation at Guard Bridge Paper Mills, St. Andrews, of a new paper making machine in which the electric drives are all electronically controlled. This machine is of the Fourdrinier type and is designed for the production of ledger and fine writing papers. It has a wire width of 122in and an operating speed range of 40 to 400ft/min.

The whole electrical and electronic control apparatus was designed by the British Thomson-Houston Company Ltd.

Power to the coach, presses, drivers and calenders is applied at no less than seventeen points or sections. Each section is driven by a D.C. motor with constant excitation, the armature being supplied from its own Ward-Leonard generator which is excited by a grid controlled bi-phase thyratron rectifier.

A small tachometer generator is coupled to the section motor and generates a voltage proportional to the section motor speed. This generator is of special design with a permanent magnet field, monel metal commutator and silver graphite brushes. The voltage from this generator is compared with the voltage of a reference busbar and the difference, if any, is applied to a valve amplifier which controls the firing of the generator field thyratrons.

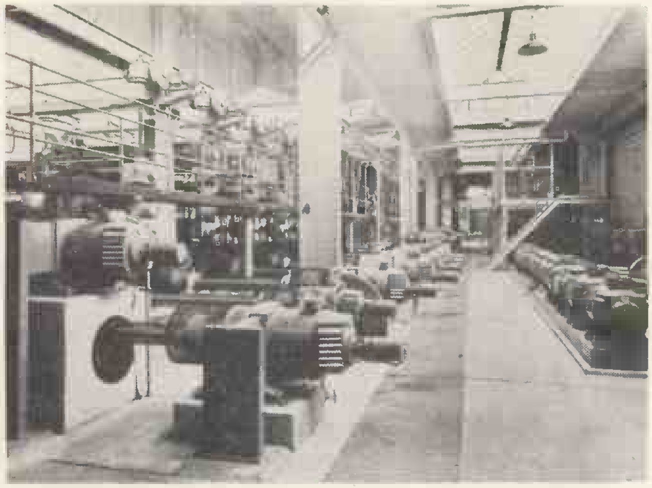
The voltage of the reference busbar which is common to all sections forms the link which rigidly ties together the speeds of the various section motors and the speed of the machine as a whole is controlled by a master rheostat which supplies the reference busbar voltage.

The speed of an individual section motor relative to the remaining section motors can be adjusted by a separate rheostat.

An exciter and alternator each controlled by its own electronic voltage regulator provide the stabilized direct and alternating current supplies for the operation of the speed control regulators.

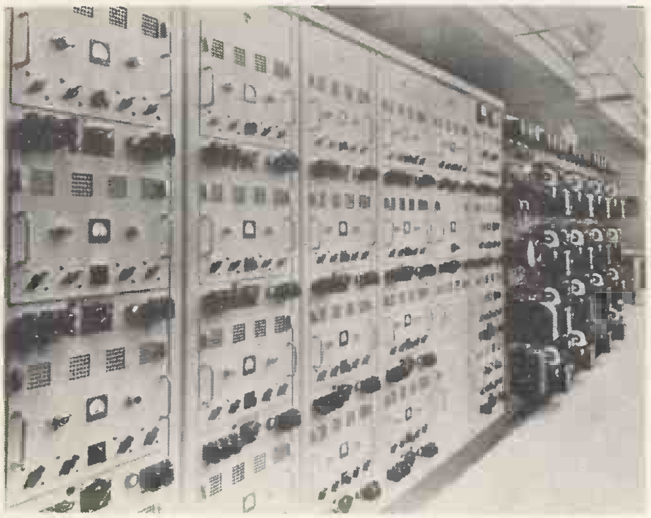
The speed regulating action is not accompanied by any mechanical movement and consequent inertia effects so that its response is very rapid, and changes in speed due to load variations are reduced to a negligible amount by ensuring a high gain of the amplifiers.

The sensitivity of the control equipment is such that a change of 100 per cent in load produces a speed change



View of drive from dry end of paper-making machine.

The tachometer-generator mounted on the end of each drive unit can be clearly seen.



Electronic and contactor control gear for speed regulation.

of less than 0·03 per cent. In practice load changes are considerably less than this amount so that variations in speed are in general limited to those produced by changes of resistance values of the components in the speed measuring circuit and of the output voltage of the tachometer generator.

By careful design these effects have been reduced to a minimum and the greatest possible speed variation from all causes is less than 0·1 per cent.

Considerable attention has been paid to the reliability of the equipment and all components have been liberally rated. By running the valves well below their normal ratings, a valve life of 30 000 hours can be expected.

AN-5 CASTING RESIN *

A new casting resin for embedding or "potting" electronic components and assemblies has recently been developed by the U.S. National Bureau of Standards' Ordnance Development Division. Easily manufactured at low cost from available raw materials, the new casting resin has good storage properties even without refrigeration. Its physical properties and electrical characteristics at both high and low frequencies are in general as good as those of the more expensive and hard-to-store casting resins. Designated as the Bureau's AN-5 Casting Resin, its unique ingredient is a small quantity of acrylonitrile monomer.

This new casting resin was developed at the Bureau to meet the need for a material that could be economically manufactured from available ingredients. It is essentially a modified styrene. To the styrene are added distilled acrylonitrile and divinylbenzene (from which nearly all the inhibitor has been removed by passing it through a column of activated alumina), hydrogenated terphenyl, polystyrene, and cobalt drier (see Table 1). Refrigeration is not necessary unless an unusually long storage life is required—storage in a cool, dark place (about 60°F) gives a shelf-life of about 8 to 18 months. Just before use, small amounts of benzoyl peroxide and "Pro-A", a polymerization promotor, are added to catalyze the curing process.

The acrylonitrile monomer is very reactive and helps speed up the polymerization of the casting resin. The divinylbenzene, a cross-linking agent, imparts certain thermosetting properties

TABLE I

Composition of AN-5 Casting Resin

COMPOUND	PER CENT BY WEIGHT
Styrene	50
Acrylonitrile	5
Divinylbenzene (a 40 to 50 per cent solution) ..	2
Hydrogenated terphenyl	13
Polystyrene	30
Cobalt naphthenate drier (6 per cent metal)	2 drops/100 grams casting resin
Catalysts, added to above just prior to curing :—	
Benzoyl peroxide ..	0·4
Pro-A.*	14 drops/100 grams casting resin

* 10 *per cent solution of N-hydroxyethyl-*1, 2, 3, 4-*tetrahydroquinoline in hydrogenated terphenyl.*

to the cured resin, reducing the tendency of flow at high temperatures. The hydrogenerated terphenyl is added to improve the impact resistance, and the polystyrene to improve the body, pouring consistency, and final mechanical properties; both of these inert ingredients also reduce shrinkage and absorb heat of polymerization. The small amount of cobalt drier counteracts the inhibiting effect of air on the surface of the casting resin during curing and results in a hard non-tacky surface. The composition can be varied considerably when it is desired to maximize specific characteristics.

Although satisfactory cures can be obtained at room temperature within several days, the curing rate is hastened by the application of a moderate amount of heat. At 50°C the resin gels in about two hours and is almost completely cured in about 20 hours. The choice of curing technique depends on the type of component being embedded and on the amount of resin required.

The physical and electrical properties of AN-5 Casting Resin are approximately the same as those of other low-loss casting resins, and in general compare favourably with some grades of polystyrene. Shrinkage during cure, though about the same (9·8 per cent) as that of many polyester and modified styrene casting resins, is still excessive for many applications. Vulnerable components like glass vacuum tubes can be protected from the crushing action of the shrinkage resin by coating them with foamed silicone rubber.

The dielectric constant of AN-5 varies from 2·6 at 10^4c/s to 2·55 at $8{\cdot}6 \times 10^9$c/s, while the dissipation factor (dry) is of the order of 0·001. The maximum thermal coefficient of expansion (linear) is 11×10^{-5}/°C.

* *Communication from M. Lorant.*

Step to Frequency Response Transforms for Linear Servo Systems

(Part 1)

By L. C. Ludbrook*, B.Eng., A.M.I.E.E.

The step response approach offers the most convenient way of synthesizing a linear servo system to meet a given specification of dynamic accuracy while subject to given aperiodic input disturbances, but since undesired periodic disturbances are also encountered in practical systems the frequency response must also be found as a check on the risk of resonant amplification and on the efficiency of smoothing.

A critical discussion of five existing methods for finding frequency response from a given graph of linear mode step response is given, followed by the theory and practical computing routines of a new method based on straight line segment approximation of the given step response graph. This combines the best features of three earlier methods and clarifies the physical picture of the relation between step function and frequency responses.

A number of analytic examples are given showing that the new method deals correctly with Bode non-minimum phase networks and that approximation of practical servo responses by about four straight lines gives acceptable accuracy up to frequencies such that about four cycles would occur in the step response settling time.

The theoretical and empirical basis is formulated for three simple approximate rules relating the cut-off frequency and the frequency and amplitude of maximum response to the shape and time scale of the step response.

In conclusion the results predicted from the new transformation method and approximate rules are compared with the actual test results previously recorded on a variety of R.P.C. servos. There is fair agreement on resonant and cut-off frequencies, but not on resonant amplification. While experimental errors are not negligible it is believed that the major discrepancies are due to the non-linearity of practical systems.

PRACTICAL continuous control servo systems can only be considered linear to a first approximation over a narrow range of input signal amplitudes bounded by the effects of friction, backlash and random noise at low level and by saturation of one or more control elements at high level. Nevertheless the art of system design has been considerably advanced by application of linear theory and it is still expedient to use these powerful informative methods for general guidance. The theoretical work in this article is valid only for the idealized linear system whose performance can be represented by a single graph of per unit output as a function of time independent of the magnitude of the Heaviside step function signal applied to the input. While primarily undertaken to help the step response approach to servo synthesis, the work may be of wider use because it covers also systems having band-pass or high-pass characteristics and is not restricted to Bode minimum phase structures[1].

The given graph of step response may have been found by measurement on an actual or simulated system (using a range of signal amplitudes to confirm linearity) or it may have been synthesized at the design stage to give specified dynamic and steady state accuracy with due regard to the fundamental "enclosed area" requirements[2,3] and to the practical restrictions imposed by the time delays in the chosen sequence of amplifiers and convertors, by the characteristics of the controlled load and available power supply, and by the linear range needed. In this process of synthesis it is often found that, from the point of view of dynamic accuracy during aperiodic input disturbances, the response to input position step must have at least 50 per cent overshoot followed by an appreciable undershoot. Before such a system can be safely applied, its frequency response must be considered together with the frequency spectrum of undesired periodic components of input signal. Even in industrial practice most servo designers could quote examples of maintained oscillation initially attributed to "servo hunting" but ultimately traced to resonant amplification of a periodic component of input signal contributed, say, by unbalanced reciprocating or eccentric rotating members outside the servo loop.

The objective of this work is to find the simplest method for deriving frequency response from a given graph of linear mode step response using only apparatus readily available to the average designer.

Existing Alternative Methods

Two of the existing transformation methods necessitate as a first stage differentiating the given graph of step response to find the system response to hammer blow disturbance. The Furth and Pringle Photoelectric Fourier Transformer[4] will accept a cut-out silhouette of the hammer blow response and give a direct display of real and imaginary parts of the frequency response on a cathode-ray tube; this machine is not generally available and appears barely accurate enough in its present stage of development. Following his ingenious solution of the inverse transformation problem Floyd[5] outlines a method of computing the real part of the frequency response from a straight line segment approximation to the hammer blow response and suggests that the imaginary part be then derived from the real part of the response on the assumption that the network is of Bode minimum phase type. The need for this assumption indicates that the information contained in the hammer blow response is not being fully

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

utilized; for example a complete treatment would show quantitatively the excess phase lag over the Bode minimum in a servo having finite dead time at the toe of its step response. A further disadvantage common to both these methods is that—at least to engineers—there is a lack of reality in the concept of an infinite amplitude disturbance adding finite energy to a system in zero time; even lightning surge phenomena are usually analysed by superposition of finite amplitude disturbances at finite time intervals.

Three other existing methods work directly from the given graph of step response and therefore have more physical appeal. Henderson's method[6] involves setting the constants of a mechanical harmonic analyser to a selected frequency, moving a tracer point around the given step response graph, reading direct from the machine the real and imaginary parts of the response at that frequency, and repeating the sequence for other frequencies until the complete response locus has been defined in sufficient detail. Instrumental and curve tracing errors should be much less than those introduced by graphical approximations and the sole objection to this method is that suitable harmonic analysers are not generally available. Bedford and Fredendall[7] working in the microsecond-Mc/s region for television amplifiers and concerned with distortion rather than delayed receipt of the picture, approximate the given graph of step response by an equi-interval staircase and give convenient direct reading charts for the corresponding components of the total vector generating the frequency response locus. With the obvious minor modification to allow for the effects of finite dead time this method has been successfully applied to servo systems but has the practical disadvantage that staircase intervals fine enough to resolve the initial steep rise give unnecessary accuracy and computing labour for the subsequent slow runback encountered in typical servo responses. Samulon[8] also concerned primarily with television apparatus, retains the principle of sampling at equal time intervals but uses a function of type $\sin xt/xt$ instead of a simple step at each interval, and shows in one example that the maximum error is thus reduced from 30 per cent to 2 per cent. This method could also be modified to allow for the effects of finite dead time, and the design tables and nomographs could be rescaled for the second-c/s region appropriate to servo problems, but the restriction to equal intervals would still give unnecessary accuracy and computing labour for responses with slow runback.

These methods have the Fourier transform as common mathematical background, but use different approximations and/or computing aids to simplify the practical calculations. The general picture is that Henderson's method would be preferred if a suitable harmonic analyser were immediately available, but as these are rarely found in design offices there is scope for a new method combining the computing simplicity of Floyd's method with the direct approach and ability to handle non-minimum phase networks of the modified Bedford and Fredendall or Samulon methods.

List of Major Symbols

θ_i = Input signal, θ_o = Output signal } of four terminal network

$(\theta_o/\theta_i)_{(1)}$ = Per unit response to step function input signal (given as a graph).

t = Time in seconds.

Characteristic features of step response $(\theta_o/\theta_i)_{(1)}$:

- A_o = Amplitude of immediate response
- T_o = Dead time before next significant change.
- $A_{(n)}$ = Amplitude, $T_{(n)}$ = Time } at end of n^{th} linear segment approximating the given graph.
- $(\Delta A)_n = A_{(n)} - A_{(n-1)}$, $(\Delta T)_n = T_{(n)} - T_{(n-1)}$, $(\Sigma T)_n = T_{(n)} + T_{(n-1)}$ } functions of the n^{th} linear segment
- T_C = Time to first cross correspondence
- A_{max} = Amplitude, T_{max} = Time } at peak of first overshoot.
- T_R = Time of runback from peak overshoot

$(\theta_o/\theta_i)_{(H)} = d/dt(\theta_o/\theta_{i(1)})$ = Per unit response to hammer blow input signal.

f = Actual frequency in cycles/second.

$\omega = 2\pi f$ = angular frequency in radians/second.

$(\theta_o/\theta_i)_{(j\omega)} = X_{(f)} + jY_{(f)}$ = Per unit response to sinusoidal input signal.

$Q = \sqrt{(X^2_{(f)} + Y^2_{(f)})}$ = Per unit amplitude response.

$\phi = \tan^{-1} Y_{(f)}/X_{(f)}$ = Phase angle.

Q_{max} = Amplitude, F_{max} = Frequency } at crest of amplitude response curve

$F_{0 \cdot 5}$ = "Cut-off" frequency at which the low-pass filter response has fallen to half the D.C. value.

The subscript $_{(1)}$ denotes the Heaviside step function.

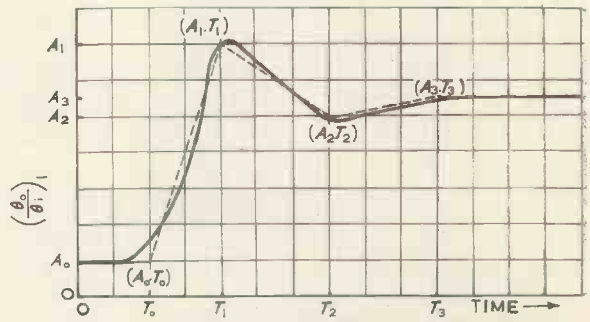


Fig. 1. Approximation of step response —— by linear segments - - -

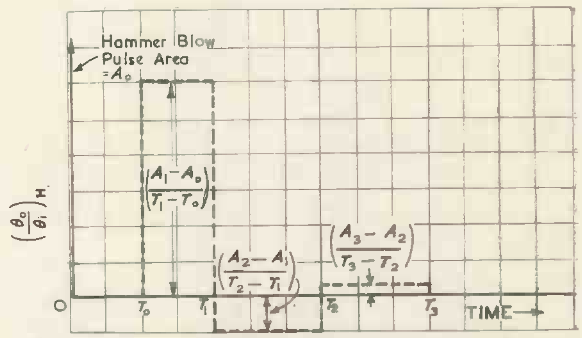


Fig. 2. Hammer blow response of system having the approximated step response

Theory of the Method

The arbitrary step response graph $(\theta_o/\theta_i)_{(1)}$ of Fig. 1 contains all the features (particularly the combination of immediate response and subsequent dead time) needed to handle low-pass, band-pass or high-pass networks, and as indicated it can be approximated by straight line segments to any desired degree of accuracy.

The hammer blow response $(\theta_o/\theta_i)_{(H)}$ of a linear system is the time derivative of the step response; hence a system with step response as approximated by straight lines in Fig. 1 has a hammer blow response as in Fig. 2 consisting of an initial pulse with area equal to the immediate component A_o, followed by a sequence of rectangles with heights and durations respectively equal to the slopes and durations of the straight line segments.

The frequency response of a linear system is the Fourier transform of the hammer blow response, and since the hammer blow (as the time derivative of a step function applied at $t = 0$) has no component prior to $t = 0$, the lower limit of integration is zero and the desired frequency response is given by:

$$(\theta_o/\theta_i)_{(j\omega)} = \int_0^\infty (\theta_o/\theta_i)_{(H)} e^{-j\omega t} dt \quad \ldots\ldots\ldots \ (1)$$

Substituting for $(\theta_o/\theta_i)_{(H)}$ the values from Fig. 2:

$$(\theta_o/\theta_i)_{(j\omega)} = \int_0^{0+} (A_o)_{(H)} e^{-j\omega t} dt + \int_{0+}^{T_0} 0 \, . \, e^{-j\omega t} dt$$

$$+ \int_{T_0}^{T_1} \frac{(A_1 - A_o)}{(T_1 - T_o)} e^{-j\omega t} dt + \int_{T_1}^{T_2} \frac{(A_2 - A_1)}{(T_2 - T_1)} e^{-j\omega t} dt +$$

$$+ \text{Similar terms until } A_{(n)} - A_{(n-1)} = 0 \quad \ldots\ldots\ldots\ldots \ (2)$$

Integrating, and manipulating by routine trigonometric methods into a form containing tabulated functions and giving the response in terms of actual rather than angular frequency:

$$(\theta_o/\theta_i)_{(j\omega)} = X_{(f)} + jY_{(f)} =$$

$$= A_o +$$

$$(A_1 - A_o)\left(\frac{\sin \pi(T_1 - T_o)f}{\pi(T_1 - T_o)f}\right)\left(\cos \pi(T_1 + T_o)f - j \sin \pi(T_1 + T_o)f\right) +$$

$$(A_2 - A_1)\left(\frac{\sin \pi(T_2 - T_1)f}{\pi(T_2 - T_1)f}\right)\left(\cos \pi(T_2 + T_1)f - j \sin \pi(T_2 + T_1)f\right)$$

$$+ \text{Similar terms until } A_{(n)} - A_{(n-1)} = 0 \quad \ldots\ldots\ldots\ldots \ (3)$$

Physically equation (3) means that each of the straight line segments contributes a component vector (Say $x_{(f)} + jy_{(f)}$) having the following characteristics:

(1) At zero frequency its amplitude equals the vertical span of its parent segment and its phase lag is either zero or 180° according as its parent segment had positive or negative slope. (Note that a horizontal segment has zero vertical span and therefore contributes no component vector.)

(2) With increasing frequency, its amplitude first falls to zero, then builds up with reversed sign to about 20 per cent of initial value, finally executes a slowly decaying oscillation about zero. The frequency at which its amplitude first reaches zero is exactly the reciprocal of the time duration of its parent segment, and sucessive zeros occur at integral multiples of this frequency. (Note that a vertical segment has zero duration, and therefore contributes a vector of constant amplitude.)

(3) It rotates clockwise in the $(X_{(f)} + jY_{(f)})$ plane through an angle of lag proportional to frequency, each complete revolution being covered for a frequency increment exactly equal to the reciprocal of the average time of occurrence of its parent segment as measured from zero time to the midpoint of the segment. (Note that the vertical segment occurring at zero time contributes a vector having infinitely slow rotation, i.e. the constant term A_o of equation (3)).

While the individual component vectors are thus very simply related to their parent straight line segments, the resultant of three or four such vectors with different rotation and attenuation rates traces rather complex patterns; the physical picture is used again later for qualitative derivation of the $Q_{max} - A_{max}$ relation, but so far it has not revealed any simpler form than equation (3) for computing particular cases.

Computing Routines

The individual terms to be computed and added are of form:

$$(x_{(f)} + jy_{(f)}) = (\Delta A)\left(\frac{\sin \pi \Delta T f}{\pi \Delta T f}\right)(\cos \pi \, \Sigma T f \; - \; j \sin \pi \, \Sigma T f) \quad \ldots\ldots\ldots\ldots \ (4)$$

in which the coefficients ΔA—the vertical span, ΔT—the time span, ΣT—the sum of start and finish times, are written down from inspection of the straight line segments, and frequency f is allocated a series of values from zero upwards to define the frequency response in the required detail.

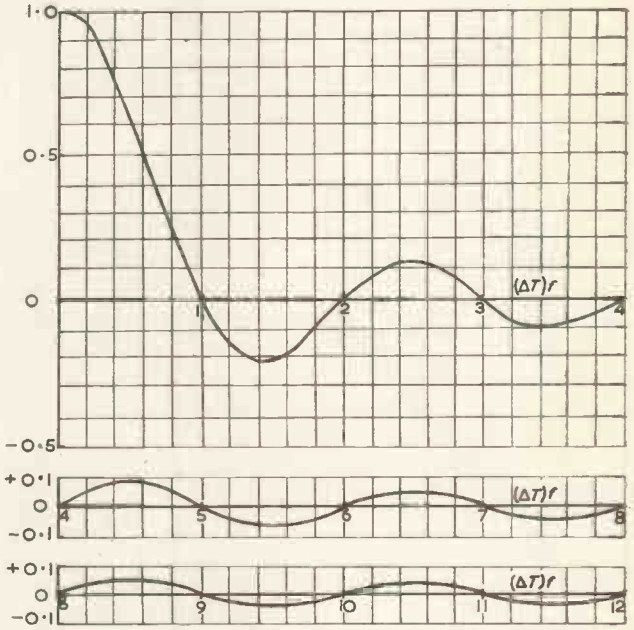


Fig. 3. Graph of $\frac{\sin \pi(\Delta T)f}{\pi(\Delta T)f}$ as function of $(\Delta T)f$

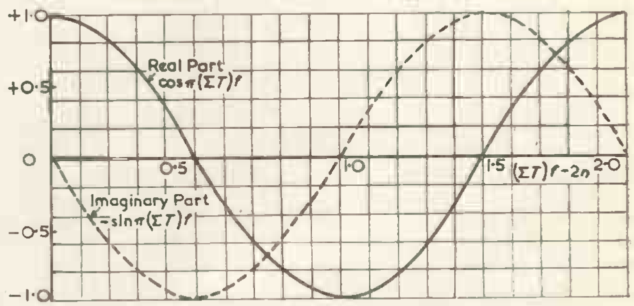


Fig. 4. Graphs for evaluating [cos $\pi(\Sigma T)f$ − j sin $\pi(\Sigma T)f$] as function of $(\Sigma T)f$

Note: n is a positive integer chosen to bring the actual value of $(\Sigma T)f$ within the plotted range of these periodic curves.

Existing tables give $\sin \alpha / \alpha$ for $0 < \alpha < 40$ radians[9] and conveniently side by side $\cos \beta$ and $\sin \beta$ for $0 < \beta < 10$ radians[10]. The precision of these tables is rarely justified in engineering work where linear theory is being applied to systems of dubious linearity. The curves of Fig. 3 and Fig. 4 give adequate precision in such cases and save a good deal of page-turning; Fig. 3 gives the factor in the second bracket for $0 < \Delta Tf < 12$ and Fig. 4 gives the real and imaginary parts of the factor in the third bracket for any value of ΣTf; in the latter case integral multiples of two are to be deducted from the actual value of ΣTf until the remainder is within the plotted range $0 < \Sigma Tf < 2$ of the periodic functions.

If the transformation from step response to frequency response is often needed, the simple computing device of Fig. 5 is worth constructing. The factor $(\cos \pi\Sigma Tf - j \sin \pi\Sigma Tf)$ in equation (4) is simply a unit amplitude vector lagging by $\pi\Sigma Tf$ radians or $0{\cdot}5\Sigma Tf$ revolutions behind the input signal vector on the positive real axis, and the factor $\dfrac{\sin \pi\Delta Tf}{\pi\Delta Tf}$ in equation (4) is simply a scalar multiplier numerically less than unity. The computing device has a graph of this multiplier as a function of (ΔTf) within a transparent cursor which can be rotated relative to the Cartesian gridded baseboard, the origin of the multiplier graph coinciding with the origin of the Cartesian grid. To find the component vector $(x + jy)$ from equation (4) as set up with numerical values of ΔA, ΣT, ΔT and chosen frequency, the procedure is:

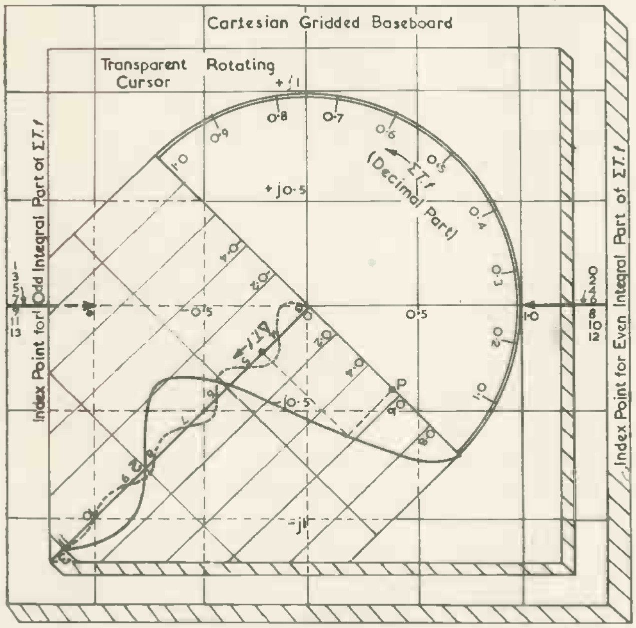


Fig. 5. Simple computing device accepting ΣTf and ΔTf and evaluating $(\cos \pi\Sigma Tf - j \sin \pi\Sigma Tf)\dfrac{\sin \pi\Delta Tf}{\pi\Delta Tf}$ as $(a + jb)$

Setting shown is for $\left\{\begin{array}{l}\Sigma Tf = 0{\cdot}25 + 2n\\ \Delta Tf = 0{\cdot}55\end{array}\right\}$ giving $(+0{\cdot}4 - j0{\cdot}4)$ as read on the Cartesian grid below point P

(1) Compute ΣTf (slide rule or mental arithmetic if suitable values of f are chosen) and rotate the cursor until the decimal part of ΣTf as read on the divided semi-circle lies over the positive real axis when the integral part of ΣTf is even, or over the negative real axis when the integral part of ΣTf is odd. The edge OD of the contained chart has now been set to the required angle of lag $\pi\Sigma Tf$ radians behind the input signal on the positive real axis.

(2) Compute ΔTf (slide rule or mental arithmetic again) and locate this value on the ΔTf axis of the contained chart. Project up to the curve and then across to the amplitude axis of the contained chart, thus defining point P on line OD. Line OP now has the direction and magnitude of the vector representing the product of the second and third brackets of equation (4).

(3) Read $(a + jb)$ the Cartesian co-ordinates of point P from the baseboard grid, and multiply these by ΔA (slide rule) obtaining $(x + jy)$ the desired solution of equation (4).

For each chosen frequency, pairs of real and imaginary parts are thus found corresponding to the component vectors originated by the straight line segments, and their algebraic sum gives the total response vector at this frequency. Most of the examples in this article were worked out on the first model of this computing device made with commercial squared paper to scale of five inches per unit amplitude, and the average computing time was about one minute per straight line segment per point on the frequency response locus. Thus, including the time to fit the straight line segments and read off the coefficients into the corresponding term equations, it takes about an hour to transform a typical four-segment step response. Particularly in the early stages of synthesizing a step response to meet given requirements, it would be useful to have a set of simple rules relating, say, the cut-off frequency and the frequency and amplitude of maximum transmission to the geometry of the step response. Use of these rules would minimize the number of abortive transformations and save appreciable computing time. In Part 2 after examining the fundamental and practical limitations of the graphical approximation method, some simple rules for R.P.C. servos are sought and tentative proposals put forward.

Finally in Part 3, the graphical method and the approximate rules are checked against the measured performance of practical servos.

(To be continued)

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TELEVISION IN THE ISLE OF MAN

The Postmaster-General announced some months ago that the Government had authorized the construction of a television transmitting station in the Isle of Man. Special efforts have been made by the BBC to find a site and install the equipment so that a low-power temporary station may be working in the Island before Christmas, a site having been chosen on the south-western outskirts of Douglas.

The temporary station will serve the town of Douglas and its immediate neighbourhood and part of the south-east coast of the Island—probably from Clay Head to Castletown. The transmitters will work on the same channel as the existing stations at Wenvoe and Pontop Pike, the frequencies being 66·75Mc/s for vision and 63·25Mc/s for sound. The asymmetric side-band method of transmission will be used, with vertical polarization.

The temporary station will remain in use until the permanent station is completed. The latter will be erected on a site not yet selected and will have higher power and greater range; it will use the same channel as the temporary station, so that receivers will not have to be altered.

A Set of 0·4 Octave Bandpass Filters

for Frequencies between 10 and 200kc/s

By H. Pursey*, B.Sc.

The set of filters to be described has been designed for the elimination of harmonics from waveforms between 10 and 200kc/s. The main requirement was that the filters should have a fairly rapid cut-off, especially on the high-frequency side, and in order to achieve this with the minimum of expense and weight, use has been made of the fact that a sharper cut-off may be obtained by operating the filters into an abnormally low impedance, in this case about a tenth of the mid-pass-band impedance. The improved frequency characteristic is obtained at the cost of a few decibels extra insertion loss, but this can generally be made up by additional amplification in the system.

EACH unit has a band-pass of about 0·4 octaves, eleven units covering the required range. Fig. 1 shows the circuit of a single unit.

An advantage of using a capacitive shunt element is that the greatest attenuation is obtained on the high-frequency side. The attenuation one octave above the lower cut-off frequency is rather more than 35 decibels, while the maximum slope of any part of the curve is 100db/octave. It is shown in the Appendix that this will modify the apparent resonant frequency of the system (defined as the frequency of maximum response) by 0·01 per cent when the element being measured has a Q of 200, and that the error decreases as $1/Q^2$ for larger values of Q.

The Zobel-type filter is obtained from the formulæ:

$$L = \frac{R}{\omega_2 - \omega_1} \qquad C_1 = \frac{\omega_2 - \omega_1}{\omega_1^2 R} \qquad C_2 = \frac{2}{(\omega_1 + \omega_2)R}$$

where R is the terminating resistance, ω_1 and ω_2 are cut-off

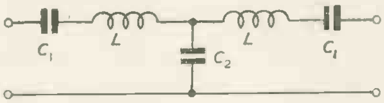


Fig. 1. Circuit of a single unit

pulsatances. The attenuation of such a filter terminated with R at both ends (i.e., fed from a source of impedance R into a load R) is given by:

$$10 \log_{10}\left[f^2\left\{1 + \frac{2}{(\alpha-1)^2} + \left(\frac{\alpha+1}{\alpha-1}\right) - f^2\left(\frac{\alpha+1}{2\alpha-2}\right)^2 - \left(\frac{1}{f^2}\right)\left(\frac{4}{\alpha^2-1} + \frac{4}{\alpha^4 - 2\alpha^2 + 1}\right)\right\}^2 + f^3\left\{\left(\frac{\alpha+1}{\alpha-1}\right) - \left(\frac{1}{f^2}\right)\left(\frac{4}{\alpha^2-1} + 2\right)\right\}^2\right]$$

where $\alpha = \omega_2/\omega_1$ and $f =$ frequency $\times\ 4\pi/(\omega_1 + \omega_2)$.

For a nominal bandwidth of 0·4 octaves (necessary if the required range is to be covered by eleven units) this expression becomes (putting $\alpha = 1{\cdot}32$):

$$10 \log_{10}\left[f^{2\,2}\left\{27{\cdot}85 - 13{\cdot}15f^2 - 12{\cdot}68/f^2\right\}^2 + f^3\left\{7{\cdot}26 - 7{\cdot}41/f^2\right\}^2\right]$$

The attenuation curve is shown in Fig. 2, and it is clear that the cut-off is not sharp enough to satisfy the above requirements. A sharper cut-off may be obtained by designing the filter for a higher characteristic impedance and working it into the same load. In this case it is found necessary to take account of the dissipation of the coils.

If the coil resistance is R_L, the attenuation is given by:

$$10 \log_{10}\left[4(1 + k - f^2)^2/k^2 + \left\{f/x - x(f^2-1)(f^2 - 1 - 2k)/f\right\}^2/k^2\right] + 20 \log_{10}\left\{(R + R_L)/R\right\}$$

where $f =$ frequency $\times\ 2\pi\sqrt{(LC_1)}$

$$x = \frac{\sqrt{(L/C_1)}}{R + R_L}$$

$$k = C_1/C_2$$

$R =$ terminating resistance.

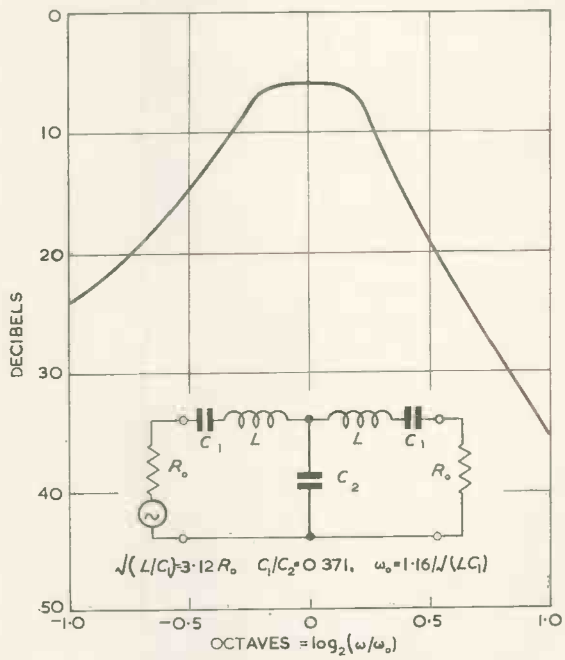


Fig. 2: Calculated response of Zobel-type 0·4 octave filter

Suitable values are found to be $x = 10$, $k = 0{\cdot}235$, $R_L = 300\Omega$. With this value of R_L, the Q of the coils is 30, and this can be realized in the case of the largest coil (129mH) by filling a Gecalloy No. 6 former with No. 44 D.S.C. wire. The other coils have higher Q's, and it is necessary to add resistance in series with them. The response is given by:

$$10 \log_{10}\left[\left\{8{\cdot}5(1{\cdot}235 - f^2)\right\}^2 + \left\{4{\cdot}25f[0{\cdot}1 - (10/f^2)(1 - f^2)(1{\cdot}47 - f^2)]\right\}^2\right] + 20 \log_{10} 1{\cdot}5$$

The curve is shown in Fig. 3. It will be seen that it is displaced to the right compared with Fig. 2: this is due to the different way of defining f in the two cases. The

* *National Physical Laboratory.*

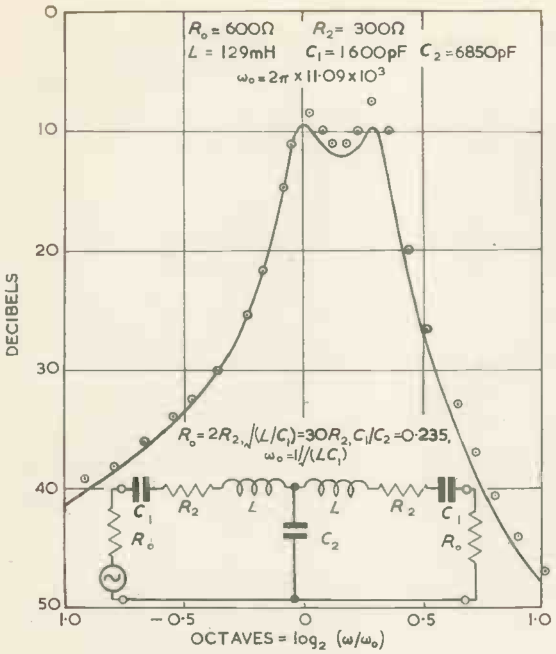


Fig. 3. Response curve for a typical filter

The full line represents the calculated response, while the points were obtained from measurements on the lowest frequency unit.

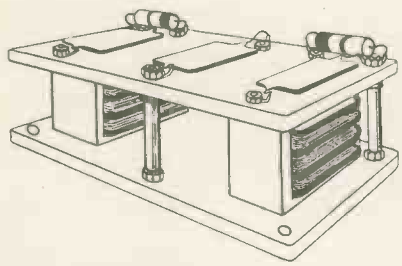
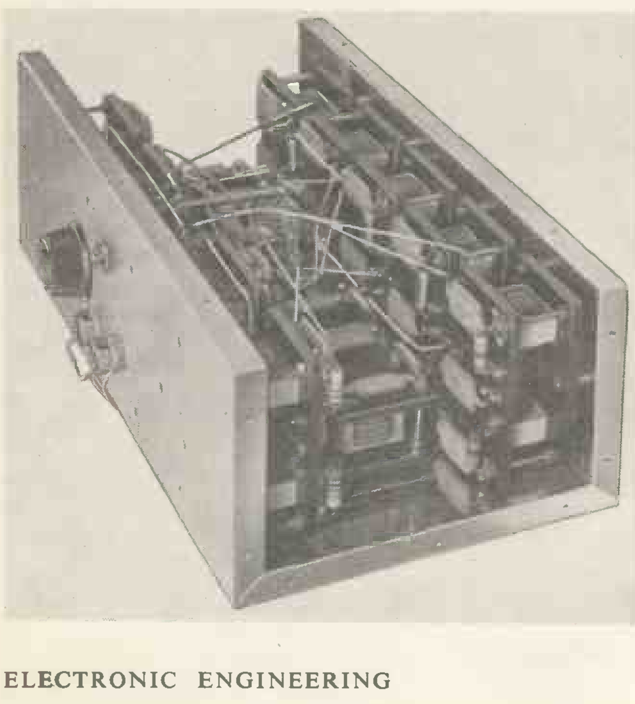


Fig. 4. A single filter unit

Fig. 5. A complete set of filters.



measured response for the lowest frequency filter is shown on the same graph. The attenuation is much more rapid than in the case of the Zobel filter, and this is obtained at the cost of an additional attenuation of about 5db in the pass-band.

A set of eleven filters has been built, designed to cover the range of from 10·5 to 220kc/s. The switching is by means of two 11-way Yaxley type wafers, with a ring on each wafer which connects to earth all contacts except those actually in use at any time. Each unit measures 3½in by 2in by 1¾in, and consists of the coils mounted between two slabs of 3/16in ebonite, while the remaining components are mounted on top. The coils are stuck to one of the slabs, and held firmly by the pressure between the two slabs. A sketch of the unit is shown in Fig. 4, while the complete set of filters and switching is shown in Fig. 5.

Table 1 gives the component values. Tolerances are ±½ per cent on reactive elements, while the total resistance (external resistance + dynamic resistance of coils at mid-pass-band frequency) is within ±5 per cent of 300 ohms.

Since the preparation of this article a further three sections, of the type described above, have been built extending the range of the unit to 420kc/s. These sections

TABLE 1
Component Values

FILTER NUMBER	FREQUENCY RANGE (3db down) (kc/s)	L (mH)	C_1 (pF)	C_2 (pF)
1	10·4 – 14·0	129·0	1 600	6 850
2	13·6 – 18·4	98.0	1 220	5 200
3	18·2 – 24·3	74·0	920	3 930
4	23·9 – 32·0	56·0	700	2 980
5	31·5 – 42·3	42·5	530	2 260
6	41·5 – 56·0	32·2	400	1 700
7	55·0 – 73·5	24·4	300	1 300
8	72·0 – 96·0	18·53	230	980
9	95·0 –126·0	14·07	170	750
10	126·0 –168·0	10·68	120	550
11	164·0 –216·0	8·06	95	430

behave satisfactorily, although it was not possible to achieve quite as wide a bandwidth as in the case of the lower frequency units, while at the same time maintaining a flat response inside the pass-band.

Acknowledgments

Acknowledgment is made to Miss M. A. Trott for her assistance in construction of the units and also for much of the computing work involved in calculating responses.

The work described above has been carried out as part of the Research Programme of the National Physical Laboratory, and this article is published by permission of the Director of the Laboratory.

Appendix

The Effect of Inserting a Network of Non-uniform Frequency Response on the Apparent Resonant Frequency of a Circuit

The resonant circuit concerned may be either electrical or mechanical. For simplicity it will be assumed to be electrical: in the case of a mechanical circuit its electrical analogue could be used in the following analysis.

Consider a simple resonant circuit consisting of inductance, capacitance and resistance, with a voltage V_I injected in series.

Then V_c, the voltage across the capacitor is given by:

$$|V_I/V_c| = [(1-\omega^2LC)^2 + \omega^2C^2R^2]^{1/2}.$$

Putting $\omega^2LC = f^2$ and $\omega L/R = Q$:

$$|V_I/V_c| = [(1-f^2)^2 + f^4/Q^2]^{1/2}.$$

If γ is the gain in decibels,

$$\gamma = -10\log_{10}[(1-f^2)^2 + f^4/Q^2].$$

Now suppose the filter unit has an attenuation characteristic (measured in decibels) of $g(f)$. Then the total gain is:

$$\gamma = -10 \log_{10}[(1 - f^2)^2 + f^4/Q^2] - g(f)$$

The maximum gain occurs when $d\gamma/df = 0$,

i.e., when $(d/df)[10 \log_{10}\{(1 - f^2)^2 + f^4/Q^2\} + g(f)] = 0$

$$\therefore 10[-4f + 4f^3(1 + 1/Q^2)] \log_{10}\epsilon + [1 - 2f^2 + f^4(1 + 1/Q^2)]\, g'(f) = 0$$

Putting $P = 1 + 1/Q^2$, this equation becomes:

$$Pf^4g'(f) + 40Pf^3 \log_{10}\epsilon - 2f^2g'(f) - 40f \log_{10}\epsilon + g'(f) = 0$$

Now assume $g(f)$ is constant. When P is nearly 1 (i.e., Q is large) the equation has a root near unity. Putting $f = 1 + \delta$, and omitting terms higher than first order in δ,

$$\delta[40(3P-1)\log_{10}\epsilon + 4(P-1)g'] = (1-P)(g' + 40 \log_{10}\epsilon)$$

Now:

$$3P - 1 \approx 2 \text{ and } 1 - P = 1/Q^2$$

$$\therefore 80\,\delta \log_{10}\epsilon \approx (1/Q^2)(g' + 40 \log_{10}\epsilon)$$

$$\therefore \delta \approx 1/2Q^2 + \frac{g'}{80Q^2 \log_{10}\epsilon}$$

The first term gives the departure of the frequency ot maximum response from $1/2\pi\sqrt{(LC)}$ due to the resistive damping, while the second term is the additional change caused by the filter characteristic. The two terms are equal if $g' = 40 \log_{10}\epsilon$.

Now $g' = d\gamma_F/df$, where γ_F is the filter attenuation. If η is the frequency measured in octaves, $\eta = \log_2 f$ and $d\eta/df = 1/(f \log_\epsilon 2)$.

$\therefore d\gamma_F/d\eta = f(d\gamma_F/df)\log_\epsilon 2 \approx g' \log_\epsilon 2$ at resonance ($f \approx 1$)

$\therefore$ the two terms are equal when:

$$d\gamma_F/d\eta = 40 \log_\epsilon 2 \log_{10}\epsilon = 40 \log_{10} 2 = 12\text{db/octave}.$$

When:

$$d\gamma_F/d\eta = 100\text{db/octave}, \quad g' = 100/\log_\epsilon 2 \text{ and}$$

$$\delta = 100/80Q^2 \log_{10} 2 = 4{\cdot}15/Q^2$$

$\therefore$ when $Q = 200$, $\delta = 0{\cdot}01$ per cent.

Hence the frequency error introduced by such a filter will not exceed 0·01 per cent provided the Q of the mechanical or electrical circuit being measured is greater than 200.

Some

High Frequency Effects in Germanium Diodes

with Special Reference to Television Sound Detectors

By D. D. Jones*, M.Sc., D.I.C., and B. C. Brodribb *

Point contact germanium diodes are being increasingly used as H.F. and V.H.F. detectors, such as the second detectors in television receivers. It has been found that the performance of some diodes at these frequencies may differ appreciably from that expected from a study of their static characteristics. The R.F. input impedance of the diode detector circuit may be considerably lower than the expected value. The maximum R.F. voltage (effectively a turnover voltage) that can be developed across the diode circuit may be 50 per cent or less of the static turnover voltage. These effects, which are shown to be consistent with hole-storage effects occurring in germanium, can give rise to highly non-linear detection. The effects can be eliminated in the diodes by production techniques designed to reduce hole-life times. It is, however, desirable to form production tests, to ensure that the R.F. performance of the diodes is satisfactory.

THE germanium point contact diode is more suitable than the thermionic diode for many circuit applications on account of its small size, low capacitance (below lpF for the GEX high reverse voltage series made by the G.E.C.), and absence of heater supplies. It can be wired directly into circuits without any complications of valve holders. An example of these applications is found in modern television receiver design, where increasing use is being made of these diodes in both vision and sound detector stages. The output impedance of the final sound channel I.F. amplifier, as well as the diode load, are made high, of the order of 20kΩ and 50kΩ respectively, to obtain maximum gain. For reasons connected with interference suppression, it is often found that the output level of the I.F. amplifier is high, voltages of 10 to 40 volts peak-to-peak being encountered. The intermediate frequency employed is often 35Mc/s or greater. It has been found that a germanium diode giving linear demodulation at lower signal levels may give rise to considerable non-linearity at higher levels of input. This non-linearity takes the form of "clipping" one of the modulation peaks, and is often referred to as "crushing" of the waveform.

Many germanium diodes have been examined for this effect, and it is found that there is no correlation between these distortion effects at radio frequencies and the static

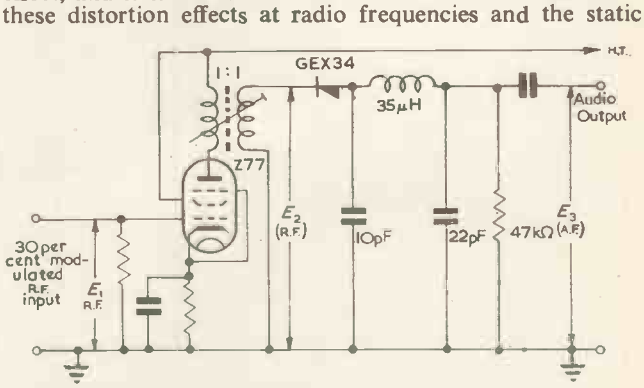


Fig. 1. Typical television sound detector circuit

characteristics of these diodes. The effects, however, are consistent with hole-storage effects, which are believed to account for the poor performance of germanium diodes in high speed pulse circuits.

Fig. 1 shows a typical sound detector circuit using a

* *The General Electric Company Ltd., England.*

germanium diode, and Fig. 2 shows the type of distortion that may occur.

In the early days of germanium devices, the main production tests applied were D.C. measurements. Table 1 gives some of the usual measurements made on four diodes, two of which had been found to give rise to R.F. limiting. Table 2 shows the performance of the same four diodes in a 35Mc/s test circuit which is described below: it is sufficient to state at this point that the linearity of E_2 and E_3 with variation of E_1 is a measure of the linearity of the diodes as detectors. Thus diodes 1 and 2 are satisfactory, whereas 3 and 4 are not: no indication of this can be seen from the D.C. measurements in Table 1.

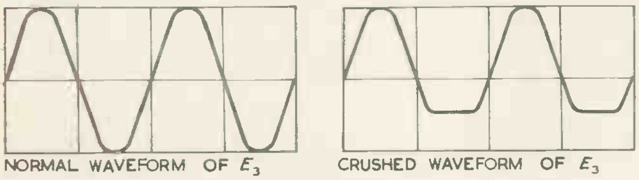


Fig. 2. Detector output waveform
(a) *Without Distortion* (b) *With Distortion*

TABLE 1
Static Characteristics of Selected Diodes

DIODE NUMBER	TURNOVER VOLTAGE	DIRECT CURRENT AT: +1 VOLT (mA)	−1 VOLT (μA)	−10 VOLTS (μA)	−25 VOLTS (μA)
1	140	4·0	2·5	11	30
2	97	2·5	3·0	67	120
3	115	2.5	10.0	21	95
4	100	1·0	2·5	86	350

TABLE 2
R.F. Characteristics of Selected Diodes

DIODE NUMBER	VOLTAGES E_2 AND E_3 CORRESPONDING TO: E_1=0·4v		E_1=1·0v		E_1=1·6v	
	E_2 (VOLTS)	E_3 (VOLTS)	E_2 (VOLTS)	E_3 (VOLTS)	E_2 (VOLTS)	E_3 (VOLTS)
1	12·0	14·0	27·5	33·0	38·0	44·5
2	11·0	13·0	26·5	31·5	34·5	40·5
3	10·0	9·0	16·0	12·5	18·5	13·5
4	11·5	12·5	17·5	17·0	18·0	17·0

R.F. Effects

Since the performance of a diode as a detector at high frequencies cannot be predicted from a study of its static characteristics, some factor must be present which is only effective at high frequencies. The most probable cause of trouble is the hole-storage effect.

When a positive potential is applied between the whisker and the block of n-type germanium used in the diodes, conduction occurs, partly due to the flow of positive charge carriers known as "holes" from the whisker to the germanium, and partly by the flow of electrons from the germanium to the whisker. If this applied potential is suddenly switched from a positive to a negative value, any holes which still exist in the germanium will be accelerated back towards the whisker. These holes are thought to have different mean life-times and mobilities in different diodes, depending upon such features as the impurity content of the material and the forming treatment applied to the contact during manufacture. Because of the finite life-time of these holes, the current I_b in the reverse direction would be expected to vary with time in the manner shown in Fig. 3.

The current will rise quickly to a peak value I_p higher than the static value I_s, because of the effect of holes, and will decay asymtotically to I_s. The time t_p taken to reach peak value has been reported to vary from a few millimicroseconds to many microseconds for different diodes.

The greater the value of the applied negative potential the greater the acceleration of holes back to the whisker, and hence the greater the value of I_p.

When the diode is used as an R.F. detector, the switch from positive to negative value does not occur instantaneously but in a time $T/2$ where T is the period of the

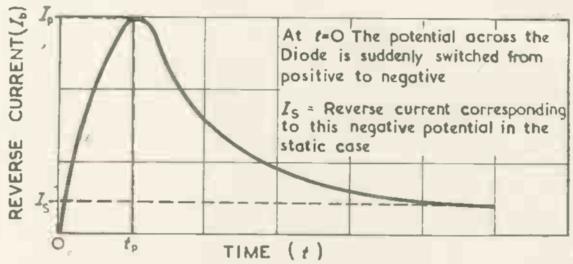


Fig. 3. Variation of reverse current with time

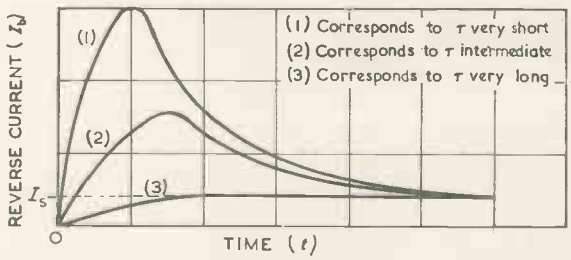


Fig. 4. Variation of reverse current with time for various switching speeds

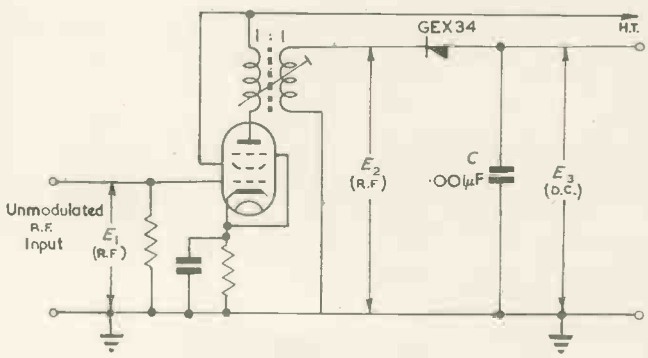
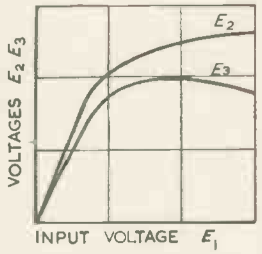


Fig. 5. Basic test circuit

R.F. voltage. The variation of the reverse current/time characteristic with τ, the time taken by the reverse potential to reach its peak value, is shown in Fig. 4.

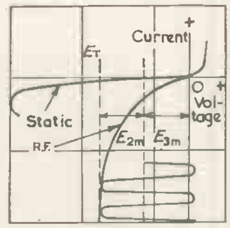
Thus the effect is expected to be more pronounced at high frequencies (i.e., τ very short), and if t_p the time to reach the peak current, is comparable with the period of the R.F. voltage, the negative voltage peaks may give rise to reverse currents considerably in excess of the value expected from the static characteristic. Thus the reverse impedance and the R.F. input impedance of the diode circuit, which largely depends on the reverse impedance, may be much less than would be expected. This effect is easily shown by a simple experiment as shown in Fig. 5.

C is a fairly large capacitor, say ·001μF, which will be charged during positive peaks through the forward impedance of the diode, and will discharge during negative peaks through the reverse impedance of the diode. If the value of the amplifier input voltage E_1 is increased, E_2 and E_3 both increase linearly at first, but eventually (due



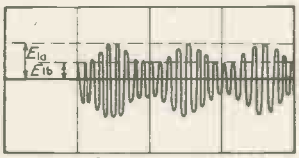
(a)

Fig. 6(a). Detector non-linearity



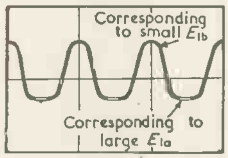
(b)

Fig. 6(b). Static and R.F. reverse characteristics for poor diode



(a)

Fig. 7(a). Input waveform to final amplifier



(b)

Fig. 7(b). Distorted output waveform from detector

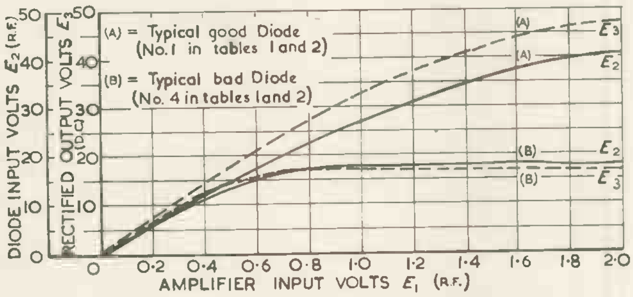


Fig. 8. Performance of good and bad diodes in R.F. test circuit

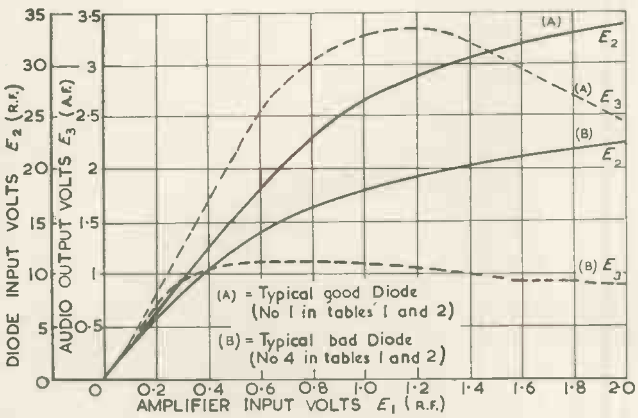


Fig. 9. Performance of good and bad diodes in television sound detector circuit

to the hole-storage effect already described) the reverse impedance of the diode will fall much more rapidly with applied R.F. voltage than expected from its static characteristics. This results in a lowering of the gain of the amplifier stage as well as a rapid discharge of the capacitor during negative peaks, leading to a flattening off, and possibly a falling off, of the curves of E_2 and E_3 with E_1. (See Fig. 6(a)).

With reference to Fig. 6(b), the sum of the peak values E_{2m} and E_{3m} of E_2 and E_3 gives E_T, which can be referred to as the dynamic turnover voltage at R.F., as opposed to V_T, the static turnover voltage.

Consider the case of a modulated input signal in a sound detector circuit such as that of Fig. 1. For low values of E_1 (see Fig. 7(a)), the reduction of reverse impedance by hole-storage effect is small and the modulation is reproduced without distortion. For large values of E_1 the hole-storage effect is much greater; hence the distortion corresponding to E_{1a} may be much greater than that corresponding to E_{1b} and this may give rise to a distorted output, as shown in Fig. 7(b).

To verify this hypothesis, a number of diodes were checked in the circuit of Fig. 5 for linearity of rectified output with R.F. input to the amplifier. They were also tested for linearity in the sound detector circuit of Fig. 1. The results, as plotted in Figs. 8 and 9, show good agreement between the tests.

Effect of D.C. Bias

If the effect is due to hole-storage, then the effective turnover voltage at R.F. should be considerably increased (in the case of a "bad" diode) if the diode is biased by a D.C. potential so that it does not conduct in the positive direction. Fig. 10 shows the effect of externally applied

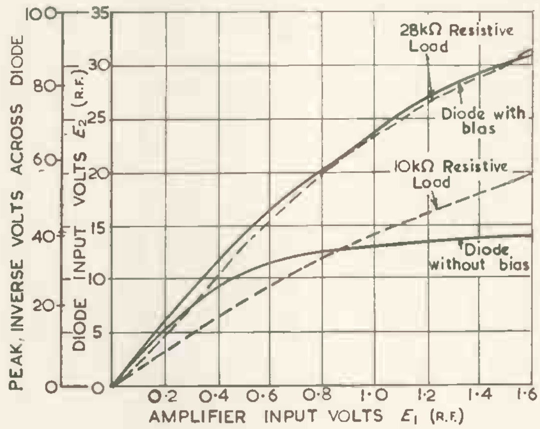


Fig. 10. Performance of a typical bad diode in the R.F. test circuit, with and without applied D.C. bias

D.C. bias on such a diode. It is seen that without this bias the turnover voltage is about 40 volts, whereas with bias there is no sign of R.F. turnover and the diode input impedance is several times higher. With the R.F. equipment used, the maximum measurable turnover voltage was 150 volts. The static turnover voltage of this particular diode was 180 volts.

Application to Production Testing

To eliminate these effects of hole-storage, special diodes are now manufactured for high frequency detection. These diodes contain specially processed germanium, and are manufactured by techniques designed to reduce hole life-time. However, to ensure that they adequately meet the requirements, they are tested by a method similar to that described.

Measurement of the R.F. turnover voltage is both laborious and unnecessary as a production test. It is adequate to compare the values of E_2 (see Fig. 5) with the diode in and out of circuit. In this way it can be ensured that the reverse impedance of the diode under R.F. conditions is sufficiently high for the applications it is designed to meet.

High Voltage Sawtooth and Rectangular Wave Pulse Generator

By W. D. Edwards*, B.Sc., Ph.D.

Vacuum tubes have been used to construct high voltage sawtooth and rectangular wave-pulse generators in order to facilitate research on the properties of dielectrics. The rectangular wave-pulse generator has an output of 0-20kV and a continously variable pulse duration of 0·05-80 microseconds. In the event of a test sample breakdown the apparatus is protected by an automatic shut-off circuit. The component circuits of the pulse generator are briefly described. The sawtooth pulse generator has a peak output of 18kV with rise times to this voltage of from 15 microseconds to 10 milliseconds. An automatic shut-off circuit is again provided and is briefly described as are the other constituent circuits of the pulse generator.

SHORT voltage pulses are being used here to measure the dependence of the electrical breakdown strength of a liquid on the time of application of the voltage [1,2]. Liquids have breakdown strengths as high as 4×10^6V/cm and the shortest sphere gaps which it is practicable to measure with the apparatus constructed are about 25 microns in length. Hence, such tests require rectangular voltage pulses of at least 10kV magnitude and duration continuously variable between 0·05 and 100 microseconds. An automatic shut-off circuit is necessary because when breakdown of the liquid sample occurs some limitation of the breakdown current is desirable. Some magnetron modulators[3] have such a voltage output but often have the disadvantages of poor pulse shape, inflexibility of pulse length and complete absence of fast automatic shut-off circuits.

Turner and Lewis[4] have made several measurements of prebreakdown currents in glasses using positive, sawtooth voltage pulses. It is desirable to extend these measurements to lower temperatures for a number of glasses. To accomplish this, standard low voltage circuits[5] for the production of sawtooth voltage pulses have been adapted for high voltages using 5D21 tetrodes. An automatic shut-off circuit which will terminate the pulse at a pre-selected amplitude or when breakdown of the sample occurs is also necessary. In the following paragraphs brief descriptions are given of the sawtooth and rectangular pulse generators which have been built.

The Rectangular-Voltage Pulse Generator

The circuit may be divided for convenience into four stages: the pulse-forming stage preceded by a driver stage, the output stage and the automatic shut-off circuit. Two thyratrons in cascade V_5 and V_6 which are normally non-conducting constitute the pulse-forming circuit (see Figs. 1 and 2).

If at time $t = t_A$ a positive pulse is applied to the grid of V_5, V_5 fires and a steep fronted positive pulse of magnitude approximating that of the supply appears across the load resistor R_L. It at time $t = t_B$ a positive pulse applied to the grid of V_6 fires V_6, then R_L is shorted and the voltage across R_L drops very rapidly to approximately half magnitude. The duration of the pulse appearing across R_L is $(t_B - t_A)$ and this may be continuously varied over a wide range of values by adjustment of the times t_A and t_B. Very rapid rates of rise and fall of the output pulse are possible as the tube and stray capacitances associated with R_L are small and as V_5 and V_6 are thyratrons large currents may be passed. These rates of voltage change and the jitter inherent in thyratrons set a limit to the shortest pulse duration which may be obtained. 2D21's used in a trial circuit produced 600V, 0·03μsec rectangular pulses with a jitter of $\leqslant 10^{-9}$sec; 0·01μsec pulses were obtainable, but greatly dependent on circuit conditions. 4C35 hydrogen thyratrons were used instead of 2D21's in the final circuit because of their zero grid cut-off voltage and the large currents and voltages necessary to drive the output stage. A vacuum tube output stage sacrifices to some extent the rapid switching given by the thyratrons, but the following points led to its adoption.

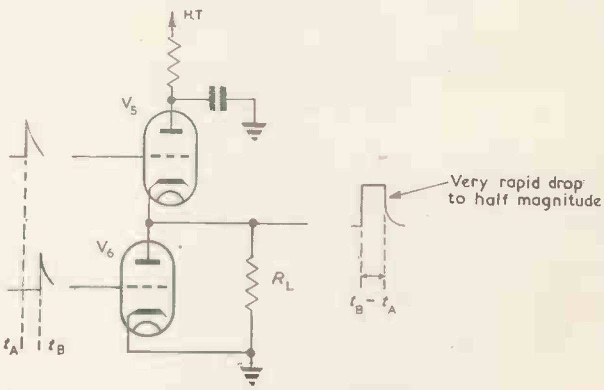


Fig. 1. The rectangular-pulse forming circuit

(a) If the output tube is normally biased beyond cut-off the comparatively slow drop from 50 per cent to zero of the trailing edge of the drive pulse (see Fig. 1) does not appear in the negative output pulse.

(b) Peak outputs of 20kV are obtainable compared with a theoretical maximum of 16kV obtainable from the drive circuit alone if 5C22 hydrogen thyratrons are used.

(c) In the event of a sample breakdown the 5D21 impedance limits the breakdown current to a few amperes and hence minimizes damage to the test electrodes.

The 5D21 anode voltage and hence the pulse magnitude was controlled by means of a Variac in the primary circuit of the transformer feeding the 20kV supply. The grid of V_7 was normally operated with 1 000 volts bias which limited the peak current to approximately 13A. However, this increased the rise time of a 15kV pulse to 0·06μsec and so for pulse lengths of less than 1μsec the bias was reduced to 500V allowing of a rise time to 15kV of approximately 0·03μsec. In the event of a sample breakdown the peak pulse current was then approximately 25A. The duration and magnitude of the negative-going output pulse was measured with an oscilloscope used in conjunction with carbon resistor potentiometers and shock excited, oscillatory calibration circuits[6]. Typical output pulse oscillograms are shown in Fig. 3.

* National Research Council of Canada, Ottawa, Ontario.

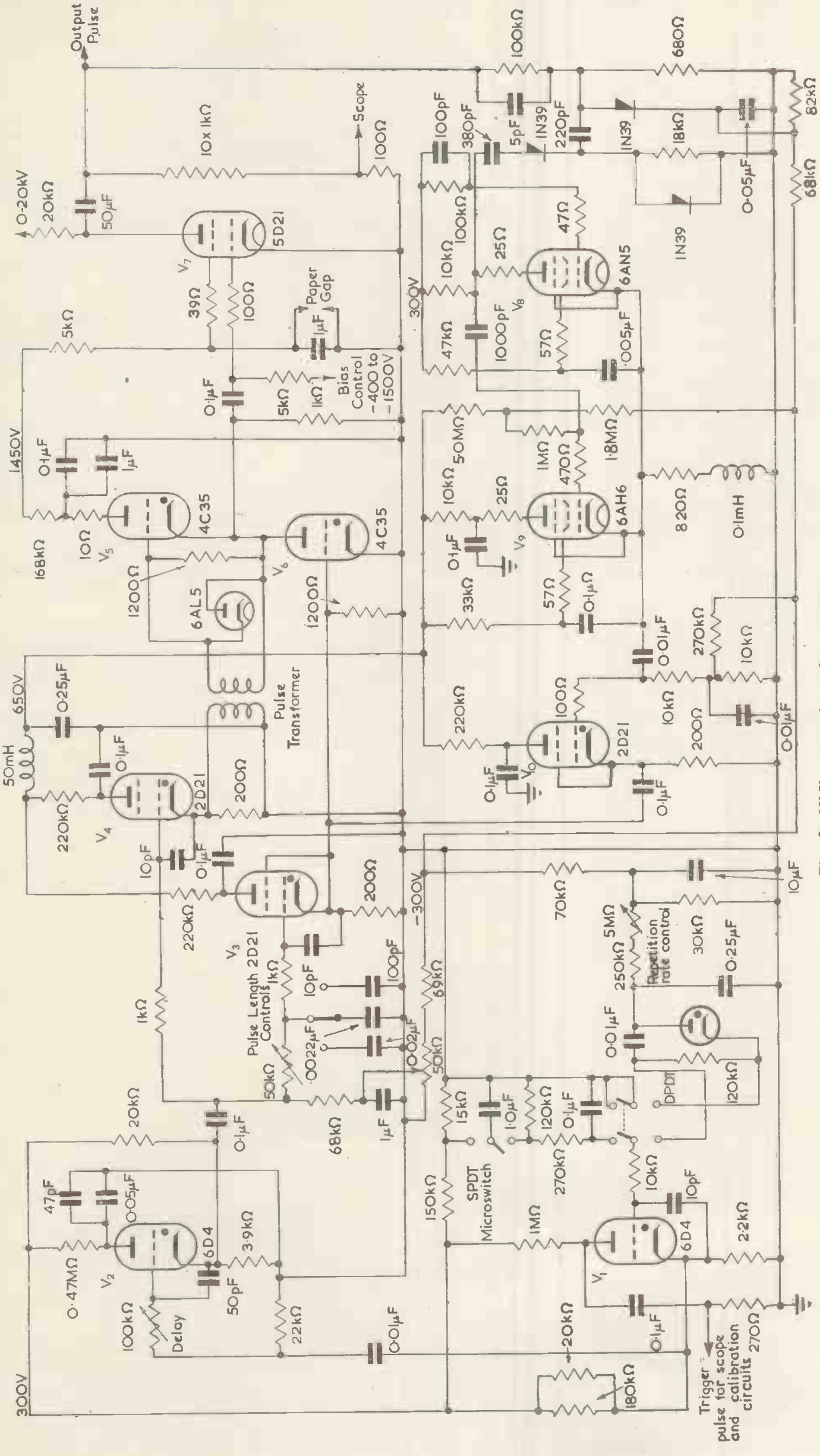


Fig. 2. 20kV rectangular pulse generator

The driver stage consists of the miniature thyratrons V_1, V_2, V_3 and V_4. Single pulses are normally required but to facilitate circuit adjustment provision is made for triggering pulses at repetition rates of 1 to 30c/s. The single pulses are provided by a manually operated microswitch which is used with an integrating circuit to ensure that V_1 fires once only each time the switch is operated. V_1 provides the drive pulse for V_2 and the negative triggering pulses for the oscilloscope and shock-excited calibration circuits. To compensate for the time delay in operation of the oscilloscope and calibration circuits a variable delay is incorporated in the grid circuit of V_2; this tube driving V_3 and V_4. The operation of V_5 and V_6 from a single trigger tube was unsatisfactory. V_6, if triggered via an integrating circuit, failed to fire properly because of its low grid to cathode impedance and caused considerable "jitter" in the output pulse length. The grid circuit of V_3 controls the firing point of V_3 and hence of V_6 and so acts as a control of pulse length. The output pulse length is continuously variable from 0·05 to 80μsec and by adjustment of the timing and other relevant circuits pulse lengths of several hundred microseconds may be obtained.

When a liquid test sample breaks down the voltage across the spark gap collapses and if a suitable differentiating circuit is connected in parallel with the spark gap a positive pulse results. The derivation of such a pulse from a resistance in series with the spark gap is impracticable because of displacement currents. In the circuit shown the positive pulse triggers the fast monostable multivibrator V_8 and V_9 in which V_8 is normally conducting and V_9 is cut off. The impedance matching tube V_{10} fires, this triggers V_6 and hence terminates the output pulse. In the absence of a breakdown the trailing

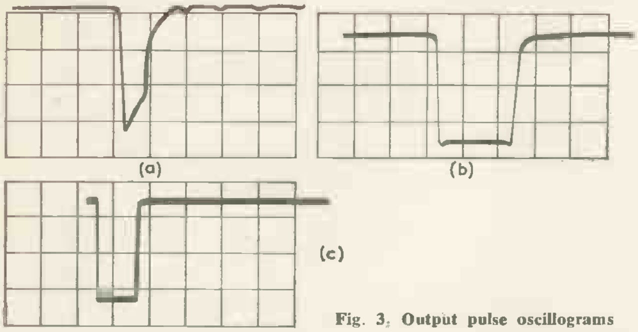


Fig. 3. Output pulse oscillograms

(a) *The shortest pulse length obtained. This figure illustrates the fastest rise and decay times possible with this circuit. Scale approximately 0·1μsec/division. 15kV pulse.*

(b) *A 15kV 1μsec pulse. Scale approximately 0·5μsec/division.*

(c) *A 15kV 10μsec pulse. Scale approximately 10μsec/division.*

edge of the high voltage pulse operates the shut-off circuit, but this is inconsequential. With this circuit the high voltage pulse was shut off 0·6μsec after a breakdown occurred; 0·5μsec of this was due to the ionization time of V_6.

All of the D.C. supplies were derived by conventional circuits from a constant voltage A.C. supply. A time delay switch was incorporated which prevented the application of H.T. until the tubes reached their ambient temperatures. 5D21's frequently flash over internally and to protect the power supply an overload relay was included in the 20kV supply.

The Sawtooth Pulse Generator

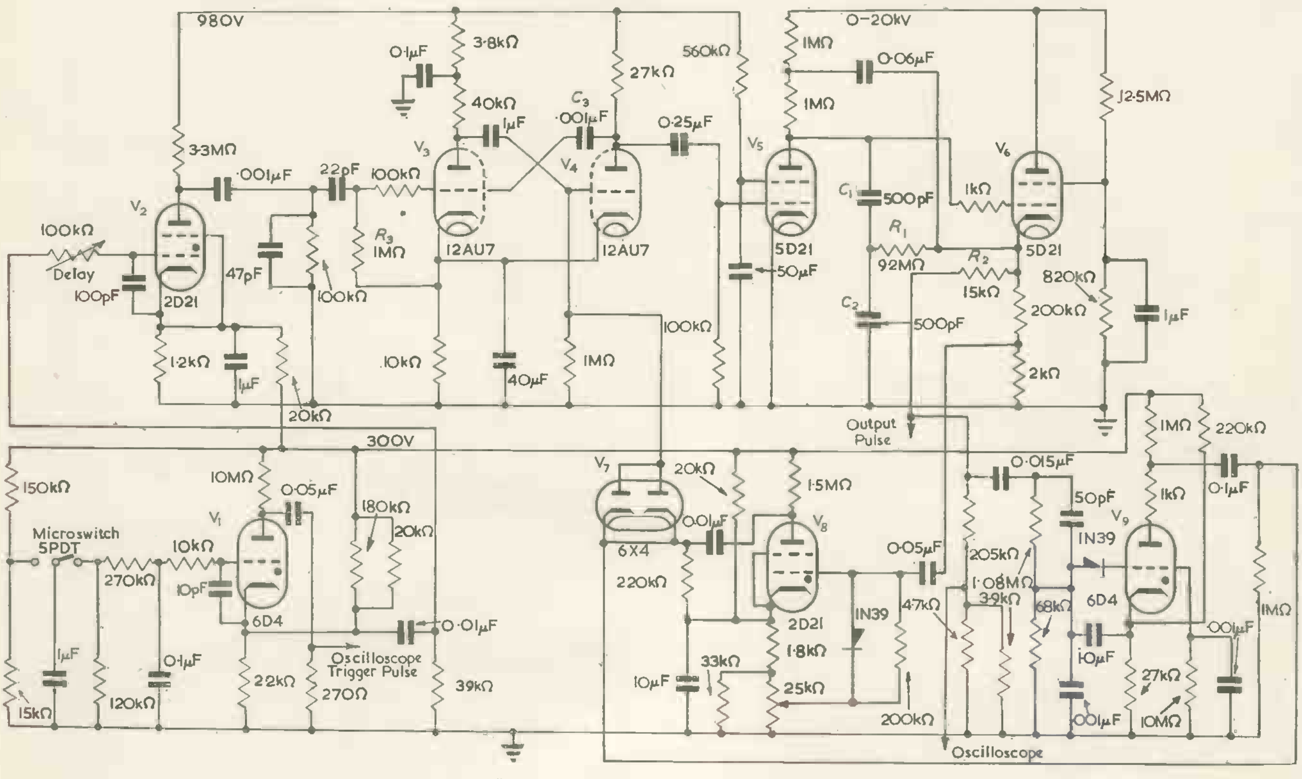
Positive, linearly rising voltage pulses with rise times to 18kV of from 15 microseconds to 10 milliseconds have been obtained with the type of circuit shown in Fig. 4. The pulse generator consists of three distinct sections, the pulse forming circuit, a driver stage and the automatic shut-off circuits.

The sawtooth pulse is produced by a bootstrap circuit in which V_5 is the switch tube and V_6 is the cathode-follower. The amplitude of the linear portion of the output pulse increases as the pre-pulse, anode-cathode voltage drop in V_6 increases. The mean power dissipation of the 5D21 is relatively small, but by tying the screen to the resistance potentiometer shown in Fig. 4 the tube was held within its rated limit. At the same time almost the maximum H.T. was maintained across the tube without introducing distortion or appreciably reducing the gain of the stage of just less than unity. With this circuit the pre-pulse currents through V_5 and V_6 were 9mA and 3·7mA respectively. The output pulse was taken from the cathode of V_6 via the series resistor R_2, of 15kΩ. If breakdown of the dielectric sample occurred, R_2 limited the heavy current which would otherwise have flown through V_6 and the sample in the interval before the automatic shut-off circuit came into operation. To test for non-linearity of the output waveform the pulse was differentiated and the resultant rectangular waveform displayed directly on the Y_1 trace and after inversion on the Y_2 trace, of a double beam oscilloscope as shown in Fig. 5(b). Observation of the parallelism of the "tops" of the two pulses was a sensitive test for non-linearity of the sawtooth and this was corrected for by adjustment of R_1* (Fig. 4). C_1 and C_2 controlled the rise time of the output pulse and each time C_1 and C_2 were changed R_1 was readjusted. The pulse amplitude was controlled by varying the drive pulse duration with the H.T. set at a fixed voltage. Variation of the H.T. voltage as with the rectangular pulse generator varies the rate of rise of the pulse as well as its amplitude, and this is not always acceptable.

This last fact made it desirable to derive a pulse amplitude limiting signal from the output pulse and not to rely on a timing circuit in the driver stage. The power required from the driver stage to cut off the 5D21 is small. The two halves of a 12AU7, V_3 and V_4, connected as a monostable multivibrator proved satisfactory, V_1 and V_2 operate

* *See Reference 5.*

Fig. 4. 18kV sawtooth voltage generator



as did V_1 and V_2 of the rectangular pulse generator and they produce an oscilloscope locking pulse and a single, negative tripping pulse for the multivibrator. When the multivibrator is triggered, V_5 receives a negative drive pulse of 900 volts amplitude from the anode of tube V_4 which is normally cut off. A fraction of the resultant high-voltage sawtooth appearing on the cathode of V_6 is fed to the grid of V_8, the amplitude discriminator tube. V_8, which is normally cut off, fires and a negative pulse reaches the grid of V_4 via the diode V_7. V_4 is hence cut off, V_5 becomes conducting and the output sawtooth is terminated. Control of the grid bias of V_8 and hence of

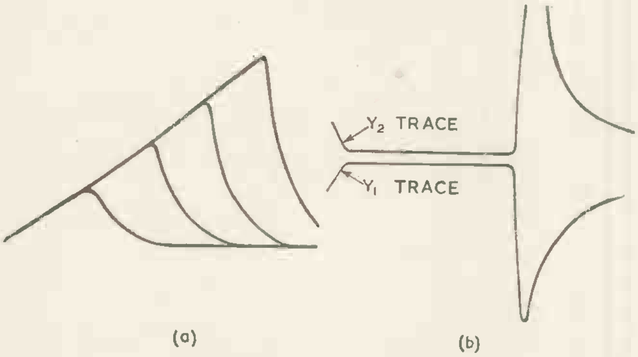


Fig. 5. Output pulse oscillogram

(a) *Typical sawtooth pulses obtained for different settings of the discriminator control.*

(b) *Normal and "inverted" derivative of a sawtooth pulse displayed on a double beam oscilloscope to facilitate detection of non-linearity of the sawtooth.*

its firing point is sufficiently accurate as a sawtooth-pulse amplitude control. Should V_8 fail to fire because of too great a grid bias or insufficient sawtooth triggering voltage, the monostable multivibrator returns to the stable position after a time determined by C_3 and R_3.

When breakdown of a test sample takes place the voltage across the sample drops to zero. This fast, negative-going "edge" is detected by V_9, the automatic shut-off tube, which fires and terminates the negative pulse applied to the grid of V_5. This leads to a reduction of the grid potential of V_6 and considerably reduces the heavy discharge current which flows through V_6 on breakdown.

All power supplies were derived by standard circuits from a constant voltage A.C. line. Voltage regulated supplies were not considered necessary. The filament supply for V_6 was taken from a low insulation-voltage filament transformer immersed in a beaker of oil.

The circuit shown in Fig 4 produced 18kV pulses with rise times of 1 millisecond. Alternative rise times of from 15 microseconds to 10 milliseconds have been obtained by adjustment of C_1 and C_2 and R_1. The multivibrator and associated amplitude and automatic shutoff circuits operated in negligible time ($< 1\mu$sec) compared with the decay time to zero of the positive sawtooth which is controlled by the discharge time of C_1 and C_2.

While it may be possible to effect some economy in components in both pulse generators this has not been attempted as both generators have given satisfactory service. Oscillographs representative of their operation are shown in Figs. 3 and 5.

Acknowledgments

The author is indebted to the National Research Council of Canada for the award of a Post-doctorate Fellowship during the tenure of which these circuits were constructed.

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SWISS ORDER V.O.R. BEACON

The Swiss civil aviation authorities have recently ordered a very high frequency omni-directional radio range (V.O.R.) beacon from Marconi's Wireless Telegraph Co., Ltd. This will probably be the first production model manufactured by any British company.

The V.O.R. beacon will be delivered early in the year, but the Swiss authorities have not yet announced at what airfield they intend to erect it. Because of the difficult conditions of wave propagation in Switzerland, careful V.O.R. tests have to be made with this first equipment.

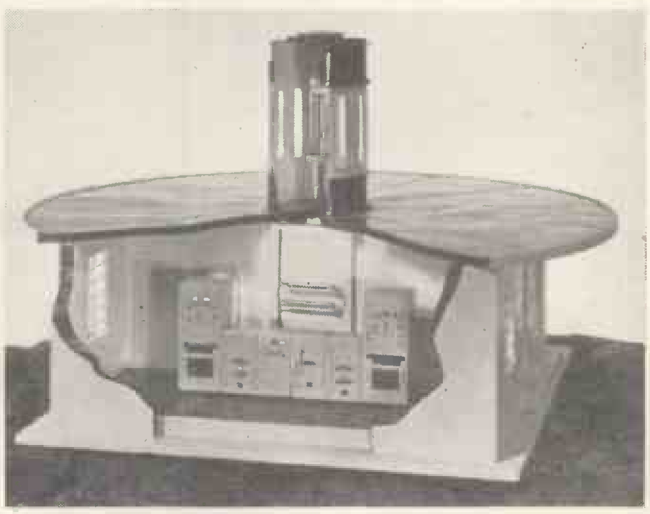
Beacons are already in operation at a few European centres, and in this country. A Marconi prototype model was installed at a base in the West of England where it has been tested successfully by American authorities.

V.O.R. is a system, standardized by ICAO (International Civil Aviation Organization), which enables aircraft within range of the beacon to take an immediate bearing without having to contact the ground and which will also tell the pilot when he has reached a selected bearing and if he deviates from it, despite such hazards as wind drift.

The range of these beacons is similar to that obtained on V.H.F. communications, up to or just beyond line of sight. In other words, the greater the height the greater the range. As an example, ranges of over 200 miles have been obtained in aircraft flying at 36 000 to 38 000ft.

Marconi's have developed a unique fixed (non-rotating) aerial with low octantal error. It has the great advantage of being completely rigid in high winds and is weather proof, and also permits the fixing of D.M.E. (distance measure equipment) or other aerials and obstruction lights with minimum difficulty.

A scale model of a Marconi very high frequency omni-directional radio (V.O.R.) beacon.



BOOK REVIEWS

The Instrument Manual

628 pp. 300 figs. Demy 4vo. United Trade Press Ltd. 2nd Edition. 1953. Price 84s.

THE ambitious title conjures up the idea of a large and comprehensive volume dealing with almost all types of measuring instruments and their applications. A close investigation shows that the work very largely achieves this aim: it is comprehensive and thorough, though note must be taken of the introductory statement that the task of preparation of this second edition was becoming more what was to be left out than what was to be included. A detailed list of contents would be too long for inclusion here, but a brief summary is that about 600 quarto pages include measuring instruments for nearly all mechanical, physical, chemical and electrical quantities. Some of the range of physical quantities includes optical; surface texture; quality; time and speed; weight; pressure and vacuum; liquid level, volume and flow; specific gravity; temperature; viscosity; humidity and instruments for meteorological, aeronautical and survey work, etc. Electrical and electronic engineers will at first perhaps be surprised to find that of twenty-five sections dealing with specific classes of instruments only three are obviously in this category. They are: Electrical measuring instruments; General electronic instruments and Nuclear instruments. But on examination of the text in detail it will be seen that electrical and electronic methods are included in many other sections, such as the resistance strain gauge (physical and mechanical properties), the flying-spot microscope (optical instruments), photocells (photometry and spectroscopy), flaw detection by X and γ rays and by ultrasonics (mechanical properties), electronic counters (time and speed), electrical and ionization gauges (pressure and vacuum), electrical and electronic thickness meters (engineering precision instruments and gauges) and many others throughout the work. The treatment of the sections is excellent, and combines theory and practice. Generally definitions are given first, correctly based on British Standard practice, so that no ambiguity in meaning should arise. Following the main text is a bibliography quoting appropriate British Standards, H.M.S.O. publications, foreign Standards, etc., then a list of text-books and articles. The choice of the last always requires considerable care to balance the inclusion of relevant articles with the exclusion of those of little interest. Finally, a Buyer's Guide concludes each section and should prove of great assistance; it is not based on the useful advertising pages which comprise some hundred pages additional to the text pages. The section on electrical instruments might be improved by the deletion of some detailed references to sources of error, etc., which are primarily of interest to the designer, which would allow more space to be devoted to modern designs which are not given. While the D.C. 'Tong-Test' is mentioned, the more widely-used transformer-operated type of hand-portable A.C. ammeter is not. As with nearly all text-books, reference to induction instruments gives only the obsolete shaded-pole type and omits the present forms of British first-grade instruments which have been available for over twenty years. Many obsolete forms of integrating house service meter are described but are not quoted as such, and some of the specialized grid-metering instruments might well have been included. In the section on electronic instruments the main forms included are the C.R.O., electron-microscope, valve voltmeter, vibration generator and frequency standards, but as has been stated many others are covered under the respective sections on specific measurements. Nucleonic instruments are well covered in the space available. There are a few lapses in terminology. It is not always easy to decide if "capacity" refers to physical capacity, or to electrical capacitance, which latter correct term is not used. Similarly, the "condenser" in the optical and liquid sections could well be discriminated in the electrical sense by the use of the standard term "capacitor" to avoid confusion. In one diagram ω is wrongly used for λ for wavelength, but in general B.S. symbols are used. Among the few mistakes or printers' errors noted was torque, a product, expressed as gm/cm. In some cases a very brief description of an instrument given in a hundred words or so could well have been amplified by a line drawing or a diagram, but of course such additions would increase the size and cost other than by the deletion of some of the more abstruse design points.

E. H. W. BANNER.

Magnetic Amplifiers

By George M. Ettinger. 88 pp., 50 figs. Crown 8vo. Methuen & Co., Ltd. 1953. Price 6s. 6d.

THE author has been fortunate in having the opportunity to study his subject first in London and now in the U.S.A., and has avoided the national bias or limited viewpoint that mars certain texts in the field of electronics. Bearing in mind the introductory role for which their Monograph series is designed, his elementary treatment of magnetic amplifiers is well balanced and should serve its purpose for anyone who is determined to go on and read the subject seriously in the original literature. Over a thousand papers on magnetic amplifiers have been published, and the author has reasonably restricted his references to 82, and given a very useful short "reading list" of 13 items, together with a short survey of other ferro-magnetic devices.

A second purpose is claimed for the book, as a guide to the practising engineer who wishes to select the most suitable device for a specific application. In Chapter VI there is a useful discussion of the factors limiting the performance, and the typical design data is more than one might hope to find in a book of only 80 pages. The treatment falls short of what could be achieved in a small compass in comparing the various classes of magnetic amplifier, the essential data being spread through several chapters of text.

One would like to see the compact presentation of such relative data, classified and in tabular form, such as appears in the discussion of Gale and Atkinson's paper (included in the reading list). It is a real contribution to the art of a subject such as this, if an author can reduce the essentials to one page, which the user can retain as a visual image. One misses any reference to the very useful "peak-height" or "flux-gate" type of sensitive amplifier.

The author deals with a number of applications of ferro-magnetis devices in computor systems, and here a tabular comparison of some of the signal parameters would be useful. In a future edition he might also consider including data on that important class of magnetic energy-amplifiers, the electro-mechanical transducers, for an efficient input transducer or outdoor motor can greatly reduce the amplification requirements in between if the servo system is thought about as a whole.

D. L. JOHNSTON.

Radio Engineering

By E. K. Sandeman. 613 pp., 207 figs. Demy 8vo. Vol. II. 2nd Edition. Chapman & Hall Ltd. 1953. Price 55s.

IN spite of the author's claim that this second edition is a thoroughly revised version of the first edition, it is, with the exception of two new sections on "Noise Factor" and "Transmission Line Filters" virtually a reprint of the first edition, including its errors.

The book contains nine chapters, nine appendices and a voluminous bibliography. The level of treatment varies widely; the first four chapters and Chapter XXII contains information which may be of value to operational personnel, while the remainder are written for the more advanced engineer.

The first Chapter (XVII) is short and deals with balanced and unbalanced circuits. Fig. 1/XVII:5f has not been corrected. Interference and noise form the subjects of the next chapter, and in the first two sections their sources are discussed and suggestions given for minimizing their effects. The remainder of this 47 page chapter deals with signal to noise ratio and noise factor and the available methods of improving the former, such as single sideband operation and directional receiving aerials. While useful information is given on these subjects, so wide a field is covered that readers who are seriously interested in these matters will probably wish to refer to works dealing specifically with them.

There follows a 60 page chapter on radio receivers, which includes an essentially descriptive treatment of detectors, T.R.F. and superheterodyne receivers. Brief references are made to super-regenerative and sideband receivers and descriptions with performance figures are given in tabular form for six commercial receivers. The procedure is described for the measurement of receiver performance, according to the R.M.A. recommendations, and the chapter concludes with brief notes on the principle of diversity reception and on fault finding.

"Measuring Equipment" is the title of Chapter XX in which 13 pages are devoted to detailed consideration of the peak voltmeter. Other equipment described includes transmission measuring equipment, ohmmeters, the Megger, various types of bridges for use at D.C., A.F., and R.F., the C.R.O. for phase and frequency measurement and as a modulation monitor, harmonic analyser, modulation meter, peak programme meter, receiver alignment equipment and absorption wavemeter.

Chapter XXI shows an engineering approach to and treatment of equalizer design and with the 13 generalized curves which are given, includes the majority of networks in general use.

The next chapter is very short and contains a descriptive treatment on level range compression and expansion and level limitation. It includes an account of a limiter circuit in use in the BBC and of the actual procedure employed in lining up the programme chain at a transmitter.

The chapter on feedback is not well presented. As a result of too much compression of the material covered, it would be difficult to follow by a reader new to the subject and yet the use of terms not well-established would inevitably confuse those with some previous knowledge. Some reference might well have been made in the text and bibliography to Bode's outstanding contribution to network analysis and feedback amplifier design.

The seventy page chapter on steady stage passive network theory employs methods which avoid the use of hyperbolic functions. While this may prove an initial simplification to the student, it it may well later become an embarrassment where the classical treatment is used. The application of matrix algebra to network problems is well illustrated by tables and examples.

The descriptive sections of the last chapter (XXV) deal with "easy methods of design and of predicting the performance of basic filter types". It contains information on characteristics and design formulae in tabulated form and sections on filter refinements. The new section on "Transmission Line Filters" is poor and it is difficult to appreciate to whom it would prove of value. It includes stubs, re-entrant filters, quarter wave networks and loaded lines. The expression "leak loading" is particularly misleading since pure shunt susceptance is meant and the treatment is based entirely on this.

The subjects covered in the nine appendices are mathematical and physical formulae; tuning problems; impedance matching in long wave aerial circuits; conditions in modulated amplifiers; calibration of aerial power distribution diagrams; Fourier analysis; representation of complex quantities, and coupling circuits as band-pass filters.

The extensive classified bibliography is probably the most valuable portion of the work, although there is only one reference to it in the text. While the author admittedly has a wide experience of radio engineering, it cannot be said that in this compendium he has added significantly to information already available, or that he has been wise in covering so wide a field.

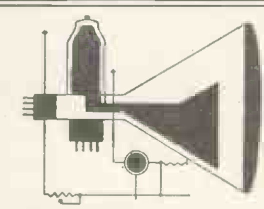
L. G. DIVE

Nuclear Physics

By W. Heisenberg. 225 pp., 40 figs. Crown 8vo. Methuen & Co., Ltd. 1953. Price 12s. 6d.

THIS book begins with a short and interesting history of the views about atoms in antiquity and also of the development of atomic theory till the close of the nineteenth century. The next chapter is devoted to molecules and atoms. The main subject of the book includes radioactivity, the binding energy of nuclei, nuclear structure, artificially induced nuclear transmutations and the methods of observation and producing nuclear transmutations. The work concludes with some account of the practical applications of nuclear physics.

The author manages to convey a great deal in simple language and without mathematics. He does not claim to have brought the subject quite up to date, but regards his book as a tentative and provisional introduction to a subject which is being continuously expanded.



LETTERS TO THE EDITOR

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

Noise-Free Instrument Cable

DEAR SIR,—In your September, 1953, issue, there appeared a letter by Mr. D. Pollard commenting on an article entitled "Noise-free Instrument Cable," by M. Lorant (*Electronic Engineering*, July, 1953). This article was based on a release by the National Bureau of Standards describing a development more fully covered in my paper entitled "Electrical Noise from Instrument Cables Subjected to Shock and Vibration," *J. Appl. Phys. (U.S.)*, *23*, 647, (1952). Previous work by others on low-noise cables is fully acknowledged in that paper, but it appeared at that time that the detailed explanation of the mechanism of noise generation in cables had not previously been formulated. To date such a formulation has not come to my attention.

My main purpose in writing this note is to draw attention to the fact that there are two fundamental differences between the cable construction which I proposed and that used until very recently by commercial manufacturers of low-noise cables. I have pointed out in my publications that this latter construction involves "mechanical considerations" which "impose limitations on the smallness" (and flexibility) of the usual low-noise cables. However, the noise can be eliminated by providing that there be absolutely no separation between the conductive coatings and the surfaces of the dielectric. This is achieved by integrally bonding the coatings to the dielectric, as against the usual sprayed or wrapped conductive coatings. Furthermore, the conductors themselves may fit quite loosely with respect to these conductive coatings, so that the construction of the cable need not be bulky or rigid. Typical laboratory-made samples of this cable used with piezo-electric accelerometers and for other high-impedance applications are 0·080in in diameter. Their flexibility is such that they may readily be wrapped around a wire 0·030in in diameter without either generating noise or suffering other ill-effects, permanent or temporary.

I am glad to acknowledge the limitation on smallness imposed by considerations of cable capacitance, which is also pointed out in my above-referenced publication.

Yours faithfully,

THOMAS A. PERLS.
National Bureau of Standards,
Washington, D.C., U.S.A.

Characteristics of the Temperature-Limited Diode Type 29C1

DEAR SIR,—I should like to make the following comments on the article by Dr. Benson and Mr. Seaman on the Mazda 29C1 saturated diode[1].

The life test figures are given in terms of the percentage change in emission current at a given filament voltage. However, the diode is normally used in stabilizer or measurement circuits, and here we are concerned with the percentage change in filament voltage required to maintain a constant emission. This latter figure gives a more realistic measure of the diode stability and permits direct comparison to be made with related reference-voltage devices, such as neon-tubes and non-linear bridges. Where the evaporation is small the equivalent filament-change is found by dividing the percentage fall in emission current (at constant filament voltage) by a constant which may be taken at 8·5. For 1 000 hours (42 days) the emission drop was 15-22 per cent (Fig. 4 & 5, ref. 1) giving an estimated increase in filament voltage, at constant emission, of 1·8-2·6 per cent. In most practical applications of the saturated diode the filament is heated from an A.C. supply. The present tests were made with the filament run from D.C. and it is important to note that the life on D.C. is not necessarily the same as on A.C. This is because tungsten filaments which have been run on D.C. exhibit a step-like surface structure, whereas A.C. heated filaments are uniformly smooth except for a restricted region near the supports where, in any case, the filament is too cool to give significant emission. The effect is fully discussed by Johnson[2] who suggests that it is caused by surface migration of tungsten ions.

The results may be compared with stability measurements made in this laboratory[3]. Here, two 29C1 diodes run on A.C. showed an equivalent change in filament voltage of 1·3 per cent over 1 000 hours at 5mA emission; another diode, the Ferranti GRD6, gave an equivalent change of 2 per cent. In view of the somewhat different test conditions these results check quite well with those of Messrs. Benson and Seaman.

The test figures given[1] only deal with the filament *voltage*, which is disappointing, as it has been found[3] that the drift of the filament *current* is very much smaller than that of the filament voltage. The improvement is quite remarkable, especially at low emissions where a 1 000-hour stability of 1 part in 500 was obtained on each of the three diodes measured. This suggests that, for best stability, the emission current should be small and the diode filament should be run under approximately constant-current conditions, i.e., with a fairly large ohmic resistance in series.

Yours faithfully,

V. H. ATTREE.
Fluid Motion Laboratory,
University of Manchester.

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The Author replies:—

DEAR SIR,— In reply to Mr. Attree we should first like to justify the procedure adopted in our examinations of the valves.

The life tests were carried out with the filament voltages maintained constant since this is the logical method and is very convenient. The degree of constancy of voltage that can be achieved depends on the source of supply but by using secondary cells of large capacity a good degree of constancy can be maintained, especially during periods when the tests cannot be monitored, such as at weekends. To maintain constant emission some form of automatic control of the filament voltage would be necessary as well as stabilized H.T. supply. In our life tests the anode voltages were allowed to vary according to the voltage of the mains supply from which they were obtained. As Mr. Attree points out, if the relationship between filament voltage and emission current is known, then changes in one can be expressed as equivalent changes in the other.

We agree with Mr. Attree that the life of such a diode when the filament is operated from a D.C. supply is not necessarily the same as when it is operated from an A.C. supply. Whether different results would have been obtained by us had A.C. heating been used is, however, doubtful because, as Mr. Attree states, his life-test figures agree reasonably well with ours and he used A.C. heating. Incidentally, Johnson's paper deals with lamp filaments only and field conditions inside a saturated diode will not be the same as in a lamp. It is thought that the anode voltage of the diode may play some part in the deterioration of the filament; it is interesting to note that the equivalent decrease of emission in Mr. Attree's tests is less than in ours. The anode voltage in Mr. Attree's case was 50V, while in ours it was 90V for three valves and 200V for the other one.

We have recently considered carrying out life tests on these diodes with A.C. heating. A difficulty which arises is that of measuring the heater voltage accurately. The use of an A.C. potentiometer is one solution but such an instrument presents certain problems with regard to power supplies. The other alternative is to run the tests on A.C. but switch over to D.C. for taking readings as was done by Mr. Attree.

Although, as pointed out, A.C. heating of the diode would be used in most applications, there are others, such as low-voltage stabilizers, which we are developing at present, where D.C. heating is employed.

Mr. Attree's remarks regarding operation at low emissions are interesting. In any practical circuit an additional variable resistor would need to be inserted in the filament circuit to allow for ageing and also for the change in characteristics should the diode be replaced. The value of the resistance that could be used would depend on the permissible losses, since the filament current is of the order of 0·8A.

Yours faithfully,

F. A. BENSON.
M. S. SEAMAN.
Department of Electrical Engineering,
University of Sheffield.

Short News Items

The Scientific Advisory Board of the Journal *Research*, under the chairmanship of Lord Waverley, is sponsoring a competition for an essay describing a recent scientific advance and its applications. The choice of subject is left entirely to competitors. The essays which should be about 3 500 words in length, must be received at the offices of Research, 88 Kingsway, London, W.C.2, on or before 30 June, 1954. Whatever the subject chosen, it may be assumed that the argument is to be placed before a board of directors or a committee, some of them with technical training and knowledge, but without specialist knowledge of the subject matter.

Prizes of £100 and £50 will, at the discretion of the judges, be awarded to those entries adjudged to be best and second best. In order to encourage the younger scientist, it has been decided to offer an additional sum of £50 to the best entry from a competitor under 30 years of age.

The *Sunday Times* has agreed to cooperate in the competition, and is offering two extra prizes of £100 and £50 for entries suitable for publication in a general newspaper and relating to one of the following general subjects: Application of Atomic Energy; Aerodynamics; Conservation or Utilization of Fuel; Use of Electronics in Business Efficiency.

Further details mav be obtained from of the offices of *Research*.

The Radar Association has formed a new division, membership of which is open to those who are, or have been engaged on radar, either within industry or H.M. Forces. The Association was originally formed as a social club for ex-wartime members of the Royal Air Force who had been engaged as radar mechanics and operators. As a result of the formation of the new division, a series of lectures has been arranged as part of the 1954 winter programme. Details and information may be obtained from The Radar Association, 83 Portland Place, London, W.1.

Mr. H. G. Nelson, Deputy Managing Director of The English Electric Company Limited and Director of a number of other companies has been asked to join the Board of the Canadian Marconi Company. English Electric recently acquired controlling interest in the Canadian Marconi Company and Mr. Nelson's appointment will assist in creating a close liaison between these two organizations.

The Institute of Radio Engineers, New York, have elected 76 radio engineers and scientists for Fellow Awards for 1954. The grade of Fellow is the highest membership grade offered by the Institute. The two recipients from this country are Dr. D. C. Espley, Chief Engineer (Telecommunications), Research Laboratories of the General Electric Company, and Mr. A. W. Montgomery, Joint Manager, Standard Telephones and Cables Ltd.

The British Thomson-Houston Co. Ltd, announce that Mr. G. W. P. Page has retired after 30 years' service with the company. Mr. P. C. Spencer is now acting as Press Liaison Officer, vice Mr. Page.

The Royal Society of Arts will reach its bicentenary in March this year, and the Council is arranging, in addition to various forms of celebration which will commemorate the Society's past achievements, a competition which will focus attention upon the future. The Society accordingly offers prizes totalling £500, the largest being £250, for conceptions of life in this planet in the year 2 000, and forecasts (in visual or written form) are invited of the future developments which may be looked for in some particular aspect of life related to Arts, Manufactures and Commerce, the field of the Society as defined in its full and original title. The chief criterion in assessing the entries will be originality. Terms and conditions relating to the competition, together with registration forms may be obtained from the Secretary, Royal Society of Arts, John Adam Street, London, W C.2. Registration forms must be completed and returned together with an entry fee of 1s. by 15 February, and the actual competitive material submitted by 30 June.

The Ninth Annual Electronics Exhibition, organized by the North-West Branch of the Institution of Electronics, will be held at the College of Technology, Manchester, during the period 14-20 July, 1954. The Exhibition will include a Scientific and Industrial Research Section, a Commercial Section —displaying manufacturers products, and a programme of lectures on subjects allied to electronics.

W. Edwards & Co. (London) Ltd, have now completed their move to new premises in Crawley and all future communications should be addressed to Manor Royal, Crawley, Sussex. The Scottish branch is still at 44 West George Street, Glasgow.

The Ministry of Supply announce that the following courses have been arranged by the Harwell Isotope School. Course 22. 1-26 March (latest date of application 1 February); Course 23. 30 August-24 September (latest date of application 2 August); Course 24. 4-29 October (latest date of application 6 September); Course 25. 15 November-10 December (latest date of application 18 October). Lectures and instruction are given by members of the Atomic Energy Research Establishment Staff. The school building, which is equipped with the latest apparatus, is just outside the security fence at Harwell. Students pay a fee of £40 for the course, and living accommodation in the neighbourhood can be arranged at moderate charges. Applications should be sent to the Isotope School, AERE, Harwell, near Didcot, Berks.

Electrical Instrument Co. (Hillington) Ltd, Boswell Square, Hillington, Glasgow, S.W.2, are manufacturing deflexional frequency meters up to 5 000 c/s ± 20 per cent of any given frequency, for instance, 800-1 200c/s or 4 000-6 000c/s. The current consumption of these instruments is in the order of 10μA only.

The Radio Industry Council announces that the Trade and Navigation Accounts for October, 1953, showed the highest monthly total of British radio exports ever recorded. The figure is £2 470 000, the previous highest being £2 310 000 in July, 1952.

Hifi Ltd, manufacturers of precision instruments, electronic controls, chokes, transformers and high fidelity sound reproducing equipment, have changed their address to Derry Works, Derry Street, Brierley Hill, Staffs. Telephone Brierley Hill 7604.

Decca Radar Ltd, announce that, following the introduction of the new Decca 45 Marine Radar in September last, the total number of ships for which the company's radar has been ordered now exceeds 3 000. This achievement has been accomplished in less than four years and ships of every class and tonnage are represented in this total.

MEETINGS THIS MONTH

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 6 January. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Gower Street, London, W.C.1.
Lecture: Engineering Design of V.H.F. Multi-Channel Telephone Equipment.
By: W. T. Brown.

Scottish Section

Date: 7 January. Time: 7 p.m.
Held at: The Department of Natural Philosophy, The University, Edinburgh.
Lecture: Radio Astronomy.
By: H. Seddon.

North Western Section

Date: 7 January. Time: 7 p.m.
Held at: The College of Technology, Manchester.
Lecture: The Manchester University Computor.
By: D. B. G. Edwards.

Merseyside Section

Date: 7 January. Time: 7 p.m.
Held at: The Electricity Service Centre, Whitechapel, Liverpool, 1.
Lecture: Interlacing Problems in Television Receivers.
By: G. N. Patchett.

West Midlands Section

Date: 26 January. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: Microwave Measuring Instruments.
By: P. M. Ratcliffe.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Date: 7 January.
Parsons Memorial Lecture: Continuity of Electricity Supply.
By: H. Leyburn.

Measurements Section

Date: 5 January.
Lectures: A Scaling Unit Employing Multi-Electrode Cold-Cathode Tubes, and A Sensitive Pulse Trigger Circuit with a Stable Threshold.
By: K. Kandiah; and
The Development of a Neutron Spectrometer for the Intermediate Energies.
By: F. S. Goulding, J. C. Hammerton, M. G. Kelliher, A. W. Merrison and E. R. Wiblin.
Date: 19 January
Discussion: High-Sensitivity Wattmeters.
Opened by: A. H. M. Arnold.

Radio Section

Date: 13 January.
Lecture: A Single-Sideband Controlled Carrier System for Aircraft Communication.
By: G. W. Barnes.
Date: 25 January.
Discussion: Should Sound Broadcasting of the Future be entirely on the V.H.F. Band?
Opened by: A. J. Biggs.

Informal Meeting

Date: 18 January.
Discussion: The Rôle of the Consulting Engineer.
Opened by: T. G. N. Haldane.

Education Discussion Circle

Date: 20 January.
Discussion: The Teaching of Magnetic Materials.
Opened by: Professor F. Brailsford.

East Midland Centre

Date: 19 January. Time: 6.30 p.m.
Held at: Loughborough College.
Lecture: Technical Colleges and Education for the Electrical Industry.
By: H. L. Haslegrave.

Mersey and North Wales Centre

Date: 4 January. Time: 6.30 p.m.
Held at: The Liverpool Royal Institution, Colquitt Street, Liverpool.
Lecture: Electronic Telephone Exchanges.
By: T. H. Flowers.
Date: 18 January. (Time and place as above.)
Lecture: Some Researches on Circuit-Breaking by means of High-Voltage Circuit-Breakers, with special reference to Current-Chopping.
By: A. F. B. Young.

North-Eastern Radio and Measurements Group

Date: 18 January. Time: 6.15 p.m.
Held at: King's College, Newcastle-upon-Tyne.
Discussion: Will Transistors Oust Receiving Valves?
Opened by: E. H. Cooke-Yarborough.

North Midland Centre

Date: 5 January. Time: 6.30 p.m.
Held at: Offices of The British Electricity Authority, Yorkshire Division, 1 Whitehall Road, Leeds.
Lecture: Voltage Transformers and Current Transformers associated with Switchgear.
By: W. Gray and A. Wright.
Date: 14 January. Time: 7.15 p.m.
Held at: Offices of the Yorkshire Electricity Board, Ferensway, Hull.
Lecture: Inherent Current, Voltage and Speed Control in Dynamo-Electric Machinery.
By: J. C. Macfarlane, J. W. Macfarlane and W. I. Macfarlane.

North Midland Utilization Group

Date: 19 January. Time: 6.30 p.m.
Held at: The British Electricity Authority, Yorkshire Division, 1 Whitehall Road, Leeds.
Lecture: The Royal Festival Hall: Electrical Installation.
By: J. G. Hunter.

Sheffield Sub-Centre

Date: 20 January. Time: 6.30 p.m.
Held at: The Grand Hotel, Sheffield.
Lecture: Colour Television.
By: L. C. Jesty.

North-Western Centre

Date: 5 January. Time: 6.15 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: The Co-ordination of Insulation of High-Voltage Electrical Installations.
By: J. S. Cliff.

North-Western Measurements Group

Date: 26 January. (Time and place as above.)
Lecture: Alternating-Current Instrument Testing Equipment.
By: A. H. M. Arnold.

North-Western Radio Group

Date: 6 January. Time: 6.30 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: Connexion between Closed-Loop Transient Response and Open-Loop Frequency Response.
By: J. C. West and J. Potts.

North-Western Supply Group

Date: 12 January. Time: 6 15 p.m.
(Place as above.)
Lecture: Transformer-Analogue Network Analysers.
By: M. W. Humphrey Davies.

South-East Scotland Sub-Centre

Date: 19 January. Time: 7 p.m.
Held at: The Heriot-Watt College, Edinburgh.
Lecture: Design Features of Certain British Power Stations.
By: S. D. Whetman.

South-West Scotland Sub-Centre

Date: 20 January Time: 7 p.m.
Repeat of South-East Sub-Centre lecture.

South Midland Centre

Date: 14 January. Time: 7.15 p.m.
Held at: The Winter Gardens Restaurant, Malvern.
Lecture: Transmission of Pictures by Radio.
By: A. W. Cole.

South Midland Radio Group

Date: 25 January. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: A Method for the Synthesis of Speech.
By: W. Lawrence.

South Midland Supply and Utilization Group

Date: 11 January. Time: 6 p.m.
Held at: The Imperial Hotel, Birmingham.
Lecture: An Assessment of the Impregnated Pressure Cable.
By: L. G. Brazier, D. T. Hollingsworth and A. L. Williams.

Southern Centre

Date: 6 January. Time: 6.30 p.m.
Held at: The S.E.E.B. Headquarters, Hove.
Lecture: Operation of Power Transformers.
By: E. T. Norris.
Date: 13 January. Time: 7.30 p.m.
Held at: The R.A.E. Technical College, Farnborough.
Lecture: Electronic Speed Control of Motors.
By: J. C. Rankin.
Date: 20 January. Time: 6.30 p.m.
Held at: The Technical College, Southampton.
Lecture: A.C. on Ships.
By: B. C. Pyle.
Date: 22 January. Time: 6.30 p.m.
Held at: The Technical College, Weymouth.
Lecture: Receiving Aerials for British Television.
By: F. R. W. Strafford.

South-Western Sub-Centre

Date: 27 January. Time: 3 p.m.
Held at: The Electric Hall, Torquay.
Lecture: Electronic Telephone Exchanges.
By: T. H. Flowers.

Ipswich District

Date: 4 January. Time: 6.30 p.m.
Held at: The Crown and Anchor Hotel, Ipswich.
Lecture: Modern Developments in Atomic Energy.
By: T. E. Allibone.

Reading District

Date: 25 January. Time: 7.15 p.m.
Held at: The George Hotel, Reading.
Lecture: Training in Electronic Fire Control in R.E.M.E.
By: R. A. Middleton.

THE INSTITUTE OF NAVIGATION

Date: 15 January. Time: 5 p.m.
Held at: The Royal Geographical Society, 1 Kensington Gore, London, S.W.7.
Lecture: Methods of Air and Surface Navigation.
By: E. W. Anderson, R. B. Mitchell, D. H. Sadler, E. Fennessy.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Ordinary Meeting

Date: 6 January. Time: 5 p m.
Held at: The Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2.
Lecture: Telex Service.
By: R. N. Renton.

Informal Meeting

Date: 20 January. Time: 5 p.m.
Held at: The Conference Room, 4th Floor, Waterloo Bridge House, London, S E.1.
Lecture: Facilities and Novel Features of the Standard P.A.B.X.
By: P. E. Pettimore and W. Sheldon.

RADIO SOCIETY OF GREAT BRITAIN

Date: 29 January. Time: 6.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London. W.C.2.
Lecture: Art and Science in Sound Reproduction.
By: F. H. Brittain.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

Date: 14 January. Time: 6 p m.
Held at: Manson House, Portland Place, London, W.1.
Annual General Meeting of Control Section followed by
Lecture: The Theory and Design of Compound Action Automatic Controllers with particular reference to the Interaction Factor.
By: H. Williamson.

THE TELEVISION SOCIETY

Date: 6 January. Time: 6.45 p.m.
Held at: Film House, Wardour Street, London, W.1.
(Joint meeting with the British Kinematograph Society.)
Lecture: The Marconi Television O.B. Unit.
By: C. Carrington.
Date: 29 January. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: American Television.
By: J. Sieger.