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Commentary

IT is a striking coincidence that news of two important submarine cable developments should have been announced within a month or so of each other.

The first of these is the proposal by the Central Electricity Authority in Britain and the Electricité de France to link the electricity networks of the two countries by means of a high power submarine cable.

It may not be clear at first sight why, with the rapid expansion of electricity generating systems in both countries, the interchange of power on a large scale should be seriously considered, but a systematic study of the forecasts of the two electricity loads has disclosed the existence of important diversities—principally the time difference between peak demands—which would justify such a scheme.

When the proposal was first studied by the two authorities in 1951 effort was concentrated on the transmission of power at 132kV by a three-phase alternating current system, but the advance of high voltage direct current transmission submarine cable schemes, notably that between Sweden and the Island of Gotland, and the development of high power mercury vapour mutators caused the alternating current cross-channel scheme to be abandoned in favour of a direct current system capable of transmitting some 120-150MW at 200kV, for which a single core cable would suffice. Among the advantages of the direct current system are the ease with which the frequencies of the two networks can be controlled and the precise control of the load interchange.

The electricity authorities concerned are now convinced of the practicability of such a scheme and work is to start next year on the project which it is planned will be completed by 1960 at a cost of about £4m.

The second event is the official opening of the trans-Atlantic telephone cable on 25 September in commemoration of which the British Post Office has issued a small publication containing the highlights of this supreme technical achievement, which far exceeds in scope anything of the kind that had been attempted before.

The Trans-Atlantic Cable Construction and Maintenance Contract—as it is known—was signed in November 1953 by the Post Office, the American Telephone and Telegraph Company, the Canadian Overseas Telecommunication Corporation and the Eastern Telephone and Telegraph Company, a Canadian subsidiary of the A.T. and T. The contract, although cold in its legal phrasing, was a tremendous act of faith in the ability of engineers on both sides of the Atlantic who were required, in the short space of three years, to create an entirely new medium of communication between the New World and the Old.

In those three years a cable and repeaters, designed to an accuracy and expectation of life never before attempted, had to be manufactured and laid across 2 250 miles of ocean at depths of up to $2\frac{1}{2}$ miles where the pressure is almost $2\frac{1}{4}$ tons to the square inch.

Along with this was the vast interconnected system of land cables and radio links at the shore ends of the cable by means of which London and New York could be connected.

Meetings between the technical representatives of the Post Office and the Bell Telephone Laboratories began in 1952 and consisted of a critical examination of each other's submerged repeaters and transmission techniques.

These techniques although different had many common features in the development of components and manufacturing and testing methods designed to ensure a very high order of reliability over a long period of at least twenty years. The emphasis was on reliability, regardless of cost, and it was agreed during these preliminary discussions that design, circuits and components should not deviate from previous tried practice without adequate justification.

The outcome of these discussions is now well known. All but 400 of the 4 500 nautical miles of cable required were to be made in Britain and laid across the Atlantic by the cable ship *Monarch*—the only cable ship in the world capable of undertaking this task. Deep-sea repeaters of American design and manufacture were to be used on the main section of the cable, while British repeaters were to be used in the shallow water Newfoundland section of cable which was to be of American manufacture.

These were but the main outlines of the scheme, for when the contract was signed in 1953 no detailed specifications existed.

The British factory which was to manufacture the deep-sea cable was little more than a collection of sheds severely damaged by bombing during the war, while the American factory for the manufacture of the repeaters was a shell needing considerable reconditioning. Yet the whole vast undertaking was complete in three years; two months ahead of schedule.

Waveguide Hybrid Circuits and their Use in Radar Systems

By J. W. Sutherland*, M.A., A.M.I.E.E.

The principal types of waveguide hybrid are described and their properties are compared. Balanced duplexers, balanced mixers and wave-guide switching, adding and subtracting circuits are discussed.

A WAVEGUIDE hybrid circuit can be regarded as a passive network, in which there is coupling between certain pairs of terminals of the network and isolation between other pairs. The most commonly used form of hybrid is a network having four pairs of terminals, represented in Fig. 1.

If radio frequency energy is applied to terminals 1 from a matched source, with terminals 2, 3, and 4 matched, the energy divides equally between 3 and 4. There is a high degree of isolation between terminals 3 and 4, and between 1 and 2. Alternatively, energy applied to terminals 2 will divide equally between 3 and 4, while the same isolation is maintained. Practical forms of hybrid include the 'magic-T', the 'rat-race', the short slot hybrid and the branched guide hybrid. Each will be considered in some detail.

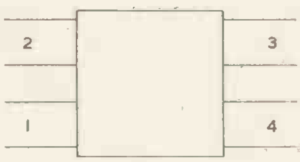


Fig. 1. Eight-terminal network

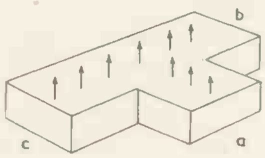


Fig. 2. H-plane or shunt-T junction

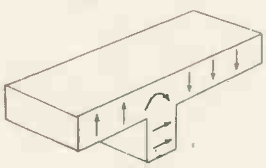


Fig. 3. E-plane or series-T junction

The 'Magic-T' Junction

In a waveguide H-plane or shunt-T junction, shown in Fig. 2, energy applied at (a) divides equally in (b) and (c), and the energy is in phase in each arm. There is no isolation between arms (b) and (c). A qualitative, and simplified explanation of this behaviour is given by considering the electric vectors shown diagrammatically in Fig. 2. In the E-plane or series-T of Fig. 3, equal division also takes place, but in this case, the power in the side arms is in antiphase.

The magic-T consists of a series and shunt-T combined into a single junction, indicated in Fig. 4. It possesses the hybrid properties described in the first paragraph. Power injected at 1 will split equally in amplitude and phase between 3 and 4, while 1 and 2 and 3 and 4 remain isolated. Power applied at 2 will split equally in amplitude, but in antiphase between 3 and 4, isolation remaining as before.

To obtain the best performance from a magic-T constructed in this way, extreme precision of manufacture is necessary and precautions must be taken to match each arm of the junction. This matching is usually achieved by a combination of posts and diaphragms (Fig. 5), or by the use of steps. The step-matched hybrid-T (Fig. 6) is more difficult to manufacture, but gives more nearly ideal performance. Since the magic-T is a relatively narrow band device (say 2 to 3 per cent bandwidth for reasonably good performance), it is necessary to match it at the particular operating frequency. The layout lends itself to use as a balanced mixer, impedance bridge, power divider etc., and these applications will be discussed later, Table 1 shows typical performances.

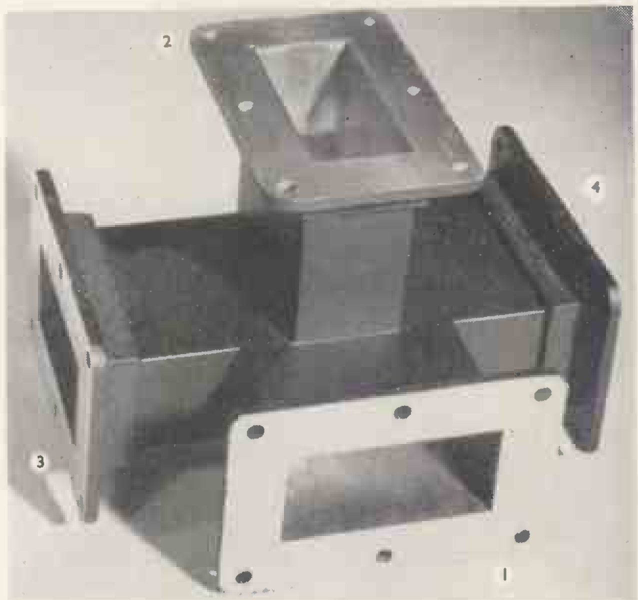


Fig. 4. 'Magic-T' junction (4 000Mc/s)

'Rat-Race' or Hybrid Ring

The waveguide rat-race consists of a ring of rectangular waveguide into which auxiliary waveguides are coupled, in the E-plane. The hybrid properties are achieved by the choice of spacing between the side arms. A typical ring is shown in Fig. 7.

Power injected into 1 will divide between 3 and 4, but will cancel in 2. This power will be in phase in 3 and 4. Power injected at 2 will divide between 3 and 4, but in antiphase, while remaining isolated from 1. Many other arrangements are possible. For matching purposes, the impedance of the waveguide forming the ring is made $Z/\sqrt{2}$, where Z is the impedance of each side arm.

* *Marconi's Wireless Telegraph Company Limited.*

Short Slot Hybrid

A directional coupler possesses certain of the properties of a hybrid junction, i.e. an eight-terminal network, with coupling between particular arms and isolation of others.

If means can be found to achieve a coupling of 3dB (equal division between 3 and 4 of energy applied at 1) and high directivity (isolation of 1 and 2), a conventional hybrid will have been made. One such coupler is the short slot hybrid (Fig. 8), developed by Riblet[3]. The waveguides to be coupled lie with their narrow walls in contact, and this interface is cut away for a specified length. Matching is completed by slightly reducing the broad dimension of the

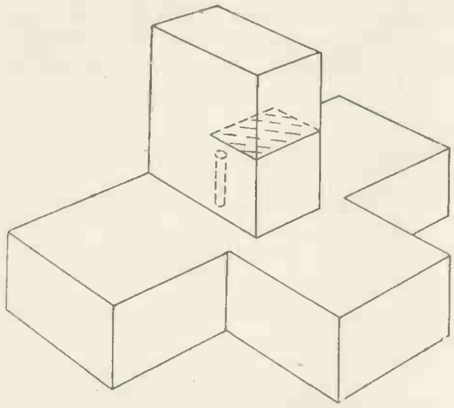


Fig. 5. Post and iris-matched 'magic-T'

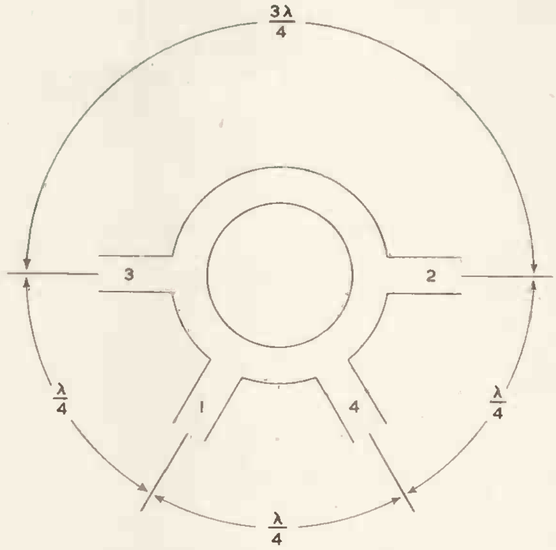
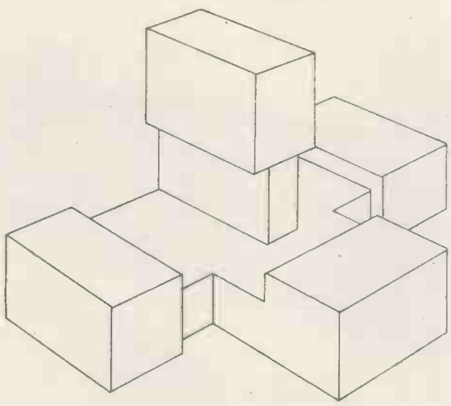


Fig. 7. The 'rat-race' or ring hybrid

Fig. 6. Step-matched 'magic-T'



Waveguide 'rat-race' (10 000Mc/s)

TABLE 1

Comparison of Hybrid Properties

TYPE OF HYBRID	MANUFACTURE	COST	MAXIMUM DIRECTIVITY (dB)	BANDWIDTH OF DIRECTIVITY* (per cent)	BANDWIDTH OF BALANCE* (per cent)
Post Matched Magic-T	Fabricated from tube	Moderately expensive	39	5	5
Step Matched Magic-T	Fabricated from solid	Expensive	50	6	5
Short-Slot ..	Fabricated	Fairly cheap	41	5	5
	Electroformed	Cheap for quantity			
Rat-Race ..	Electroformed	Fairly cheap for quantity	46	4	4
Binomial Slot ..	Fabricated from solid	Expensive	52	10	7
	'Milled Block' technique	Fairly cheap for moderate and large quantities			

* *Bandwidth in this case cannot be defined in absolute terms and these figures are considered for one type of hybrid relative to another.*

two waveguides for the length of the cut away, or 'short slot' and by introducing a small amount of capacitive reactance in the centre of the slot.

The phase relations in this hybrid are interesting; the energy injected at 1 splits as before between 3 and 4—the phase of the voltage at 3 leads that at 4 by $\pi/2$. The actual phase at 4 leads the phase of the voltage which *would* appear in the absence of a slot by $\pi/4$. A simplified vector

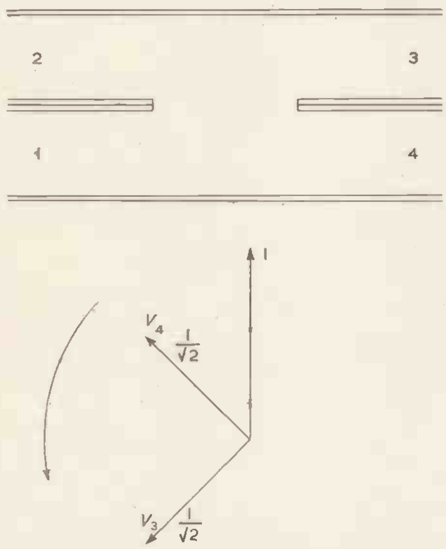
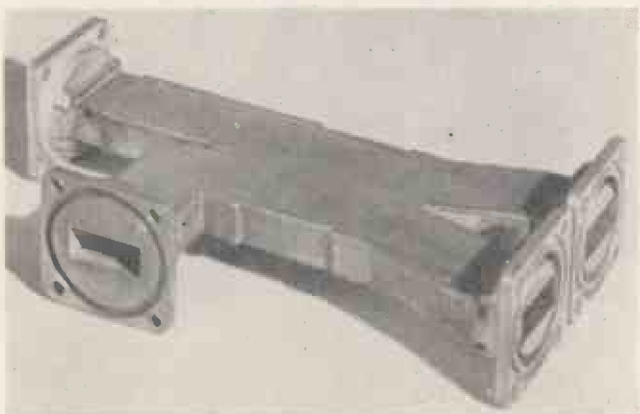


Fig. 8. Short slot hybrid phase relations



Electroformed waveguide section incorporating short slot hybrid (10 000Mc/s)

diagram can therefore be drawn. The junction is symmetrical and reciprocal relations hold.

The Binomial Slot Hybrid

Another form of directional coupler which can be adapted for use as a waveguide hybrid is shown in Fig. 9 and is sometimes known as the branched guide coupler.

To produce a coupling of 3dB, the branches would have to be increased to such a size that the directivity would be unacceptable. Therefore, the number of branches is increased and the amplitude of voltage coupling due to each slot is arranged to be proportional to the coefficients of a binomial expansion, in order to maintain high directivity.

The design figures for a five slot coupler are as follows: the individual slots are arranged to have coupling of 20·8dB, 8·5dB, 5·01dB, 8·5dB and 20·8dB. This gives voltage coupling for each slot in the ratio 1 : 4 : 6 : 4 : 1; the successive power coupling for unity power input is ·008, 0·139, 0·268, 0·081 and ·004 giving a total of 0·5, which corresponds to the correctly balanced hybrid. In practice, if a hybrid is constructed to these figures, slight 'trimming' may still be necessary to optimize the performance. The phase relationships are opposite to those of the short slot hybrid. The phase at 3 lags relative to that at 4 by $\pi/2$. The hybrid is symmetrical and reciprocal relations hold.

Balanced Duplexer

The isolation between the transmitter and receiver of a radar equipment provided by a conventional single t.r. cell duplexer is normally only just sufficient to prevent

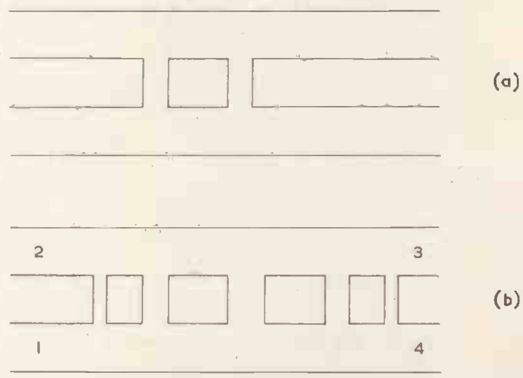
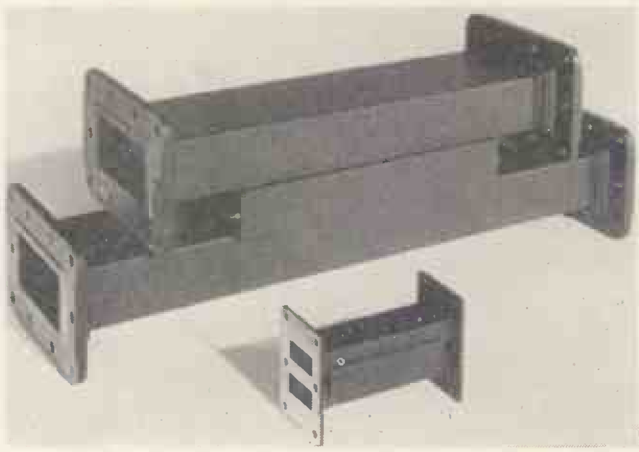


Fig. 9(a). Branched guide coupler; (b) binomial slot hybrid



Binomial slot hybrid

permanent damage to the receiver, and the temporary deterioration of receiver noise figure causd by crystal paralysis must be tolerated. With a 'spike' break through of about 0·1 erg, which will not be sufficient to cause permanent damage to the mixer crystal (i.e., burnout), paralysis may be caused which will deteriorate the receiver performance by as much as 1dB. If this spike is reduced by a further 10dB or more, to 0·01 erg or less, the paralysis is no longer observed. If a balanced duplexer is introduced, this further isolation can be achieved. For maximum isolation, the two t.r. cells used in a balanced duplexer must have closely matched characteristics; the electrical lengths, reflection coefficient, and breakthrough powers must be as nearly equal as possible. Fig. 10 is a general diagram of a balanced duplexer. When the transmitter fires, the power divides equally in hybrid 1, and breaks down the two t.r. cells; the main power is therefore reflected from the front windows of the cells, recombines in hybrid 1 and passes to the aerial. The two break-through powers from the t.r. cells combine in hybrid 2, being absorbed in the load. Any break-through power to reach the receiver is further reduced by the isolation of the second hybrid, although in practice the full isolation is not effective. In order to obtain the best performance from a balanced duplexer, twin t.r. cells have been developed, in which two complete sets of windows, electrode structures, etc., are enclosed in the same unit, sharing the same gas filling. The electrical characteristics of the two halves of the unit are matched on test; phase differences between the paths through two halves can in this way be reduced to 5° or less, whereas the phase difference between random samples of single cells may

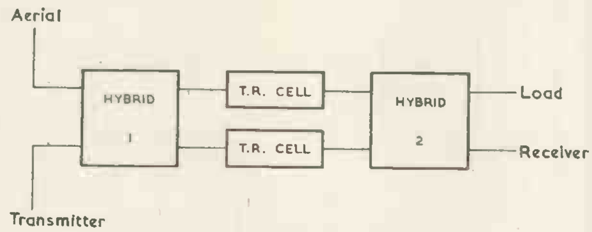


Fig. 10. Balanced duplexer

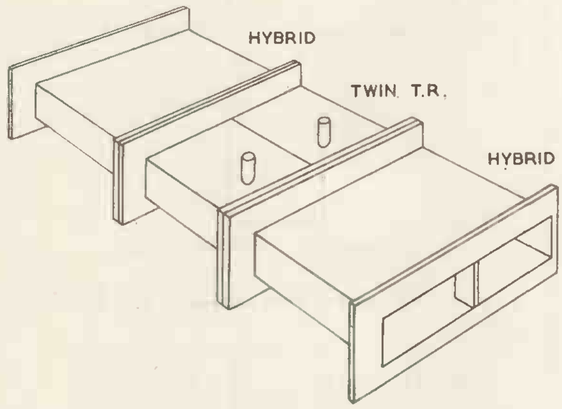


Fig. 11. Balanced duplexer using short slot hybrid

spread over as much as 35°. Secondly, since the gas filling of the twin cell is common to both halves, gas clean-up and ageing will probably occur to the same extent in each half and the pair will remain matched throughout the useful life of the cell.

Two arrangements of twin t.r. cell have been produced, a side-by-side system, in which the narrow waveguide walls are in contact, for use with short slot hybrids, and one cell over the other, where the two halves are spaced by an amount suitable for use with a binomial slot hybrid. Figs. 11 and 12 show these arrangements.

Hybrid Phase Shifter

If a hybrid is connected as shown in Fig. 13, with a short circuit in arms 3 and 4, power entering at 1 will split into 3 and 4, be reflected and recombine to emerge at 2. If the electrical lengths of arms 3 and 4 are equal and the hybrid is matched, all power entering at 1 will emerge at 2. If now the two short circuits are moved together, the electrical length of the whole system can be varied, and consequently the device can be used as a phase shifter.

In certain high power applications, it is possible to place a discharge tube across arms 3 and 4, at a fixed distance in front of the two short circuits. If high power is passed through the system, the tube breaks down, behaves like a short circuit in the new position, and shifts the phase of the high power relative to the phase of low power. This arrangement can form part of a 'pre-t.r.' in duplexer systems.

Hybrid Waveguide Switch

Fig. 14 shows a combination of a hybrid phase shifter and two other hybrids used as a waveguide switch. The method of operation is as follows: power entering at A will divide equally into the two channels B and C. If the difference in path length between B and C is an integral number of waveguide wavelengths, the two signals will combine at X. If now the short circuits in the phase shifter are each moved a quarter of a wavelength, so that paths B and C differ by an odd number of waveguide half-wavelengths, the signals in B and C will combine in Y. Therefore this operates as an effective changeover switch. For rapid successive operation of the switching cycle, a rotating shutter may be introduced into arms 3 and 4, located

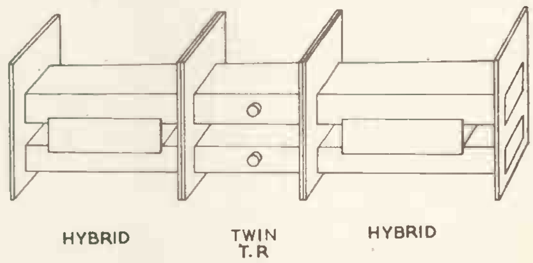


Fig. 12. Balanced duplexer using binomial slot hybrid

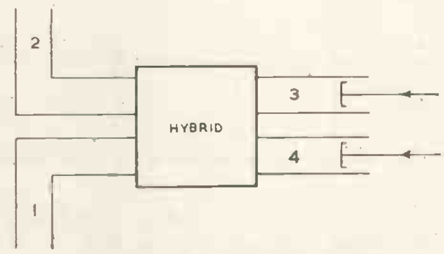


Fig. 13. Hybrid phase shifter

a quarter of a wavelength in front of the short circuits, and this has been found to give satisfactory results. For simple amplitude comparison of two coherent signals, the signals are applied to D and A in the phase appropriate to the type of hybrid. In one position of the shutter, D is passed to Y and A to X, and in the other position, the D to X and A to Y. If a detector is now placed at X or Y, a square wave output will be obtained, whose amplitude is proportional to $|D-A|$ and whose sign depends on the sign of $(D-A)$.

The Addition and Subtraction of Signals Using Waveguide Hybrids

It is a simple property of a waveguide hybrid, that signals in the phase appropriate to the type of hybrid applied to arms 1 and 2 will give rise to the difference and sum of these two signals in arms 3 and 4. Fig. 15 shows a simple qualitative analysis of this for a short slot hybrid, where signals A and B must be applied in quadrature.

Thus signal $A\angle 0°$ at 1 produces voltages $A/\sqrt{2}\angle -45°$ at 3 and $A/\sqrt{2}\angle +45°$ at 4;

and signal $B\angle +90°$ at 2 produces voltage $B/\sqrt{\ }\ \angle -45°$ at 3 and $B/\sqrt{2}\angle -45°$ at 4.

Thus signals at 4 are in phase and will add and signals at 3 are in antiphase and will subtract. Thus $(A+B)$ appears at 4 and $(A-B)$ at 3.

This hybrid property has a number of applications in radar, particularly in aerial systems where continuous comparison of signals is required. In a typical arrangement, shown schematically in Fig. 16, an aerial with four beams arranged in a square is used. By the use of hybrids, signals proportional to $(A+B)-(C+D)$, $(A+C)-(B+D)$ and $(A+B+C+D)$ can be obtained, enabling continuous information of elevation bearing and range to be provided.

Balanced Mixer

The waveguide hybrid lends itself particularly well to use in a balanced mixer in radar receivers. In a receiver using a single mixer, the local oscillator output contains components of noise at signal and image frequencies, which contribute to the noise generated by the receiver and de-

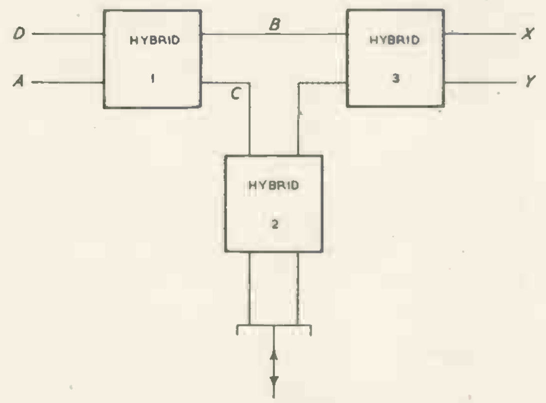


Fig. 14. Hybrid waveguide switch

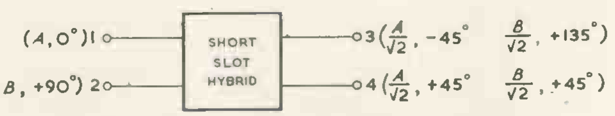


Fig. 15. Addition and subtraction in the short slot hybrid

grade the overall noise performance. A balanced mixer is arranged as shown schematically in Fig. 17. Signal and local oscillator are applied to inputs 1 and 2 and mixers are connected to outputs 3 and 4. Phase relationships are such that the intermediate frequency outputs due only to signal are in antiphase at 3 and 4, but i.f. outputs due to noise components of local oscillator at signal and image frequencies are in phase. Thus by combining the two i.f. outputs through a 180° phase-shift network, or by using crystals of opposite polarity, the wanted signals can be added in the first stage of the receiver, but the noise components are cancelled. Complete cancellation depends on the balance and isolation of the hybrid and on similarity of the two mixers and crystals. Modern mixer crystals have a remarkable consistency of r.f. impedance and quite a small spread of conversion loss; consequently the balanced mixer can be effective in suppressing local oscillator noise. The noise spectrum of a klystron local oscillator follows the same shape as the resonance curve of the klystron cavity. Consequently a cavity with a high loaded Q will contribute less noise than one with a lower loaded Q for measurements made at the same i.f. But the high Q cavity will have a smaller electronic tuning range than the lower Q cavity, and consequently there must be a compromise between noise and electronic tuning range. The higher the intermediate frequency, the less will be the effect of local oscillator noise; so once again, the choice of i.f. must be a compromise. The improvement in overall noise factor to be expected by the use of a balanced mixer as opposed to a single-ended mixer will be from 0·5dB to 1·5dB, according to local oscillator and i.f. Reference 7 gives a detailed analysis of balanced mixer operation, with particular emphasis on the symmetry of the system.

Other Uses of Hybrids

The full isolation of a waveguide hybrid depends, as was stated previously, on all arms of the hybrid being matched, including the source. However, the isolation between arms 1 and 2 will still be maintained if arms 3 and 4 present an identical impedance. If 3 and 4 do not present the same

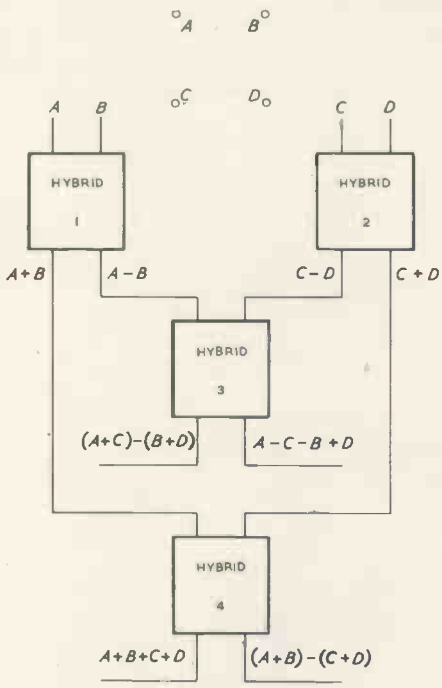


Fig. 16. R.F. head of radar equipment, using hybrids

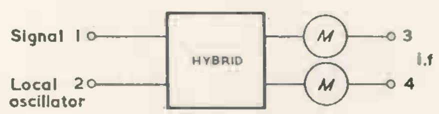


Fig. 17. Balanced mixer

impedance, there will be coupling between 1 and 2. This property forms the basis of the impedance bridge. A matched source is applied at 1 and a matched detector at 2. A known impedance (in many cases a matched load) is connected at 3 and the unknown impedance at 4. If the unknown impedance is adjusted for zero signal at 2, the impedances will be equal at 3 and 4. The use of a source swept in frequency and an oscillographic display of the detector output provides a most useful laboratory instrument.

A further application of this hybrid property is the "Pound" frequency stabilizer described in reference 10. In outline, the oscillator whose frequency is to be stabilized is applied to 1, a matched load is connected to 3 and a cavity tuned to the desired frequency is connected to 4. When the oscillator is on the correct frequency, the cavity 'looks like' a matched load, and there is no output from the detector at 2. If the frequency drifts, the apparent impedance of the cavity changes and the detector gives an output which is used to correct the frequency of the oscillator.

A similar device used to detect small changes in frequency is the microwave interferometer described by Harvey[11]. Arms 1 and 2 of the hybrid have a source and detector as before. Arms 3 and 4 each have a short circuited waveguide, one of which is approximately one wavelength long and the other of the order of 250 wavelengths long. Therefore the bridge will be sharply balanced for a number of closely spaced frequencies (corresponding to the fringes of an optical interferometer), and consequently will behave as a sensitive detector of frequency change. The whole set of fringes can be moved in frequency by slight variation of the length of the shorter waveguide. The necessity for a waveguide 200 to 300 wavelengths long restricts the convenient use of this interferometer to the millimetre wavelengths.

A further use of hybrids is to divide power equally between two feeders, where it is important that there shall be no coupling between them. A typical example is in the automatic frequency control of radar receivers, where it is necessary for the local oscillator to feed signal and a.f.c. mixers, but essential that there shall be no direct coupling between them.

Acknowledgment

Acknowledgment is made to Marconi's Wireless Telegraph Company Limited for permission to publish this article.

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CALDER HALL

On 17 October Her Majesty Queen Elizabeth visited Cumberland for the official opening of Calder Hall. The ceremony consisted of diverting some of the electrical energy produced by the nuclear power station of the United Kingdom Atomic Energy Authority into the grid system of the Central Electricity Authority.

Thus, the first nuclear power station operating on a commercial basis in Britain—and indeed in the world—was opened.

The main purpose of Calder Hall, however, is not the generation of electrical energy, but the production of plutonium, and for this reason it will still continue to be operated by the Atomic Energy Authority. It is anticipated that it will not be until 1961 that nuclear power stations designed specifically and solely for the generation of electrical energy will be in operation by the Central Electricity Authority.

Preliminary details of Calder Hall have recently been released to coincide with the official opening by the Queen, but fuller information will be given at the British Nuclear Energy Conference to be held in London from 22-23 November.

A general description of the Calder Hall installation was given in *Electrical Energy* (October 1956) and a description of the burst slug detection equipment is given on page 487 of this issue.

In June of last year, it was announced that increased supplies of plutonium were required and, to avoid interference with the existing nuclear power station programme and the investigation already being undertaken by the Atomic Energy Authority, it was decided to build three more units of the Calder Hall type, each consisting of two reactors and their associated generating plant.

One of these units is located alongside the present Calder Hall unit and will be known as Calder "B". Work has already started on the new site and at the moment the building and civil engineering work is approaching completion. It is planned that Calder "B" will be in operation by the summer of 1958.

The remaining two units are to be erected at Chapel Cross, near Annan, Dumfriesshire, where preliminary work has already started.

The installation of Calder Hall—known as Calder "A"—is illustrated below. It consists essentially of:

(a) 2 Gas-Cooled Graphite-Moderated Reactors.

(b) 8 Heat Exchangers.

(c) 8 Gas Blowers.

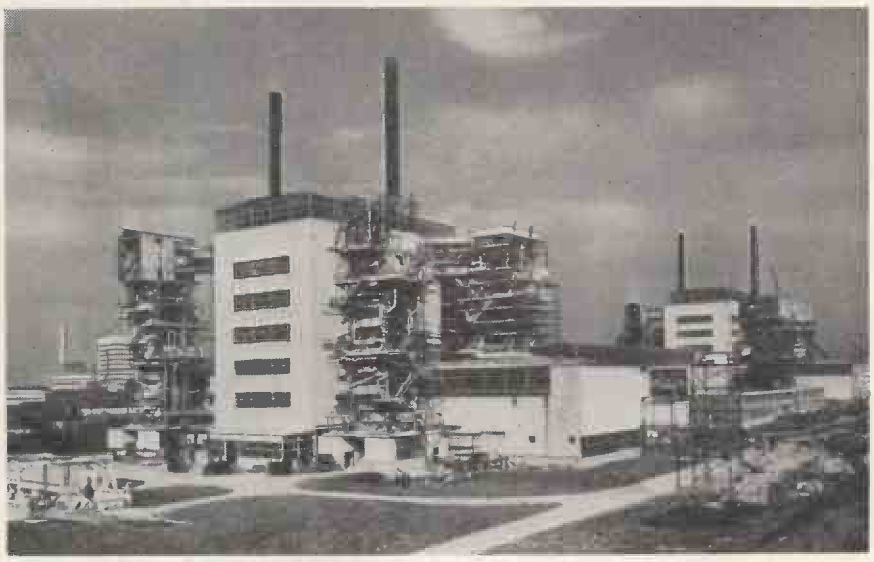
(d) 4 Turbo-alternator sets and Condensing Equipment.

(e) 2 Dump Condensers.

(f) 2 Cooling Towers.

The alternators are air-cooled, have a maximum continuous rating of 23MW at a power factor of 0·8 (lagging) and a speed of 3 000rev/min. The generator voltage is 11·5kV three-phase, 50c/s.

General view of Calder "A" showing the turbine hall flanked at each side by a reactor building—the heat exchangers are located at the corners of the reactor building. Behind Calder "A" and to the extreme left can be seen the Windscale factory.



The Control of Thermionic Valve Envelope Quality by Thermal Shock Testing

By G. D. Redston*

The factors controlling the failures of all-glass valves in thermal shock tests are discussed. It is pointed out that the different thermal shock tests bring out different defects.

Evidence is given showing that the downward thermal shock test results are more nearly correlated with service life than those of the upward thermal shock test. Tempering the valve base improves resistance to downward thermal shock, but has failed to produce a significant change in the number of failures on life test.

In a previous paper[1], the writer has described the glass working processes in the manufacture of the all-glass thermionic valve. While glass has many virtues as an envelope material for valves it has the disadvantage of being mechanically weak, and the valve manufacturing processes can leave the glass envelope in a strained state so that it cracks spontaneously after some time in service. By careful factory control cracking in service is minimized, but the responsibilities of some modern valves are such that this matter cannot be left to chance. Tests for proneness to spontaneous failures must be applied before batches of valves are dispatched.

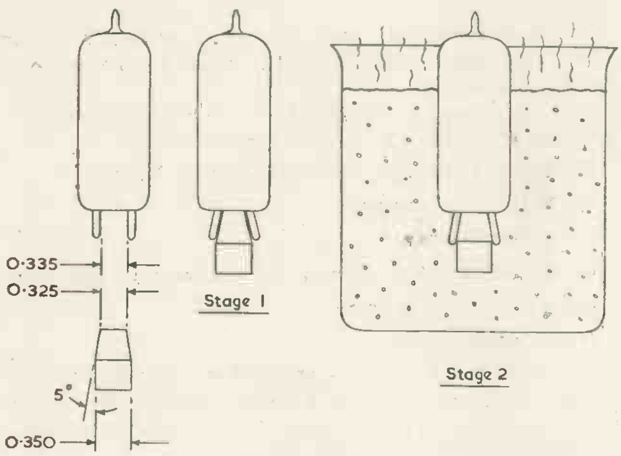


Fig. 1. The 'B' test

The usual procedure is to apply a thermal shock test on a representative sample group of valves, accepting or rejecting the batch according to an acceptance quality level with the help of statistical tables[2].

Thermal shock tests consist either of sudden heating or sudden cooling and are called respectively upward or downward thermal shock tests. The most widely used test for valves is the 'B' test which is a form of upward thermal shock test.

In this article the function of such thermal shock testing is examined with the following questions in mind:

(a) Do thermal shock tests adequately measure the "spontaneous failure proneness" of a group of valves?

If so;

(b) Which form of thermal shock test gives the most helpful indications?

To do this, the well known 'B' test is examined. It is pointed out that the types of failure which occur on B test do not necessarily occur in service; a better test is the downward thermal shock test.

The Upward Thermal Shock Test

The specification for the B test[3] states "The test shall consist of forcing the pins of the valve over the specified cone (Fig. 1) and then completely submerging the valve and cone in boiling water for a specified time . . . the water container shall be at a temperature between 97°C and 100° . . . the water container shall have a minimum capacity of 2 litres per 15 valves and shall be at least three-quarters full . . . valves before test shall be at room temperature and shall have been submitted to approved pin-straightening.

(a) Align the axis of the valve with the axis of the specified deflexion cone and carefully push the small end of the cone into the circle formed by the valve pins until the cone lies firmly against the valve bottom.

Note. If some pins are bent more than others the test is being made improperly.

(b) Place the holder of valves into boiling water so that the valves and cones are completely submerged for a period of 10 seconds.

(c) Remove the valves from the water and allow to cool to room temperature on a wooden support."

The defects which the B test detects are, in common with all upward thermal shock tests, those on the inside surface of the envelope. In addition, the test detects many of the defects associated with the contact pins.

For convenience in fault diagnosis a classification of types of crack was built up on the observation that the majority of base failures occurred in one of four ways:

Type 1. Crack tangentially at the seal.

Type 2. Crack radially across the base through a pin.

Type 3. Crack radially across the base *not* through a pin.

Type 4. Crack tangentially at the pin circle, through a pin.

This classification was used in the experimental work described.

Those defects which were observed to cause B test failures in the factory practice were:

(a) Crizzles and laps on the inside surface of the base.

(b) Under-melted seal.

(c) Over-melted seal.

(d) Pin circle diameter errors.

(e) Pins too stiff.

(f) Incorrect tempering.

(g) Mismatched bulb and base glass.

** Brimar Valve Works, Standard Telephones and Cables Ltd.*

(a) *Crizzles and Laps on the Inside Surface of the Base*

Cracks, crizzles and laps on the inside surface of the base acted as fracture origins for cracks which were usually of Type 3.

(b) *Under-melted Seal*

If there were internal crevices between base and bulb after sealing there was a tendency to Type 1 failures on B test.

(c) *Over-melted Seal*

When the seal was over-melted the bulb adhered to the stem fillets, forming a crevice as well as thickening the glass in the region of the seal. Both of these factors led to Type 1 failures on B test, although Type 2 and Type 3 cracks sometimes occurred in this case.

(d) *Pin Circle Diameter Errors*

If the pin circle diameter was too small the B test plug

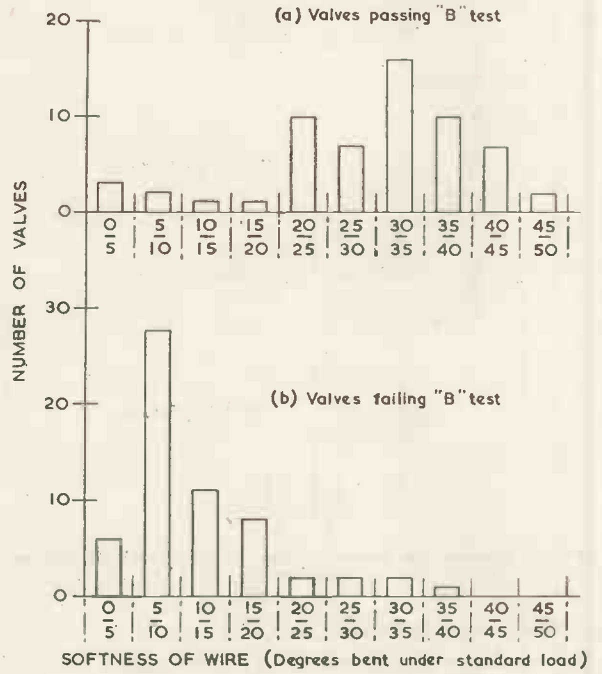


Fig. 2. Effect of softness of wire on 'B' test failure rate

would tend to break the base by forcing the pins apart. In a particular case, where 700 valves with p.c.d. 0·005in below standard were tested, 15 failed on B test with typical Type 4 cracks. When 102 valves with p.c.d. correct at 0·375in were made up similarly and tested there were no failures from this cause.

(e) *Pins Too Stiff*

Even with correct p.c.d. the insertion of the B test plug forces the pins out somewhat so that pin stiffness plays a part in the strain transmitted to the base.

Taking some envelopes which had passed and some which had failed the B test, their pins were carefully extracted and the stiffness measured. For this purpose a "Bendometer" constructed to ASTM Specification B113-41 was used. The results (Fig. 2) showed that the proportion of hard pins in valves which had failed was very high. The failures in this case were also of Type 4.

(f) *Incorrect Tempering*

Tempering is one of the less understood factors affecting the B test results.

Laboratory tests established a broad relationship between 'radial stress' (observed by looking through the base) and failure rate (see Table 1).

TABLE 1

Relationship of Apparent Stress and B Test Failure

APPARENT STRESS ($m\mu$)	FAIL B TEST (per cent)
68–91 T	60
23–45 T	43
Nil	0
0–23 C	0
23–45 C	0
45–68 C	50
68–91 C	66

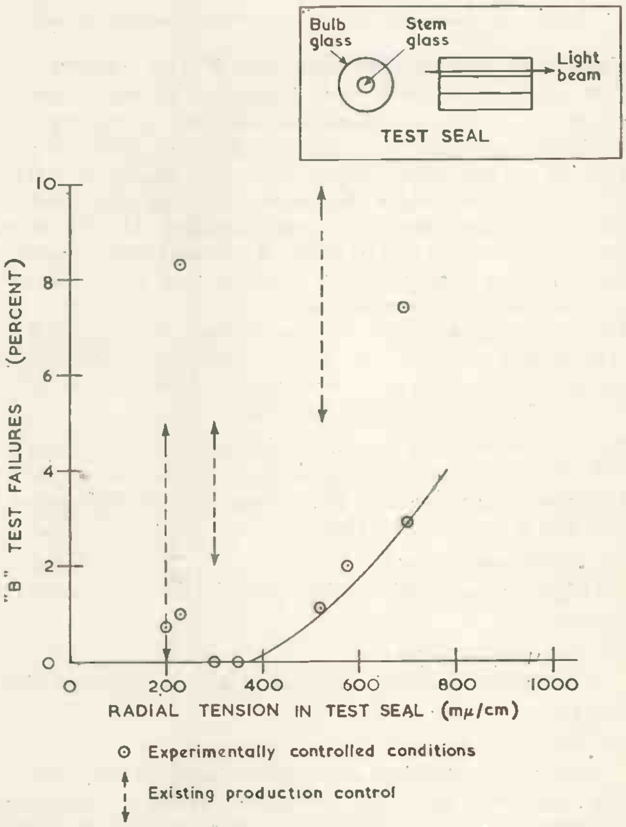


Fig. 3. Dependence of minimum failures 'B' test on degrees of mismatch

This indicated that tempering should aim for the 'low compression' condition.

More will be said later on tempering.

(g) *Mismatched Bulb and Stem Glass*

The effect of bulb-stem expansion mismatch was found to be a most important factor in controlling B test failures. Valves were made with a number of pairs of glasses using the best available glassworking technique and were then B tested. The mismatch of each sealing pair was measured in a standard manner using the stress in an annealed bead seal of bulb glass on stem glass cut for axial sighting[4]. The results were as shown in Fig. 3 and Table 2.

The cracks were in general of Types 3 and 2.

As is seen from the table some of the better pairs were also tried in production where manufacturing errors apparently caused increased B test losses.

From this work (which summarizes observations taken over many months) it was concluded that as the mismatch increased and the stem became increasingly under tension the difficulty of setting the sealing-in machine increased. Beyond a mismatch represented by $500m\mu/cm$ in a standard seal no sealing technique which gave low B test results was discovered.

TABLE 2

Effect of Mismatch on B Test Results

STRESS IN SEAL ($m\mu$/cm.)	FAIL B TEST (per cent)	
	EXPERIMENTAL	PRODUCTION
200 T	0·7	0 to 5
230 T	8·4 & 1·0	—
300 T	0	2 to 5
350 T	0	—
520 T	1·1	5 to 10
580 T	2·0	2
700 T	2·9	—
700 T	7·5	—

Failures in Service Compared with B Test Failures

Having now outlined those factors which were found to affect the B test, attention was directed to the types of valve failure which occurred in service. In general, glass failures did not figure largely in service returns on valves, but those valves which did come back from the field for this reason were examined and classified. Of the valves which failed for cracked base, the cracks were classified into the four types and it was found that Type 2 largely predominated. This was not at all the case with B test failures! Fig. 4 shows the comparison. Further analysis showed that many of the defects which the B test singled out in the valves did not ever cause field failures. Consideration of the individual defects verified this.

(a) *Crizzles and Laps on the Inside Surface of the Base*

In fact approximately 20 per cent of cracked bases did fail from internal crizzles in the base. For this defect therefore the B test was of value.

(b) *Under-melted Seals*

Origins from under-melted seals did not occur in practice.

(c) *Over-melted Seals*

Over-melted seals did not fail in service, although they looked very poor.

(d) *Pin Circle Errors*

There was insufficient evidence to deduce whether valves had failed from pin circle errors. Such errors are influenced by socket design, as failures can be caused by inaccurate dimensioning and also by inadequate float of the contact clips in the socket[5]. If the socket clips were absolutely rigidly fixed then the base would tend to split (probably Type 2 cracks), but the forces needed to split the base would be relatively large. Normal valves fail at over 9lb radial force per pin on a test of a few seconds' duration; allowing a factor of three for the strength-time effects the socket must therefore exert a steady lateral force of at least 3lb to produce failures in service. Most modern sockets are designed with adequate freedom of movement for the contact clips. Furthermore, limits of the order of a few pounds weight have been set for the insertion and withdrawal forces of a gauge with standard pin dispositions[6]. These indirectly limit the lateral force that the socket can exert on the pin.

Thus only a bad displacement of the valve pin from its true geometrical position is likely to cause failures in modern sockets. On the other hand, the B test produced copious failures when the p.c.d. was too small. For other errors of the p.c.d. the B test was completely insensitive.

(e) *Pins Too Stiff*

Imperfections of pin circle and socket design and also bad handling were more likely to crack a valve with stiff rather than soft pins. It was frequently found that an envelope returned for cracked base showed vestiges of severe bending of the originating contact pins caused probably by careless extraction from the valve socket. Stiffness of pins in itself was not a disadvantage in valve service. The B test, however, always tended to produce failures with stiff pins.

(f) *Incorrect Tempering*

Experiments have indicated that there is no variation of service failures within a wide range of degrees of tempering. B test results were markedly affected by degree of tempering.

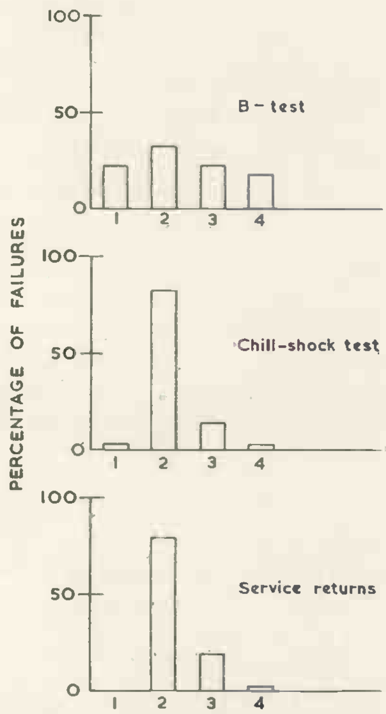


Fig. 4. Fracture type-distribution for 'B' test chill-shock test and service returns

(g) *Mismatched Bulb and Stem Glass*

No serious study of the effect of mismatch on service failure had so far been made.

On this analysis it was clear that although the B test was useful for controlling a number of possible manufacturing errors in the factory a test must be sought which would give a more realistic indication of the service life of a batch of valves.

The Downward Thermal Shock Test

The chill-shock test, i.e. a test in which the valve is suddenly cooled from some higher temperature, has the advantage of being more in line with modern ideas of glassware testing. In contrast to the upward thermal shock which probes the inside surface of the glass envelope, the chill-shock test puts the outside surface of the glass into sudden tension and any weak spots are thus made to break. It is the outside surface of the glass article which suffers impact, atmospheric corrosion and stresses from any mechanical loads such as clips etc., and it is the outside surface which is therefore really interesting.

Early Chill-shock Trials

The original chill-shock tests carried out by the writer were made at a time when it was necessary to detect fierce strain rings in valve envelopes. After some preliminary tests it was found that strain rings could be identified by a chill-shock from an air oven at 175°C to water at 20°C:

Good envelopes:	22 tested,	3 failed.
Strained envelopes:	44 tested,	19 failed.

Further work showed that the test sensitivity was poor, in that slightly strained envelopes were not detected, so this work was discontinued.

Serious interest was directed to the chill-shock test when it was found that the type of cracked base obtained corresponded with the type of crack obtained in service failures. (See Fig. 4). As a result of this observation it was decided to investigate the chill-shock test thoroughly, with the object of using it side by side with the B test as a quality control.

Chill-shock Tests with B Test Plug

Tests were carried out by first inserting the B test plug then slowly heating the valve in water and finally giving it the chill-shock. Increasing temperature differences were used for the shock; the " hot " bath was water and the cold was either water or, for 0°C, salt water and ice[7].

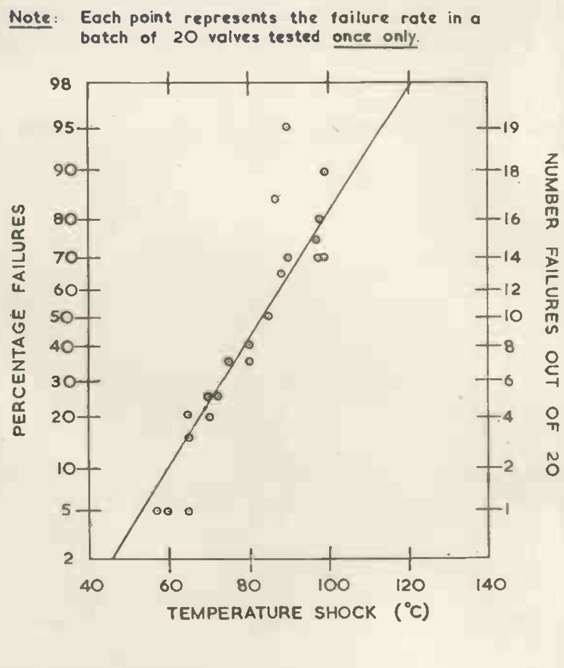
A large group of valves was tested to establish the statistical variation to be expected. Groups of 20 valves were selected at random from a bulk sample of several hundred valves and were subjected one at a time to the chill-shock routine, care being taken to keep the motion and time of transfer the same for all valves. The results are shown in Fig. 5. For the particular batch of valves tested the temperature for 50 per cent failures was 84°C. The cumulative failures curve fitted the Normal (Gaussian) distribution, giving a coefficient of variation of 22 per cent. (But the present data could be fitted almost equally well to a straight line. The straight line has no obvious physical significance however.)

Using the same batch of valves, experiments were made utilizing an oil bath for heating. In this way chill-shock temperatures of 110°C and 120°C were reached, but the failure curve then gave a higher 50 per cent failure value. It was thought that the thin layer of oil carried over with the valves impaired the heat transfer efficiency of the cold water. A detergent was added to the water in an attempt to remove the suspected oil film, but no significant improvement resulted.

Chill-shock Tests Without B Test Plug

With further valves from the same batch chill-shock tests

Fig. 5. Probability plot of effect of temperature shock on proportion of failures on a chill-shock test which used the ' B ' test plug



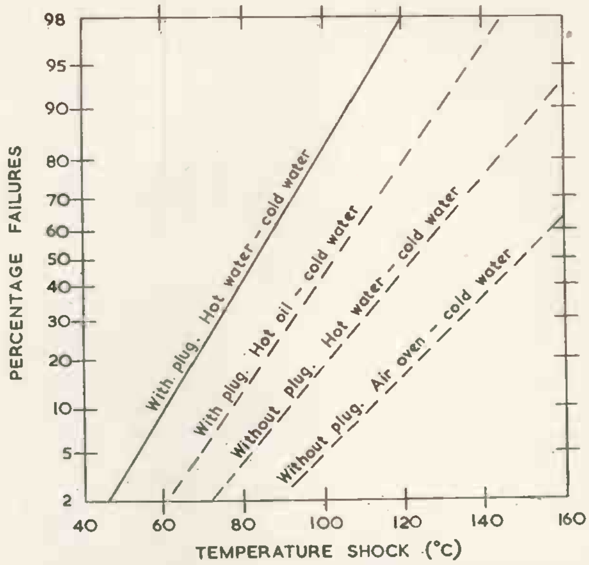


Fig. 6. Relationship between results of various types of chill-shock test on miniature valves sampled from a uniform batch

were carried out using hot water for shocks up to 90°C and an air oven for the remaining shocks up to 180°C.

The results for the various forms of chill-shock tests are shown in Fig. 6. It was found that the use of the plug increased the failure rate and that there was a greater breakage in water to water tests than in the other forms of test.

The defects which might be expected to cause failures on chill-shock tests are:

(a) Crizzles on the outside surface of the base.

(b) Absence of gas cushion where the contact pin enters the glass.

(c) Mismatched bulb and base glass, and

(d) Incorrect tempering.

Crizzles and gas-cushion defects are, in effect, small cracks which extend on chill-shock test and probably in service. In manufacturing practice, therefore, care was taken to avoid them. The effect of varying mismatch between bulb and base glass on chill-shock test was not studied because those glasses selected for B test suitability also proved satisfactory for chill-shock and in service. As chill-shock testing gave conflicting results, a closer study of tempering and its effects was necessary.

Analysis of Layer Stress

Tempering sets up a sharp stress gradient in the base of the valve. In the first place a more detailed experimental analysis of this ' layer ' stress was required. Photoelastic measurements on normal valves using suitable immersion liquid (bromoform) established that the apparent layer stress passed from quite strong tension on the inner surface to a compression on the outer surface of a similar order (Fig. 7(b)). By varying the tempering the stresses were increased (Fig. 7(c)) and by annealing they were almost eliminated (Fig. 7(d)).

The curves were made more complicated, however, by the presence of depressions in the centre of the button stem. Special stems with no depressions were therefore prepared. These gave clearer layer stress curves, the shapes of which could be due to:

(a) Tempering effects.

(b) Differing amounts of highly stressed bulb glass through which the light beam has to pass, and

(c) Slight bending action by the bulb on the base.

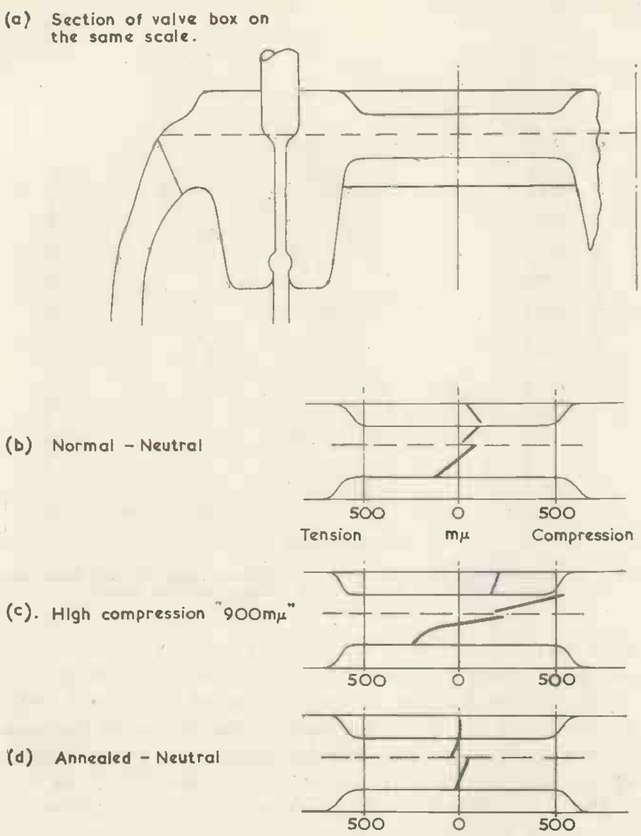
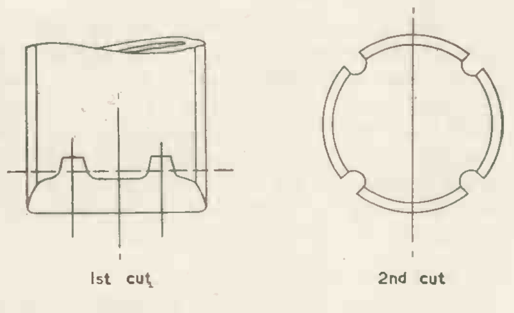


Fig. 7. Retardation gradients resulting from various strain-setting arrangements

Fig. 8. Stages in cutting away bulb glass to determine the effect on layer stress



The relative importance of these three factors was investigated by examining the stress pattern after deliberately removing a source of stress by cutting the glass bulb away from the stem. In the first cutting (Fig. 8) the bulb glass was sliced off at the level of the top of the stem. Although this reduced bending of the stem, it left a ring of stressed bulb-glass around the stem. In the second cutting, this ring of bulb-glass was slotted radially to break its continuity and to reduce substantially any circumferential stress present in the bulb. Those two cuts successively reduced the apparent internal tension layer. The external compression layer was almost unchanged (Fig. 9).

Although annealing reduced the retardations considerably (Fig. 7(d)), a detail which still remained was the step in the stress gradient at the extremity of the bulb seal. The slight step was consistent with the bulb being in either circumferential compression or in radial tension.

From these observations it was concluded that the apparent layer stress distribution due to tempering itself was of the same form as in any other force-cooled article and that the bulb stresses had a considerable influence only on the internal apparent stress, although in practice the internal centre depression in the stem masked off the region in which this influence was greatest.

For routine investigations of layer stress it was decided to choose the centres of the top and bottom depressions for the sighting points.

Effects of Layer Stress on Chill-shock Results

A series of experiments was carried out[7] to determine the influence of the layer stress on chill-shock results. Ten groups with twenty dummy valves in each group were sealed-in, each group with a different sealing technique. For each group, average values (Table 3) were found for the layer stress at the two sighting points.

TABLE 3

Effect of Layer Stress on Thermal Endurance (Average of 6 out of 20)

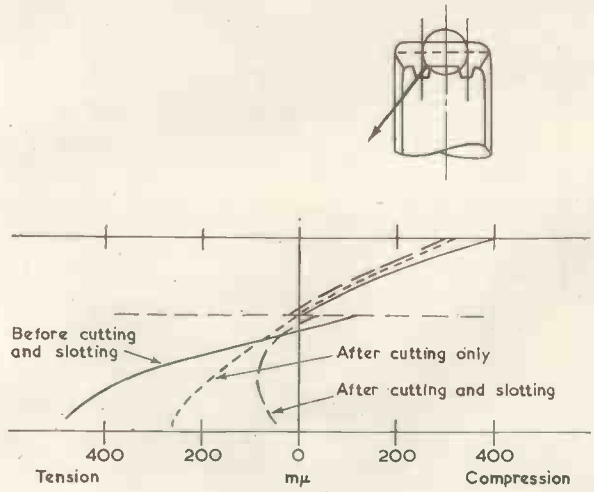
EXTERNAL (C) (mμ)	INTERNAL (T) (mμ)	T + C (mμ)	50 PER CENT FAILURE TEMPERATURE (°C)
630	270	900	64
290	318	608	85
210	420	630	90
176	273	449	93
198	220	418	84
104	170	274	95
162	30	192	64
96	101	197	47·5

Each group was chill-shock tested to find the 50 per cent failure temperature at temperature intervals increasing by 10°C until all failed. It was found that correlation between the individual external or internal layer stress reading and the 50 per cent failure temperature was poor. However, on the theory that the *sum* of the two represented the slope of the layer stress gradient and hence the degree of tempering, an interesting correlation was obtained (Fig. 10).

This was later confirmed by production experience, Fig. 11 showing the form of correlation obtained. It appears that the tempering at first rapidly increases the strength, but that further tempering leads to a decline.

From the results shown in Table 4 it was seen that over the range of layer stresses studied there were no failures on life. For practical purposes then layer stresses had little or no effect on life, although they did affect the chill-shock test results.

Fig. 9. Effect on retardation gradient of cutting away glass bulb



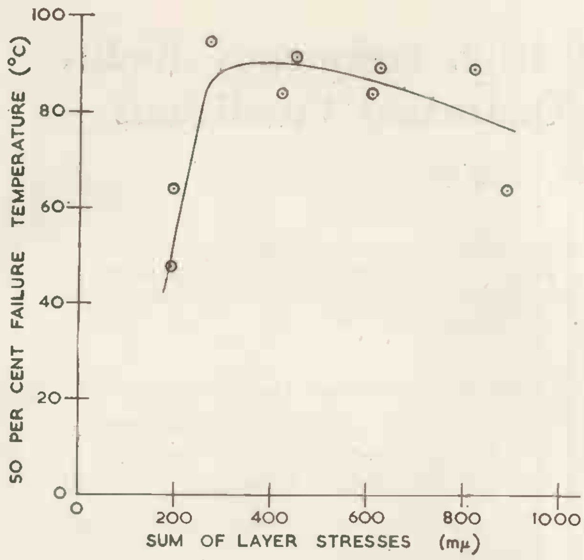


Fig. 10. Effect of layer stress on thermal endurance of miniature valves

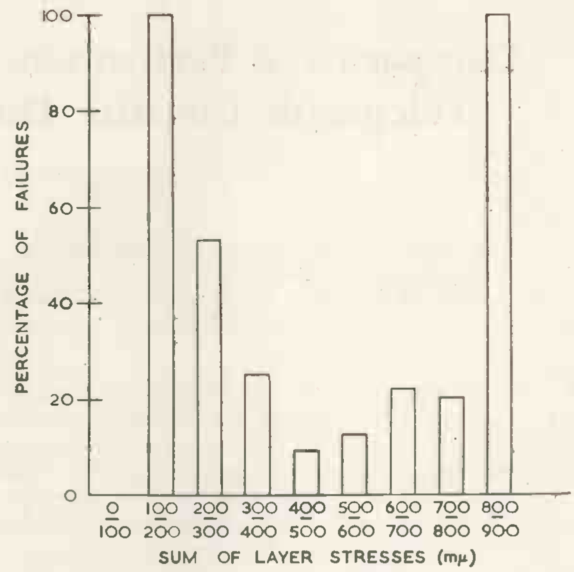


Fig. 11. Relationship between layer stress and failures in chill-shock testing noval valves

TABLE 4
Failures in Service Compared with Chill-Shock Failures

RANGE OF SUM OF LAYER-STRESSES IN GROUP	CHILL-SHOCK TEST		LIFE TEST	
	NUMBER TESTED	NUMBER FAILED	NUMBER TESTED	NUMBER FAILED
150 to 300mμ	30	15	30	0
300 to 450mμ	62	9	62	0
450 to 600mμ	42	4	42	0
600 to 750mμ	30	2	30	0
750 to 900mμ	40	4	40	0
Over 900mμ	28	8	28	0

Conclusions

The chill-shock test has been found to reproduce more nearly the types of base failures which are obtained in service and hence its results are considered more functionally significant than those of the B test.

The B test does, however, detect one type of fault also found in service failures, internal crizzles, but the relative occurrence is low. The main purpose of the B test is to detect manufacturing errors.

A direct correlation between the layer stress in the valve base and sensitivity to chill-shock has been found. But direct experiments have yielded no failures on life test, regardless of degree of layer stress. Thus no corresponding correlation has been found between layer stress and failures on life test. The point merits further consideration.

Acknowledgment

Acknowledgment of permission to publish this article is extended to the Directors of Messrs. Standard Telephones and Cables Limited.

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Production Facilities for Computers

A factory which will shortly provide some 300 000sq ft of floor space has recently been undergoing conversion at West Gorton, Manchester, and will allow a production line assembly for the Ferranti Pegasus and Mercury computers. It is claimed that this factory will be, when complete, one of the largest computer production lines outside the United States of America.

The manufacturing activity is supported by research laboratories both in the factory and at Bracknell, near Ascot. These laboratories employ more than 100 development engineers. In addition, there is a Computer Centre at 21 Portland Place, London, where the main body of the Sales Division and a Computing Service are located.

Ferranti Ltd have already manufactured ten large electronic computers and now have additional orders for some thirty-four of these machines of a total value of over £2m. Most of these orders are for the Pegasus and Mercury computers. Some of these are ready for immediate delivery and it is planned that about 25 computers will have been completed and delivered by the end of next year. The rate of output will be more than doubled in the following year. Although this production is at the moment devoted mainly to computers for use in Science Departments of Universities and Research Departments of Industrial Firms and other organizations, the extensive production facilities will make an important contribution to the manufacture of the joint Ferranti/Powers-Samas Integrated Data Processing equipments.

The new West Gorton factory showing Pegasus and Mercury Digital Computers under construction.

Comparative Performances of High Frequency Radio Telegraph Circuits During Disturbed Conditions

By R. J. Hitchcock*, M.A., A.M.I.E.E.

A statistical examination, covering both sunspot maximum and minimum, shows the varying effects of ionospheric disturbances on several important radio routes. Although high latitude routes deteriorate during these disturbances it is shown that the performance of others may be virtually unaffected and, in some cases, may even improve under such conditions.

It is well known that on any radio path operated on high frequencies by way of the ionosphere there are upper and lower limits to the frequency that can be successfully employed. The upper limit (maximum usable frequency or m.u.f.) is dependent on the geographical position of the radio path, the time of day, the season and the phase of the solar cycle. The lower limit (lowest usable high frequency or l.u.f.) is, in addition to the above factors, dependent on the effective power radiated and the signal-to-noise ratio required at the receiving point.

In general the characteristics of the parameters governing high frequency propagation during normal or quiet ionospheric conditions are reasonably well known and under such conditions, both the m.u.f. and l.u.f. for any particular radio circuit can be calculated with a reasonable degree of accuracy. However, from time to time there occur abnormal conditions resulting in severe interruptions to radio communication between certain parts of the world. On such occasions it is believed that, due to severe reduction in the ionization density of the 'F' region coupled with an apparent increase in height of the layer, the m.u.f.'s on certain circuits are considerably reduced while, at the same time, the l.u.f.'s may substantially increase due to an increase in the attenuation exerted by the 'D' region.

From the practical point of view the communications engineer responsible for the operation of radio circuits is interested in having advance knowledge of:—

(1) The probable day of onset and the duration of disturbed conditions.

(2) The probable effect of such periods on the radio routes operated by his particular communication agency.

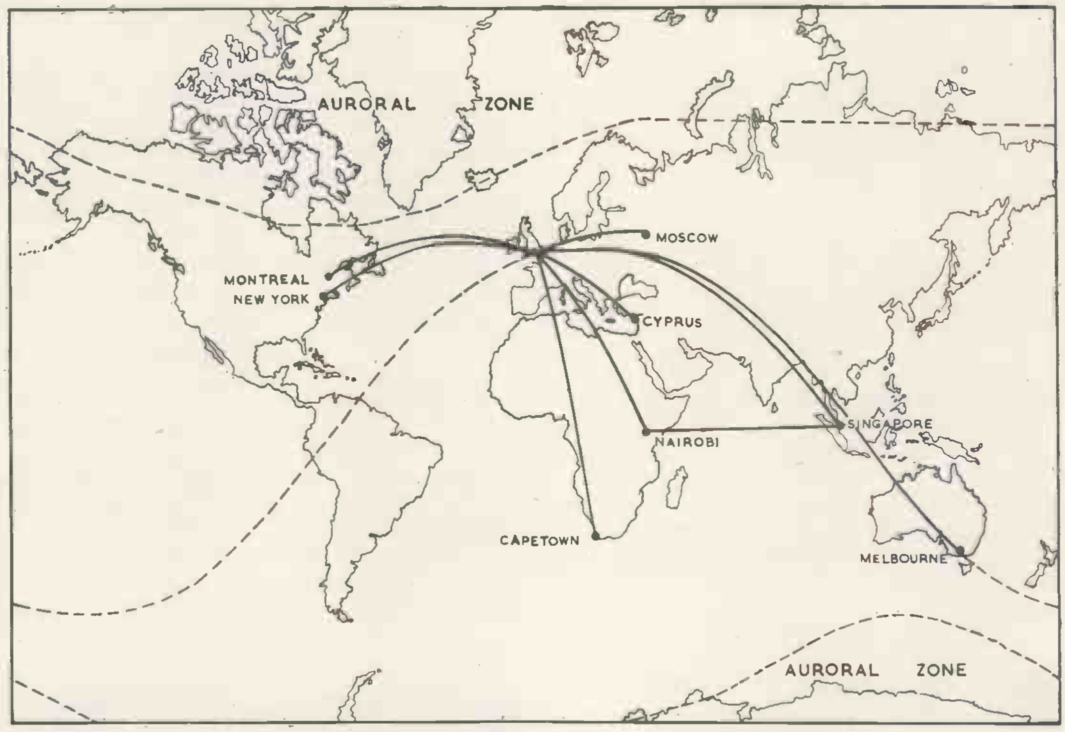


Fig. 1. Great circle paths of the radio telegraph circuits discussed

Research into (1) is currently being pursued in several countries and it is not proposed to discuss this subject here beyond noting that the state of knowledge of solar-terrestrial relations is not sufficiently advanced to ensure any great success at present[1]. Due both to their commercial importance and to the severity and frequency with which they are interrupted, information bearing on (2) has in the main been confined to the performance of circuits between North America and the United Kingdom[2,3]; little information being available on variations in performance experienced by other important radio circuits during these disturbances. It is considered that such information, related to the relevant ionospheric and geomagnetic parameters, is essential if basic storm warnings are, in the future, to

* *Cable and Wireless Ltd.*

be usefully applied to operational needs. In this article a statistical examination is made of the degree of departure in performance experienced by certain important radio circuits during disturbed North Atlantic conditions and the results presented in relation to variations in geomagnetic and ionospheric characteristics, where this latter information was available.

General Considerations

For successful communication on any particular circuit it is necessary to have a 'pass-band' between m.u.f. and l.u.f. in which an operating frequency can be chosen. The pass-band which has diurnal, seasonal and solar cycle variations is naturally different for each circuit. For example, a long distance circuit having considerable differences in local time between the terminals will have a narrower pass-band than a relatively short distance single-hop circuit. Similarly circuits in tropical areas, where m.u.f.'s are very high, have wider pass-bands than circuits of similar length in high altitudes. It therefore follows that, even if the magnitude of an ionospheric disturbance was equal throughout the world, circuits with the narrowest pass-bands under normal conditions would be the most affected and the first to fail when conditions were disturbed. As is well known the severity of ionospheric disturbances is not equal throughout the world and therefore the performance of a radio circuit under storm conditions will depend on both the amount of ionospheric disturbance on its path and the basic characteristics of the pass-band between m.u.f. and l.u.f. under quiet conditions. As this latter factor is obviously unique to each circuit it is difficult to obtain any quantitive values on the relationship between the two effects and in this article no attempt at any such division has been made. A further consideration is of course the relationship between the operating frequency and the m.u.f. and l.u.f.; but this may be an almost negligible factor if alternative frequency allocations are available.

The circuits analysed here are all important traffic carrying systems and therefore, within the limits of present knowledge, well-engineered and equipped. It can be assumed that they are representative of current point-to-point practice in that they have day, intermediate and night frequencies reasonably near to the optimum for the route and that, if during disturbed conditions the normal frequencies fail, efforts would be made to work on other allocations. The only qualification to this is that on the Melbourne-London route, owing to the very restricted nature of the pass-band between m.u.f. and l.u.f. and the fact that large aerial arrays exist for a narrow range of frequencies, changes from the usual 11Mc/s transmissions are infrequent.

Owing to the variations in the diurnal performance of circuits, it was considered desirable to use, as far as possible, only data relating to radio circuits operating for 24 hours per day. Thus, not only was the choice of circuit limited, but as the traffic requirements, in terms of hours per day, tend to vary over a period of years, it was seldom possible to analyse the same circuit at the two extreme phases of the sunspot cycle. Fig. 1 shows the great circle paths of the radio telegraph circuits referred to in this article.

Basis and Method of the Analysis

As radio circuits between North America and the United Kingdom are extremely susceptible to interruption from ionospheric causes the radio engineer has tended to relate his degree of 'disturbance' or 'storminess' to the amount of interruption suffered by one or other of these major radio circuits. The term ionospheric storm has no precise scientific definition and therefore for the purposes of this analysis it was decided to use the yard stick of the radio engineer and consider the performance of the New York-London radio telegraph circuit as the basis upon which to define disturbed conditions.

On all radio circuits it is customary for the receiving station watchkeepers to record the merit of their circuits. In the case of the circuits considered here the various receiving stations all used the following code:—

ZSO—Signals readable without repetition (sent once).

ZST—Signals readable only when repeated (sent twice).

ZSU—Signals unreadable.

The daily reception logs of the circuits under consideration were examined and values of 10, 5 and 0 assigned to each hour of ZSO, ZST and ZSU respectively and an average Figure of Merit 'M' produced for each day. The onset of disturbed conditions was taken as the day of commencement of a major departure from the monthly mean value of M for the New York radio telegraph circuit.

For each disturbance the values of ΔM, i.e. the departures of the daily values of M from the monthly mean, were tabulated for the four days prior to a disturbance, the day of onset and the seven days following. The day of commencement of each storm was designated zero day and by means of the superposed epoch method the average values of ΔM for the commencement days of all the storms (zero days) and for all the days ± 1, ± 2, ± 3 etc. relative to this were plotted for each circuit.

Results of Analysis

PERFORMANCE OF NORTH ATLANTIC PATH DURING DISTURBED CONDITIONS

Superposed epoch curves of the performance of the New York-London radio telegraph circuit covering 19 storms between September 1952 and January 1955, a period near sunspot minimum, and 16 storms between July 1946 and October 1949, a period near sunspot maximum, are shown in Fig. 2(a). It can be seen that in so far as this route is concerned, the effect of disturbances in minimum sunspot years is both more severe and more prolonged than in maximum years.

It is interesting to compare these results with the superposed epoch curves for departures, Δf_oF_2, from monthly mean values of midnight f_oF_2 at Slough for the same series of disturbances and same zero days (Fig. 2(b)). It is seen that although the depression of f_oF_2 is more prolonged in the minimum period the greatest magnitude of the depression occurs at sunspot maximum. It is well known that the poorer performance of the radio path in sunspot minimum is due to the already depressed state of the m.u.f., accompanied by a very much smaller reduction of the l.u.f., normally associated with this phase of the solar cycle. Under these conditions the radio circuit is unable to accept very great changes in either of these parameters.

Superposed epoch curves for departures, ΔC, from the monthly mean of the Abinger daily magnetic value C based on the same zero days as for the New York circuit are shown in Fig. 2(c). The relationship between the mean values of ΔM and the mean values of ΔC is shown in Fig. 2(d).

PERFORMANCE OF OTHER LONG DISTANCE RADIO PATHS DURING DISTURBED CONDITIONS ON THE NORTH ATLANTIC

For the sunspot minimum period September 1952 to January 1955, the performance of the radio telegraph circuits from Melbourne to London (both short and long routes) and from Nairobi to London were analysed by the superposed epoch method using the same zero days as for

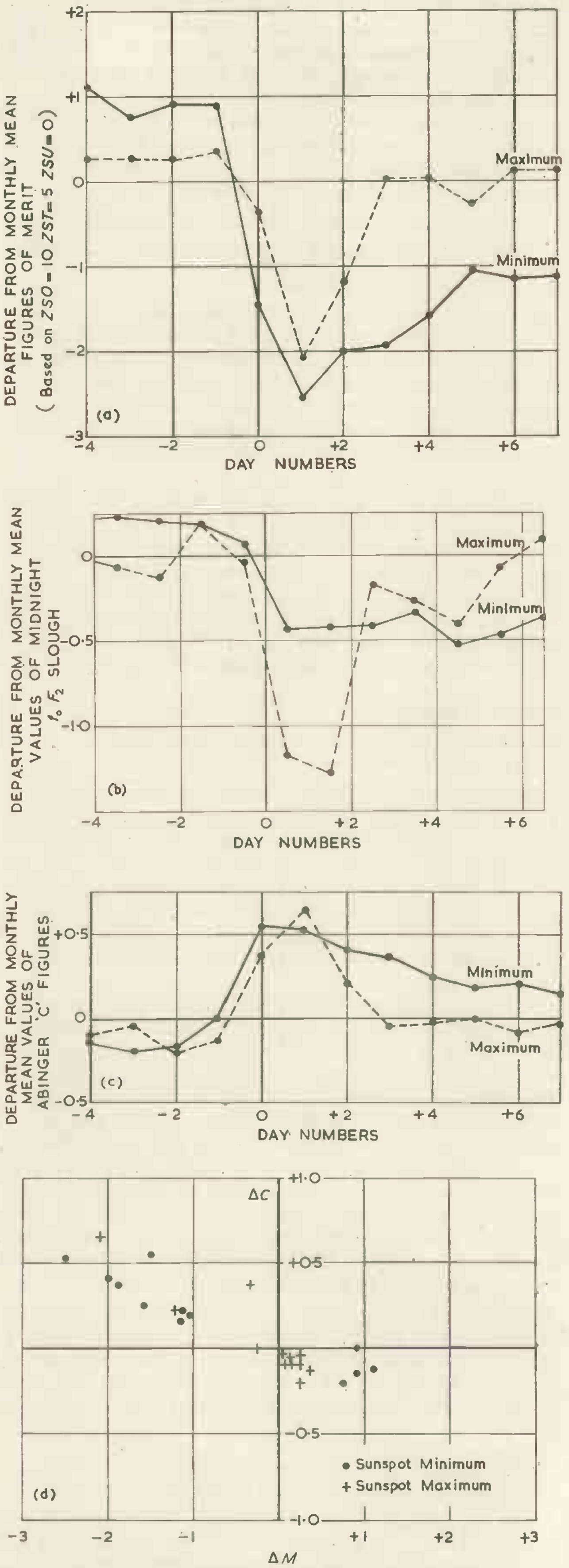


Fig. 2. Superposed epoch curves for the New York-London radio telegraph circuits and the relationship between mean values of $\triangle M$ and mean values of $\triangle C$ for each day number

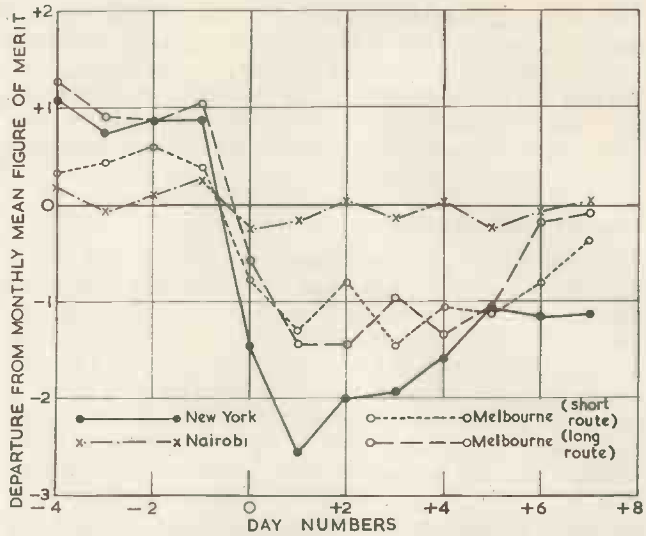


Fig. 3. Superposed epoch curves for the New York to London, Melbourne to London, and Nairobi to London radio telegraph circuits based on the first day New York showed a storm effect

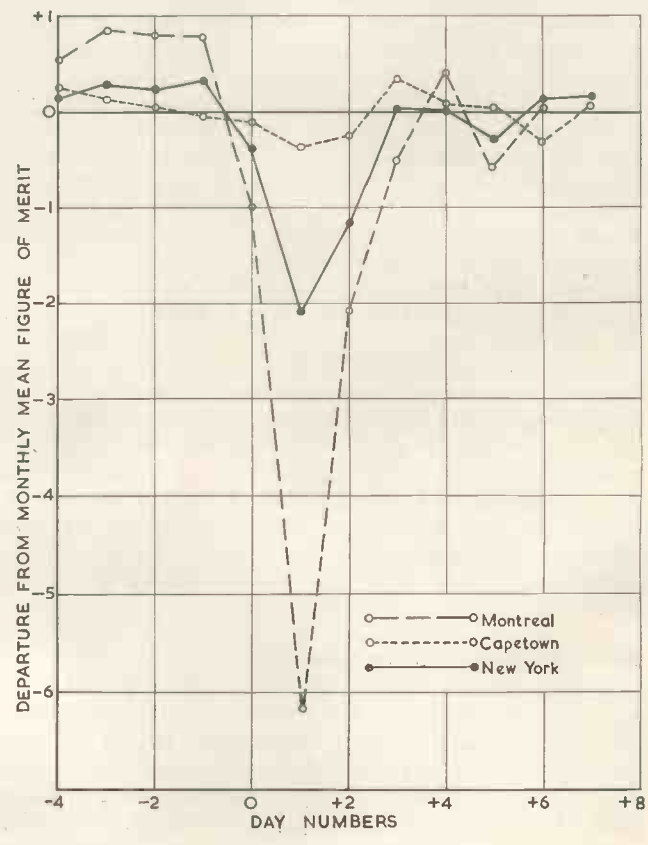
Data covers 19 ionospheric disturbances during the sunspot minimum period (September 1952-January 1955)

New York. In Fig. 3 the results have been compared with the characteristics of the New York circuit illustrated in Fig. 2. It will be seen that both the Australian circuits follow much the same trends as the New York circuit although the magnitude of the effect is somewhat less. The Nairobi circuit on the other hand is almost unaffected.

For the sunspot maximum period, July 1946 to October 1949, similar curves (Fig. 4) were drawn for the Montreal-

Fig. 4. Superposed epoch curves for the New York to London, Montreal to London, and Capetown to London radio telegraph circuits based on the first day New York showed a storm effect

Data covers 16 ionospheric disturbances during the sunspot maximum period (July 1946-October 1949)



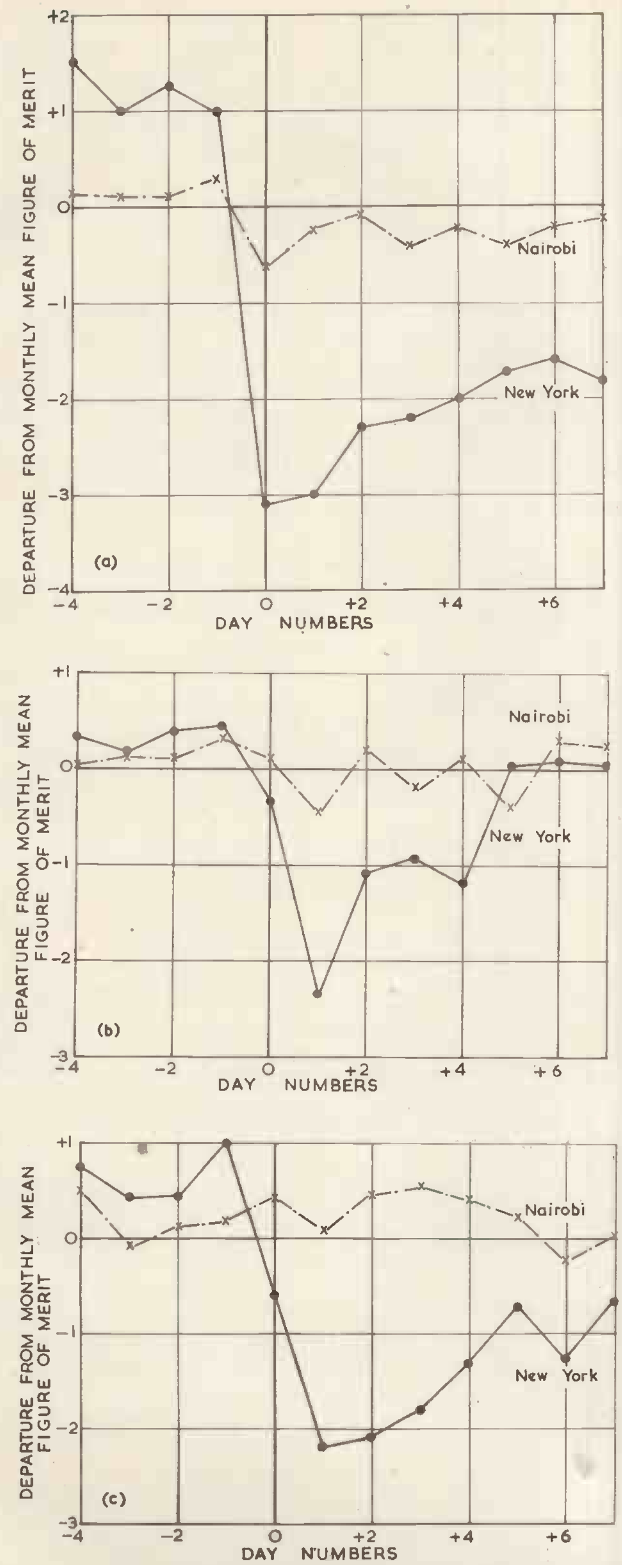


Fig. 5. Superposed epoch curves for the New York-London and Nairobi-London radio telegraph circuits based on the first day New York showed a storm effect (September 1952-January 1955)

(a) EQUINOX (February-March-April) (August-September-October) 9 storms

(b) SUMMER (May-June-July) 4 storms

(c) WINTER (November-December-January) 5 storms

London and Capetown-London circuits. The Montreal circuit, with its more northerly great circle path and greater auroral absorption, is considerably more affected by these disturbances than is the New York circuit, while, as with the Nairobi circuit in the sunspot minimum, the Capetown circuit is almost unaffected.

Seasonal Variations

In view of the fact that in a yearly analysis the Nairobi-London circuit appeared to show practically no storm effects (Fig. 3), a further series of superposed epoch curves for this circuit and the New York circuit, based on the same zero days, were plotted in terms of winter (November, December, January) summer (May, June, July) and equinox (February, March, April, August, September and October). This unavoidably reduced the number of disturbances associated with any one curve; in the case of the summer for instances, only four disturbances were available.

It will be seen from Fig. 5(a) that during equinox months the Nairobi circuit does, in fact, deteriorate at the same time as the New York circuit and that in summer (Fig. 5(b)) there is also a tendency for the circuit to become erratic. When the winter months were analysed, however, it was found that out of six major disturbances on the

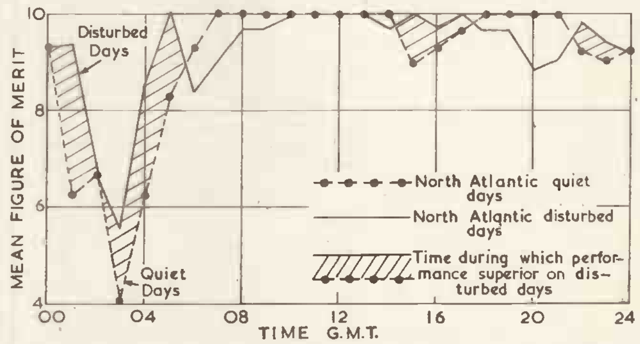


Fig. 6. Diurnal performance of the Nairobi-London radio telegraph circuit during North Atlantic quiet and disturbed days, winter, sunspot minimum

North Atlantic route the performance of the Nairobi circuit improved in all but one case, the single exception being the disturbance commencing 13th November, 1953. The characteristics of this circuit for the five remaining storms are shown in Fig. 5(c). The average hourly figures of merit for the three days prior to these five winter storms and the three days after onset are shown in Fig. 6. It will be seen that when the North Atlantic route is disturbed, the enhancement in performance of the Nairobi circuit is mainly centred on 03.00 GMT, a period when the winter values of f_oF_2 at Nairobi reach a minimum and the southern end of the circuit is controlling. An increase in circuit performance at this time would infer an increase in f_oF_2 (i.e. m.u.f.) at the Nairobi end. Such a phenomenon has been reported by Appleton and Piggott[4], who have shown that equatorial stations exhibit a quite well-marked increase in f_oF_2 roughly simultaneous with the temperate zone depression. As the characteristics of the Nairobi-London circuit are influenced by more than one ionospheric reflection point, the apparent lack of any enhancement in performance during other seasons may be due to the controlling effects of the more northerly parts of the ionosphere. Unfortunately insufficient data was available from ionospheric stations in Central Africa to make any comparison with this circuit performance possible.

The station nearest to the geographic and geomagnetic equators for which the author had detailed ionospheric information was Singapore. Superposed epoch curves for departures from the monthly mean noon and midnight values of f_oF_2 for this station covering the same disturbance and zero days as for the Nairobi circuit are shown in Fig. 7. The curves have been corrected to GMT day numbers and it will be seen that although there is no obvious enhancement after day 0 there appears to be a general increase from (D – 2) to (D + 2). (It would appear that the sudden drop in f_oF_2 commencing (D + 2) to (D + 3) is the late manifestation of the North Atlantic disturbances.)

Unfortunately, data was not available for a circuit operating over a totally equatorial path for 24 hours per day, but only for a circuit Singapore to Nairobi, operating during the hours of local daylight. Probably due to the limited hours of this service, a detailed analysis failed to show any seasonal effects; the performance of the day

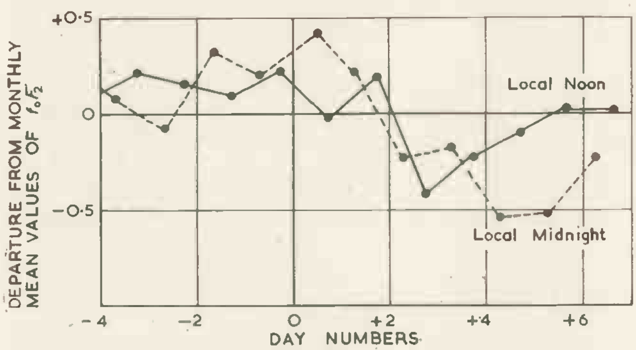


Fig. 7. Superposed epoch curves for $\triangle f_oF_2$ at Singapore with zero day based on the first day the New York-London radio telegraph circuit showed a storm effect

Data covers 19 disturbances during the sunspot minimum period

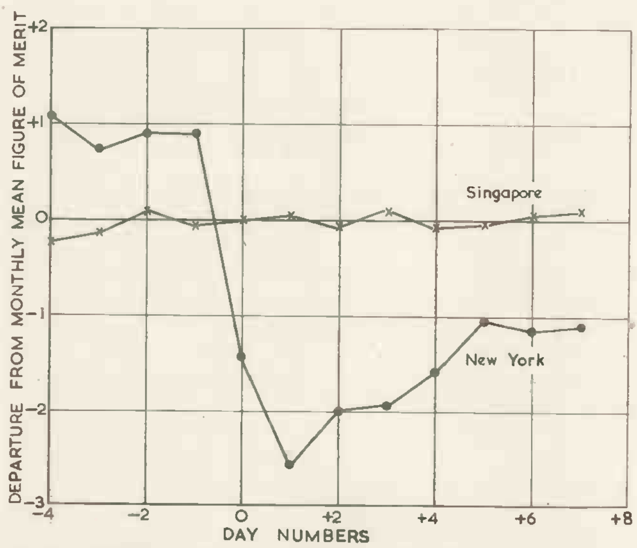


Fig. 8. Superposed epoch curves for the New York to London and Singapore to Nairobi (day frequency only) radio telegraph circuits based on the first day New York showed a storm effect (September 1952-January 1955)

frequency only, without regard to season is shown in Fig. 8.

Performance of Shorter Circuits During Disturbed Conditions on the North Atlantic

The circuits so far considered have been sufficiently long for it to be assumed that propagation is maintained by more than one reflection from the ionosphere and in the cases of New York, Montreal and Australia the local time differences between the terminals together with the relatively high latitudes of the radio paths have resulted in circuits with comparatively restricted 'pass-bands' between m.u.f. and l.u.f.

Fig. 9 shows superposed epoch curves for the shorter distance single-hop circuits, Cyprus-London and Moscow-London, related to the disturbed periods on the North Atlantic over the last sunspot minimum. It will be seen that the Cyprus circuit is almost unaffected. This is a general characteristic of short and middle distance European routes where the pass-band between m.u.f. and l.u.f. is normally sufficiently wide to allow considerable fluctuations in either parameter without degrading circuit performance. A seasonal analysis of this circuit failed to disclose any significant effects.

The performance of the Moscow circuit shows the effect of moving the ionospheric reflection point some 10° northwards, to approximately 55°N, the control point now being comparatively near the north auroral zone (Fig. 1). It will be seen that although there is now a definite correlation in performance with the North Atlantic route, the magnitude of the deterioration is still relatively small.

The performances of the Moscow and Melbourne (short route) circuits are particularly interesting in that they afford a comparison of the effects of the basic quiet state characteristics of a circuit with the effects of geographical

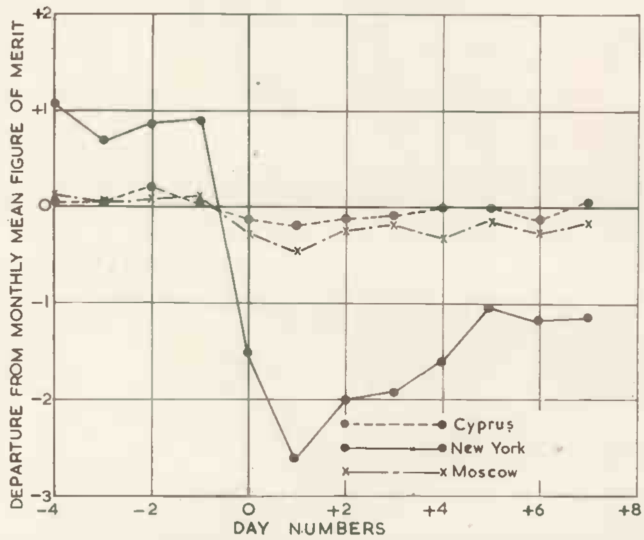


Fig. 9. Superposed epoch curves for the New York to London, Cyprus to London, and Moscow to London radio telegraph circuits based on the first day New York showed a storm effect

Data covers 19 storms during the sunspot minimum period (September 1952-January 1955)

location. Both routes enter the United Kingdom from an easterly direction, and while the Moscow circuit is the more northerly and nearer to the auroral zone it is relatively short, with a broad 'pass-band' between m.u.f. and l.u.f. under quiet conditions. The Melbourne circuit on the other hand is considerably more difficult under normal conditions with a low m.u.f. and high l.u.f. The latter characteristic would appear to be responsible for the greater degree of deterioration suffered by the Melbourne circuit (Fig. 3) compared with the more northerly Moscow circuit.

Conclusions

The degree of deterioration suffered by the New York-London radio-telegraph circuit during disturbed conditions on the North Atlantic has been compared with the departure in performance experienced on other important radio circuits and with variations in certain ionospheric and magnetic characteristics over the same series of storm epochs.

From the limited number of circuits analysed it may be concluded that although certain circuits deteriorate when the North Atlantic Routes are disturbed, many are virtually unaffected and some appear to improve under these conditions. This improvement in performance is found on the Nairobi to London circuit during the winter (northern hemisphere) and may well be a feature of other circuits with control points in the equatorial zone.

It is believed that information on the comparative performance of circuits during ionospheric disturbances and the fact that many do not necessarily exhibit the same characteristics as the North Atlantic route, is not only of value to communication engineers engaged in such tasks as the planning of relay routes but, assuming an advance in the basic forecasting of storms, should also be of value in the successful application of these forecasts to actual circuits.

Acknowledgments

The author wishes to thank the Post Office for making available the reception logs of radiotelegraph circuits received in the United Kingdom and for permission to publish data relating to the New York, Melbourne and Moscow circuits for the years after 1951. He also wishes to acknowledge the assistance received from colleagues, particularly Mr. H. M. Smith, in the preparation of the analyses and is indebted to the Engineer-in-Chief of Cable and Wireless Ltd. for permission to publish the article.

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Jig Boring Machine with Electronic Positioning

The new 'Matrix' No. 150 jig boring machine is a development of the well-tried 'Matrix' No. 50. Two important features of the new machine are the use of an electronic linear measuring system for positioning the table and the incorporation of extensive push-button controls.

These new features, which have been developed by Mullard Research Laboratories, make setting up simpler than before, and are also a necessary step in the evolution of a machine with fully automatic positioning. Settings can be made in either direction to within 0·001in.

The Electronic Positioning System

The internal standards of length in the 'Matrix' No. 150 take the form of steel buttress scales manufactured by the Coventry Gauge and Tool Company Ltd. Each buttress scale consists of a solid steel bar having a sawtooth form cut into it of accurate 0·1in pitch. A step, machined lengthwise, reveals the form consisting of upright and sloping edges. The maximum adjacent pitch error is 0·00002in.

Interpolation between the 0·1in steps of the master scale is accomplished optically.

A glass scale a little over 2in long carries upwards of 1 000 alternate opaque and transparent bars, equally spaced. These are scanned by a moving line of light and the projected image reduced optically by a factor of 20 so that 1 000 bars fit exactly between adjacent vertical edges of the buttress scale. As the table is driven, the bars are counted as they emerge from behind a vertical edge by means of photo-multiplier tubes and an electronic counter. The number counted is subtracted electronically from the required number and the error displayed on a row of dial lamps. When the error is reduced to zero the central lamp lights and the table is clamped by operating a pushbutton.

Since 1 000 units are counted when the table moves 0·1in each unit represents 0·0001in. To set the table it is only necessary to traverse within 0·1in of the required position set three dials marked 'hundreds', 'tens', and 'units', to the required number of ten-thousandths of an inch, and traverse slowly to reduce the displayed error to zero. The scales for both directions of table positioning are identical. Provision has been made so that the buttress scale can be shifted through 0·1in to allow an arbitrary origin of co-ordinates to be selected when the work piece has been clamped to the table.

Flying Line Scanner

The opaque and transparent bars on the glass interpolating scale are scanned four times per second by a line of light produced by a specially designed scanning mechanism. The light originates from a line filament prefocus car headlamp bulb. A condensing lens provides a beam of parallel light wide enough to illuminate the glass scale over its entire length. The scale is, however, shielded from the light, which can only reach it via one of four glass rod lenses mounted in the centres of four planet gears which form part of an epicyclic gear train.

The scanner makes use of the fact that in an epicyclic gear train, if the sun gear and carrier are driven at the correct relative speeds, the planet gears while moving round the sun, do not rotate about their own axes. With large diameter ball races for the planet bearings, the centre of the gears can be bored out and rod lenses mounted in them at right-angles to the gear axes. While the rod lenses move in a circular path, they remain upright, and by using a lens longer than $\frac{3}{8}$in, an arc of motion can be selected such that the 2in × $\frac{3}{8}$in scale is fully scanned from end to end.

Electronic Equipment

With the exception of the photo-multiplier tubes and their associated circuits all the electronic part of the system is mounted in a console separate from the jig boring machine.

In operation of the positioning system the last three digits of a dimension are set up on three 10-position dials. Each dial switch is connected to a diode matrix which applies to the appropriate electronic counter the voltages necessary to set it to the number dialled. (The counters are of the four-stage binary type, with feedback).

Pulses from either of the photo-multiplier tubes are fed to the counter which, at the end of a scan, registers the difference between the number set up on the dials and the actual number of ten thousandths of an inch through which the table has moved, i.e. it registers the error in the position of the table. At each scan a fresh measurement of position is made, independent of the previous scan. There is no loss of accuracy or of origin of co-ordinates due to power failures or when the measuring system is switched off overnight.

The output of each counter passes via another matrix to a series of cold-cathode trigger tubes which are used to switch on indicating lamps. The purpose of these second three matrices is to break down the error numbers into a series of groups. Thus if the error is 100 or more units one of the lamps lights. If between 50 and 99 units, the adjacent lamp lights, and so on in steps of decreasing size until, as the correct setting is approached, indication is given in steps of one unit, i.e. 0·0001in. The error is in fact displayed on a row of 17 lamps. The centre lamp is for zero error, and the end lamps for an error of 100 or more units in either direction.

Push Button Controls

The electronic console also houses contactor mechanisms which, in conjunction with push buttons on the jig borer itself, control certain of the machine functions. For the rapid motion of the table in either ordinate, non-locking push buttons are provided which give a table speed of 36in per minute. On release of these buttons the table automatically stops and is clamped up. For slow motion, locking buttons are provided which give an adjustable table speed from 0·6in to 6in per minute. Locking buttons provide hand control of the table from a hand wheel giving 0·01in of table motion per turn. A stop-and-lock button takes the table out of slow motion or hand control and clamps the table. In hand control, and when the positional check buttons are pressed, the indications of the measuring system are displayed to the operator to allow setting and checking of table position. The boring spindle is started and stopped by push buttons. A master stop button switches off the entire console and machine with the exception of the table clamps.

The table setting motor is coupled to the tables by magnetic clutches. The speed change gear box is of constant mesh type with magnetic clutches coupling in high or low speed.

Waveguide Surface Finish and Attenuation

(Part 1)

By J. Allison*, B.Sc., Ph.D., and F. A. Benson*, M.Eng., Ph.D., A.M.I.E.E., M.I.R.E.

It has recently been established that the discrepancies between theoretical and experimental performances of waveguide surfaces are largely due to the effects of surface roughness. Information is presented here on the surface finishes of drawn, mechanically-lined, sprayed, cast, electroplated, electropolished, chemically-polished and electroformed waveguides.

LOSSES in waveguides and cavity resonators produced by the flow of induced currents in the metal walls can be calculated from formulæ derived on the assumption that the surfaces are perfectly smooth. Several investigators[1-16] have reported, however, that at microwave frequencies measured losses are considerably greater than those calculated from the formulæ. For example, Vivian[1] found that the attenuation produced by brilliantly-finished copper waveguide at 9 000Mc/s was about 30 per cent higher than calculated and he reports losses in standard copper waveguide in the region of twice the theoretical values. Again, Morgan[2] states that experimental studies on cavity resonators and waveguides have revealed losses 10 to 60 per cent higher than calculated values.

It has now been established that the losses in a conductor at microwave frequencies can be correlated with the surface finish of the metal. In recent years the authors, and a few other investigators, have been concerned with the effect of surface roughness on the attenuation of certain drawn, polished, electroplated, electroformed, mechanically-lined, sprayed and cast waveguides. The important results obtained here.

Attenuation Measurements

The attenuation of a length of waveguide may be determined by measuring the voltage-standing-wave ratio (v.s.w.r.) in the guide when terminated by a short-circuit. The attenuation may be expressed as

$$\alpha = 10 \log [(r + 1)/(r - 1)] \text{ dB} \quad \ldots\ldots\ldots (1)$$

where r is the v.s.w.r. measured as a quantity greater than unity.

The simplest system (Fig. 1) employs an unmodulated source, the probe signal from the standing-wave indicator being detected by a crystal and recorded on a d.c. meter. The sensitivity may be increased by means of a d.c. amplifier between the detector and the meter or more conveniently by modulating the source and feeding the rectified signal to a selective amplifier. Such a scheme enables small attenuations to be measured, thus eliminating the need for excessively long lengths of waveguide which are necessary with the simple method.

The difficulties encountered in measuring high v.s.w.r.'s by the direct method may be overcome by using a technique described by Roberts and von Hippell[17,18]. The distance w between two points of equal field strength on either side of a minimum in the standing wave is measured accurately. If the value of the field strength at these points is k times the minimum, k being chosen arbitrarily to provide a convenient meter deflexion, the v.s.w.r. can be expressed as

$$r = [k^2 - \cos^2 (\pi w/\lambda_g)]^{\frac{1}{2}}/(\sin \pi w/\lambda_g) \quad \ldots\ldots (2)$$

where λ_g is the guide wavelength.

Calculation of Attenuation in Waveguides Neglecting Surface Roughness

A general formula for calculating waveguide attenuation has been given by Kuhn[19]. For the normal H_{01} mode in a rectangular waveguide, which has been used during the present investigations, the attenuation is given by

$$\alpha = [(c/\sigma) \cdot (\mu_1/\mu) (\epsilon/\mu)^{\frac{1}{2}}]^{\frac{1}{2}} \frac{\lambda_g}{b \cdot (\lambda_e)^{3/2}} \cdot [(\lambda_e^2/\lambda_{cr}^2) + (b/2a)] \quad \ldots\ldots\ldots\ldots (3)$$

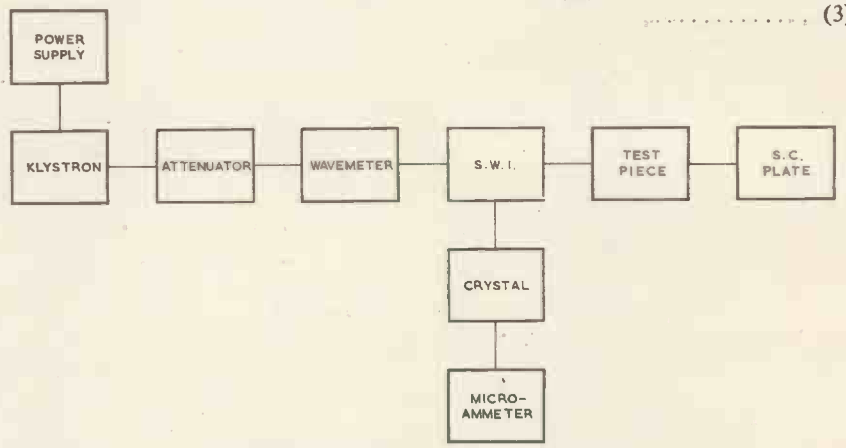

Fig. 1. Equipment for attenuation measurements

where a and b are the short and long internal dimensions of the waveguide respectively, c is the velocity of light, σ and μ_1 are the conductivity and permeability of the waveguide wall metal respectively, ϵ and μ are the permittivity and permeability of the dielectric (usually air) and λ_g, λ_e and λ_{cr} are the guide, free-space and critical wavelengths respectively.

In deriving equation (3) Kuhn assumes that the surfaces are smooth and that the disturbance of the fields due to losses in walls of relatively high, but finite conductivity is so small that it can be neglected.

Thus, if α is measured at a given frequency, by the method already outlined, the effective value of σ can be calculated for a non-magnetic material and compared with the measured d.c. value.

Measurement of D.C. Resistivity of a Waveguide Sample

To compare theoretical and experimental performances of waveguide surfaces, figures for the d.c. resistivities must be known. Reliable values can only be obtained from measurements where resistances of known lengths together with areas of cross-section are determined. It is difficult to find the cross-sectional area of a long length of commercial guide, where dimensions have been found sometimes to be as much as 0·015in under nominal. It can be estimated, however, by determining volumes of several short known lengths[14].

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Surface-Roughness Measurements

The internal surfaces of some tubes which had their attenuations measured were examined to try to establish a correlation between experimental and theoretical attenuation values in terms of surface roughness. Such examinations also reveal information about surface irregularities of standard waveguides and give some idea of the degree of surface finish desirable.

Several ways of evaluating surface roughness have been described previously[16]. There are objections to the use of either the stylus type of instrument or the optical methods in the case of waveguide specimens. They do not give the ratio of the actual to ideal length of a surface which is shown later to be the important information required for the present studies. Consequently the surfaces were examined by polishing and microscopic inspection where short lengths are taken from the waveguide sample and are electroplated with a metal contrasting in colour without damaging the original surface. The lengths are then cut along the particular sections required and mounted in 'Perspex' moulds to facilitate polishing of the edges.

A projection microscope is used for the examinations. The ground-glass screen of the microscope is replaced with clear glass and tracing paper so that the whole surface contour of one side of a waveguide section can be traced and accurate measurements can then be made of the length of the contour for comparison with the corresponding ideally smooth surface. For plated surfaces the corresponding base-metal surface tracings can also be obtained, thus giving a measure of the levelling, if any, produced by the plating. Such measurements need to be limited to a small number of cross-sections, but different sections from the same length of waveguide give surprisingly similar results.

Skill and experience are required to produce relatively scratch-free surfaces for examination. A different technique using electropolishing has, therefore, been adopted recently where the protective plating need not be contrasting in colour, but it must electropolish in the same bath as the sample. After the usual mounting and a preliminary polishing with emery paper, the specimen is electropolished in a suitable bath (Fig. 2).

Attenuation Formula Which Takes Account of Surface Roughness

Since, in general, many irregularities on the waveguide surfaces have magnitudes much greater than the corresponding skin depths it seems reasonable to assume that the increase in measured attenuation over the calculated value for a particular waveguide is due to the increases in lengths of path traversed by currents. Making this assumption, equation (3) may be modified to account for surface roughness by introducing three additional factors K_{T1}, K_{T2} and K_p. K_{T1} and K_{T2} are the ratios of the actual lengths to the ideal lengths for the long and short sides of the waveguide respectively, transverse to the axis. K_p accounts for the effects of surface roughness in the longitudinal direction and is defined in a similar manner to K_{T1} and K_{T2}.

The modified attenuation expression for the H_{01} mode is

$$\alpha' = [(c/\sigma) \cdot (\mu_1/\mu) \cdot (\epsilon/\mu)^{1/2}]^{1/2} \cdot \frac{\lambda_g}{b \cdot (\lambda_e)^{3/2}} \left[\left\{K_{T2} + (b/2a) \cdot K_{T1}\right\}(\lambda_e^2/\lambda_{cr}^2) + \frac{K_p \cdot b}{2a}\left\{1 - (\lambda_e^2/\lambda_{cr}^2)\right\}\right] \quad \ldots\ldots\ldots (4)$$

The ratio of the calculated attenuation which takes account of surface roughness (α') to that calculated assuming perfectly flat surfaces (α) becomes therefore for the H_{01} wave,

$$\alpha'/\alpha = \frac{[K_{T2} + (b/2a)\, K_{T1}](\lambda_e^2/\lambda_{cr}^2) + bK_p\left\{1 - (\lambda_e^2/\lambda_{cr}^2)\right\}/2a}{(\lambda_e^2/\lambda_{cr}^2) + (b/2a)} \quad \ldots\ldots\ldots (5)$$

Thus, if the values of the surface-roughness factors are determined for a specific guide the actual attenuation in it can be accurately estimated. Results of attenuation and roughness measurements have repeatedly agreed very closely with equation (5).

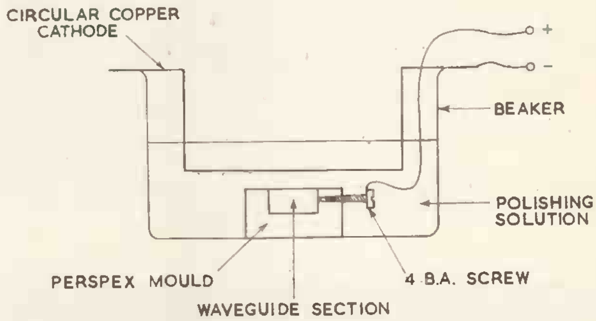


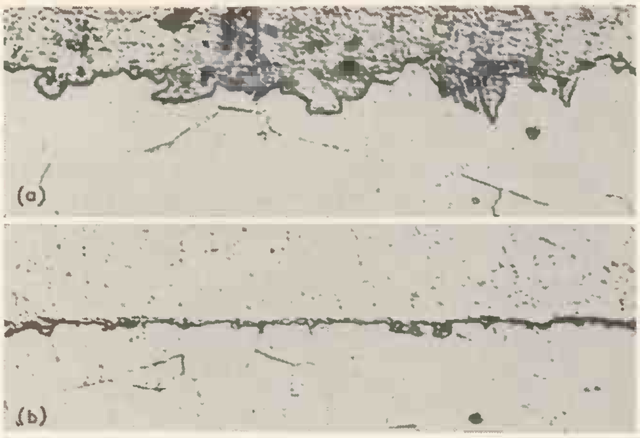
Fig. 2. Electropolishing apparatus for microscopic examination

Morgan[2] has determined theoretically the power dissipated by eddy-currents in a metallic surface at microwave frequencies in the presence of regular parallel grooves or scratches whose dimensions are comparable with the skin depth. Experience has shown that the special kind of surface studied by Morgan is not found in practice and as expected his results are higher than experimental values obtained for surfaces with random irregularities.

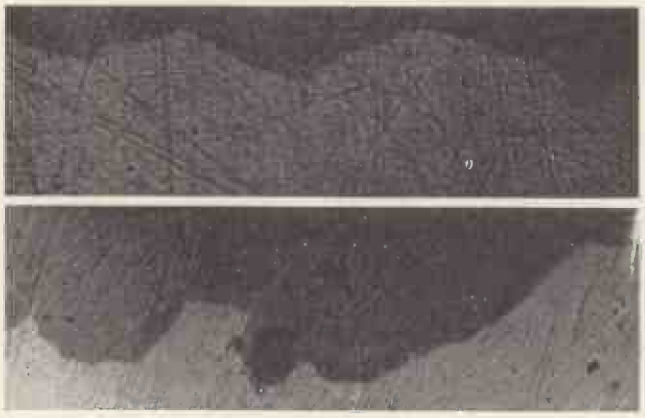
Drawn Tubes

The authors original investigations were concerned with drawn surfaces of brass and copper at frequencies near 10 000Mc/s. The effective resistivities of the surfaces were found from attenuation measurements and compared with their measured d.c. resistivities.

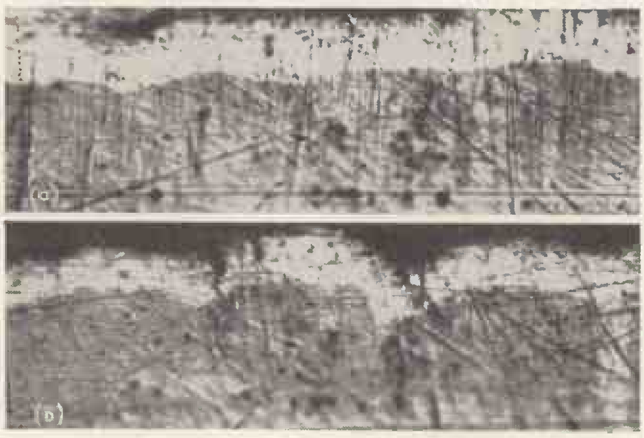
Measured values of d.c. resistivity for brass waveguides were always near the expected figure, but for copper, measured d.c. resistivities were sometimes nearly three times the order expected. Spectrographic analysis of one sample of such waveguide having high resistivity showed that the material was deoxidised arsenical copper. That the sample showed high resistivity is, therefore, not surprising. Some early reports about copper waveguides having exceedingly high ratios of measured-to-calculated attenuations probably resulted because d.c. resistivities had high values and were not measured. In fact, the ratios of measured-to-calculated values recorded by the authors were far from being as large as anticipated from such reports. For normal-drawn copper tubes the measured values were from 10·2 to 12 per cent greater than the theoretical values, the corresponding figures for normal-drawn brass tubes being 6 to 7·4 per cent. Precision-drawn copper and brass tubes showed discrepancies of only 3·4 per cent and 1·4 per cent respectively. Brass tubes always give attenuations nearer to the calculated values than copper tubes for a particular kind of guide, possibly due to the softer nature of the copper and its consequent greater liability to scratching and/or to the higher friction between copper and the drawing dies.



Figs. 3(a) and (b). Normal drawn brass waveguide surface from 1in × ½in guide
Sections perpendicular to waveguide axis. Magnification: 2 000. Brass in lower parts of photographs.



Figs. 4 and 5. Normal drawn brass waveguide surface from 3in × 1in guide
Sections perpendicular to waveguide axis. Brass in lower parts of photographs.



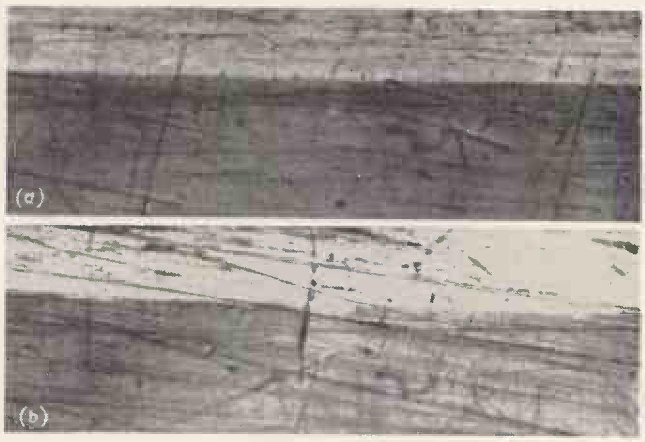
Figs. 6(a) and (b). Precision drawn copper waveguide surface from 1in × ½in guide
Magnification: 940. Copper in lower parts of photographs.

Typical photomicrographs for normal commercial brass waveguide with internal dimensions 1in by 0·5in are shown in Fig. 3. Irregularities having magnitudes up to about 400μin and following no definite patterns are found. A very rough part of a surface is frequently followed by a fairly smooth section.

Typical photomicrographs for 3in by 1in guides are shown in Figs. 4 and 5. For brass the greatest irregularities observed here had magnitudes of about 800μin and the ratio of lengths of actual surfaces to those of ideally smooth surfaces, obtained from measurements on photomicrographs, varied from 1·10 to 1·73. For copper guides of this size many irregularities have magnitudes of 1 300μin.

Surfaces of best precision-drawn waveguides are remarkably good, particularly brass ones, as is suggested by the results of attenuation measurements. The main surface deformities from transverse sections of precision-drawn 1in by 0·5in copper guides are 'draw' lines ranging from ·001in deep by ·0005in wide in the long sides to ·0005in deep by ·001in wide in the short sides. The photomicrographs of Fig. 6 show a typical transverse section and a section including 'draw' lines. Longitudinal sections of

Fig. 7. Precision drawn copper waveguide surface from 1in × ½in guide
Section from long side parallel to waveguide axis. Magnification: 940. Copper in lower part of photograph.



Figs. 8(a) and (b). Precision drawn brass waveguide surface from 1in × ½in guide
Sections perpendicular to waveguide axis. Magnification: 940. Brass in lower parts of photographs.

drawn tubes are much smoother than transverse ones, the only irregularities being very shallow grooves about ·001in long by ·0003in deep, although there are occasional places where the microprofile is rough (Fig. 7). Precision-drawn brass guide is the best of all drawn tubes examined. Transverse-section contours of such tubes are quite flat only undulating gently for the most part (Fig. 8(a)). Very occasionally there are more severe changes in surface level. A few 'draw' lines about ·0003in deep are observed (Fig. 8(b)). There are no grooves or elevations in longitudinal-section brass waveguide surfaces (Fig. 9).

Waveguide surfaces should be protected from corrosion. Losses in guides of brass, copper and silver-plated brass increased by up to 22 per cent after they had been placed out-of-doors in Sheffield for 19 weeks and the losses produced by electropolished copper wires left out-of-doors for several months increased by 40 per cent[6]. Lending[13] points out that although the increased losses have generally been

Fig. 9. Precision drawn brass waveguide surface from 1in × ½in guide
Section from short side, parallel to waveguide axis. Magnification: 940. Brass in lower part of photograph.

attributed to increased roughness of the base metal the corrosion product materially affects the resistivity particularly in the millimetre-wave region. Lacquers can be used for protection without appreciably increasing the loss[6].

Lending suggests that tellerium-copper should be used for millimetre-wave work in place of the common coin silver, because it is a better conductor and has better corrosion characteristics.

Drawn aluminium waveguides are light, but have a high ratio of measured-to-calculated attenuations[20]. Both aluminium and magnesium components require protective coatings because of their ageing and corrosion characteristics[13].

The properties of a sample of 1in by 0·5in German-silver drawn waveguide have been determined by the authors. The measured attenuation was 0·346dB at a frequency of 9 620Mc/s and the roughness coefficients were $K_{T1} = 1{\cdot}002$, $K_{T2} = 1{\cdot}003$, $K_p = 1{\cdot}006$. Because of its high resistivity ($30{\cdot}85 \times 10^{-6}\Omega$-cm) the waveguide has a much greater attenuation than drawn copper or brass tubes and its only advantages are its strength and non-corrosive nature.

The measured surface-roughness factors for various types of waveguide sample are given in Table 1, together with figures for the ratio α'/α which have been calculated, using the experimental figures for the roughness coefficients, for the H_{01} mode in a rectangular waveguide with internal dimensions 1in by 0·5in. The frequency has been taken as 10 000Mc/s.

Mechanically-Lined Waveguides

Nearly perfect waveguide surfaces can be obtained by rolling silver on to copper, forming mechanically silver-lined tubes. This technique eliminates all nodular effects produced by silver plating and ensures a perfectly flat, highly conducting interior surface (Fig. 10). The silver

TABLE 1

Attenuation Values and Surface Roughness Factors for Various Types of Waveguide.

TYPE OF WAVEGUIDE	K_{T1}	K_{T2}	K_p	α'/α
Electropolished copper	1·002	1·001	1·004	1·002
Electropolished brass	1·004	1·006	1·005	1·004
Normal-drawn brass*	—	—	—	1·060-1·074
Normal-drawn copper*	—	—	—	1·102-1·120
Precision-drawn copper	1·020	1·015	1·006	1·012
Precision-drawn copper*†	—	—	—	1·034
Precision-drawn brass	1·002	1·001	1·001	1·001
Precision-drawn brass*†	—	—	—	1·014
Acid-copper electroplated	1·625 max.	1·625 max.	—	—
Bright-copper electroplated	1·002	1·002	1·002	1·001
Silver-plated from cyanide bath*	—	—	—	1·580
Bright-silver electroplate	1·012	1·004	1·006	1·007
Chemically-polished copper	1·002	1·005	1·005	1·003
Chemically-polished brass	1·003	1·002	1·004	1·003
Electroformed copper on expendable alloy mandrel, Sample 1	1·018	1·025	1·013	1·017
Electroformed copper on expendable alloy mandrel, Sample 2	1·036	1·067	1·049	1·051
Electroformed copper on stainless steel semi-permanent mandrel	1·010	1·021	1·001	1·008
Electroformed copper on stainless steel permanent mandrel, Sample 1	1·006	1·005	1·003	1·003
Electroformed copper on stainless steel permanent mandrel, Sample 2	1·001	1·002	1·001	1·001
Drawn aluminium*	—	—	—	1·115
Sprayed zinc on steel mandrel	1·016	1·009	1·014	1·013
German silver	1·002	1·003	1·006	1·007

* *These values of α'/α have been determined experimentally from attenuation measurements.*

† *The measured values of attenuation cannot be compared with the estimated attenuation for similar types of guide, because the roughness measurements were carried out on guides of slightly better quality.*

lining has to be quite thick compared with the skin depth of silver, however, which results in an expensive product.

Sprayed Surfaces

Spraying techniques may be used for the fabrication of waveguide components. Metal is sprayed on to a suitable mandrel to the required thickness and the mandrel is then removed. The surface finish of the mandrel determines the roughness of the final product, provided the sprayed metal flows on the mandrel surface. The main drawbacks of the system are reductions in mechanical strength and electrical conductivity[21] of the material.

The characteristics of sprayed-aluminium components formed on mild-steel mandrels have been described by the

Fig. 10. Mechanically lined silver waveguide surface from 1in × ½in guide
Section perpendicular to waveguide axis. Magnification: 940. Silver in lower part of photograph.

Fig. 11. Acid copper electroplated brass waveguide surface from 1in × ½in guide
Section perpendicular to waveguide axis. Magnification: 950. Brass in lower part of photograph and acid copper plating in the dark band.

authors[20]. The surfaces of such components are not good, probably due to poor flowing properties of the initial layers of aluminium. Samples formed by spraying a thin layer of zinc on to a steel mandrel and backing this with a thicker layer of sprayed brass to give additional strength had a surface finish, however, comparable with that of precision-drawn copper tube. The main roughness is in the form of grooves in the transverse section which are probably due to withdrawal of the mandrel. It seems, therefore, that sprayed zinc has better flowing properties than aluminium.

Castings

Castings in aluminium or magnesium have been used for T-junctions, bends, corners and straight waveguide lengths. One major disadvantage of using magnesium is the galvanic action between this metal and others[13]. The surface finish of a cast aluminium waveguide (internal dimensions 3in by 1in) has been investigated by the authors[20]. The component was produced by using plaster of Paris mixture which was de-hydrated at a high temperature (600 to 800°C). The surface finish of the box into which the plaster is cast is a very important factor. The surfaces examined were fairly flat, but microscopically rough and contained a few blow holes. Considerable improvement in the standard of surface finish of such castings has, however, been obtained recently.

Electroplated Surfaces

It has been standard practice both in this country and elsewhere for many years to construct most microwave components from brass and then plate the resulting product with silver, which has good conductivity. Plating is also used sometimes for protection. Measurements on silver-plated brass waveguides, however, showed unexpectedly high attenuations and gave very variable and disappointing results. For example, tests on normal silver-plated brass guides, of rectangular shape, with internal dimensions 1in by 0·5in, gave an average ratio of measured-to-calculated attenuation value of 1·58. Five feet lengths of these guides were plated in a cyanide bath, giving a matt finish, using a special anode inside each to try to achieve a uniform plating thickness. Other measurements[13] have shown that silver-plated brass guides have losses 34 per cent above theoretical at 3 000Mc/s and 58 per cent higher than theoretical at 9 375Mc/s. It seemed from these measurements that electroplated surfaces could be rougher than the original drawn ones. Work was, therefore, undertaken to confirm this.

Fig. 12. Cyanide silver electroplated waveguide surface from ¾in × ¼in guide
Section perpendicular to waveguide axis. Magnification: 980. Cyanide silver plating in the light band and brass waveguide in the lower part of the photograph.

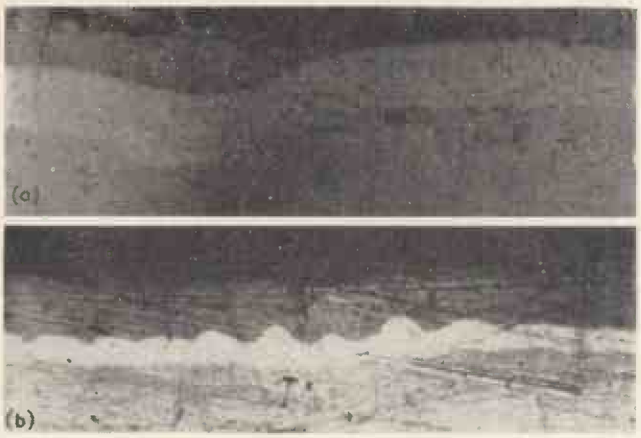
First, samples of brass waveguide only 1 to 2in long with internal dimensions 1in × 0·5in were plated in an acid-copper bath to a thickness of about 0·001in at a current density of about $7A/ft^2$. A typical photomicrograph showing that the plated surfaces can be very much rougher than, and bear no relation to, the original surface, is shown in Fig. 11. For some parts of the tube the ratio of the length of the plated surface to the ideal one may frequently be as high as 1·625.

Similar results are obtained when brass guides are plated with silver in a normal cyanide bath giving a matt finish although the plated surfaces are not as rough, in general, as acid-copper ones. Some parts have silver surfaces which do not follow the original brass ones at all, as can be seen from Fig. 12.

The inclusion of certain addition agents in silver-cyanide baths gives a much brighter silver surface with a certain degree of levelling. Using cyanide baths with additions of turkey-red oil and carbon disulphide the authors have tried to produce a plated surface smoother than the original surface. Bright-copper deposits can also be obtained by adding small quantities of colloidal material[22] to either an acid or a cyanide bath. These deposits show considerable improvements over normal-plated surfaces with regard to levelling properties[23].

Results for bright-silver surfaces are rather disappointing although the surface finish is always much better using a bright bath than an ordinary one. In general, the silver faithfully reproduces irregularities of the base metal (Fig. 13(a)), but there is not much levelling and frequently the silver is rougher than the original surface (Fig. 13(b)). Bright copper plating, however, appears to have good levelling properties (Fig. 14).

Silver-plated component losses tend to increase with age[13]. Increases of up to 160 per cent in the losses were



Figs. 13(a) and (b). Bright silver electroplated brass waveguide surface from 1in × ½in guide

Sections perpendicular to waveguide axis. Magnification: 940. Bright silver plating in the light bands and brass in the lower part of the photographs.

Fig. 14. Copper electroplated brass waveguide surface from 1in × ½in guide

Section from long side parallel to waveguide axis. Magnification: 940. Brass in lower part of the photograph and bright copper plating in the dark band.

measured at 35 000Mc/s after a few months shelf life. These increases are far more apparent at the higher frequencies. Lending[13] supports the conclusions earlier reached by the authors[15, 16], that the accepted practice of silver plating brass waveguides is probably not worthwhile. Experimental evidence suggests that in the millimetre region the practice should definitely be discouraged. A rhodium flash over silver plating will minimize the effects of ageing, but the initial losses will be somewhat higher.

Silver-plated stainless steel has been used in the lower microwave frequencies, but sometimes the silver peels from the steel[13]. Silver-plating of aluminium has also been tried[1, 13] but needs careful control and is not very successful outside the laboratory. Successful attempts to put silver on plastic waveguides have been described[13].

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(To be continued)

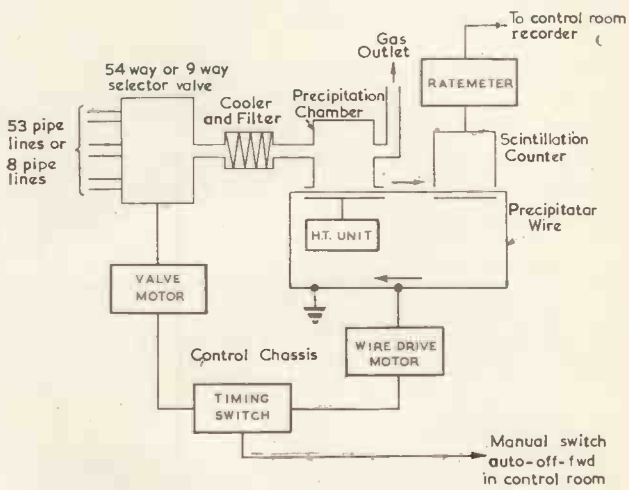
Burst Slug Detection Gear at Calder Hall

The burst slug detection gear installed at Calder Hall is provided to give automatic warning of excessive radioactivity caused by faulty fuel containers within the Atomic Pile.

In a nuclear power plant of the Calder Hall type, the reactor coolant gas (CO_2) is recirculated through a closed circuit and failure to detect faulty fuel elements would result in fission products leaking out and spreading radioactivity through the entire gas circuit.

Basically, the principle of detection involves the examination of small samples of the coolant gas circulating around the uranium containers.

The Calder Hall pile is divided into eight separate areas, each controlled by a 54-point gas sampling plant which registers back to a central control room. From each of these areas, 212 gas sampling pipelines are led through manually-operated cocks which reduce the number of channels to 53, by combining them in groups of four. Each of the 53 pipelines is connected to the inlet ports of a 54-way selector valve, which in turn links them to outlet ports. Samples of the gas are taken from each inlet port every 27min, the valve remaining stationary at each inlet for 28sec and taking 2sec to travel between ports.



54- or 9-way gas sampling system

The 54-way selector valve of each area is connected by single pipeline, via a cooler and filter, to its own precipitator unit.

When a fuel container is damaged, fission products, some of them in gaseous form, escape into the main coolant stream. The particles are arrested by the filter, mentioned above, but some of the radioactive gases pass on into the precipitator unit where they decay and become deposited by electrostatic precipitation on to a wire which remains stationary for 28sec during the sampling period.

At the end of the sampling period, the portion of wire carrying the radioactive deposit travels for two seconds to a scintillation counter, remaining there another 28sec for checking.

This scintillation counter operates a rate-meter which in turn indicates on a 54-way recorder in the control room and if a high radioactive count is registered a warning bell is operated.

During the time that this piece of wire is being checked for radioactivity, a gas sample is being taken from the next inlet port on the selector valve and the products are being deposited on another portion of the wire. No portion of the wire is used more than once in 30min, thus allowing long-lived activity to decay.

Overall sampling of the gas coolant in the pile, is governed by a 9-point gas sampling plant consisting of eight additional pipelines (four from the inlet and four from the outlet gas ducts) joined to the eight inlet ports of a 9-way selector valve. Each inlet port is connected in turn with an outlet port and the valve remains stationary for a 28sec sampling period at each inlet and also takes two seconds to travel between ports. Each inlet is checked, therefore, once every 4½min. This 9-way selector valve is connected, via a cooler and filter, by a single pipeline to a duct precipitator unit, similar in operation to the others mentioned.

Any suspect pipeline can be sampled continuously by diverting it to one of two spare precipitator units which, operating in conjunction with a single-point recorder in the control room, constitutes a single-point gas sampling plant.

54-Way Selector Valve

This valve consists of a body with 53 inlet ports, the 54th position having an internal transfer port which maintains gas flow through the valve.

Each inlet port is connected in turn with an outlet port by the operation of a rotor arm. This is coupled to a worm housing driven through a freewheel mechanism by a motor fitted an electro-magnetic brake.

As the rotor arm approaches an inlet port, a cam mechanism fitted to the housing snaps it forward into position, the freewheel mechanism being disengaged by the drive motor. The circuit to the valve motor is, at the same time, broken by a valve switch.

54-Way Selector Switch

A 54-way selector switch which is driven from the worm housing contains a synchro, together with a 6- and a 9-way switch.

This synchro operates the repeater of a 54-way dial indicator which is located in the control room and provides remote indication of the valve rotor arm position. The 6-way switch operates 54 times successively during the 9-way switch cycle.

There are nine pens located in the recorder carriage, each of which is governed by the 9-way switch. Each carriage position is registered by neon indicators situated on the recording dial and the 6-way switch controls the movement of the 54 point recorder carriage.

Precipitator Unit

This comprises a precipitator body fitted to a stand which houses a control chassis, h.t. unit, rate-meter and indicator panel.

The precipitator body contains a precipitation chamber assembly and a scintillation counter, a precipitation wire being passed through these and around driving and idling end pulleys to form a continuous loop.

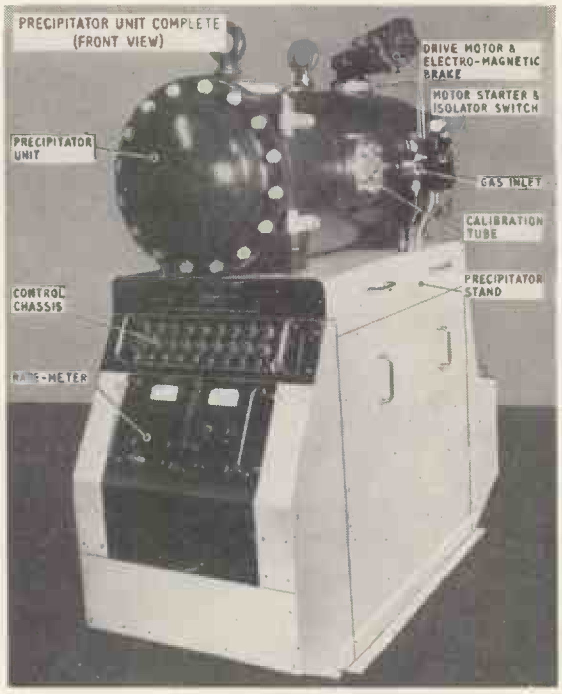
The control circuit for the wire-motor and valve-motor is housed in the control chassis and failures within the gas sampling plant are visually indicated by warning circuits.

The h.t. unit provides the high positive potential on the precipitation chamber outer electrode relative to the wire and the ratemeter is operated by the output from the scintillation counter.

A gas temperature gauge, which operates a warning circuit, and gas pressure gauge are contained in the indicator panel.

Principles of Operation of Plant

The gas enters the precipitation chamber maintained at 5kV, which creates an intense electric field between the outer electrode and the precipitator wire held at earth potential. Radioactive atoms of gas which decay into a solid state are deposited on to the precipitator wire.



At the end of the 28sec sampling period, this portion of the wire is moved to the scintillation counter, and the selector valve moves to the next inlet port.

The radioactive solids deposited on the precipitator wire decay again by emitting beta-particles which pass through a phosphor cylinder mounted concentric with the wire. Light flashes produced during this decay are detected by a photo-multiplier which is in contact with the phosphor in the cooling tube.

The rate-meter sums up or integrates the number of light flashes over the measuring period of 28sec and this is shown on the recorder in the control room.

Control and warning circuits for each plant are completed through a switchboard and any faults developing are visually indicated on the switchboard and by group warning alarms in the control room. In addition, high counts of radioactivity detected in any plant also operate warning alarms.

In the event of a serious fault occurring in a precipitator unit, one of the two spare precipitators is brought into action.

The burst slug detection gear was designed by Plessey Nucleonics Ltd. in collaboration with the United Kingdom Atomic Energy Authority.

The Design fo Cold-Cathode Valve Circuits

(Part 2)

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

Logical Operations

IN digital circuit applications, it is desired to make the firing of each cold-cathode valve dependent on a particular combination of signals being applied to it in order to perform certain logical operations. The valve itself merely acts as a power amplifier; the signals must be combined in a circuit known as a gating circuit before being applied to the trigger. Examples of gating circuits using rectifiers are shown in Figs. 15 and 16.

In Fig. 15 the firing of valve V_1 or V_2 or V_3 causes V_4 to fire: the circuit is therefore known as an OR gate. If only, say, V_1 is fired, its cathode potential rises to a positive voltage sufficient to fire V_4. Rectifier MR_1 is conducting, but rectifiers MR_2 and MR_3 are biased non-conducting so that the cathodes of V_2 and V_3 remain near earth potential and any other circuits which may also be connected to these valves are unaffected.

In Fig. 16, valve V_4 can only fire if V_1 and V_2 and V_3 are all on: the circuit is therefore called an AND gate. If only V_1 is fired, its cathode potential rises and current flows through R_1 to MR_2 and MR_3 and the cathode resistors of V_2 and V_3.

Because the resistances of MR_2, MR_3 and the cathode load resistors are small compared with R_1, the trigger potential of V_4 does not rise sufficiently to fire this valve. If only V_2, say, is fired, its cathode potential rises and rectifier MR_2 is biased off. The backward resistance of MR_2 is high compared with the resistance of R_1, so the trigger potential of V_4 again does not rise sufficiently to fire the valve. If V_2 and V_3 are both "on" then MR_2 and MR_3 are both biased off. If V_1 is also on, then its cathode potential is applied through R_1 to the trigger of V_4 and fires this valve.

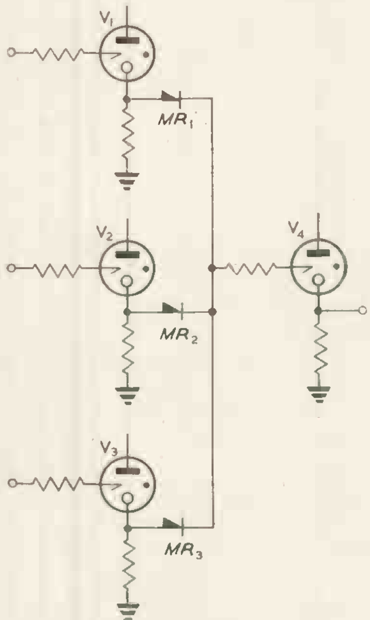


Fig. 15. Gate circuit for logical operation OR

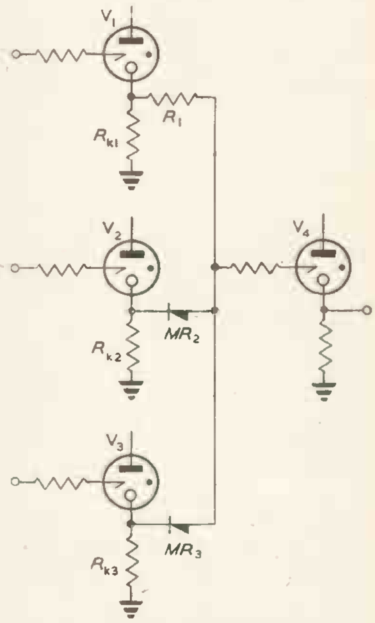


Fig. 16. Gate circuit for logical operation AND

In rectifier gating circuits of the type shown in Figs. 15 and 16 miniature selenium rectifiers are generally used. The backward resistance of these is usually so high that it can be ignored, but the forward resistance is by no means negligible. In Fig. 15, the cathode voltage required from valve V_1 must exceed the firing voltage of the valve V_4 by the amount of the p.d. across R_1 due to the current in it, which must be greater than I_t. The value of R_1 is usually high (e.g. 1MΩ) in order that a single cathode can feed several gating circuits, so the cathode output voltage V_k must exceed V_{ts} by an ample margin. Now $V_k = V_a - V_{am}$, so a high anode supply voltage V_a must also be used and a valve is required which has a high anode striking voltage (e.g. $V_{as} > 230$V). Another limitation on the performance of the gating circuit of Fig. 16 is the self-capacitance of the rectifiers. If the cathode voltage of V_2, say, rises very rapidly, a short pulse is applied to the trigger of V_4 through the capacitance of MR_2 and the valve may fire. This may be obviated by connecting a small capacitor to earth at the trigger end of MR_2. Alternatively, the rate of rise of the voltage applied to MR_2 may be slowed down until the effect of the self-capacitance of MR_2 becomes negligible by means of a capacitor across R_{k2}. Such a capacitor is often required also for extinguishing the valve, as described previously.

Another form of AND gate for use in cold-cathode valve circuits is shown in Fig. 17. If, initially, V_1, V_2 and V_3 are all "off" the cathodes of V_1 and V_2 and the trigger of V_3 are all at earth potential and the coupling capacitor C_c is uncharged. If V_1 only is fired, its cathode rises to a positive potential V_k which is applied momentarily to the trigger of V_3 via C_c.

The trigger potential thus rises instantaneously to V_k and then decays exponentially as C_c charges. However, the cathode resistor R_k is chosen so that V_k is less than V_{ts} so valve V_3 does not fire. If V_2 only is fired, its cathode potential rises to V_k and capacitor C_c is charged via R_b until the trigger potential of V_3 is also V_k; again the valve does not fire. If now, V_1 is also fired, its rise in cathode potential is momentarily applied to the trigger of V_3 in addition to the p.d. already across C_c. The trigger potential of V_3 is therefore instantaneously doubled and the valve fires. It should be noted that the circuit depends for its operation on valve V_2 being fired sufficiently early for the coupling capacitor C_c to be almost fully charged before V_1 is fired. (The p.d. across C_c will rise to 95 per cent of V_k in time $3R_bC_c$). If V_1 is fired too soon after V_2, or is fired before V_2 or simultaneously, then V_3 will not strike. Because of its method of operation the circuit of Fig. 17 is usually called a "pulse-plus-bias" gate. It must always be remembered when designing circuits that this gate is not merely an A AND B gate, but really an A BEFORE B gate. This fact is sometimes inconvenient, but often it can be an advantage, as will be seen when counting circuits are considered.

The speed of operation of the pulse-plus-bias circuit is limited by the time taken to charge the coupling capacitor

* Siemens Brothers & Co. Ltd.

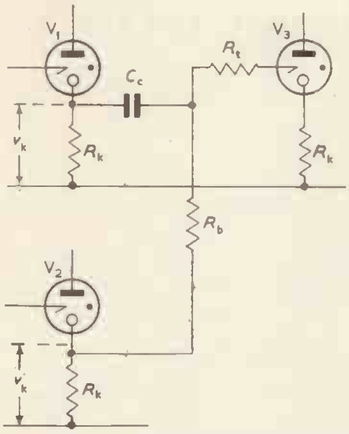


Fig. 17. Pulse-plus-bias AND gate

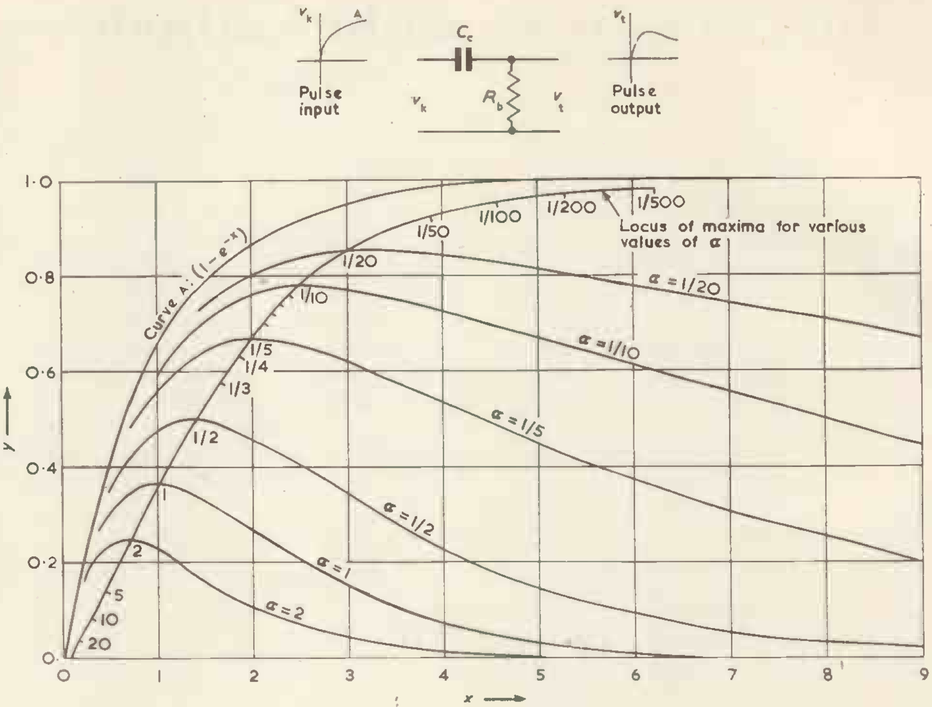


Fig. 18. Plot of functions
$y = (e^{-\alpha x} - e^{-x})\ (1-\alpha)$

This represents the output voltage due to a fixed input voltage ($1-e^{-x}$) when the time-constant of the RC coupling network is varied. (Time-constant of network = 1/a × time-constant of applied voltage)

C_c. If, in an attempt to reduce this charging time, the capacitance is made very small, it will discharge too quickly and the trigger potential of V_3 may decay below V_{ts} before the valve has fired. To prevent this, the time-constant with which C_c discharges must be sufficiently large to ensure that the trigger potential exceeds V_{ts} for longer than the ionization time of the valve. However, it is not often possible to make C_c as small as these considerations permit, as will be explained below.

So far, we have assumed that the signal voltage which is applied to the coupling capacitor rises instantaneously, with the result that its full magnitude is transferred to the trigger of the valve. However, in practice, its rate of rise is necessarily finite and less than its full magnitude is therefore transferred to the trigger.

If the voltage applied to the coupling capacitor rises exponentially then the voltage which is obtained across R_b varies as shown in Fig. 18. It will be seen that if the time-constant R_bC_c is equal to that of the input signal then only 37 per cent of the input voltage appears across R_b. Even if R_bC_c is ten times the time-constant of the input signal the output voltage only rises to 78 per cent of the final input voltage.

The curves in Fig. 18 can be applied to the design of pulse-plus-bias gates driven from valves which have cathode capacitors for deionization as described previously. The value of R_b is usually much greater than R_k, so the coupling circuit does not greatly affect the cathode voltage whose rise is given by equation (4). The curves in Fig. 18 then enable a suitable value to be determined for C_cR_b. It will thus be seen that the time-constant of the coupling circuit must not merely exceed the ionization time; it must be several times the deionization time, which is usually much longer.

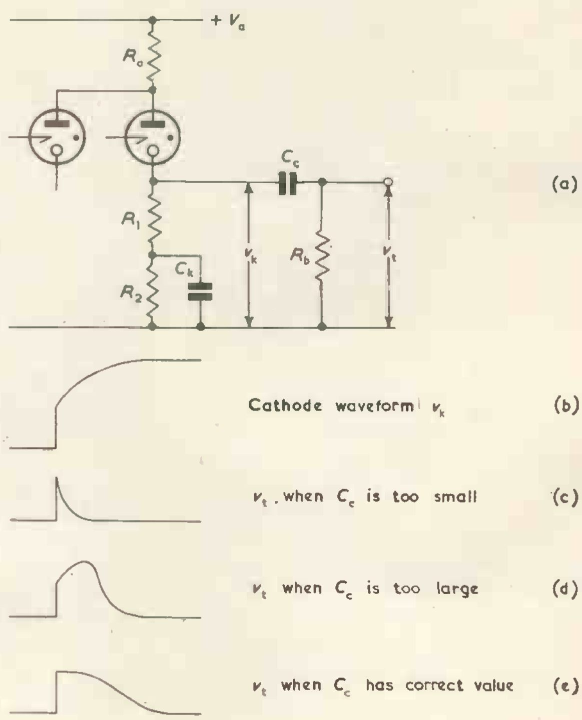
An improvement in circuits using pulse-plus-bias gating in conjunction with cathode extinguishing can be obtained by modifying the cathode circuit as shown in Fig. 14. The circuit between the cathode of one valve and the trigger of the next* is then as shown in Fig. 19(a). The voltage at the cathode of valve V_1 is then shown in Fig. 19(b). The trigger voltage of V_2 instantly rises by an amount:

$$V_t = (V_a - V_{am})R_1/(R_a + R_1) \quad \text{.........} \quad (6)$$

but its subsequent behaviour depends on the value of C_c. If C_c is small, the trigger voltage immediately begins to decay away. If C_c is large, the trigger voltage v_t first rises further and then falls fairly rapidly. However, if the value

* *This cathode circuit configuration is similar to the anode circuit of a resistance-capacitance coupled thermionic-valve amplifier with anode decoupling capacitor Cd. It is well known that the square-wave response of the RC coupled amplifier is maximally flat when the time-constants RbCc and RaCd are equal.*

Fig. 19. Use of cathode circuit of Fig. 14 to compensate for effect of the coupling circuit



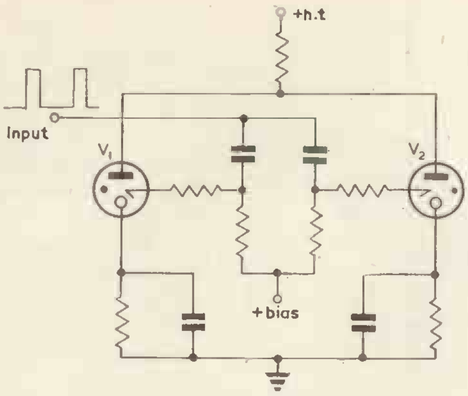


Fig. 20. Binary counter stage

of C_c is chosen appropriately, v_t will be stationary at $t = 0$ and then decay fairly slowly. It can be shown that the condition for this is:

$$C_cR_b = C_kR_1(1 + R_1/R_a) \quad \ldots\ldots\ldots\ldots \quad (7)$$

It will be noticed that this condition is independent of R_2, which can therefore be chosen solely to suit C_k and R_a with regard to the conditions for extinguishing the valve. R_1 is then chosen from equation (6) to give the required value of trigger voltage for valve V_3. The required value of coupling-circuit time-constant is then given by equation (7). The trigger voltage remains above 95 per cent of its peak value V_t for a time $\frac{1}{2}C_cR_b$, so it will exceed V_{ts} for sufficiently long to ensure that the valve fires.

The gating circuits described above perform the logical operations of A AND B and A OR B, but another operation is often required, namely A AND NOT B. This requires an output signal to be produced only when a particular cold-cathode valve is *not* fired. One method of doing this when rectifier gating is employed is to use a pair of valves, with cathode extinguishing capacitors as shown in Fig. 9. One valve is normally operated to hold open an AND gate of the type shown in Fig. 16. Extinguishing this valve by firing its partner thus prevents the gate from being open. It will be thus seen that the NOT gates are more complicated and expensive than AND gates. The design of economical circuits therefore usually hinges on the designer's ability to rearrange the logical requirements of his circuits so as to avoid the use of NOT gates wherever possible. When pulse-plus-bias gating is used a NOT gate can be simply obtained by using diodes, as will be described later.

Some Basic Circuits

A binary counter based on the use of pulse-plus-bias gating with cathode capacitors for extinguishing the valves is shown in Fig. 20. If valve V_1 is initially "on", then V_2 is "off" but its trigger is biased positive. When a positive pulse is applied to the input terminal the additional voltage applied to the trigger of valve V_2 causes it to fire. When V_2 fires, the anode potential falls and extinguishes V_1 as described before. Now valve V_2 is "on" and V_1 is "off". The next pulse which is applied to the input terminal therefore fires V_1 and extinguishes V_2 and so causes the circuit to revert to its original state.

The cathode of either V_1 or V_2 can be connected to the input terminal of a second stage, which thus receives pulses at half the rate of the first stage and delivers pulses at one quarter of that rate. A chain of such stages, each of which delivers "carry" pulses to the next, can thus be used to count in the binary scale the number of pulses which are applied to the first stage. Other cold-cathode binary counter circuits are also well known[7,11,15].

When counting is required to a scale other than the binary one, the circuit of Fig. 20 can be extended as shown in Fig. 21. The trigger of each valve receives bias from the preceding cathode. If, initially, the first valve is "on" and all others are "off" the first input pulse will fire the second valve and extinguish the first. The next input pulse will fire the third valve and extinguish the second, and so on. If the counter is required to operate cyclically, the cathode of the last valve can be connected back to bias the trigger of the first valve so that the circuit forms a closed ring. Alternatively, if non-cyclic operation is required, the circuit can take the form of an open chain as shown in the

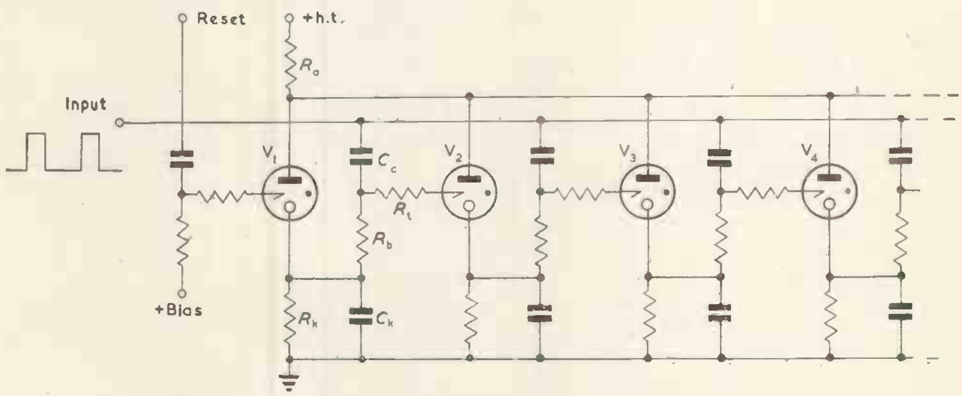
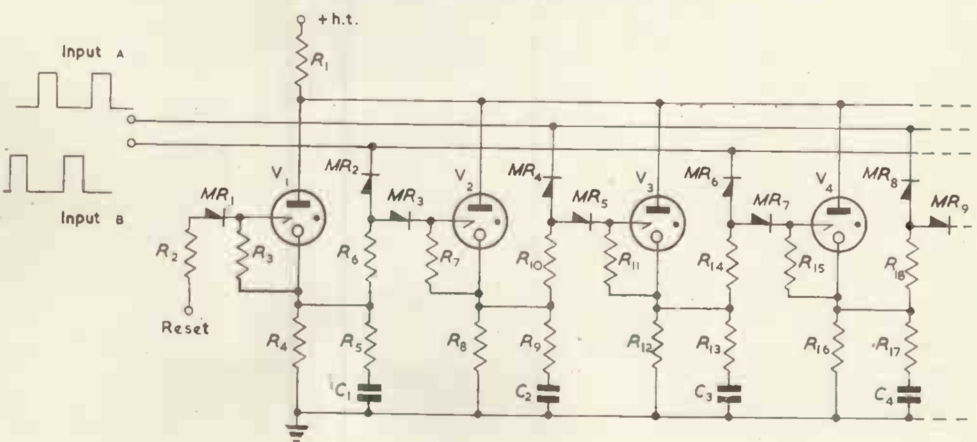


Fig. 21. Chain counter using pulse-plus-bias gating

Fig. 22. Chain counter using rectifier gating



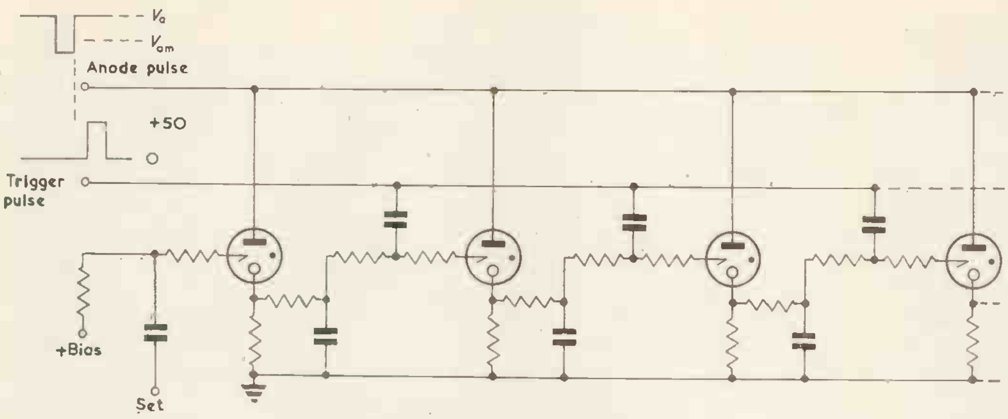


Fig. 23. Pattern shifting register

figure. The counter can be reset to its initial condition at any time by applying a positive pulse to the trigger of the first valve. Thus, the correct initial conditions can be obtained when the h.t. supply is switched on by connecting the trigger of V_1 to the h.t. supply through a small capacitor.

A counter circuit using rectifier gates between the stages[18] is shown in Fig. 22. In order to use this form of gating, the input pulses have to be arranged so that alternate pulses are applied to each of the two input terminals A and B. If, initially valve V_1 is on and the other valves are off, a positive pulse applied to input terminal B will bias off rectifier MR_2 so that the cathode voltage of V_1 is applied to the trigger of V_2 via R_6: valve V_2 then fires and extinguishes V_1. The next pulse is applied to terminal A and so fires valve V_3 which extinguishes V_2, and so on. If the input pulses were not divided into even and odd pulses as described, it would be necessary to control their length very accurately. For, if all the rectifiers MR_2, MR_4, MR_6, MR_8, were connected to a single input terminal and an input pulse of excessive length were applied, the first valve which fired could trigger the next valve, and so on down the chain until the input voltage was removed.

The use of even- and odd-pulse input terminals removes this possibility by ensuring that the input voltage can never be applied simultaneously to adjacent valves. In the pulse-plus-bias counter this difficulty is avoided because only one stage can have bias applied to its trigger before the pulse arrives and so only one valve can fire. Counter circuits such as those shown in Figs. 21 and 22 can be used for many purposes such as decimal digit stores and pulse distributors.

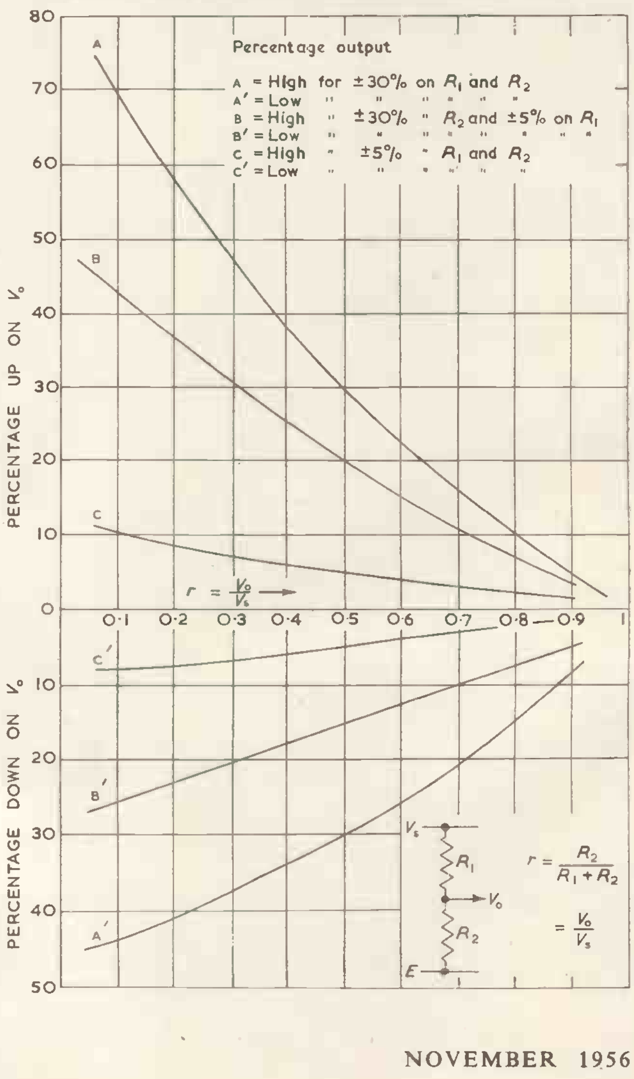
The circuit of Fig. 21 can be modified to form a pattern register or shifting store. In a counter, only one valve is "on" at a time and each input pulse fires the next valve, which extinguishes the previous one. In a shifting store, any pattern or combination of valves may be "on" and an input pulse causes the whole pattern of glowing valves to step forward one position. In order to permit any number of valves to be on and to be extinguished together, a pulsed h.t. supply is used instead of a common anode resistor. The circuit is shown in Fig. 23. Each valve that is "on" charges up its cathode capacitor which applies bias to the trigger of the next valve. The negative pulse applied to the anodes extinguishes the conducting valves, but the cathode capacitors remain charged. Consequently, the positive pulse which is applied to the triggers when the anode pulse disappears strikes those valves which are also receiving bias from the cathode capacitors of preceding valves. Thus, each time the anode and trigger pulses occur, the pattern of conducting valves steps on one position.

Circuit Tolerances

The simple pulse-plus-bias circuits described so far can only be made to operate reliably for long periods if valve and components having close tolerances are used. A typical cold-cathode valve has a trigger striking voltage V_{ts} of 60 to 80V but, in order to ensure a small ionization time even when the valve has aged, it is desirable to apply about a 100V signal to fire the valve. The pulse-plus-bias circuit must therefore ensure that the sum of pulse-plus-bias exceeds 100V, but that each separately is less than 60V. The signal voltages used must therefore have the comparatively narrow limits of 50 to 60V.

In general, both the pulse and the bias voltage are derived from the cathodes of other cold-cathode valves. In a typical circuit as shown in Fig. 17 the cathode voltage V_k is obtained from the h.t. supply voltage V_a by means of a potentiometer comprising the anode resistor R_a, the valve itself and the cathode resistor R_k. The h.t. voltage can

Fig. 24. Percentage variation of output voltage with nominal ratio of potential divider for various tolerances on resistance values



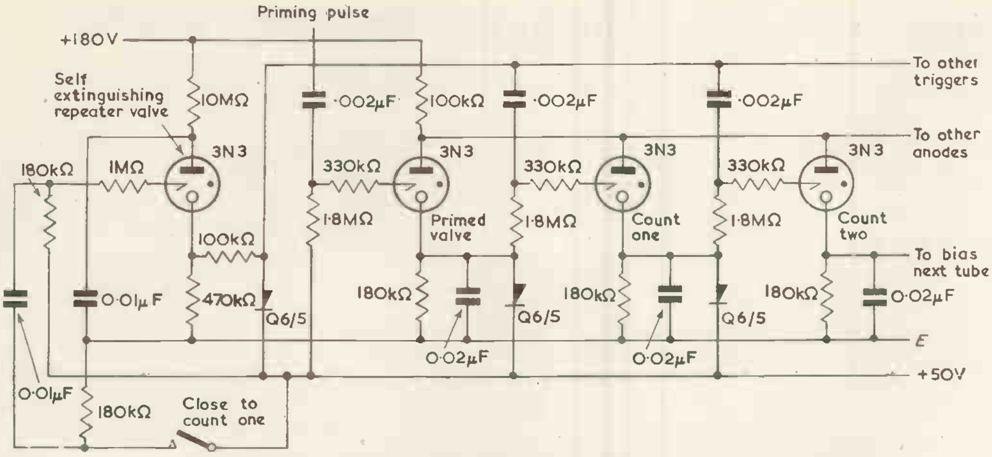


Fig. 25. Chain counter circuit arranged to count pulses from a relay contact

usually be controlled within fairly close limits, but the maintaining voltage V_{am} of the valve may vary between, say, 70 and 80V and the resistors will have tolerances. The tolerance which has been postulated on V_{am} already equals the ± 5V tolerance permissible on V_k without allowing any margin for variation of R_a and R_k. The effect of the resistor tolerances actually contributes the greatest amount to the variation of V_k, because the tolerance on the output voltage of a potentiometer exceeds that on the resistances of either of its two arms. This is shown in Fig. 24 which plots the tolerance on the output voltage of a potentiometer against its ratio for several values of resistance tolerances. Thus, the simple basic circuit can only be made to work reliably for long periods by selecting the valves to close tolerances and using close-tolerance high-stability resistors. This is clearly an undesirable state of affairs and much time and ingenuity have been devoted to developing circuit techniques which permit wider tolerances.

One method of obtaining wider tolerances is to use the method of rectifier gating previously described. It is then necessary for V_k to exceed a lower limit to ensure triggering of the next valve, but no upper limit is imposed. Consequently the components can have wider tolerances. However, the method requires a valve with a rather high value of V_{as} and it does lead to some rather clumsy circuit arrangements, as has already been seen.

A method of obtaining wider circuit tolerances is to retain pulse-plus-bias gating, but to avoid the use of cathode capacitors for deionization. An anode resistor is then no longer required and V_k is equal to the h.t. voltage minus the maintaining voltage of the valve and so independent of resistor values. This technique is useful in large equipments where the valves are extinguished by a pulsed h.t. supply.

If it is desirable to retain both pulse-plus-bias gating and the use of cathode capacitors for deionization, it is possible to obtain wider circuit tolerances by adding rectifier clamps to the cathode circuits as shown in Fig. 25. The current through each valve when it is " on " is arranged to exceed that which produces a p.d. of 50V across R_k so that the rectifier conducts and clamps the cathode potential to 50V. The cathode voltage is then determined by the 50V supply potential, which can be closely regulated, and the p.d. across the rectifier, which is small. The signal voltages are thus almost independent of variations in the resistors and valves. This rectifier clamping technique is widely used and enables very reliable performance to be obtained from pulse-plus-bias circuits. It is also sometimes possible to use cold-cathode diodes instead of metal rectifiers for performing the clamping.

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(*To be continued*)

Instrument Landing System at Waddington

Waddington is the first of a large number of R.A.F. stations to be equipped with the Pye ground instrument landing system. This new system, which uses a uni-directional localizer aerial, has important advantages over its predecessors. Difficult sites, which had many beam bends with the old type omni-directional localizer can, when fitted with the new system, provide clean approach paths free from ambiguities. The automatic beam control circuits associated with the electronic modulation ensure consistency of beam angles regardless of changing weather conditions.

I.L.S. is pilot operated in that the ground installation provides continuous information which the pilot interprets by his instruments. Once the aircraft has been permitted to enter the approach zone, the landing is effected entirely by the use of these instruments without any control or verbal communication with the ground control.

I.L.S. ground equipment comprises three distinct elements. The localizer transmitters, the glide path transmitters and the marker beacon transmitters. All these units are designed to work in conjunction with remote control apparatus so that the installation is run unattended. Supervision is exercised by the air traffic controller by means of a monitor which instantly indicates a fault condition. A feature of the system is that the courses are stabilized. This is achieved by the use of negative feedback from the course monitor to the transmitters and has the effect of automatically correcting any minor changes in the transmitters' performance.

The localizer transmitter radiates two amplitude modulated signals of 90c/s and 150c/s in the v.h.f. band, 108 to 112Mc/s. The radiation pattern of the aerials is such that a beam is directed along the runway which guides the aircraft in azimuth so that it can be brought into line with the runway. The carrier is modulated to a depth of 20 per cent by each tone and an indicator in the aircraft compares the two modulation levels. If the aircraft is on course, there is no difference in the levels of modulation and a vertical pointer on the indicator takes up a central position. If the aircraft flies off course, the receiver measures the difference in the depth of modulation which is produced and the pointer moves to the left or right. A movement to the left indicates "fly left" and vice versa.

The glide path transmitter operates on a similar principle to that of the localizer, but here the aerials are arranged so that a vertical radiation pattern is produced. Indication in the aircraft is produced on a meter in which the pointer is placed horizontally. In practice, both horizontal and vertical pointers are contained in the same instrument. The transmitter operates in the u.h.f. band (328·6 to 335·4Mc/s) and is amplitude modulated by each tone to a depth of 45 per cent.

The marker beacon transmitters operate on a frequency of 75Mc/s and the aerials radiate vertically. Three are used for each approach system and are placed at fixed intervals along the path.

The Flowsolder Method of Soldering Printed Circuits

By R. Strauss*, A.I.M., and A. F. C. Barnes*, B.Sc., A.R.S.M.

In the Flowsolder dipping unit, which avoids some of the difficulties inherent in the conventional flat dip-soldering of printed circuits, a stationary wave of molten solder is created by pumping the metal upwards through a rectangular nozzle. The pre-fluxed circuit panels are passed through the crest of the wave. The Flowsolder unit, which is used with specially developed fluxes, facilitates the soldering of printed circuits, free from faulty joints or bridging.

FULL exploitation of the possibilities of printed circuits demands a soldering technique which is quick, gives consistently reliable results and fits readily into a moving production line.

Initially the leads of components inserted into the circuit panel were joined individually to the copper conductors of the printed wiring by hand-soldering, but it was obvious that the location of all joints in one plane made dip-soldering the natural choice for the completion of all soldered connexions in one single operation.

Basic Requirements

Numerous methods have been proposed and are being used to dip-solder printed circuits. The basic requirement with all dip-soldering operations is to present a clean, fluxed metal surface to molten solder which is free from oxides and at the correct temperature.

Dip-soldering of printed circuits raises some special problems of its own:

(1) Only the underside of the printed wiring panel must come in contact with the molten solder.

(2) Dipping temperature and dipping time must both be reduced to a minimum to avoid the formation of blisters between the laminated plastic and the copper foil, and to prevent heat damage to the components mounted on the panel.

The first requirement limits the dipping technique to flat dipping which is inherently more difficult than dipping by complete immersion.

The second requirement restricts the choice of solder to an alloy with as low a melting point as possible, and as a rule solders containing not less than 55 per cent tin are being used.

Dip-soldering Methods

A number of methods have been described for carrying out the flat dipping process. The difficulty with dipping a flat, fluxed circuit panel is twofold: trapping of flux, with formation of 'icicles', or bridging between adjacent conductors.

Unless the flux is given a chance to flow away from the printed wiring and from the joints which it covers as dipping commences, the molten solder is prevented from tinning the copper and completing the joints. Fluxes used for this kind of work are almost invariably of the activated resin type, and the problem can be solved to some extent by a suitable choice of resin, solvent, and resin concentration to produce a flux which is mobile enough at soldering temperature to be displaced by the solder. Nevertheless, with large flat areas some flux is bound to remain trapped if the panel is lowered vertically on to the solder, and a number of arrangements have been described which try to overcome this difficulty. A survey of American methods[1] of circuit dipping mentions entry and exit of the circuit into the bath at an angle, or sweeping the circuit across the solder surface.

Angle entry and exit are also recommended in literature in this country[2], while a ready-made dip-soldering machine is available in which a system of pivoted arms provides entry and exit of the panel at an angle.

Another method[3] solves the problem of angled entry and exit by curving the panel against a jig so that it forms a shallow arc which is then swung across the solder surface. This latter method is proposed only for pre-tinning of the circuit panels, as it would obviously be impracticable to curve a panel loaded with components. For soldering the connexions in the pre-tinned panel, flat dipping with a straight downward action is recommended.

While the trapping of the flux is mainly controlled by the way in which the circuit enters the solder bath, the occurrence of 'icicles' or of bridging between adjacent printed conductors depends on the manner in which the panel is withdrawn from the bath. Withdrawing at an angle makes it easier for excess solder adhering to tinned areas to drain off and to run back into the bath, and this is the reason for the various endeavours to provide angled withdrawal. Vibration of the panel, while it leaves the solder bath, is sometimes recommended in order to assist draining excess solder. The difficulty with flat dipping is that no flux is left on the panel when it leaves the solder bath and dip-tinning experience shows that it is very difficult to avoid the formation of 'icicles' or tears when withdrawing a tinned article through a solder surface without a flux cover.

In all cases, it is essential that the surface of the solder bath should be free from oxide in order to ensure clean tinning and sound joints. Mechanical or hand skimming immediately before dipping is normally recommended; in addition, anti-oxidant cover oils are available to prevent dross formation on the solder bath.

The Flowsolder Dipping Method

The above considerations have led to the development of the Flowsolder method of printed circuit dipping as an alternative to flat dipping.

With the Flowsolder method, the printed circuit is passed over a stationary wave of molten solder which is formed by forcing the metal upwards through a suitably formed nozzle. The circuit panel remains horizontal and moves on a straight line, and angled entry and exit of the circuit undersides are provided by the shape of the solder wave.

Fig. 1 shows a comparison between the flat dipping method and the Flowsolder principle.

Practical trials carried out with Flowsolder dipping units have shown that many of the difficulties inherent in the flat dipping principle can be overcome with the new method. In addition, the uninterrupted straight-line movement of the panel across the bath lends itself readily to incorporation in a steadily moving assembly line.

The angled entry into the surface of the solder wave,

* Fry's Metal Foundries Ltd.

combined with the washing action of the moving solder, prevents trapping of flux or air. Contact with constantly flowing solder also provides for more rapid heat transfer from solder to panel, which reduces the dipping time and assures consistently sound joints.

The molten solder welling up through the nozzle is drawn from below the surface of the bath and is therefore at a constant, easily controlled temperature and free from oxide. Consequently no skimming or anti-oxidant covers are necessary. Constant movement of the solder ensures that the whole of the bath is kept at an even temperature.

With all flat dipping methods, careful control of the level of molten solder is essential where the panel moves along a fixed path. The panels are normally 1/16in thick;

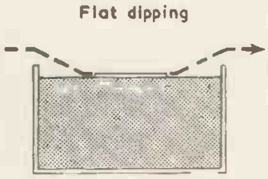
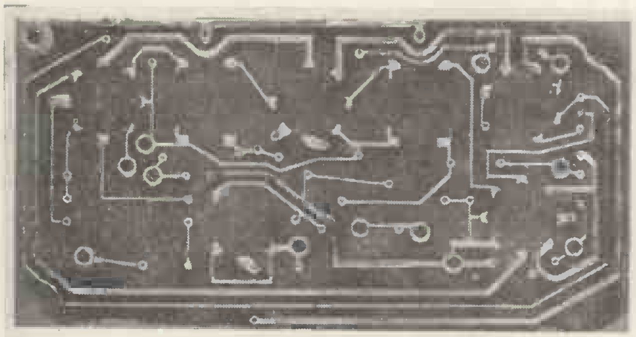
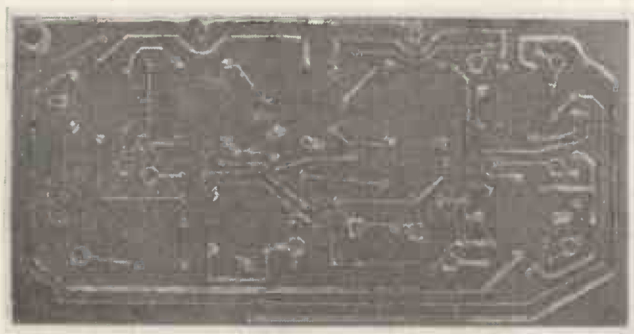


Fig. 1. Comparison of the flat-dipping and Flowsolder methods



(a)



(b)

Fig. 2. Printed circuit soldered with (a) Flowsolder 'P' flux, (b) Flowsolder 'N' flux

too high a solder level will cause flooding of the solder over the top, while a low level leads to incomplete soldering, so that the level must be controlled to within less than ±1/32in.

With the Flowsolder system, elaborate level control to close limits is not necessary because the height of the wave is easily regulated by adjusting the speed of the metal pump which forces the molten solder upwards through the nozzle. The level of the bath is kept constant within convenient limits by using an automatic ingot feeder, which provides a constant supply of pre-heated solder to the bath and avoids temperature fluctuations which would arise from adding fresh metal at intervals.

Special fluxes have been developed for use with the Flowsolder dipping unit. By adjusting the viscosity of the molten resin it has been found possible to formulate a flux (Flowsolder flux grade 'P') which remains on the circuit panel after its passage through the solder wave while being mobile enough to allow the molten solder complete access to the printed wiring and the joints to be soldered. Because of its presence on the exit side of the solder wave, it effectively prevents the formation of 'icicles' and bridging. The dipped panel is covered with a thin layer of solid resin, which is dry and hard, and conforms to all the requirements of A.I.D. specification D.T.D.599 as regards insulation resistance, freedom from corrosive action etc.

Where a flux residue is undesirable or inadmissible, another flux (Flowsolder grade 'N') is available, which is removed during the passage of the panel through the solder wave. To avoid excess solder pick-up on the exit side, vibration of the panel during its passage through the solder wave is recommended.

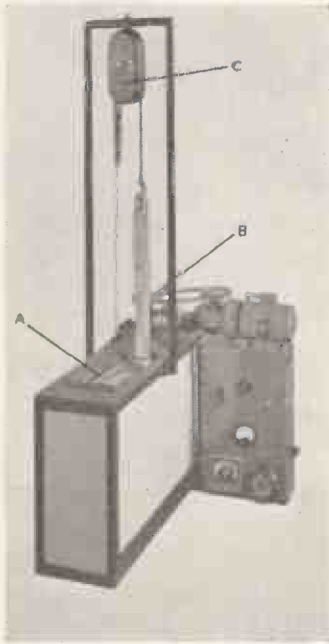


Fig. 3. The Flowsolder unit

Fig. 2(a) and (b) shows circuits soldered by the Flowsolder method with flux 'P' and flux 'N' respectively.

Other types of flux, which can be expected to simplify the production of printed circuits, are under development.

The Flowsolder Dipping Unit*

Fig. 3 shows a Flowsolder dipping unit. It consists essentially of a dipping bath holding approximately 500lb of metal. The solder wave (A) is produced by pumping molten solder through an appropriately shaped conduit by a motor-driven metal pump. The driving pulley of the pump is seen at (B). The shape and smooth surface of the solder wave are important. The close-up of the solder wave in Fig. 4 shows its freedom from oxide and the smooth flow of solder. Fresh solder is fed to the bath in the shape of elongated so called 'feeder ingots,' one of which is visible in Fig. 3. As the solder is consumed

* *The design of the Flowsolder unit is protected by patent application.*

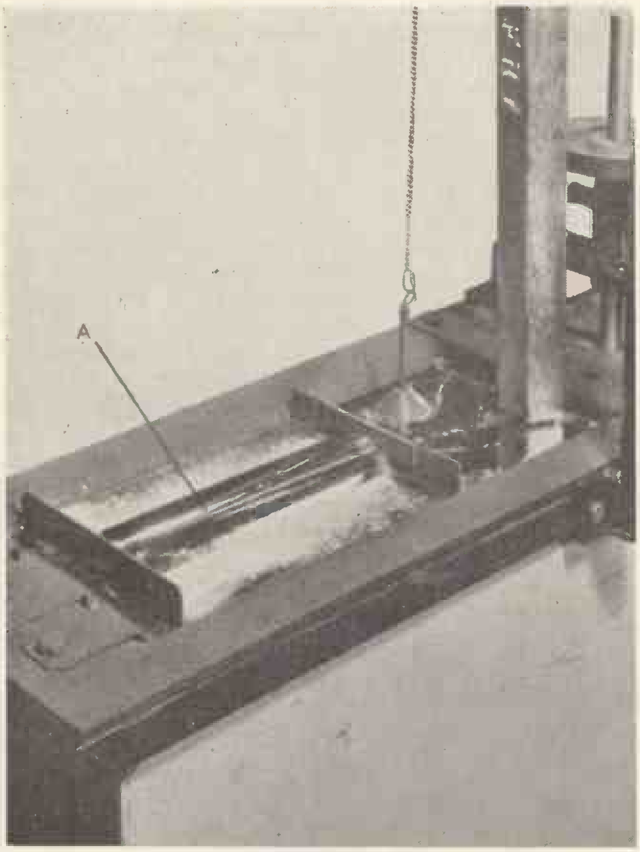


Fig. 4. Close-up view of solder wave

by the dipping process, the ingot is gradually lowered into the bath by the float controlled ingot feeder (C).

The unit shown in the illustration is electrically heated, but gas heated units can be supplied if necessary. The temperature of the molten solder is thermostatically controlled (D on Fig. 3).

The circuits are passed across the solder wave by a traversing mechanism, consisting of a trolley moving on a set of rails. The traversing mechanism is not part of the standard unit, because its design will depend on the nature and lay-out of the individual user's production line. Traversing speeds of the circuit across the wave depend on the kind of flux used; a speed of 4ft/min is suitable for grade "P" flux, while 2ft/min is required for grade "N" flux.

The width of the solder wave is 8in which is sufficient for the majority of commercially produced circuits, but nozzles up to 10in wide can readily be accommodated in the standard unit.

The metal temperature in the bath is normally controlled at 250°C. The solder recommended for the machine is a highly refined grade, specially produced to give clean, sound joints and to minimize solder consumption.

Even minute amounts of nickel, zinc, or cadmium lead to the formation of 'icicles' and to unsound joints, and therefore components with tags or leads plated with these metals should on no account be used on printed circuits which are to be dip-soldered. Leads and tags coated with tin or solder by electroplating or hot tinning are suitable.

The application of the flux to the circuits previous to dipping can be done by brushing, spraying, or flat immersion, and again it was felt that the design of that part of the equipment was best left to the individual user of the machine. Best results are obtained if the flux is allowed to dry on the panel before immersion. This can be hastened by passing the panel over a source of radiant heat or by blowing hot air against the underside of the fluxed panel.

The Flowsolder unit is designed to deal solely with the soldering part of printed circuit production; by its nature it is well suited to be incorporated in a fully or partly automatic production line.

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Synchronizing Low Frequency Pulses With a High Frequency Free-Running Time-Base

By M. V. L. Bennett*

A gating circuit is described which allows a pulsing unit driven by a low frequency oscillator to be synchronized with a high frequency free-running time-base. The device is particularly useful in electrophysiological applications where a stimulus may be applied only at intervals very long compared to response time. A new type of bistable trigger is included in the circuit.

IN the oscillographic display of electrical activity, it is frequently desired to invoke a transient effect by injecting an electrical pulse into the system under investigation. Although the pulse may be applied at random phase with respect to the oscilloscope sweep, it is usually more convenient to have the pulse synchronized, either initiating the sweep, or occurring at some definite interval along the sweep. More especially in the study of response to two pulses, it is necessary to have pulses occur in definite and regulatable phase with respect to the sweep cycle. Two methods of synchronizing pulse with sweep are generally employed: (1) pulse occurring at a definite time delay after the sweep starts, and (2) pulse occurring at a specific distance along the sweep, i.e. when a specific voltage is applied to the X-plates of the oscilloscope. It is clear that in the first system, time is independent of sweep velocity and place on the screen is not; and in the second system, place is independent of sweep velocity and time is not. Each system then has its advantages depending on what is being studied.

In electrophysiological work, one may apply an electrical pulse, i.e. a stimulus, to some tissue, and record a response either locally or in another part of the organism being studied. It may be possible to stimulate only at intervals

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very long compared with physiological response time. For example, in mammals, if a sensory nerve is stimulated, a localized potential change may be recorded on the surface of the cerebral cortex. This 'primary cortical response' lasts of the order of 50msec including latency of onset, but is altered if elicited more than once every 2 to 5sec. In the usual apparatus used to study such responses, the oscilloscope time-base gives single sweeps when pulsed by a low frequency oscillator, each single sweep triggering the stimulator. The responding tissue is unobserved for most of the time. In order to have any intervening activity visible a free running time-base may be employed, but isolated from the low frequency oscillator which now controls the stimulator. The stimulus then occurs at varying phase with respect to the sweep cycle, making the responses difficult to observe. A more compact solution than using two oscilloscopes is to have a device which synchronizes a single stimulus with the next complete sweep occurring after a pulse from the low frequency oscillator. Such a device is a gate which, while the time-base is free-running, lets through to the stimulator single signals from the time-base at a rate determined by the low frequency oscillator.

For time-delay synchronizing, the gate could be simply a bistable circuit which assumes one state for a low frequency oscillator pulse and the other for onset of time-base sweep. The transition from oscillator-induced state to time-base-induced state would provide a pulse to energize the delay circuit of the stimulator. However, for variable-position synchronizing, some sort of proportional gate must be used, that is a gate which puts out when open a signal proportional to the input. The gate should open at the beginning of the first sweep after the low frequency oscillator has fired and close at the end of the same sweep. The circuit illustrated satisfies these criteria. A type of bistable trigger, unpublished as far as the author is aware, is described and discussed in its general application.

Operation of the Circuit

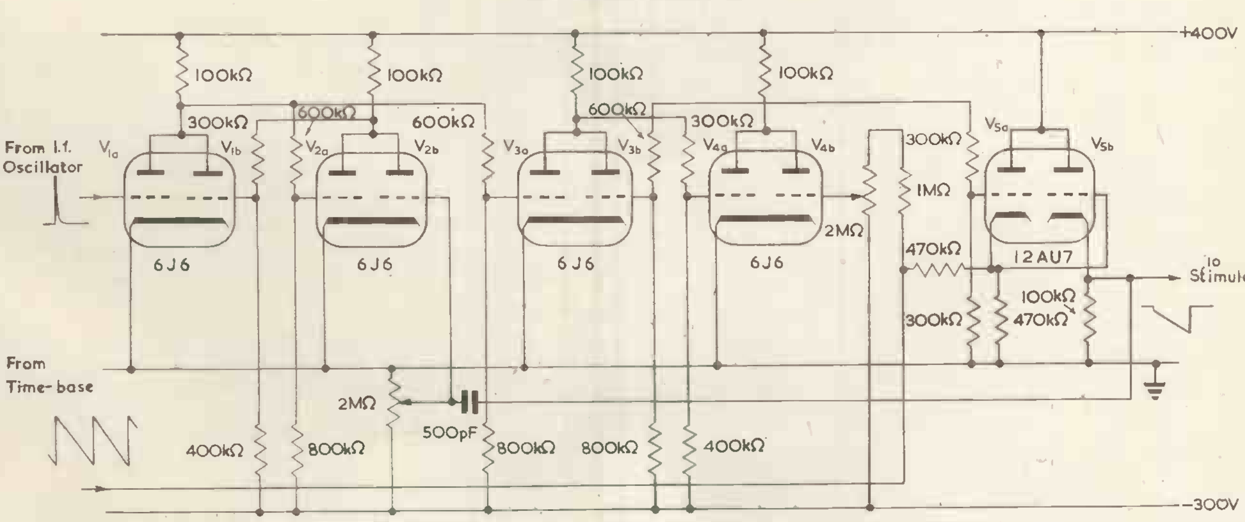
The operation of the circuit can be seen with reference to Figs. 1 and 2. Fig. 2 shows the potential changes at several points in the circuit during the operation cycle. The grid of V_{1a} and grid of V_{2b} are normally biased to cut-off. The interconnected halves of V_1 and V_2 form a bistable triode pair, V_2 normally conducting, V_1 non-conducting. When the low frequency oscillator fires, a positive pulse is transmitted to the grid of V_{1a}; V_1 anode drops; the grid of V_{2a} goes below cut-off so that V_2 anode rises; the grid of V_{1b} goes above cut-off, and V_1 is held conducting after the positive pulse to the grid of V_{2a} subsides. The bistable combination V_1V_2 assumes its other state as a result of being pulsed by the low frequency oscillator. The grid of V_{3a} is now below cut-off.

The grid of V_{4b} is adjusted by the preset potentiometer so that V_4 is just conducting at the end of the flyback and the beginning of the sweep. (The use of d.c. control on this grid assures that the gate does not open until the flyback is actually completed.) As the grid of V_{4b} goes above cut-off, V_4 anode starts to drop, and the grid of V_{3b} approaches cut-off. If the V_1V_2 combination has been pulsed, the grid of V_{3a} is below cut-off. As V_4 anode drops, the grid of V_{3b} drops; V_3 anode rises bringing up the grid of V_{4a}. The regenerative condition is reached and the circuit transfers to its other state, so that V_4 continues to conduct throughout the sweep cycle. Thus after the V_1V_2 combination has been pulsed, the return of the sweep to its starting position on the screen transforms the V_3V_4 combination so that the anode potential of V_4 is near earth. It is possible to adjust the values of the grid resistors so that the fall in V_4 anode potential before change of state occurs is small. Then the fall in V_4 anode potential at the beginning of the sweep when the grid of V_{3a} is above cut-off is also small.

$V_{5a}V_{5b}$ constitutes the proportional gate. When V_4 is non-conducting, the gate is closed. The grid of V_{5a} is more positive than the cathode socket would be if the valve were removed from the circuit. V_5 cathode follows and holds its cathode up so that no signal appears at the grid of V_{5b}. When V_4 anode is near earth, the grid of V_{5a} is also near earth; the gate is open. The cathode of V_{5a}, very positive to the grid, merely acts as stray capacitance, and the time-base signal attenuated by half passes through the cathode-follower, V_{5b}, that provides a low impedance output. The receipt of a low frequency oscillator pulse followed by return of the sweep to its starting point opens the gate.

The output of the gate is differentiated by the *RC* network, so that flyback produces a positive pulse on the grid of V_{2b}. V_2 anode drops; the grid of V_{1b} goes below cut-off; V_1 anode rises; the grid of V_{2a} opens, so that V_2 remains conducting and V_1 anode is high. V_3 remains conducting, as the grid of V_{4b} goes negative. The gate is restored to its initial state, having passed one time-base sweep to the stimulator.

Fig. 1. The complete circuit



Some Details of Construction

The bias of the grid of V_{1a} is controlled at the time-base. It will be seen that if V_1 is kept conducting, the gate is unable to shut itself off. The switch which controls the operation of the time-base also earths the grid of V_{1a} during other modes of operation of the time-base; that is when the time-base is running free, giving single sweeps triggered by the low frequency oscillator, or giving single sweeps externally triggered, every time-base sweep is accompanied by a stimulator pulse. The time-base employed has been slightly modified after Attree[1]. The fly-back of this time-base is exceedingly rapid so that no difficulty is had in differentiating the gate output to get a positive pulse at fly-back. The time-base signal is taken via a cathode-follower from the charging cathode-follower

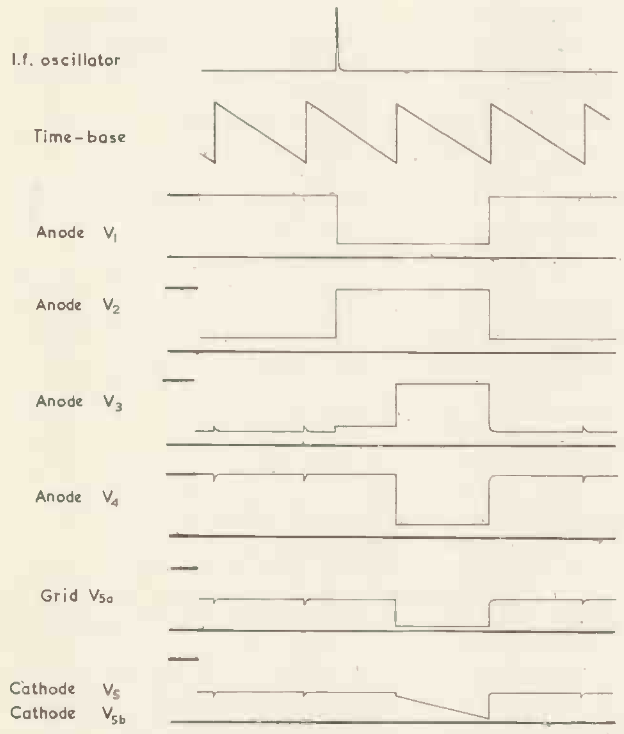


Fig. 2. Waveforms showing potential changes at several points in the circuit during the operation cycle

for the Miller circuit capacitors. The limits within which the Miller circuit runs are quite independent of sweep velocity, a necessary condition if d.c. level is to signal the beginning of time-base sweep.

The component values do not appear to be critical, although no systematic test of tolerance has been made. The h.t. voltage may be changed 50V or more without preventing the operation of the circuit. The drop in anode potential of V_4 before change of state in V_3V_4 takes place is approximately 100V. The drop could be greatly reduced by adjusting the coupling resistors to the grids, but in this circuit it is of no importance. The change in V_4 anode potential is much larger than necessary for the gating pulse. As long as the potential change at V_4 anode is larger than the waveform to be transmitted through the gate, the voltage-dividing resistor to the grid of V_5 can be adjusted so that the gate does not open until actual transition in V_3V_4 occurs. (The only limitation is the maximum permissible heater-to-cathode voltage in the cathode-followers.) In point of fact it does not matter if the gate opens at the beginning of the sweep, provided it closes before the stimulator is triggered.

The valves are all miniature types, and with no difficulty all components were mounted on a 2in by 4in by 7in chassis, attached to the back of the time-base. The time-base power pack provides the necessary h.t. and filament supplies: approximately h.t.+ 20mA, h.t.− 10mA, filament 6V at 2·1A.

The Paralleled Triode Bistable Circuit

The bistable circuit uses a double triode as a single valve, anodes and cathodes tied together. Both grids may pass current. This use of triodes is not new, but as far as the author knows, their use in a bistable circuit has not been published. As a trigger valve the operation is exactly opposite to that of the double-pentode trigger. In the pentode circuit the suppressor grid is also used as a control grid; either grid may cut the valve off. With one grid at earth in each valve, the anodes are coupled to the other grids so that a bistable pair is formed, only one valve conducting at a time. A negative pulse to the grid at earth of the conducting valve cuts off the valve causing the second valve to conduct, cutting off the other grid on the first valve. The trigger is refractory at the first grid to further pulses of either sign. In the 'paralleled' triode, either grid makes the valve conduct. One grid in each double valve is biased to cut-off; the others are coupled to the opposite anodes to form a bistable pair of triodes, as in V_1V_2. A positive pulse to the biased grid of the non-conducting double valve causes the other double valve to cut-off, opening the parallel grid on the first. The trigger is refractory at the first grid to further impulses of either sign. (There is a slight difference in anode voltage when one or when both grids are open, but it is insufficient to effect the anode voltage of the nón-conducting valve.)

The number of components required in each case is approximately the same. Although the pentode requires a screen resistor, the externally-controlled grid is normally at earth potential; only a single grid leak resistor is required. The externally controlled grid of the paralleled triode requires negative bias, usually achieved with a two-resistor voltage-divider chain. It would appear that the 'paralleled' triode bistable circuit has the same useful qualities as the pentode circuit; those of high impedance a.c.-d.c. input, and of refractoriness to subsequent pulses; but is activated by positive pulses. It should therefore be applicable wherever those properties are desired, and where positive control pulses are more readily available than negative.

The Circuit in Use

The visual display using the circuit is greatly improved. The ability to see all activity on one screen with the infrequent stimulus occurring at the same place each time it is fired is very convenient. There is some loss of discrimination of small responses in the presence of noise, but then one returns to single sweep operation. It is still advantageous to be able to switch rapidly back and forth without over-stimulating between continuous and single-sweep observation. Eye strain is much reduced as the level of illumination of the retina is more nearly constant.

Acknowledgment

This article represents work done at the Oxford University Laboratory of Physiology during the tenure of a United States Public Health Service Predoctoral Fellowship from the National Institute of Neurological Diseases and Blindness.

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A Wide Band Differential Amplifier of Unity Gain

By J. C. S. Richards*, B.Sc., Ph.D.

Many electronic instruments will only measure potentials with respect to earth. To enable such instruments to measure the alternating potential difference between two terminals, neither of which is earthed, a differential amplifier which will handle large input signals and has a single-ended output has been devised. The rejection ratio is greater than 500 and the gain, which is stabilized, is within 5 per cent of unity over the frequency range 5c/s to 500kc/s.

It is frequently necessary to observe the potential difference between two terminals neither of which is at earth potential. Unfortunately, many commercially available electronic instruments—cathode-ray oscilloscopes, valve voltmeters, etc.—have only two input terminals, one of which must be earthed. A number of differential amplifiers having single-ended output have been described, but most of them are intended to deal with signals of a few millivolts at frequencies less than 10kc/s. It was decided to develop a unit which would convert push-pull to single-ended signals and which would handle a sufficiently wide range of signals to make it useful as a general purpose instrument.

Such an instrument should have its gain stabilized at some convenient value (1, 10, or 1/10, say) over a wide range of frequencies (at least 10c/s to 100kc/s). It should be capable of handling signals as large as 100V r.m.s., and should give an output of at least 5V r.m.s. The rejection ratio need not be very much greater than 100 (500 is adequate), since the in-phase and out-of-phase signals are expected to be of the same order of magnitude, and few electronic instruments have an accuracy greater than 1 per cent. The unit must be portable, and any preset controls should need only occasional adjustment.

The Basic Circuit

The minimum number of stages in a feedback amplifier is a function of the final gain and degree of stability required. If the gain with feedback must remain within 2 per cent of unity, then a single stage should be sufficient. The basic circuit of a suitable single valve differential amplifier is shown in Fig. 1. It consists of a conventional cathode-coupled valve pair in which a fraction of the voltage of each anode is fed back to the corresponding grid. The output signal v_o is taken from the anode of V_1 only.

The grid to earth voltages (v_1 and v_2) of the valves may be written as:

$$v_1 = k_1 e_1 - k_1' R_1 i_1$$
$$v_2 = k_2 e_2 - k_2' R_2 i_2$$

where i_1 and i_2 are the anode currents of V_1 and V_2 respectively, and k_1, k_1', k_2 and k_2' are determined by the feedback networks R_6, R_7 and R_8, and R_9, R_{10} and R_{11}. The output voltage v_o may then be written:

$$v_o = \frac{-\mu_1 k_1 R_1 e_1 \{R_3(\mu_2+1) + R_2 + r_{a2} + \mu_2 k_2' R_2\} + \mu_2 k_2 R_1 e_2 R_3(\mu_1+1)}{(R+r_a+\mu k'R)\{R+r_a+\mu k'R+2(\mu+1)R_3\}} \quad \ldots\ldots\ldots (1)$$

In the denominator of this expression R, r_a, μ, k and k' have been written for the average values of R_1 and R_2, r_{a1} and r_{a2}, μ_1 and μ_2, k_1 and k_2, and k_1' and k_2' respectively, since small differences between R_1 and R_2 etc., are not important. Even if R_3 is not large, it is possible to make v_o zero when $e_1 = e_2$ by choosing the values of k_1 and k_2 suitably, but the required values of k_1 and k_2 depend on the value of r_{a3}, a parameter not likely to remain constant.

However, if R_3 is made much larger than $(r_{a2} + R_2)$ then one can write with sufficient accuracy:

$$v_o = \frac{-R_1\{\mu_1(\mu_2+1)k_1e_1 - \mu_2(\mu_1+1)k_2e_2\}}{2(\mu+1)(R+r_a+\mu k'R)} \quad \ldots. (2)$$

If k_1 and k_1 are now adjusted to make v_o zero when $e_1 = e_2$, the maintenance of balance depends only on μ_1 and μ_2, two notably constant valve parameters. There are two additional advantages: (1) if, for any reason μ_1 or μ_2, k_1 or k_2 changes in value, the magnitude of v_o predicted by equation (2) is in general much less than that predicted by equation (1); (2) the balance is independent of the actual value of R_3, provided only R_3 is large. This latter proviso is important at higher frequencies where R_3 is likely to be shunted by various stray capacitances.

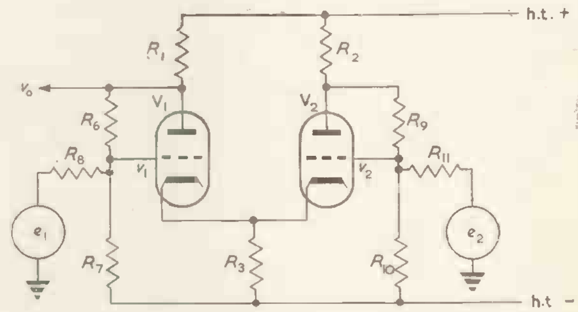

Fig. 1. Cathode-coupled valve pair with feedback

The in-phase gain M_i is defined as $v_o/(e_1 + e_2)$ when $e_1 = e_2$, and is obtained directly from equation (2) by putting $e_1 = e_2$. The differential gain M_d is defined as $v_o/(e_1 - e_2)$ when $e_1 = -e_2$. Since small differences between μ_1 and μ_2, k_1 and k_2 etc are not important in the expression for M_d we may write the differential gain as:

$$M_d = -(1/2) \cdot \frac{\mu k R}{R + r_a + \mu k' R}$$
$$= -(1/2)\, k \cdot \frac{A_o}{1 + k' A_o} \quad \ldots\ldots\ldots\ldots (3)$$

where $A_o = \mu R/(R + r_a)$, the gain of the valve used as a simple amplifier. For a pentode, $A_o \simeq gR$.

If $k'A_o \gg 1$, then $M_d \simeq (1/2)k/k'$. If A_o changes by p per cent say, then M_d will change by $p/(1 + k'A)$ per cent.

The Cathode Circuit

If equation (2) is to represent the behaviour of the circuit then the common cathode load (R_3 in Fig. 1) must be very much larger than $r_{a2} + R_2$. If V_1 and V_2 are triodes, this may be achieved by replacing R_3 by the anode impedance r_{a3} of a pentode. If the circuit analysis given so far is to be directly applicable when V_1 and V_2 are pentodes, the

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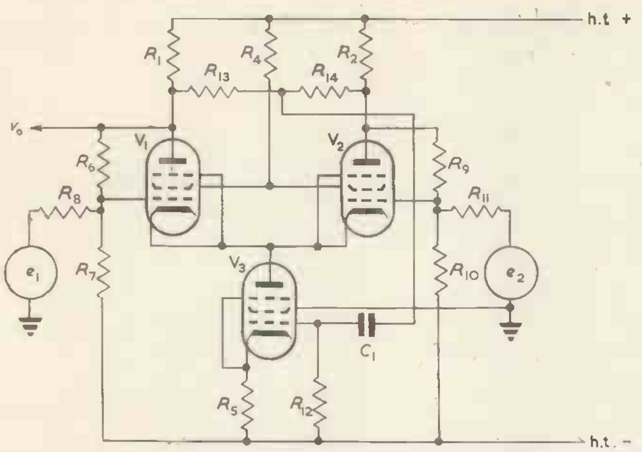


Fig. 2. Cathode-coupled valve pair using pentodes, with common mode feedback via R_{13}, R_{14}, C_1 and R_{12}

screen grids of V_1 and V_2 must be decoupled to their cathodes, in which case the effective resistance in the cathode circuit is r_{a3} in parallel with the common screen resistor (R_4, say) of V_1 and V_2. This value of resistance is too low. As a means of obviating this difficulty, a common mode feedback circuit[1,2] has been described, in which a voltage equal to the mean potential of the anodes of V_1 and V_2 is fed to the grid of the pentode (V_3, say), whose anode is connected to the common cathode of V_1 and V_2. The effective value of the cathode load then becomes $R'(1 + (1/2)\,g_{m3}R)$ where R' is the impedance of r_{a3} and R_4 in parallel, R is the mean value of the anode loads of V_1 and V_2, and g_{m3} is the mutual conductance of V_3. Unfortunately, this does not give a value of effective cathode impedance sufficiently large for the present purpose.

The circuit adopted is shown in Fig. 2. V_1 and V_2 are pentode valves, but their screens are not by-passed to the cathode. This circuit has been analysed, but the complete expression for v_o is rather lengthy and will not be given here. It turns out that, with any values of circuit parameters likely to be met in practice, the differential gain M_d is given with sufficient accuracy by equation (3). The in-phase gain M_i is proportional to $\mu_1'(\mu_2'+1)k_1 - \mu_2'(\mu_1'+1)k_2$, where μ_1', μ_2' are the inner amplification factors of V_1 and V_2. Hence, if M_i is made zero by adjusting k_1 and k_2, the maintenance of balance depends only on μ_1' and μ_2', which are reasonably constant valve parameters.

The Complete Circuit

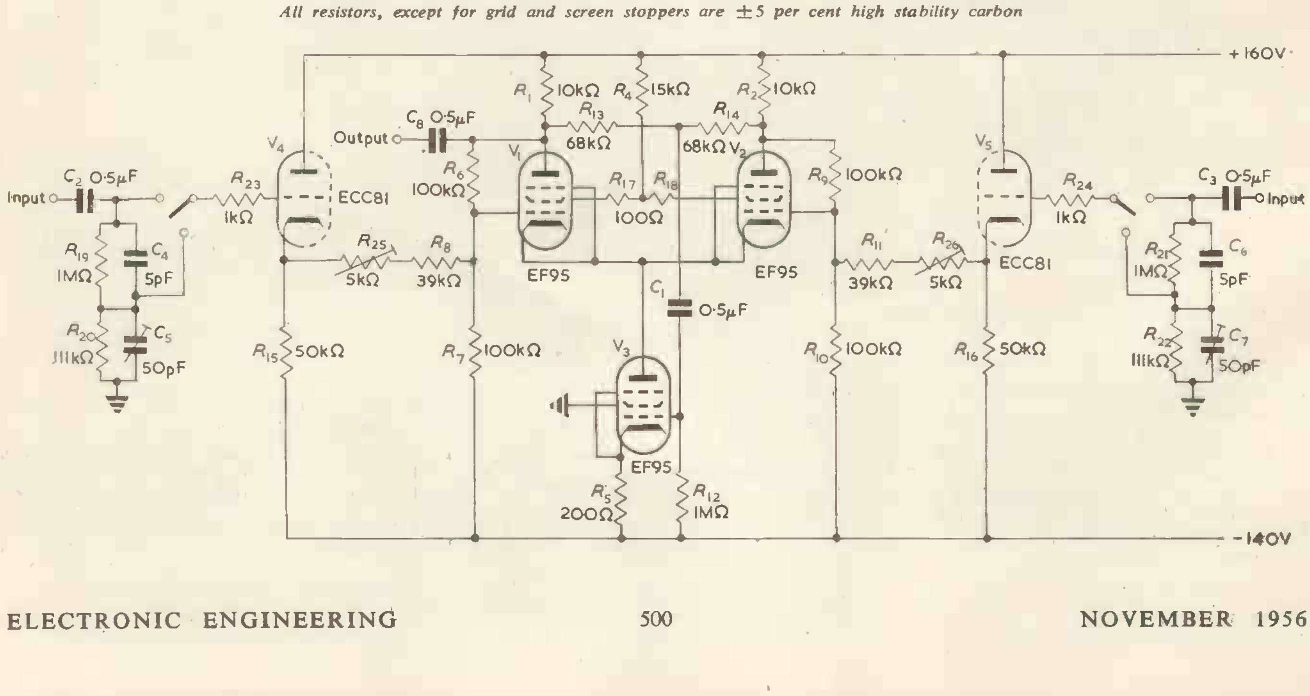
The complete circuit of a differential amplifier of unity gain is shown in Fig. 3. The cathode-follower valves, V_4 and V_5, give a high input impedance and isolate the feedback networks from the circuit under test. The differential feedback networks contain no blocking capacitors, so that the quiescent anode currents of V_1 and V_2 tend to remain the same when the heater voltage is varied or the valves are changed. It is well known that the mutual conductance of a valve is very closely related to its anode current, and only small variations are to be expected in the mutual conductances of V_1 and V_2. Hence the feedback network is much more effective in stabilizing the gain than would appear from equation (3) alone. In the present circuit, $1 + g_mRk'$ is only about 9, but experimentally it is found that the variations in M_1 caused by changing the valves (without adjusting the feedback network) rarely amount to more than ± 1 per cent.

Two variable resistors R_{12} and R_{25} are included in the circuit. R_{12} is used to set the gain exactly at unity. If a long-term variation in gain of ± 2 per cent can be tolerated, it may be replaced by a suitable fixed resistor. R_{25} is adjusted to make the output voltage a minimum when equal in-phase signals are applied to the input terminals. If a rejection ratio of about 100 is adequate, R_{25} may be replaced by a fixed resistor, but even when the rejection ratio must be more than 1 000, R_{25} requires only occasional readjustment. Both R_{12} and R_{25} must be carbon potentiometers, since the inductance of the wire wound variety makes an appreciable contribution to the impedance at higher frequencies. As the carbon potentiometers form only a small part of the feedback resistance network, and only a small steady current passes through them, their tendency to drift slightly in value is unimportant.

The resistors forming the input attenuation are chosen to be within $\pm\frac{1}{2}$ per cent of their nominal value, and the attenuation of the two networks is matched to within $\pm 0{\cdot}1$ per cent by adding a small series resistor to either R_{20} or R_{22} as required. The compensating capacitors C_5 and C_7 are adjusted in the usual manner with a square wave input signal. The input coupling capacitors C_2 and C_3 are made

Fig. 3. Circuit of unity gain differential amplifier

All resistors, except for grid and screen stoppers are ± 5 per cent high stability carbon



equal within ± 1 per cent, either by selection from ± 20 per cent tolerance stock components, or by adding suitable small capacitors in parallel as required.

It is necessary to provide adequate screening between input and output circuits; in particular, either the input or output lead to the amplifier must be screened. In the present instance the output is taken via a screened coaxial cable having a capacitance of about 60pF which is permanently connected to the amplifier. Most cathode-ray oscilloscopes, etc., have an input capacitance of between 10 and 50pF, so that the total capacitance across the amplifier output is of the order of 100pF. The performance figures given below were obtained under this condition.

Performance of the Amplifier

If the antiphase gain is set to unity at a frequency of 1kc/s, it remains within 5 per cent of unity over the range 5c/s to 500kc/s; it falls by 3dB at about 0·7c/s and 1·2Mc/s. With an in-phase input signal of 5V r.m.s. at a frequency of 1kc/s, the balance control R_{25} may be adjusted to give an output of the same order of magnitude as the hum and noise (about 1mV r.m.s.) so that the rejection ratio is about 5 000. Varying the heater voltage by ± 10 per cent will not cause the rejection ratio to fall below 1 000. Changing all the valves in the amplifier may cause the ratio to fall as low as 200, but it takes only a few seconds to readjust the balance controls to restore the rejection ratio of 5 000. In normal day-to-day use, if the balance control is only occasionally adjusted, a rejection ratio of more than 500 may be expected.

At low frequencies the rejection ratio falls because of the difference between the input time-constants. Typically, it is greater than 1 000 at 5c/s and does not fall below 300 at 1c/s.

The rejection ratio at high frequencies depends largely on the balance between the various stray capacitances which shunt the feedback resistors. The degree of balance required is very high—for a rejection ratio of 1 000 at 100kc/s the stray capacitances associated with the feedback network of V_1 should be within about 0·1pF of those of the feedback network associated with V_2. It is almost impossible to achieve this precision merely by arranging the relative position of components. The performance with a typical layout is given in line 1 of Table 1.

Fortunately the performance can be much improved by adding at most two (in practice often only one) compensating capacitor—one between one anode and its corresponding grid, and one between one grid and earth. These additional capacitors are generally less than 1pF. They are best made by soldering one end of a short length of 20 or 22 s.w.g. insulated copper wire to one of the two electrodes between which the capacitance is to be increased. The free end of the wire is brought near the wiring associated with the second electrode, and its exact position is adjusted to give the best possible rejection ratio at a frequency of 100 or 200kc/s. The choice of electrodes between which the capacitance is to be increased is made empirically. In practice, the adjustment is very simple and it takes only a few minutes to obtain a performance similar to that shown in line 2 of Table 1.

TABLE 1

Frequency (kc/s)	1	10	100	500	1 000
Rejection ratio (uncompensated)	5 000	1 500	150	40	15
Rejection ratio (compensated)	5 000	5 000	3 000	500	50

With the attenuators in the $\times 1$ position the amplifier will handle an antiphase input signal as large as 10V r.m.s. in the presence of an in-phase signal of 10V r.m.s. The attenuator increases the signal handling capacity to 100V r.m.s., which should be adequate for most purposes. When the in-phase input signal at the grids of V_1 and V_2 is made greater than 5V r.m.s. the harmonics generated in the valves reduce the rejection ratio, but it is still greater than 2 000 at 10V r.m.s. input.

Power Supplies

The h.t. power supplies must have a noise and ripple content of less than 1mV. Stabilization against long-term variations in the mains supply need not be very great, but the internal impedance of the h.t. supplies, particularly that of the negative line, must be less than about 20Ω over the amplifier pass band. This fairly stringent requirement on the power supply impedance arises because of the relatively low impedance level in the amplifier itself.

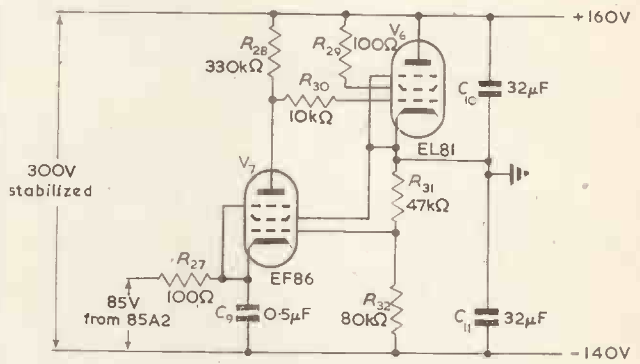


Fig. 4. Power supply for amplifier

A suitable power supply consists of a transformer with a 350-0-350V h.t. winding feeding a conventional full wave rectifier circuit with a capacitor input filter. The resulting voltage (of the order of 450V) is reduced to 300V by means of the cascode amplifier degenerative stabilizer described by Attree[3], with the minor modification that, since the output current is never more than 30mA, an EL81 is used as series valve in place of a 12E1. Neither the positive nor the negative end of the 300V supply is connected to earth; by means of the circuit shown in Fig. 4 it is split into +160V and −140V supplies.

Conclusion

By using two types of feedback and simply adjusted preset controls, it is possible to make a differential amplifier of unity gain which can be used as a push-pull to single ended convertor over the frequency range 5c/s to 500kc/s. The amplifier complete with power pack may easily be accommodated in a case 7in by 14in by 10in high, and weighs about 20lb, so that it is readily portable.

Some consideration has been given to the possibility of producing a modified form of this amplifier with direct coupling throughout, but in the author's experience the limitations of direct-coupled amplifiers are such that general purpose instruments are of relatively little use, and it is usually best to construct a special amplifier for the job in hand.

REFERENCES

1. Balanced Output Amplifiers of Highly Stable and Accurate Balance. Communication from E.M.I. Laboratories. *Electronic Engng. 18*, 189 (1946).
2. RICHARDS, J. C. S. An Improved Type of Differential Amplifier. *Electronic Engng. 28*, 302 (1956).
3. ATTREE, V. H. A Cascode Amplifier Degenerative Stabilizer. *Electronic Engng. 27*, 174 (1955).

Short News Items

Ardente Acoustic Laboratories Ltd have now moved to their new factory at 8-12 Minerva Road, North Acton, London, N.W.10. Telephone Elgar 3923.

Mr. W. M. York of E. K. Cole, Chairman of the Public Relations Committee of the Radio Industry Council, has resigned from the Panel of Judges appointed by the R.I.C. for awarding premiums for technical writing. His place has been taken by Mr. Arthur Clarkson of the General Electric Company.

The Pullin Group of Companies have recently opened new West End Showrooms at Electrin House, 93-97 New Cavendish Street, London, W.1.

The University of Sheffield and The United Steel Companies Ltd have come to an agreement for the purchase and installation of a Ferranti 'Pegasus' digital computer. The £50 000 computer, which has already been ordered from Ferranti, is to be delivered in March 1958, and will ultimately be installed in a new mathematics block at the university, for which approval has recently been given. Sheffield University and United Steel will each use the machine for their own research programme.

The Council of the British Institution of Radio Engineers has announced the award of prizes to candidates who took part in the Graduateship Examination during 1955. These are as follows. The President's Prize to A. C. Dev, a Junior Commissioned Officer of the Indian Electrical and Mechanical Engineers; The S. R. Walker Prize to G. R. Tyler, a medical laboratory technologist, in charge of the biochemistry laboratory at Sarnia General Hospital, Ontario; The Audio Frequency Engineering Prize to N. G. Lolayekar, who is in charge of the Audio Engineering, Measurement, and Radio Sections of Eastern Electric and Engineering Co, Bombay; The Electronic Measurements Prize to K. K. Nambiar, who has recently completed a post-graduate diploma course in electronic engineering at the Madras Institute of Technology.

Redifon Ltd announce that they have been awarded contracts worth nearly £500 000 for the supply and installation of a complete radio communications system for Iran's Police Forces. The project involves the supply of complete station equipment including transmitters, receivers, aerial systems, power plants and ancillary gear, together with the services of engineers to organize and supervise the installation programme.

Mr. J. M. Bedford, late Chief Engineer of the Radio Section of Messrs. Igranic Ltd, has recently joined the Research and Development Department of Parmeko Ltd.

Fielden Electronics Ltd have recently celebrated their tenth anniversary. Since October 1946 the company has been transformed from a small concern, manufacturing one electronic instrument for the textile industry, to an expanding international organization. The company now produces instruments to cover the fields of level, temperature, pressure, moisture and micro-measurement as well as textile and laboratory equipment.

A new Johnson Matthey associate company came into existence recently with the inaugural general meeting of Etablissements Johnson Matthey et Cie, S.A. in Paris. The issued capital is Frs. 40 000 000 of which Johnson, Matthey & Co Ltd hold $81\frac{1}{2}$ per cent. The company has been formed to develop Johnson Matthey sales in France. It has taken over the merchanting business of M. Pierre Motton, who has worked in close collaboration with the London company since 1917 when his family became agents for the sale in France of Johnson Matthey products for the ceramic industry. The new company has also absorbed the business of Maurice Carriere et Cie, who acted as correspondents in France for the Bullion Department of Johnson Matthey. The offices are at 76 Boulevard Haussmann.

The Third Instrument Display, organized by the Scientific Instrument Manufacturers' Association was recently opened at SIMA House, 20 Queen Anne Street, London, W.1. This latest display shows a further cross-section of the British instrument industry and will remain open until the end of the current year. Although a permanent exhibition, it will, however, allow some interchange of equipment during this period.

The Radio Industry Council have announced that the Scottish Radio and Television Exhibition will be held at Kelvin Hall, Glasgow, from 22 May to 1 June, 1957.

The Export Panel of the British Standards Institution has recently issued a statement on the policies it advocates with regard to international work on standards.

The Paper Making Section of the College of Science and Technology, Manchester, has recently been given laboratory plant and equipment worth about £5 000 by the Reed Paper Group.

Mr. G. A Marriott has been nominated as next President of the British Institution of Radio Engineers. He is Manager of the Osram Valve and Electronics Department of the General Electric Co, and is a Director of the Marconi-Osram Valve Co Ltd. He has served on the board of the British Radio Valve Manufacturers' Association since 1940, and has been Vice-Chairman and Chairman of the Association for several years. He has also been Chairman of the Radio Industry Council, and is still a member of the Executive Committee.

The BBC are presenting one or two research scholarships each year, valued at £385 per annum, to university graduates in Electrical Engineering or Physics who obtain good Honours Degrees, giving them the opportunity to work for a higher degree at any university in the United Kingdom, not necessarily at the one where they graduated. The scholarships are for two years in the first instance with the possibility of extension in suitable cases, if necessary. The scholarships are limited to male British subjects normally resident in the United Kingdom. The only condition applying to the subject for research is that it must be in those fields of telecommunications or physics which have an application to sound or television broadcasting. The Corporation has given one research scholarship only this year and that was to Mr. P. C. J. Hill who graduated at Birmingham University with 1st Class Honours in Electrical Engineering. Mr. Hill will conduct his researches in the Department of Electrical Engineering at the Imperial College of Science and Technology, under the supervision of Dr. Gabor.

Antex Ltd of 3 Tower Hill, London, E.C.3 are United Kingdom agents for the Victoreen Instrument Co, Ohio, U.S.A. This American company are manufacturers of voltage regulators and glass sealed resistors.

Atkins, Robertson & Whiteford Ltd of Glasgow and London have moved to a larger new factory in the Thornliebank Industrial Estate, Glasgow. Telephone Giffnock 1031/2.

Lion Electronic Developments Ltd have appointed Mr. L. E. Moore as Sales Manager for both the Electronics Division (Development and Production) and the Leocast Division (Resin Encapsulation Techniques). Mr. Moore was previously in charge of Technical Liaison.

Winston Electronics Ltd of Govett Avenue, Shepperton, Middlesex, have changed their telephone number from Walton-on-Thames 2732 to Walton-on-Thames 6321 (5 lines).

Notes from ——— NORTH AMERICA

'Avalanche' Diode Development

Bell Telephone Laboratories have developed techniques for making diffused junction silicon "avalanche" diodes having lower impedance and higher power capability than have been previously available. The "avalanche" diode is a special type of diode which exhibits a very sharp reduction in impedance at a specific inverse voltage. Devices having this characteristic are useful in voltage regulation or control devices, as voltage reference elements, and in signal circuits as surge protective elements. The breakdown voltage can be controlled over a range of about 5 to 500V by controlling the junction impurity gradient, and the breakdown voltage of specific units can be predicted quite closely from a knowledge of this gradient. Lowering the impurity gradient by increasing the depth of penetration increases the breakdown voltage.

Fabrication technique consists of diffusing phosphorus into a p-type silicon wafer to form a p-n rectifying junction. Careful control of depth of diffusion results in units having predictable characteristics. The ohmic junction on the other side of the wafer is formed by diffusing boron into the silicon, then plating with nickel. With this technique, contact resistances as low as $10^{-3}\Omega/\text{cm}^2$ can be achieved.

In the range of 10 to 100V, breakdown voltage can be predicted and controlled in production to within about 5 per cent.

Current prior to breakdown is of the order of 1μA or less for units rated at 10V and above. At breakdown, the impedance is reduced to a few ohms for currents in the milliampere range, and can be of the order of a fraction of an ohm for high current surges.

Temperature coefficient of voltage breakdown is about 0·07 per cent/°C for the higher voltage diodes, but substantially zero for units in the 6V region.

Slow-Scan Television

A compatible system now making it possible to economically combine slow and fast scan television has been developed by the General Electric Company. The system changes fast television to slow scan television by means of an electronic convertor.

Slow scan television produces one picture every 4 or 5 seconds and has its greatest possibilities in uses not requiring transmission of motion. Distortion of the picture results from movement of the subject, but pictures without much action, such as a picture of a cheque for signature comparison, can be sent over lines that approach the telephone line bandwidth and can be received by slow television monitors.

It is also possible to receive a fast television picture transmission without a direct hookup to the fast camera and convert it for slow-scan television.

The use of a television picture tube with a long-decay phosphor preserves the picture as the scan line sweeps slowly across the screen in forming the image.

Computer Reliability Conference

It is announced that the 1957 Western Joint Computer Conference will be held in the Statler Hotel in Los Angeles from February 26-28. The conference is to be under the joint sponsorship of the I.R.E., A.I.E.E., and A.C.M. The theme of the meetings will be "Techniques for Reliability".

'Picture-Phone' Using Telephone Wires

The Bell System's research and development organization has used an experimental picture-phone system to transmit recognizable pictures over distances as far as from New York to Los Angeles.

A technical description of this system was recently given at a joint meeting in Los Angeles of the Institute of Radio Engineers and the West Coast Electronic Manufacturers' Association.

Experimental pictures vary in size from 1 × 1½in to 2 × 3in and are viewed from one to two feet away. Unlike television, a new picture is displayed every two seconds. It has good black-and-white contrast and the person at the other end of the line is recognizable. Head and shoulders can be seen and facial expressions are readily apparent.

The picture-phone described is the first system of its kind to use a pair of ordinary telephone wires and therefore has promise of being commercially feasible. Only one other line, consisting of a pair of wires like the regular telephone line, would be installed on the customer's premises to carry the picture.

It will be possible for a caller's picture to be 'dialed' like an ordinary telephone call, provided the switch on the picture equipment is turned on at both ends of the line. It the switches are off, the telephone call will be completed without pictures.

Toroidal Coil Winder

The 'Midjet', produced by Electro Devices Company, Inc. Wilmington, Mass., is a versatile machine that permits winding finished coils having an inside diameter (hole) as small as 3/32in. The unique method of winding off the inside of the shuttle instead of the top makes this possible. Maximum finished coil diameter is 1in and maximum height, measured along the toroid axis, is ½in. The machine can use wire sizes from 30 to 46 s.w.g. and provides 360° coverage of the core, the wire being accurately wound on the core in any type of sector build-up required by the particular application. Multiple windings are also readily wound, with the multiple leads separated.

The winding speed, throughout the entire range of wire sizes, covers a range from 0 to 200 turns/min, the cam arrangement practically eliminating tensile shock to the wire, the major cause of wire breakage.

Polished winding guides confine the wire to a simple plane between the shuttle and the core; the wire flows through a hardened and polished vent in the shuttle, preventing kinks and loops and avoiding scarfing or abrasion to the inner surface of the coil and insulation on the wire.

Transistorized Amplifier

Baird Associates of Cambridge, Mass. announce the availability of the Model JG2 transistorized mixer amplifier, which is said to be ideal for remote-pickup recordings, outside public-address systems and 'on-the-spot' interviews. Model JG2 is self-contained and includes a 13V power supply which will accept any of a wide variety of standard mercury cells. These batteries, which have a life in excess of 100 hours are housed in spring-loaded holders that facilitate rapid replacement.

The equipment consists of two low-noise preamplifier stages, one low-noise mixer stage, one amplifier stage and one output stage, pnp junction transistors being used in all stages except the amplifier or driver stage, where use of an npn type stabilizes overall gain with temperature. The 85dB gain permits the use of a low-gain microphone. Choice of 50 or 250Ω input impedance, balanced or unbalanced, is provided and the standard 600Ω output impedance also balanced or unbalanced, permits direct coupling to telephone lines, tape recorder and other devices. The output level is 1mW average, 10mW peak, the signal-to-noise ratio 60dB for a-60dBm microphone input and the peak and average distortions are 2·5 per cent and less than 1 per cent respectively.

Snap-in Controls for Printed Circuits

A self-supporting, snap-in variable resistor for printed wiring has recently been announced by the Electric Component Division of Stackpole Carbon Company. This control, type LR-70, measures 57/64in diameter and stands 7/8in off the mounting board. It is supported by four legs—the three regular voltage taps and a larger, case-ground leg. No further mounting is required since the legs merely snap into the printed wiring board to form a strong support.

The shaft is supported by a 3/32in extruded bushing and may be left round or equipped with flat, screwdriver-slot, split knurl, diamond or straight knurl, or tongue. Single or double-pole snap switches are available with ratings from 15A, 15V d.c. to 6A at 125V a.c. or d.c.

The LR-70 is rated at 0·75W for values above 10kΩ and 0·5W below 10kΩ.

LETTERS TO THE EDITOR

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

Measurement of Self-Capacitance

DEAR SIR,—In his article in the August, 1956 issue (p. 350) Mr. J. P. Newsome recommends what he states to be a novel method, which is a modification of the self-resonance method. It is, however, identical with variety (b) of the previous method, which is well known and which Mr. Newsome justifiably describes as tending to give low accuracy. Except for different symbols, equation (9) relating to the recommended method (after it has been corrected by inverting the factor $(f_{sr}'^2 - f^2/f^2)$) is the same as equation (5).

Although described as a detailed survey, the article does not refer to what is believed to be the only method capable in ordinary practice of high precision[1]. In it, the answer is given as one third of the difference in capacitance between two settings of a variable capacitor; no absolute values of capacitance or values of frequency are required.

Yours faithfully,

M. G. SCROGGIE

Bromley,
Kent.

1. SCROGGIE, M. G. A Method of Measuring the Self-Capacitance of Coils. *Wireless Eng.* 10, 477, (1933).

The Author Replies:

DEAR SIR,—In reply to Mr. Scroggie, the method to which equation (5) has reference is based on resonating the inductor at two frequencies below the self resonant frequency whereas the method to which equation (9) refers is based on a determination of the self-resonant frequency (f_{sr} or f_{sr}'). Since differing experimental procedures are involved, the methods cannot be indentical.

When C_s is small, the term $(C_{T_1} - n^2 C_{T_2})$ in equation (5) can only be determined with adequate accuracy if a precision variable capacitor and wavemeter are employed; the quantities f_{sr} and f_{sr}' can be quite accurately measured without such resources. It is upon this fact that the Q meter method of measuring self-capacitance (using equation (11)) can be expected to give results of adequate accuracy without the need to employ precision measuring equipment.

I regret the absence of Mr. Scroggie's method of measuring self-capacitance from the article and take the liberty of mentioning a form of it in this correspondence. With reference to Fig. 3(a) of the article, C_T is replaced by a fixed capacitor of capacitance C_F and the oscillator adjusted to give resonance; if C_F is now replaced by an incrementally calibrated capacitor which to restore resonance is adjusted to a value C_T then $C_{T_1} = C_F$. The capacitances C_{T_1} and C_F are placed in series across the test inductor and frequency adjusted to give resonance at a value f_r; then

$$f_r = \frac{1}{2\pi\sqrt{[L(C_s + C_{T_1}/2)]}} \qquad \text{(A)}$$

The frequency is then adjusted accurately to a value $f_r/2$ and the two capacitors are paralleled with the test inductor: to restore resonance the variable capacitance is now set to a value C_{T_2}, thus

$$f_r/2 = \frac{1}{2\pi\sqrt{[L(C_s + C_F + C_{T_2})]}} = \frac{1}{2\pi\sqrt{[L(C_s + C_{T_1} + C_{T_2})]}} \qquad \text{(B)}$$

From equations (A) and (B) we have,

$$C_s = 1/3\,(C_{T_1} - C_{T_2}) = \Delta C_T/3 \qquad \text{(C)}$$

where C_T is the necessary alteration in C_T in passing from the first to the second measurement.

Yours faithfully,

J. P. NEWSOME

The University of Nottingham,
Nottingham.

DEAR SIR,—In an interesting article published in your August 1956 issue, Mr. J. P. Newsome describes various procedures for evaluating the self-capacitance C_s of an inductor. One of these methods utilizes the Q-meter circuit of his Fig. 3(b) to obtain the values of the tuning capacitor C_t required for resonating the circuit (maximum reading on V) at a series of known operating frequencies, f, which are less than the resonant frequency f_{rs} of the inductor. Then, on plotting the reciprocal of f^2 versus C_t, a linear graph is obtained, which has a slope equal to the product $4\pi^2 L$, an intercept with the vertical axis equal to the reciprocal of f_{rs}^2, and an intercept with the horizontal axis equal to the negative value of C_s.

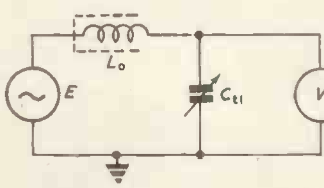


Fig. A. The Q-meter circuit

E = constant voltage source, frequency adjustable and calibrated

V = high impedance electronic voltmeter

For precise work, this method is limited by the fact that C_t has a minimum value substantially greater than zero. This means that the highest frequency at which a datum point can be taken will be substantially less than f_{rs}. Consequently, the graph must be extrapolated downwards to the left in order to determine both the $1/f_{rs}$ and C_s intercepts. Observation errors will cause a certain degree of uncertainty in the exact location of the best representative straight line and in the precise evaluation of these intercepts.

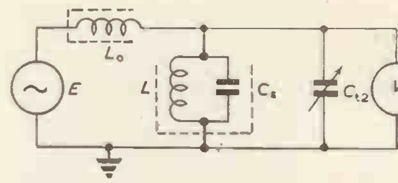


Fig. B. The unknown inductor, added in parallel with C_t

As an appendage to Mr. Newsome's article, I would offer a two-step technique which I have found to give more reliable results. Referring to Fig. A, a suitable reference inductor L_o, preferably shielded, is first introduced alone into the Q-meter circuit, and the resonating value C_{t_1} of the tuning capacitor obtained in the usual manner.

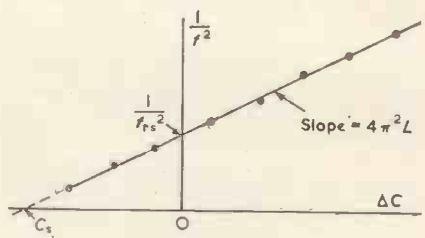


Fig. C. The linear graph obtained

The unknown inductor L, and C_s (which may or may not be shielded) is then added in parallel with C_t, giving Fig. B. A new value C_{t_2} of the tuning capacitor is found which will give a new maximum reading of V at the same operating frequency. For each value of f, the algebraic difference $\Delta C = C_{t_2} - C_{t_1}$ is thus obtained. A plot of $1/f^2$ versus ΔC then determines a straight line, Fig. C, having the equation:

$$(1/f^2) = 4\pi^2 L\,(\Delta C + C_s) = 4\pi^2 L \Delta C + (1/f_{rs}^2)$$

At frequencies less than f_{rs} C_{t_2} will exceed C_{t_1}, and the inductor will have a positive phase-angle and net inductive reactance At frequencies greater than f_{rs}, C_{t_1} will exceed C_{t_2} (ΔC being negative), and the inductor will function with a negative phase-angle and possess a net capacitive reactance. Oviously, if f is adjusted to equal f_{rs} the unknown inductor will have a zero phase-angle and ΔC will be zero, thereby affording a direct empirical evaluation of f_{rs}.

By this procedure, datum points for the linear graph may be obtained in both the first and second quadrants, resulting in more accurate and precise intercept values. If the operating frequencies can be duplicated accurately, the initial tests for the C_{t_1} values need not be repeated when measuring a series of unknown inductors.

The reference inductor L_o must be chosen to have its natural frequency above the operating frequency and its inductance of a value which permits both C_{t_1} and C_{t_2} to lie within the range of the tuning capacitor. It is not necessary to use the same L_o for all the datum points in Fig. C. A series of high Q shielded reference inductors, covering the entire oscillator frequency range, may be purchased with most commercial Q-meters.

Yours faithfully,

H. W. LAMSON,
General Radio Company,
Cambridge,
Massachusetts.

The author replies:

DEAR SIR.—Mr. Lamson has quoted a very interesting extension of the Q meter method of determining the self-capacitance of an inductor in which the test inductor is placed in shunt with the Q meter tuning capacitor, which is already adjusted to give resonance with a suitable screened inductor.

Mr. Lamson's technique represents an improvement in that (1) the method is graphical and should yield greater accuracy than single results, (2) it eliminates the need for a low frequency measurement of inductance in order to determine C_s.

In undertaking measurements with $f > f_{sr}$ the test inductor may be working beyond its normal working frequency range: this may result in (1) a high loss component which will make ΔC difficult to determine accurately, (2) a possibility that the equivalent circuit of L in shunt with C_s, both having fixed values, may not be correctly representing the behaviour of the inductor; such a condition would probably cause departure from linearity of the graph shown in Fig. C. It is, of course, not essential to undertake measurements with $f > f_{sr}$ to obtain good accuracy: in this case points on the negative part of the ΔC scale would be absent.

I have always found the Q meter method as quoted in my article to give reliable and consistent results. Equation (2) of the article is capable of graphical treatment, but it is not in such a convenient form as Mr. Lamson's equation.

Yours faithfully,

J. P. NEWSOME,
The University of Nottingham,
Nottingham.

The Nth Power of a 2 × 2 Transfer Matrix

DEAR SIR.—2 × 2 transfer matrices have often been used in this journal and, particularly in analysing iterated networks, the n^{th} power of such a matrix is frequently required.

The purpose of this note is to point out the relationships which exist between various useful forms which can be obtained.

Reciprocity will not be assumed; that is to say, the results apply generally and not only when the determinant is unity.

The standard method of raising a matrix A to its n^{th} power is that of diagonalization. The matrix is first diagonalized, i.e. expressed, as $A = V\Lambda V^{-1}$ where Λ is the diagonal matrix of eigen-values and the columns of V are eigen vectors of A. Then $A = V\Lambda^n V^{-1}$ intermediate $V^{-1}V$s' cancelling out.

This gives an expression for A^n in terms of the eigen values λ_1, λ_2 of A, which satisfy

$$\lambda^2 - (a_{11} + a_{22})\lambda + \Delta = 0 \quad \ldots.(1)$$

where $A = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix}$ and $\Delta = \det A$ $= a_{11}a_{22} - a_{12}a_{21}$. and their corresponding eigen vectors $[Z_{1,1}]$ and $[Z_{1,2}]$ say. The expression is

$$A^n = \begin{bmatrix} Z_1\lambda_1{}^n - Z_2\lambda_2{}^n & -Z_1Z_2(\lambda_1{}^n - \lambda_2{}^n) \\ (\lambda_1{}^n - \lambda_2{}^n) & -Z_2\lambda_1{}^n + Z_1\lambda_2{}^n \end{bmatrix} . 1/(Z_1 - Z_2) \quad \ldots.(2)$$

This result has been given by Fisher[1], with the above reasoning couched in more physical terms; he shows that Z_1 and Z_2 are the iterative impedances of A, and that λ_1 and λ_2 are the phase factors for the two eigen-vector "waves". Evaluating Z_1 and Z_2.

$$Z_1 = \frac{\lambda_1 - a_{22}}{a_{21}} = \frac{a_{12}}{\lambda_1 - a_{11}}.$$

$$Z_2 = \frac{\lambda_2 - a_{22}}{a_{21}} = \frac{a_{12}}{\lambda_2 - a_{11}}$$

Substitution of these values in equation (2) leads to:

$$A^n = \frac{\lambda_1{}^n - \lambda_2{}^n}{\lambda_1 - \lambda_2} A - \frac{\lambda_1{}^{n-1} - \lambda_2{}^{n-1}}{\lambda_1 - \lambda_2} . \lambda_1\lambda_2 I \ldots.(3)$$

This result, which was obtained in a rather different way by Shen[2], expresses A^n in terms of A and its eigen values only.

Now $\lambda_1\lambda_2 = \Delta = \det A$ from equation (1), so putting $\sqrt{(\lambda_1/\lambda_2)} = e^{j\phi}$,

$$\lambda_1 = \Delta^{\frac{1}{2}}e^{j\phi}; \ \lambda_2 = \Delta^{\frac{1}{2}}e^{-j\phi} \quad \ldots.(4)$$

On substituting these values into equation (3), a very convenient form for A^n appears:—

$$A^n = \Delta^{(n-1)/2} \frac{\sin n\phi}{\sin \phi} A - \Delta^{n/2} \frac{\sin (n-1\phi)}{\sin \phi} I \quad \ldots(5)$$

Where $\cos\phi = \frac{1}{2}(e^{j\phi} + e^{-j\phi})$

$$= \tfrac{1}{2}\Delta^{-\frac{1}{2}}(\lambda_1 + \lambda_2) \text{ by (4)}$$

$$\therefore \cos\phi = (t/2\Delta^{\frac{1}{2}}) \quad \ldots.(6)$$

where $t = \text{trace } A = a_{11} + a_{22}$ since $\lambda_1 + \lambda_2 = a_{11} + a_{22}$ by equation (1); the trace of a matrix is an invariant and is equal to the sum of its eigen values and to the sum of its diagonal terms.

Now the Chebyshev function $U_n(z) = \sin n\phi$ where $\cos\phi = z$ (associated with the Chebyshev polynomials $T_n(z) = \cos n\phi$ where $\cos\phi = z$) is $\sqrt{(1-z^2)}$ times a polynomial of degree $n - 1$. In fact

$$U_n(z) = \sqrt{(1 - z^2)} \sum_{r=1}^{/(n_{+1})[2]} \binom{n-r}{r-1} (2z)^{n-r}/(-2z)^{r-1}$$

If $F_n(t,\Delta)$ is now written

$$F_n(t,\Delta) = \Delta^{(n-1)/2} \frac{U_n(z)}{\sqrt{(1-z^2)}} \quad \ldots.(7)$$

where $z = (t/2\Delta^{\frac{1}{2}}) = \cos\phi$ by equation (6),

$$F_n(t,\Delta) = \Delta^{(n-1)/2} \frac{\sin n\phi}{\sin\phi}$$

and equation (5) becomes:

$$A^n = F_n(t,\Delta) A - F_{n-1}(t,\Delta).\Delta I \quad \ldots.(8)$$

This result expresses A^n in terms of trace A and det A, both fundamental invariants of A, through the polynomial F_n which is closely related to the Chebyshev functions. Fekhikher[3] first obtained this result by other means.

To sum up, equation (2) gives A^n in terms of physical constants of the structure. Equations (5), (6) have proved to be more useful in the theoretical analysis of structures, in particular when mirror symmetric sections have been considered.

i.e. $a_{11} = a_{22} = (t/2) = \Delta^{\frac{1}{2}}\cos\phi$

$$\therefore A = \Delta^{\frac{1}{2}} \begin{bmatrix} \cos\phi & j\sqrt{(a_{12}/a_{21})}\sin\phi \\ j\sqrt{(a_{21}/a_{12})}\sin\phi & j\cos\phi \end{bmatrix} \quad \ldots.(9)$$

which, substituted into equation (6), gives:

$$A^n = \Delta^{n/2} \begin{bmatrix} \cos n\phi & j\sqrt{(a_{12}/a_{21})}\sin n\phi \\ j\sqrt{(a_{21}/a_{12})}\sin n\phi & \cos n\phi \end{bmatrix} \quad \ldots.(10)$$

Equations (3) and (8) are of mainly theoretical interest; equation (3) gives A^n in terms of A and all (i.e. both) the eigen values; equation (8) gives A^n in terms of A and all the spurs (i.e. trace A and det A) of A.

The above methods break down when the eigen values are equal.

i.e. $\lambda_1 = \lambda_2 = (t/2) = \pm\Delta^{\frac{1}{2}}$

i.e. $\cos\phi = \pm 1$; $\sin\phi = 0$.

In this case the limits of equation (5) as $\cos\phi \rightarrow \pm 1$ are required.

$$\cos\phi = 1: A^n = \left\{nA - \Delta^{\frac{1}{2}}(n-1)I\right\} (\Delta^{\frac{1}{2}})^{n-1}$$

$$\cos\phi = -1: A^n = \left\{nA - (-\Delta^{\frac{1}{2}})(n-1)I\right\} (-\Delta^{\frac{1}{2}})^{n-1} \quad \ldots.(11)$$

D. T. SWIFT-HOOK,
Research Laboratories of The General Electric Co. Ltd.,
Wembley,
Middlesex.

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BOOK REVIEWS

Automatic Digital Computers

By M. V. Wilkes. 305 pp. 50 figs. Demy 8vo. Butterworths Scientific Publications. 1956. Price 32s.

Automatic Digital Calculators

By A. D. and K. H. V. Booth. 261 pp. 60 figs. Academic Press Inc., New York. Butterworths Scientific Publications. 1956. Price 32s.

BOTH of these books are intended to serve as an introduction to the rapidly expanding field of automatic digital computing. The authors of both books were among the pioneers in the development of this field in this country, and are thus well qualified to write such an introduction to the subject. Although the ground covered in each book is similar, there are differences in the treatment of the subject and in the emphasis placed on certain sections. The interest of both books would have been considerably widened if a somewhat broader survey of the present position had been included.

Automatic Digital Computers by M. V. Wilkes starts with an interesting chapter devoted to the history of the development of automatic digital computing from the early ideas of Babbage (which were never practically realized until the inventions of electronics were applied in the field) up to the more recent developments since 1945. There is a clear discussion of the elements of logical design and a chapter on programming is included. Of more direct interest to the electronic engineer are the chapters on storage systems and switching and computing circuits. These chapters serve as a general introduction to the physical principles which have been employed in digital storage systems and to the types of circuit which are used. The use of Boolean algebra in the design of switching circuits is treated briefly with some indication of its limitations. The provision of a complete chapter on computers built with electromagnetic relays seems to place undue emphasis on this early development. It would have increased the value of the book, if, instead, this space had been devoted to a fuller discussion of the advances in logical design which have been made in the last few years. The presentation is lucid and the book is easily read.

There is an error of fact on page 197 where the current necessary to switch one of the cores in the MIT matrix store should be 1A instead of 1·6A. There are also a few misleading errors in the text. On page 192 capacity has been written where inductance is surely meant and on page 193 the turns ratio in the shifting register being described must be greater, not less than unity as stated. An error on page 234 in the formula for one of the input levels to the discriminator, $rV/(R+r)$ instead of $rV/(R+2r)$ leads the author to take, instead of the usual limiting case $R \ll r$, the other extreme where $r \ll R$. This leads to the approximate values given in the test which, of course, all tend to zero.

The second edition of *Automatic Digital Calculators*, the first edition of which was reviewed in these columns in February 1954, has been revised but, unfortunately, not as completely as one would have wished. For example, the survey of computing machines given on pages 16 to 20 is very incomplete and the figures quoted for the operating times of several machines are incorrect. Operating times without the addition of store access time are very misleading and of limited use in assessing the relative performance of machines. Again, since the ferrite core matrix store of Whirlwind I has been operating continuously at MIT for the last three years, and similar systems have been incorporated in several other machines, a clear statement that this form of store was no longer in the experimental stage could have been included.

This book covers roughly the same ground as the previous one, with rather more space devoted to programming and the applications of machines. The book is divided into small sections, there are six chapters with less than ten pages, and the presentation suffers from a lack of continuity. The section on logical design is very specialized. dealing only with the slower type of serial machine with a magnetic drum store, the logical design of which differs radically from the fast parallel machines now in operation. Because of the confusion between the terms immediate access or delayed access storage, and the parallel or serial operation of a store, the statement on page 24 that the organization of a machine depends to a large extent upon the choice of store, is misleading. Surely the designer has to choose the mode of operation of the store to suit the type of machine required. In the chapters dealing with circuit design "rise time" is invariably used when "time constant" is meant.

The circuit design of digital computers is treated very differently in the two books. In *Automatic Digital Computers* an attempt has been made to indicate some of the problems peculiar to this branch of electronic engineering and to provide an introduction to the philosophy governing the subject. The method adopted by the authors of *Automatic Digital Calculators* on the other hand. is to give specific circuits, with a description of their operation. Unfortunately there is no indication of the limitations of the circuits given. When circuit component values are quoted and stated to be suitable, the reader must remember that their suitability is, in general, limited to the slower type of machine.

Both books conclude with a section on the applications of computing machines and a discussion of the relationship between machines and intelligence, a subject leaving scope for a good deal of speculation. The bibliography and index appended to each book is adequate to guide the reader requiring more detailed information. One point that emerges from reading both books is the need for standardization of terminology and of the graphical symbols used, in this important new field.

W. RENWICK

Television Engineering—Principles & Practice
Volume Two
Video-Frequency Amplification

By S. W. Amos and D. C. Birkeinshaw. 272 pp. 156 figs. Demy 8vo. Iliffe & Sons Ltd. 1956. Price 35s.

THIS book is one of a series of BBC engineering training manuals and deals solely with video frequency amplification. It is written for and is mainly useful to the engineers working on the transmitting end of television. Most receiver engineers are unlikely to need to know so much about video frequency amplification. An exception would be those engineers concerned with distributing or relaying television at video frequencies.

As would be expected from the two authors, the book is extremely workmanlike and covers the subject thoroughly and in a readable and interesting manner. After a list of the symbols used throughout the book (a very good practice), the requirements of a video amplifier for British television standards are explained. A number of actual photographs of test card C are shown illustrating the effects of various distortions and one which to all intents appears to be a perfect reproduction of the C card is labelled "... reasonably free from distortion". This is, to say the least, an understatement! If only our receivers were all as good.

Simple *RC* coupled valves are next discussed followed by circuits to improve their frequency and phase response. The series shunt and combined series and shunt inductance methods are described in detail.

The next chapter concerns circuits employing cathode components for frequency correction by means of current feedback. It may have been appropriate in this chapter to point out that where valves are worked over large parts of their grid bases, quite large changes of g_m occur and correction methods such as these, which involve g_m, are often a compromise between correction at high and low signal levels. Possibly the BBC do not use large parts of grid bases. Certainly Gamma change and correction is not dealt with at all in the book; a notable omission.

Chapter VI deals with the relation between passband and gain, figure of merit of valves and limiting values of gain and frequency.

Cascaded and distributed amplifiers are next described. Chapter IX deals most interestingly with phase equalizers. As in all the other chapters it is plentifully illustrated by curves and examples and where mathematics occur they are clear and easy to follow.

This reviewer would have liked more practical details of Derivative Equalizers, but Lattice, T and Bridged-T structures are very fully covered.

Intervalve couplings and the troubles they bring follow, and anode, screen grid and cathode decoupling circuits are next discussed, including methods of compensating for imperfect decoupling. Valves used purely for decoupling anodes or cathodes, such as common cathode circuits, are not mentioned, but screens fed from a cathode follower are briefly mentioned. D.C. coupling problems are next briefly dealt with. As feedback is so useful it is not surprising that the authors chose to devote a large section to the subject and cover it very neatly.

The book ends with a short chapter on circuit noise and a useful bibliography.

C. H. BANTHORPE

Electronics and Electron Devices

By A. L. Albert. 572 pp. 70 figs. Demy 8vo. 3rd edition. The Macmillan Co. New York and London. 1956. Price 44s.

THE author of this book is Professor of Communication Engineering at Oregon State College. It is the third edition of "Fundamental Electronics and Vacuum Tubes, the title having been modified in accordance with present usages. It has been rewritten almost entirely, and much of the original descriptive material has been eliminated to provide space for new material.

An important feature is the chapter on Magnetic Amplifiers. A chapter is also included on Wave-Shaping and Control Circuits, covering differentiating, integrating, limiting, clipping, multivibrator and similar circuits. As in previous editions, the Standards of the American Institute of Electrical Engineers and of The Institute of Radio Engineers have been followed closely.

Abacs or Nomograms

By A. Giet. 235 pp. 152 figs. Demy 8vo. Iliffe & Sons Ltd. 1956. Price 35s.

THIS book has been translated from the French by Miss Helen Phippen, a Senior Research Assistant at the Science Museum, South Kensington, and the editor, Mr. J. W. Head, is mathematical consultant to the Research Department of the British Broadcasting Corporation.

The edition has been specially adapted for English readers and will prove of practical interest to engineers and physicists, and others who require time-saving methods when performing repetitive and complicated calculations.

ELECTRONIC EQUIPMENT

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

TUNABLE X-BAND MAGNETRONS

(Illustrated below)

Mullard Ltd, Century House, Shaftesbury Avenue, London, W.C.2

THE JPT9-01 is a tunable X-band integral magnet continuous wave magnetron with waveguide output. The valve delivers a c.w. output of 5 to 10W over a 450Mc/s band, centred on 9375Mc/s in the 3cm band. Tuning is by a single control, with a total range of 800Mc/s, including the stated 450Mc/s, band.

The JPT9-02 is a similar valve but intended for pulsed operation.

45kW DIELECTRIC HEATING GENERATOR

(Illustrated below)

Radio Heaters Ltd, Eastheath Avenue, Wokingham, Berkshire.

A 'RADYNE' high power dielectric generator unit has just been released, which is designed to be used singly or in batteries feeding on to an oven with a conveyor running behind a row of units. Each unit is capable of delivering 45kW on a continuous basis at a frequency of 18 to 20Mc/s and is capable of removing 120lb of moisture per hour from materials which are almost impossible to dry in any other way.

The equipment will feed conveyors up to 5ft wide and is of high reliability. On many difficult drying operations that previously took hours to be completed, electronic dielectric heating permits drying times of a few minutes.

It is claimed that the new 'Radyne' 45kW unit is the largest unit of this capacity to be in regular production outside the United States of America.

VOLTAGE REGULATORS AND GLASS SEALED RESISTORS

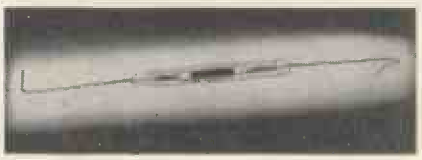
(Illustrated below)

Distributed by Anglo-Netherland Technical Exchange Ltd, 3 Tower Hill, London, E.C.3

A SERIES of corona regulator and negative glow discharge tubes is manufactured by the Victoreen Instrument Co., Cleveland, U.S.A., typical applications being:

(a) Voltage regulators of low current power supplies for geiger counters, photomultiplier tubes, c.r.t. second anode, focus and convergence, infrared tubes and klystrons.
(b) Reference tubes in series regulated power supplies.
(c) Voltage limiters.
(d) Trigger tubes, connected directly in h.t. circuits.
(e) Oscillators with a high stable output.
(f) Coupling elements in h.t. circuits.

Of special interest to the television industry is the series of Victoreen very high voltage regulators, with operating voltages from 15 to 27kV. Regulators can be made to special order for any operating voltage between 50V and 50kV. The same manufacturer also produces glass-sealed resistors covering the range 100 to $10^{14}\Omega$. These are of high accuracy and good long-term stability. Two types of resistor are produced; the HI-MEG resistor, originally developed for use in electrometer circuits and having a carbon coated element, vacuum-sealed in a glass envelope, and the deposited carbon resistor, the element of which is sealed in a glass envelope which contains an inert gas.



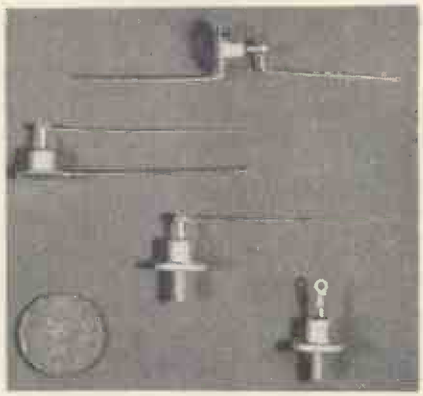
SILICON JUNCTION RECTIFIERS

(Illustrated above right)

The British Thomson-Houston Co. Ltd, Rugby.

A SERIES of low-power silicon junction rectifiers has been developed, which will offer many advantages where high temperature working and small size are required. With these silicon rectifiers, a maximum junction temperature of 200°C is possible, allowing operation in ambient temperatures up to 175°C. At these temperatures, inverse voltages up to 300V are possible, but if the maximum junction temperature is limited to 120°C, allowing operation in ambients up to 100°C, inverse voltages up to 400V are possible.

Another important application of the silicon rectifier is as a voltage reference (Zener Diode) particularly in the range of 4 to 8V and suitable for junction temperatures up to 250°C.



The units are all encapsulated in identical fashion. They are hermetically sealed by means of a projection weld and are being made in three forms, wire-ended, stud-mounted with top flying lead and stud-mounted with top soldering tag.

TEMPERATURE SENSITIVE RESISTORS

Welwyn Electrical Laboratories Ltd, Bedlington, Northumberland.

UNDER the proprietary name 'Welistors' these components are at present being manufactured in rod form and are provided with wire terminals for soldering direct into electrical, radio or electronic circuits. However, the method of manufacture is such that elements can be made having various shapes, sizes and characteristics, while constant research is being carried out in order to increase the range.

'Welistors' lend themselves to a number of applications in various forms of circuits. They can be used either as an element sensitive to changes in ambient temperature or as a self-heating component, enabling a time factor to be introduced.

Under the former category fall such applications as the temperature compensation of coils, transformer windings, focusing and picture-height coils in television sets, moving-coil meters and transistors.

The second effect is widely used for the limitation of current surges where valve heater filaments are connected in a series chain, as a shunt across indicator lamps where these are used in series with other components and also as a form of time delay usually in connexion with one or more electro-mechanical relays.

AUDIO OUTPUT VALVE

The General Electric Co. Ltd. Magnet House, Kingsway, London, W.C.2.

THE KT88, with an anode dissipation of 35W has been introduced as being a higher-power version of the KT66, although smaller in size. It does not replace the KT66, but is complementary to it for output powers in excess of those readily available from existing KT66 circuits.

At a supply voltage of 500V, with auto-bias operation, the available power output is 50W or twice that obtainable from a pair of type KT66, while at a supply voltage of 560V, with fixed bias, an output power of 100W is available.

The KT88 has a larger cathode, allowing for a higher mutual conductance, and a more modern type of construction, permitting the use of higher anode voltages and dissipations. It is designed for use mainly in push-pull circuits and will operate satisfactorily as either a triode or a pentode.

FREQUENCY METER

(Illustrated below)

The British Thomson-Houston Co. Ltd, Rugby.

TYPE JT2 frequency meter consists basically of a saturating transistor amplifier which charges and discharges a capacitor on each cycle of the input voltage. As the mean current through the capacitor is proportional to the input frequency, it can be rectified and shown directly in terms of cycles per second on the scale of a d.c. moving-coil meter.

Provision is made on a rotary switch for selecting any one of six frequency ranges with full-scale deflexions of 300c/s, 1kc/s, 3kc/s, 10kc/s, 30kc/s, or 100kc/s, respectively. On all but the highest range the accuracy of measurement is better than 1 per cent of full-scale deflexion; even on the highest range this accuracy holds good up to about 70kc/s, but because of limiting transistor characteristics, measurements above this frequency may be subject to slightly greater inaccuracy, but not more than 5 per cent at 100kc/s.

To maintain the inherent accuracy of the equipment over a wide input voltage range, another switch is provided to select one of two appropriate inputs to the amplifier to suit the applied voltage, thus the instrument can be used to measure frequencies within the accuracies mentioned above with inputs ranging from 0·1V to 500V.

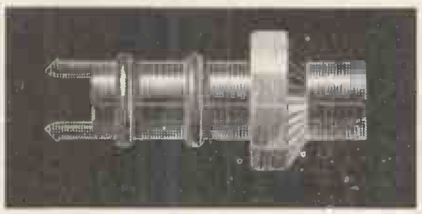
The instrument is essentially a current-operated device and takes 40μA r.m.s. for a sine wave input and up to 110μA peak-to-peak for other waveforms.

The long-term stability of the meter is excellent and when using the internal battery as a source of reference voltage the calibration drift is negligible. The meter is primarily designed for use with the battery as a reference (a life of about 2 000 hours being achieved under continuous operating conditions) when the drift is less than 1 per cent for a 10 per cent variation in the mains supply voltage.

The 'zero-set' control for compensating small variations is brought out in the centre of the front panel.

Another interesting application of the instrument is for measuring the speeds of rotating shafts. As well as the two voltage positions and 'calibrate' on the input switch on the right-hand side of the panel, another position 'P' is available for use with a semiconductor photocell input. A combined photo-electric head incorporating a light source and photocell will be available shortly as an accessory, and by focusing this on to a suitably marked rotating shaft, a direct reading will be obtained of its speed.

The instrument is compact and weighs only 9lb.



TURRET TERMINAL

(Illustrated above)

The Plessey Co. Ltd, Ilford, Essex.

A NEW type of rolled turret terminal, which combines similar efficiency with lower cost when compared with conventional types made from solid materials, is now being manufactured.

A series of serrations has been added to the underside of the flange, thus preventing the terminal from rotating when overheated or knocked.

Designed to meet the requirements of the Joint Services Radio Component Specification, these turret plugs are made from brass to B.S.267, are silver-plated, and can be applied to tag-boards and terminal strips made to manufacturers' individual requirements.

They are available in the following two sizes:

X.T.1 with a base diameter of 0·078in, for use where maximum wire size for centre hole is no greater than 0·028in diameter (22 s.w.g.). X.T.2 with a base diameter of 0·1in, for use where maximum wire size for centre hole is no greater than 0·048in diameter (18 s.w.g.),

POTENTIOMETERS FOR PRINTED CIRCUITS

A. B. Metal Products Ltd, 17 Stratton Street, London, W.1.

THESE controls, of the composition element type are available for mounting parallel to the plane of the printed panel and for mounting at right-angles to the plane of the panel.

Both versions are available with double pole mains switch but no provision is made for incorporating the switch terminals in the printed circuit.

To avoid surface deterioration during transport and storage, all soldering surfaces are hot tin dipped and coated with a protective lacquer during manufacture. thus ensuring perfect soldering during processing.

The potentiometers have a high quality phenolic casing of sturdy construction and the solder tags are of special design, removing all danger of loosening or turning under operating conditions. A metal cover connected to the fixing bush provides automatic earthing of the cover. The resistance element is a special coating permanently bonded to a laminated phenolic base. A dual finger wiper arm of special alloy rides freely on the surface of the resistance element, giving exceptional freedom from noise.

Linear or tapered resistance range from 1kΩ to 5MΩ with tolerance of $\pm$20 per cent of stated value. Closer tolerances are available if specified.

The power rating for linear units is 0·5W and for tapered units 0·25W, at 40°C ambient temperature.

BATCH COUNTING

(Illustrated below)

Sutton Coldfield Electrical Engineers, Reddicap Trading Estate, Sutton Coldfield.

A RECENT application of electronics to counting and batching applications is in equipment dealing with metal caps after the final stage of production. Not only does this eliminate hand work, but the greatest accuracy, and therefore economy, is achieved.

In this instance, a capacitive pick-up head with its probe either touching or close to the caps is coupled to a standard batching unit. On completion of the batch a valve is operated which supplies air to actuate two cylinders. One of these traps the flow of caps while the other removes and replaces the containers. As this process only takes a fraction of a second, production flow is not interrupted.

The equipment counts, in this instance, at an average speed of 400/min, with a 'peak flow' exceeding 750/min. Accuracy is 100 per cent at this speed and tests have shown that considerably higher speeds are possible.

The batching unit can be set to batch in multiples of 100 up to 4 900, and includes a 'state of batch' indicator which automatically resets and a 'number of batches' indicator which can be hand-reset.

The capacitive pick-up head is used in this instance so that accurate signals can be obtained without separating the caps. Photo-electric, inductive or microphonic units can be coupled to the same batching unit.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 28 November. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: Colour Television.
By: G. N. Patchett.

South Midlands Section

Date: 1 November. Time: 7 p.m.
Held at: The Winter Gardens, Malvern.
Lecture: Principles of the Light Amplifier and Allied Devices.
By: T. B. Tomlinson.

North-Western Section

Date: 1 November. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester 1.
Lecture: Electronics applied to Physiology.
By: H. W. Shipton.

Merseyside Section

Date: 7 November. Time: 7 p.m.
Held at: Council Room, Chamber of Commerce, 1 Old Hall Street, Liverpool 3.
Lecture: Industrial Television.
By: J. E. H. Brace and R. Swinden.

Scottish Section

Date: 8 November. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: The Oscilloscope for Engine Testing.
By: R. K. Vinycomb.

West Midlands Section

Date: 14 November. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: The Automatic Factory.
By: J. A. Sargrove.

North-Eastern Section

Date: 14 November. Time: 6 p.m.
Held at: Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Lecture: Some Practical Aspects of Echo Sounding.
By: A. M. Sutton.

Scottish Section

Date: 23 November. Time: 7 p.m.
Held at: Department of Natural Philosophy, University of Edinburgh.
Lecture: Information Theory.
By: L. C. Stenning, P. Jones and P. Holroyd.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Ordinary Meeting

Date: 8 November.
Lecture: Visit of the British Electricity Supply Delegation to the Soviet Union.
By: J. Eccles.
Date: 29 November.
Lecture: Differences of Opinion about Dimensions.
By: R. O. Kapp.

Measurement and Control Section

Date: 6 November.
Lectures: Electric Strength of High Compressed Gases.
By: E. H. Holt (to be read by H. Tropper).
and: Insulation Properties of Compressed Electro-Negative Gases.
By: P. R. Howard.
Date: 20 November.
Discussion: Data Processing Equipment for Experimental Work; a review of Techniques and Methods.
Opened by: M. V. Wilkes.

Informal Meeting

Date: 12 November.
Discussion: Power Factor in Industrial Installations.
Opened by: C. F. Freeman.

Radio and Telecommunication Section

Date: 14 November.
Lectures: Frequency Diversity in the Reception of Selectivity Fading Binary Frequency-Modulated Signals with special reference to Long-Distance Radio-Telegraphy.
By: J. W. Allnatt. E. D. J. Jones and H. B. Law.
Investigation of the Spectra of Binary Frequency-Modulated Signals with Various Build-up Waveforms.
By: J. W. Allnatt and E. D. J. Jones.
An Improved Fading Machine.
By: H. B. Law, F. J. Lee, F. A. W. Levett and R. C. Looser.
The Detectability of Fading Radiotelegraph Signals in Noise.
By: H. B. Law.
The Signal/Noise Performance Rating of Receivers for Long-Distance Synchronous Radiotelegraph Systems using Frequency Modulation.
By: H. B. Law.

Utilization Section

Date: 15 November.
Lectures: A Self-Oscillation Induction Motor for Shuttle Propulsion.
By: E. R. Laithwaite and P. J. Lawrenson.
Brushless Variable-Speed Induction Motors.
By: F. C. Williams, E. R. Laithwaite and L. S. Piggott.

Supply Section

Date: 21 November.
Discussion: Aluminium Conductor Cables.
Opened by: G. F. Peirson.
Date: 28 November.
Lecture: The Automatic Solution of Power-System Swing-Curve Equations.
By: C. Adamson, L. Barnes and B. D. Nellist.
Lecture: Electronic Analogue Computer Study of Synchronous Machine Transient Stability.
By: A. S. Aldred and P. A. Doyle.
Lecture: Dynamic Operation of an A.C. Network.
By: S. Kaneff. (To be read by I. McDonald).

East Midland Centre

Date: 6 November. Time: 6.30 p.m.
Held at: The E.M.E.B. Service Centre, Derby.
Lecture: Potentialities of Railway Electrification at the Standard Frequency.
Date: 22 November. Time: 6.30 p.m.
Held at: Loughborough College.
Informal evening on Ultrasonics in Industry.
Talk by C. F. Brocklesby.

Cambridge Radio and Telecommunication Group

Date: 13 November. Time: 8 p.m.
Held at: The Cavendish Laboratory, Free School Lane, Cambridge.
Section Chairman's Address.
By: R. C. G. Williams.

East Anglian Sub-Centre

Date: 6 November. Time: 7.30 p.m.
Held at: The Assembly House, Norwich.
Lecture: Some Impressions of Electrical Progress in the U.S.S.R.
By: C. T. Melling.
Date: 12 November. Time: 6.30 p.m.
Held at: Crown and Anchor Hotel, Ipswich.
Lecture: Atomic Power Stations.
By: D. F. Welch.

North-Eastern Centre

Date: 5 November. Time: 6.15 p.m.
Held at: The Lit. and Phil. Society, Newcastle-upon-Tyne
Lecture: Nuclear Reactors for Power Generation.
By: B. L. Goodlet.
(Joint Meeting with the Northern Counties Association of The Institution of Civil Engineers and the North Eastern Branch of The Institution of Mechanical Engineers).
Date: 12 November. Time: 6.15 p.m.
Held at: The Neville Hall, Newcastle-upon-Tyne.
Lecture: The Control and Instrumentation of a Nuclear Reactor.
By: A. B. Gillespie.

North Midland Centre

Date: 20 November. Time: 6.30 p.m.
Held at: Offices of the Central Electricity Authority, Yorkshire Division, 1 Whitehall Road, Leeds.
Discussion: The Influence of Maintenance Requirements on the Design of Industrial Electrical Equipment.
Opened by: H. C. Fox.
Date: 22 November. Time: 6.15 p.m.
Held at Catterick Camp.
Lecture: Communication by Tropospheric and Ionospheric Scatter.
By: J. A. Saxton and W. J. Bray.

Sheffield Sub-Centre

Date: 21 November. Time: 6.30 p.m.
Held at: The Grand Hotel, Sheffield.
Lecture: Germanium and Silicon Power Rectifiers.
By: T. H. Kinman, G. A. Carrick, R. G. Hibberd and A. J. Blundell.

North-Western Radio and Telecommunication Group

Date: 7 November. Time: 6.45 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Informal evening on Electronics and Automation.
Talk on: Some Industrial Applications.
By: H. A. Thomas.

South-East Scotland Sub-Centre

Date: 6 November. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: TRIDAC—A Large Analogue Computing Machine.
By: F. R. J. Spearman, J. J. Gait, A. V. Hemingway and R. W. Hynes.

South-West Scotland Sub-Centre

Date: 7 November. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: TRIDAC—A Large Analogue Computing Machine.
By: F. R. J. Spearman, J. J. Gait, A. V. Hemingway and R. W. Hynes.

South Midland Radio and Telecommunication Group

Date: 26 November. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Informal evening on Electronics and Automation.
Talk on: Some Industrial Applications.
By: H. A. Thomas.

North Staffordshire Sub-Centre

Date: 19 November. Time: 6.30 p.m.
Held at: Duncan Hall, Stone.
Lecture: The Generation and Synthesis of Music by Electrical Means.
By: A. Douglas.

Southern Centre

Date: 26 November. Time: 7.30 p.m.
Held at: The R.A.E. Technical College, Farnborough.
Lecture: An Approximate Method for Finding the "Best Linear Servomechanism."
By: H. H. Rosenbrock.

Western Centre

Date: 12 November. Time: 6 p.m.
Held at: The South Wales Institute of Engineers, Park Place, Cardiff.
Lecture: Germanium and Silicon Power Rectifiers.
By: T. H. Kinman, G. A. Carrick, R. G. Hibberd and A. J. Blundell.
Date: 29 November. Time: 6.30 p.m.
Held at: The Colston Hall, Bristol.
Faraday Lecture: Nuclear Energy in the Service of Man.
By: T. E. Allibone.

Western Utilization Group

Date: 26 November. Time: 6 p.m.
Held at: Electricity House, Colston Avenue, Bristol.
Lecture: Future Trends in Aircraft Electrical Systems.
By: D. C. Flack.

West Wales (Swansea) Sub-Centre

Date: 8 November. Time: 6 p.m.
Held at: South Wales Electricity Board Showrooms, The Kingsway, Swansea.
Lecture: The Potentialities of Railway Electrification at the Standard Frequency.
By: E. L. E. Wheatcroft and H. H. C. Barton.
Date: 27 November. Time: 6.30 p.m.
Held at: Brangwyn Hall, Swansea.
Faraday Lecture: Nuclear Energy in The Service of Man.
By: T. E. Allibone.

THE TELEVISION SOCIETY

Date: 9 November. Time: 7 p.m.
Held at: 164 Shaftesbury Avenue, London, W.C.2.
Lecture: New Techniques in Receiver Construction (Printed Circuits).
By: W. I. Flack.
Date: 22 November. (Time and place as above).
Lecture: Alternatives to the N.T.S.C. Colour System.
By: E. L. C. White.

PUBLICATIONS RECEIVED

HOW TO SELECT AND USE YOUR TAPE RECORDER by David Mark is a recent addition to the publications of John F. Rider Publisher Inc. and includes a buyers' guide section. John F. Rider Publisher Inc, 480 Canal Street, New York 13, N.Y. Price $2.95.

'F' TYPE CARTER HYDRAULIC INFINITELY VARIABLE SPEED GEARS is a recent publication of Carter Gears Ltd, Thornbury Road, Bradford 3. Sizes F10, F12 and F14 Carter gears are covered in this publication, together with details of Carter Gears fitted with flange mounted spur reduction gears for low speed conditions. Examples on gear selection and instructions for the installation and maintenance of 'F' Type Carter Gears have also been included.