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Commentary

SCIENTIFIC 'firsts' are frequently surrounded by doubt and rival claims since, when any problem claims attention there will often be many people in divers countries working upon it and it is very probable that more than one of them will arrive at the same conclusion at about the same time, although, mayhap, by different routes. While this is true, to some extent, concerning the discovery of the electron, it is reasonably certain that Joseph John Thomson was the first to give conclusive proof of its existence, a knowledge, the possession of which has brought about such significant changes in our way of life and in the whole of scientific progress in the last sixty years. The discovery of the electron is brought to mind at this particular time since it was on 18 December 1856, exactly one hundred years ago, that J. J. Thomson was born in Cheetham, Manchester.

Thomson was without doubt, one of the great men of his time for, in addition to the vast amount of research and discovery he carried out personally, he was also a foremost teacher and leader of men. As an indication of this it may be recalled that, in addition to winning the Nobel Prize himself, no less than eight of his pupils received this award, while twenty-seven were elected Fellows of the Royal Society and some seventy-nine professorships were held by them. Thomson himself received the honours usual to a man of his achievements. He was awarded the Nobel Prize for physics in 1906 for 'the great merits of his theoretical and experimental researches on the discharge of electricity through gases'. This work included, of course, his expositions on the electron or 'corpuscle' as he called it. He was knighted in 1908 and was President of the Royal Society from 1915 to 1920. In 1918, after being Cavendish Professor of Experimental Physics at Cambridge for thirty-four years, he became Master of Trinity College.

Thomson's own contribution to science was large. In all, he published some 230 papers and eleven books, and some of these were epochs. The positive discovery of the electron can be dated from his 1897 lecture to the Royal Institution while his Dover Address of 1899 can be regarded as the manifesto for the twentieth century research in experimental atomic physics. He also forecast thermionic emission from heated metals and, in 1912, accomplished the first separation of isotopes, the existence of which had been deduced by Soddy a year previously.

Despite these great personal achievements it was, perhaps, as the leader and source of inspiration of the Cavendish Laboratory that he did his greatest work for, as J. A. Crowther has remarked, during his direction of the Cavendish Laboratory research changed from a 'personal idiosyncrasy' to the 'normal completion of a university training'. Thomson was appointed Cavendish professor in December 1884, at the age of twenty-eight, in succession to Lord Rayleigh who had accepted the professorship for five years following the premature death of Clerk Maxwell in 1879. Clerk Maxwell rarely had more than two or three students at his lectures and only a handful of experimental researchers, yet by 1890 there were twenty research students in the Laboratory. These, Thomson visited every morning to discuss their problems and suggest solutions to their difficulties; a practice which he maintained throughout his time at the Laboratory. Thomson held the Cavendish chair for thirty-four years until he was succeeded by one of his foremost students, Lord Rutherford, in 1919. During this time the output of the Laboratory was enormous; in the years 1896 to 1900 alone one hundred and four original research papers being published. It is perhaps indicative of Thomson's methods that during the whole term of his directorate the annual expenditure on special research equipment never exceeded £550, although in some years there were forty research workers in the Laboratory. The many fine investigations and discoveries made during this period were carried out with some of the simplest of equipment and, although admittedly science is becoming increasingly complicated, this point may well be borne in mind by the research worker of today. It was also a dictum of Thomson that students should, within reason, attempt to arrive at their own methods of solving problems and not be biased by the approach of previous workers.

Sir J. J. Thomson died on 30 August 1940 and his ashes were interred near those of Newton, Kelvin and Rutherford in Westminster Abbey.

As a footnote to this short account of the scientist whose birth centenary is now being celebrated it is of interest to recall that although Thomson came from a home of modest means, he left a fortune of £82 000 which he had accumulated by the judicious investment of his savings for, apart from his Nobel Prize, he had never received any large sums of money during his lifetime. He was, indeed, in many ways a fine example for the young man of today.

The Production and Testing of Potted Circuits

By T. C. B. Talbot*

The development of a technique for the production of small batches of a hundred or less potted circuits is outlined, and is illustrated by describing its application to the production of a typical unit. Some limitations of the technique are pointed out and improvements for large quantity production are suggested.

Finally, the development and application of special test gear for use with potted circuits are described.

WHILE several articles have appeared in technical journals describing the properties of casting resins and the general techniques to be used in potting circuits[1,2,3], little has been published concerning the specific techniques used in production.

It is the aim of this article to trace in detail the development of a technique used in a small production run and by way of illustration to describe its application to the production of a typical electronic unit. It must be borne in mind that the technique described has been developed specifically for small batch production, where the cost of expensive jigs and tools is not warranted; as a result, production costs are high compared with conventional assembly methods but these would be considerably reduced by application of the further development and more advanced tooling which would be necessary for large scale quantity production. In the work to be described, only the injection moulding tools for the production of the polythene moulds in which the circuits are potted are highly developed. All remaining tools are comparatively simple and cheap.

The assumed requirements for the electronic equipment were reliability, small size, and the ability to withstand the vibration and acceleration forces experienced in modern aircraft, and to operate between wide temperature limits and in humid atmospheric conditions. It has been shown that potted circuits fulfil all these requirements but at the expense of weight and the lack of economy inherent in the 'throw away' nature of the potted circuit[1].

Initial experiments with several commercially available casting resins led to the adoption of Bakelite Polyester Resin SR17449. It was found that an inert filler was necessary to prevent cracking of the potted blocks at low temperature and the following mix gives satisfactory results:—

Bakelite Resin SR17449	100 parts by weight
Catalyst Q17447	1·66 parts by weight
Accelerator Q17448	2·25 parts by weight
Filler (200-mesh mica flour)	25 parts by weight

Using this mix, the setting time is about five hours at an ambient temperature of 17°C and the exothermal temperature rise in a small block (about $4in^3$) during the polymerization reaction is of the order of 25°C.

In a complete equipment to which the technique was applied, the majority of circuits are potted. A power unit does not lend itself readily to potting owing to the bulk of its components, and this unit is built in a conventional form, using sealed components. The remainder of the equipment is split into sub-units, which, in turn, are split into potted sub-assemblies.

Sub-miniature 'wire-in' valves are used wherever possible, but these are not cast in with their associated components for three reasons; firstly because the valve is probably the least reliable component in a circuit, secondly, because its cost is high compared with its associated components, and finally, because it is normally the most significant source of heat. Transformers, with the exception of those sufficiently small in size, are also excluded from cast blocks on account of their bulk and comparative cost. It is also necessary to omit trimmer capacitors, potentiometers, relays and a small number of critical components which may be subject to replacement.

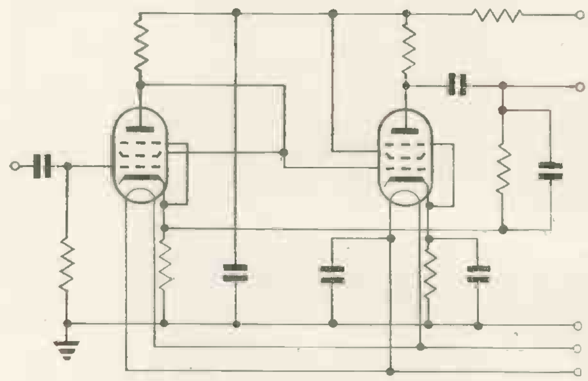

Fig. 1. Circuit of feedback pair

It is proposed to use for illustration of the technique the production of a feedback pair, the circuit of which is shown in Fig. 1.

Assembly of the Block Prior to Casting

Components are mounted between two synthetic resin bonded paper (s.r.b.p.) boards, suitably punched to accept the component wires. Sheets of s.r.b.p. are punched with a matrix of holes, symmetrical in both dimensions, spacing of hole centres being 3/16in. This entails that component wires are centred on standard dimensions in all types of block, allowing the use of certain common tools. Suitably shaped boards are then stamped from the punched stock, a pair of these being used for each block. In one of the pair (the terminal board) tinned eyelets are inserted in all holes which are to receive components or spacing wires, other holes being left blank. In the remaining board, eyelets are inserted only in those holes which accept spacing wires. The spacing wires maintain the distance between the boards after assembly and before casting. Four 18 s.w.g. tinned copper wires are normally used, but the number may vary with the size of the block.

The components and spacing wires are now assembled between the boards by hand, and the whole assembly is dropped into a spacing jig, consisting of an aluminium channel section in which are milled two slots to accept the boards, which are thus held at the correct distance until the spacing wires are soldered in.

Components are adjusted by hand to lie centrally between the boards and all wires are then soldered to their respective eyelets. The assembly may now be removed from the

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spacing jig and is sufficiently robust to withstand normal handling without damage or distortion, although components are retained only at one end.

Inter-connexions between components are now made by soldering 30 s.w.g. tinned copper wire between the appropriate component wires as close to the outer surfaces of the boards as possible. This wiring requires careful dressing after insertion to ensure freedom from short-circuits. The spacing wires are normally used to provide connexions through the block where required.

The next operation is the removal of the excess of component wires. On the terminal face, this is done by inserting the block into a pin-cropping guillotine (Fig. 2). The plate of the guillotine is drilled with the same matrix as the punched s.r.b.p. sheet, and the holes are tapered down from the entry face to facilitate insertion of the block. The knife of the guillotine operates against the rear face of the plate, and cuts the terminal wires to a standard length of 7/16in. Owing to the standard pattern of holes on the guillotine plate, this tool is common to all blocks.

Fig. 2. Pin-cropping guillotine

On the remote face of the assembly, all wires, with the exception of two of the spacing wires, are manually cut as close to the board as possible. The two remaining spacing wires are cut to a length of 3/16in from the face of the board and are later used to locate the assembly in the mould when casting.

Finally, the assembly is inserted into a pin-straightening jig which sets the terminal wires normal to the terminal face. The pin-straightening jig is again drilled to the standard matrix, but in this case the holes only just clear 18 s.w.g. wire and the entry taper is short, whereas on the guillotine the holes are slightly larger and the entry taper is longer.

Mould Production and Casting

The choice of material for the construction of moulds is subject to several requirements. The moulds must be produced rapidly and cheaply and they must be robust and free from damage during handling. For one or other of these reasons metal moulds are not particularly suitable and it was decided to use injection mouldings in polythene for the purpose. These fulfil the above requirements and, in addition, as polythene is incompatible with polyester resins, a mould release agent is not required.

A common moulding shell is used to give a common outer profile for most moulds. Extension pieces can be fitted to the shell to produce larger moulds. A variety of shell inserts are provided to give the individual internal contours of the mould. It should be pointed out that different types of cast block have individual shapes. Most are tapered to suit the draft of a sand-cast metal container, but some types are of rectangular shape and are housed in sheet metal containers in the main equipment.

Both the shell and the inserts of the moulding tool are machined from brass. Part numbers of the respective blocks are engraved on the face of the appropriate insert normal to the direction of withdrawal. These part numbers are transferred in relief to the polythene mould and thence to the cast block.

Once the initial tooling is complete, moulds may be produced in quantity with great rapidity. They are extremely shock proof and durable, and will withstand rough handling through a long life. The moulding shell and insert and a completed mould for a feedback pair are shown in Fig. 3.

In the bottom of the well of the mould are two recesses which accept the two 3/16in locating wires on the bottom face of the assembly to be cast. The assembly is located at the top by a 16 s.w.g. aluminium top plate, which picks up two of the terminal wires on the assembly and which carries two dowels which engage in holes on the top face of the mould. This accurately locates the assembly in the mould. A portion of the aluminium top plate directly above the assembly is cut away to provide an aperture through which to pour the liquid resin. The plate is coated with carnauba wax to prevent adhesion of the resin and is held in place by two battery clips.

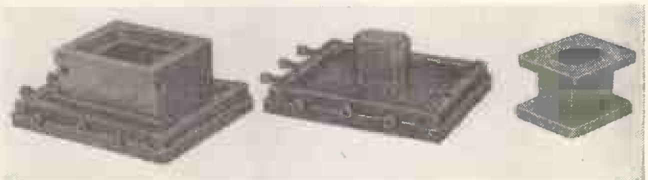

Fig. 3. Moulding shell, insert, and completed mould for feedback pair

During an early attempt to cast a block of components, it was found that the phenolic based varnish, which is normally used to seal the cut edges of s.r.b.p. board against the ingress of moisture, inhibited the polymerization of the resin, and blocks using boards treated in this way failed to set. Raw edges of board are not now sealed. Tests on samples of unsealed boards which had been open to the atmosphere for some time showed a deterioration in leakage resistance between eyelets by a factor greater than 10. Resistance between eyelets in some cases measured as low as 15MΩ. It is therefore necessary to bake the assembly in an oven at 60°C for one hour immediately prior to casting. This is effective in restoring the insulation properties of the boards.

Due to the comparatively short pot life of the mix, it is desirable to fill a number of moulds from a single mix. After preparation of the mix, it is placed for a few moments in a vacuum chamber, and the pressure reduced, to remove bubbles of air. The mix is then poured into a funnel shaped polythene hopper, and the moulds are passed under this, one by one, and filled to the upper surface of the top plate (Fig. 4). Care is taken to ensure that air bubbles are not trapped in the assembly. In the slow setting mix used, air bubbles have time to settle out without agitation of the mould, but further degassing in a vacuum chamber would be required for a quicker setting or more viscous mix. Even in the mix used, however, air bubbles do tend to be trapped under the aluminium top plate and a right angled piece of wire is used to rake the surface of the resin under the top plate to remove these.

During the polymerization reaction, the resin shrinks slightly. This is apparent as a depression in the resin surface in the pouring aperture of the top plate, giving a meniscus effect. Shrinkage, however, is not sufficient to drop the surface of the resin below the top plate and, hence, a flat ledge appears around the top surface of the block where it is in contact with the top plate. This ledge is used as a bearing surface on which to clamp the block into its case.

When polymerization is complete and the resin has set, the top plate is removed and the block is withdrawn from the mould. It is then placed in an oven to harden off for one hour at 60°C. The final product is a hard, robust, opaque block with $\frac{1}{4}$in terminal wires appearing on the terminal face (Fig. 5). It is resistant to corrosion, mould growth, and insect attack, and may be stored indefinitely without special packaging, since it is completely impervious to moisture absorption and surface condensation may be quickly driven off before installation in a unit.

Fig. 4. Filling apparatus

The Sub-Unit Case and Final Assembly

Since the terminal wires on the cast block are, of necessity, close, the insulating path between adjacent terminals is very short and the probability of their becoming bridged by moisture in a humid atmosphere is high. In order to prevent this, the use of a moisture proof container is desirable. Furthermore, cases arise where electrical screening between sub-units is necessary. To satisfy both these requirements and to assist in cooling, it is convenient to use a sealed metal container for each sub-unit. The quantity produced did not warrant the tool cost of a die-cast container, so a machined sand-casting in aluminium alloy is used. In order to save space, the walls of the container are thin and tend to be porous in places. The casting is therefore impregnated with a thermosetting resin before use to ensure adequate sealing.

To improve the cooling of the unit, the valves are housed in thin walled drawn aluminium tubes which are mounted in a flange on the casting and again sealed-in with a thermosetting resin (Fig. 6). This arrangement provides only a narrow neck of metal and consequently a high thermal resistance between the valves and the body of the casting. Cooling is enhanced by a draught of air forced along the length of the amplifiers which are so mounted that the greatest volume of air flows in the channel which is formed between adjacent units, and into which the valve tubes protrude. Bowed leaf springs are assembled into the valve tubes to prevent damage to the valves by vibration.

Signals and supplies are fed to and from the sub-unit by means of multiple glass-to-metal seals cemented into the case with thermosetting resin, the outer terminals of which are wired to miniature Jones type plugs mounted externally on the case.

All cementing operations are carried out simultaneously. The parts to be sealed in are lightly retained to the case, small self-tapping screws being used where necessary. The joints are then dusted with the thermosetting resin powder and the assembly is brought up to curing temperature on a hotplate, when the powder fuses and flows into the joints, forming, when cured, a joint which is satisfactory both as regards hermetic sealing and mechanical strength. It has been found that traces of grease inhibit the curing of the resin, so the casting and parts to be sealed-in are thoroughly degreased before assembly.

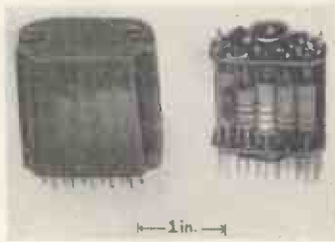

Fig. 5. Feedback pair before and after casting

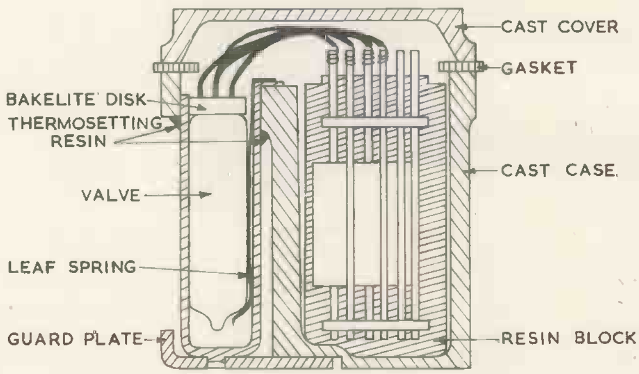

Fig. 6. Cross-section of sub-unit

Sealing of the cover of the sub-unit to the main body of the case is provided by a flat rubber gasket compressed between the two. Considerable pressure is required to obtain a satisfactory seal and this is provided by a number of socket head cap screws around the unit which are captive in the cover. Plugs screwed down on rubber washers are provided on the cover to give access to the preset controls inside the unit and also to provide a thread into which an adaptor may be screwed for pressure testing. In operation, the amplifier is sealed at atmospheric pressure and is designed to withstand the pressure differentials experienced at high altitude.

In the final assembly of a sub-unit, the cast blocks are dropped into place in the container and held in place with small clamps, and the valves are inserted into their tubes. In order to prevent the valve wires being bent close to the envelope, with the resultant risk of cracking the seal, small bakelite disks, 3/32in thick, containing holes on the same pitch circle as the valve terminations are threaded on to the wires and seated down on the base of the valve. The valve wires are now sleeved, and terminated by soldering them to the appropriate pins on the resin blocks. The remainder of the wiring is then carried out and any external components added to complete the sub-unit. A typical sub-unit containing four cast blocks and ten valves measures 11in by 2in by 2 7/16in and its weight is $2\frac{3}{4}$lb.

Future Development

The process outlined above has been in operation for some months and has been successfully used in a small production run. The process, however, is slow and costly and improvements are required before embarking on large quantity production.

About 70 per cent of the assembly time of the blocks prior to casting is taken up with the soldering of component wires to their eyelets and the adding of inter-component wiring on the boards. It is clear that the use of printed wiring on the boards would give a material improvement in assembly time and dip soldering follows automatically as an additional improvement. Some form of jigging of the boards and components will be necessary to maintain the shape of the assembly before and during dip soldering; suitable cranking or upsetting of the component wires may be sufficient.

Use of one of the modern 'foam' resins for casting the blocks would serve to reduce the weight of the equipment, but there appears some doubt, as yet, about the 'wetting' properties of these resins.

Die-cast cases would undoubtedly be superior to the present sand-castings, but the tools would be expensive and would only be economical for a very large quantity. In a die-cast case, the valve tubes would probably be cast in, thereby considerably reducing assembly time.

Production Testing

It will be apparent that the nature of the potted circuit demands a different approach to production testing from the conventional circuit assembly. In a conventional unit, faulty components can be changed at a very late stage in the assembly and it is common practice, in fact, to test nothing, apart perhaps from transformers, until the final assembly has been completed. With potted circuits, however, testing must be carried out stage by stage in the interest of economy. The block must be tested after assembly of components but before casting, so that faulty components may be replaced at this stage. In fact, only the outer components of the block can be changed at this stage, and if a component in the centre of the assembly is faulty it is generally more economical to reject the assembly rather than attempt a repair. Similarly, in view of the fragility of sub-miniature valves, it is desirable to test the block after casting and before the addition of the valves.

A stage-by-stage test procedure has therefore been evolved. It is obviously undesirable to solder and unsolder valves to the terminals of the block at each stage, both in the interest of time saving and of valve preservation. A 'plug and socket' technique has been devised, using the cast block as its own plug, the terminal wires being the plug pins. It is clear that a multiple socket, similar in pattern to the punched s.r.b.p. from which the terminal boards are stamped, will accept any type of block. A tool was therefore made to mould such a socket in the polyester resin, the resultant socket being a flat 'biscuit' 3in square by $\frac{1}{4}$in thick, and containing 200 recesses on the standard matrix. These recesses are so shaped that small sprung socket contacts of the type used in miniature valve-holders may be inserted and locked in. Socket contacts are inserted into the recesses appropriate to the terminal wires of the particular block and the biscuit, thus assembled forms an effective multi-pole socket for the block. The socket is mounted on a small box which contains the valves appropriate to the block under test and any other necessary components, all of which are wired to the appropriate socket contacts. A variety of test jigs of this form, one for each type of cast block, has been produced, the differences being only in the disposition of the socket contacts on the biscuit and the contents and internal wiring of the jig.

Each jig is fitted with a 33-pole plug on one end face, which mates with a socket on a versatile test set. By using standard pins for such functions as power supplies on all jigs, it is possible to provide enough facilities on one 33-pole connector to test all blocks.

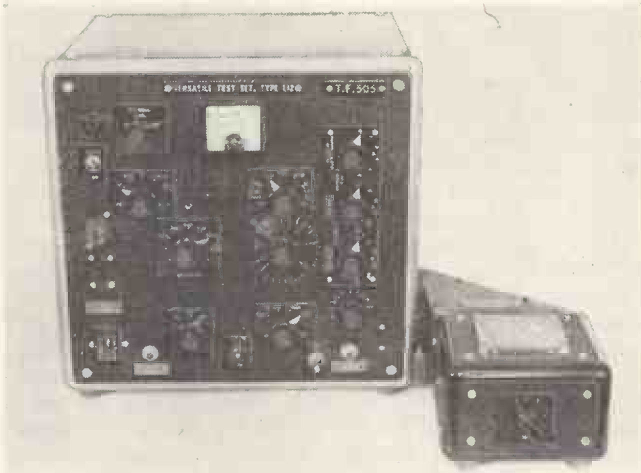

Fig. 7. Test rig for feedback pair

The test set contains power supplies, oscillators, attenuators and a metering circuit to enable frequency response measurements to be made over a wide range. The meter can also be switched to measure other circuits currents and voltages as desired. In the case of the feedback pair, the test set, together with the appropriate test jig, is sufficient to test the block, a measurement of h.t. current consumption, mid-band gain and frequency response being adequate. Where the block contains pulse circuits, trigger circuits, blocking oscillators and the like, further test gear is required. It has been found convenient for the batch testing required in production to develop a special pulse generator and oscilloscope, in addition to the test set, with facilities peculiar to the cast blocks under test. These are used in conjunction with the test set for testing more complicated blocks, though in certain of the tests, the test set acts merely as a junction box to couple the test jig with the pulse generator and oscilloscope. The test rig is shown in Fig. 7 as laid out for the testing of a feedback pair. The test jig with socket on top is seen on the right-hand side of the test set.

It has proved convenient to further extend the use of the test rig to the testing of complete sub-units and to the testing of video transformers. For the testing of complete sub-units, a test jig is used where the socket is replaced by connectors on flying leads which mate with the plugs and sockets on the sub-unit under test. In the case of video transformers, a comparison bridge has been built into a test jig box, the transformer under test being compared with a standard. In this case the test set acts as source and detector for the bridge.

It can be seen from the foregoing that, by the use of a highly developed and fairly complex test rig, the testing of potted circuits is greatly facilitated and may be rapidly carried out, at the same time ensuring a high degree of reliability in the finished product.

Acknowledgments

The author is indebted to the Directors of Messrs. Bush Radio Ltd for permission to publish this article and to Messrs. W. M. Lloyd, C. W. Ward and R. E. Gardner of Bush Radio who initiated much of the development leading to the techniques described.

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A 'True Motion' Radar System

A new system of marine radar presentation known as True Motion Radar T.M.46 has recently been introduced by Decca Radar Ltd and will be available early in 1957.

This system, it is claimed, is the first of its kind to show directly on the face of the p.p.i. the true motion of all objects within radar range as opposed to the conventional relative movement.

The display of true motion is obtained by feeding compass and speed information into a small resolver in a Trackmaster unit which is either fixed to the display unit or sited immediately adjacent to it. Speed and direction of 'own' ship are here converted into movements East-West and North-South. These movements control the amount of current in the off-centring coils of the display itself and affect the position of the electrical centre, that is, the position of 'own' ship about which the trace rotates. On a compass stabilized display when the electrical centre moves in harmony with the course and speed of 'own ship', the effect on the p.p.i. is to present true motion, that is a 'bird's eye' view of the situation, since the component of 'own' ship's course and speed normally present in the apparent motion shown on a conventional display has been entirely removed.

The Decca 'True Motion Radar' display and the Trackmaster unit.

The amount of movement imparted to the display depends on the radar scale in use and is automatically adjusted when range scale is altered.

The speed of 'own' ship is fed into the Trackmaster unit either by hand or automatically from the ship's log. Speed must be known to the same accuracy as is required when constructing a proper motion plot by hand, so that the accuracy requirement is not high.

The estimation of the course and aspect of other ships is made by observing the afterglow trails behind those ships' echoes. A mechanical bearing cursor, with parallel lines, can be used to read these courses when a high degree of accuracy is required.

Special facilities are provided to read both range and bearing regardless of the position of 'own' ship on the face of the p.p.i. Range rings and a variable range marker appear as concentric circles around 'own' ship and have normal radar accuracy, being independent of any movements imparted to the display.

An electronic bearing marker is provided to measure bearings of other vessels irrespective of the position of 'own' ship and to measure these bearings with a greater accuracy than is possible by conventional mechanical methods. The marker appears as a radial electronic line centred on the position of 'own' ship and bearings are read from a calibrated dial.

Since 'own' ship is continually moving over the face of the tube in the direction of the ship's course, the range of warning ahead is steadily being reduced and would eventually be nil unless the position of 'own' ship were reset. Resetting is effected simply and speedily by shift controls and 'own' ship can be positioned anywhere on the face of the tube. After resetting, the true motion picture is again developed almost instantaneously.

Speed of 'own' ship is fed into the Trackmaster unit by setting the estimated speed on a calibrated dial. Alternatively, the output from certain types of transmitting logs may be used if preferred.

A small error in speed will not materially affect the accuracy of range and bearing measurement.

The display may be operated to show either a conventional 'ship's head up' picture, a conventional compass stabilized picture, or the true motion picture.

It is believed that the system described can be a major step forward in marine radar development, in giving to a navigating officer at a glance other vessels' true course and speed without the need for any calculations with their possibility or error, and therefore can make its contribution to greater safety at sea, especially in view of the increased traffic density in the world's sea lanes in recent years and the speed and size of modern ships.

Cathode-Follower Type Power Supplies

By B. J. Perry*, B.Sc., Ph.D., A.R.C.S.

The use of triode valves as rectifiers in variable voltage power supplies is discussed and compared to the simple series stabilizer. The performance of a combination of these circuits is given and made the basis of a practical circuit which is of particular use in electrophoresis applications. A form of low frequency instability sometimes associated with the circuit is noted.

THE use of grid-controlled valves for rectification is well known[1], though it is less usual to find hard valves so employed. When a triode valve is used as a rectifying element, with its impedance controlled by its grid voltage (Fig. 1), the arrangement is usually referred to as a cathode-follower type rectifier[2]. This particular circuit is similar to

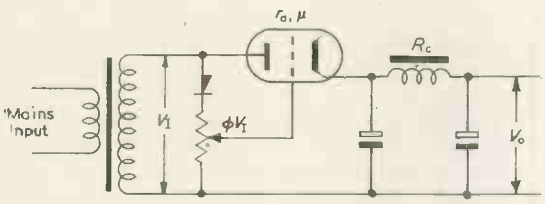


Fig. 1. Simple cathode-follower power supply.

the simple series stabilizer (Fig. 2). When these two circuits are combined a power supply of medium stability and considerable flexibility is obtained.

The Cathode-Follower Type Rectifier

The main advantage of a cathode-follower type power supply lies in the wide range of output voltage which is available[2], and the ease with which it can be controlled. The output impedance varies little with output voltage, and for the circuit of Fig. 1 is given by the expression:

$$Z_o = \frac{r_a}{1+\mu} + R_T \cdot \frac{1+\phi\mu}{1+\mu} + R_c \quad \ldots\ldots\ldots (1)$$

where r_a and μ are the usual valve constants, R_T the transformer secondary winding resistance, R_c the smoothing choke resistance and ϕ the fraction of V_I used to control the valve. (In this case ϕ is approximately equal to V_o/V_I).

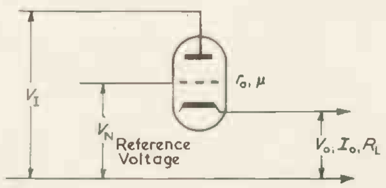


Fig. 2. Simple series stabilizer

At maximum output voltage ($\phi = 1$), equation (1) is very similar to that for the output impedance of an ordinary fixed voltage power pack, with the impedance of the rectifier exchanged for an impedance equal to approximately $1/g_m$ of the triode. In practice, the value of $1/g_m$ may well be less than 100Ω.

As ϕ tends towards zero the second term in equation (1) is reduced from R_T towards $R_T/(1+\mu)$ and the effective Z_o is thus reduced. However, of the three terms in the equation, the largest is usually the resistance of the smoothing choke R_c. This can be reduced in the same manner that R_T is reduced, by placing the choke on the input side of the control voltage potentiometer in the negative line (Fig. 3). The output impedance then becomes:

$$Z_o = \frac{r_a}{1+\mu} + (R_T + R_c)\frac{1+\phi\mu}{1+\mu} \quad \ldots\ldots\ldots (2)$$

As Z_o is in part dependent on the expression $1/g_m$ the regulation of the power pack will be poorer at small values of anode current when the value of g_m is also small.

The Series Type Stabilizer

Expressions for the stability of the series type valve stabilizer are well known[3,4], the simple incremental input-output ratio for the circuit of Fig. 2 being:

$$S_o = dV_I/dV_o = (r_a/R_L) + 1 + \mu \quad \ldots\ldots\ldots (3)$$

For normal stabilizer loads this approximates to μ. The percentage stabilization ratio for this circuit is:

$$\frac{dV_I/V_I}{dV_o/V_o} = 1 + \mu \cdot (V_N/V_I) \quad \ldots\ldots\ldots (4)$$

In this equation (as for all stabilization circuits[4]), when the

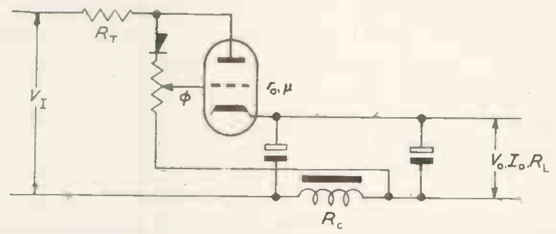


Fig. 3. Cathode-follower power supply with improved output impedance

reference voltage approaches zero the stabilization ratio approaches unity.

The principle of the series stabilizer can be incorporated into the cathode-follower power supply circuit by using a reference voltage to control the grid of the rectifying triode (Fig. 4). The output impedance is then further improved, being given by the expression:

$$Z_o = \frac{r_a}{1+\mu} + \frac{R_T + R_c}{1+\mu}$$

If a neon tube is used for the reference voltage a further term has to be added due to its internal impedance. Z_o thus becomes:

$$Z_o = \frac{r_a}{1+\mu} + \frac{R_T + R_c}{1+\mu} + (R_N/R)(R_T + R_c)\phi \quad \ldots. (5)$$

Where R_N is the internal impedance of the neon tube and R the neon load resistor which is assumed large in comparison with R_N, R_T or R_c.

The percentage stabilization ratio for the complete circuit of Fig. 4 is then:

$$\frac{dV_I/V_I}{dV_o/V_o} = 1 + (V_N/V_I)\,\frac{R_c + R_T + \phi\mu(R + R_c + R_T)}{R + R_N + \phi \cdot \mu \cdot R_N} \quad (6)$$

which reduces to equation (4) if R is greater than R_c, R_T, R_N and $\phi\mu R_N$.

* *Physics Department, Westminster Hospital.*

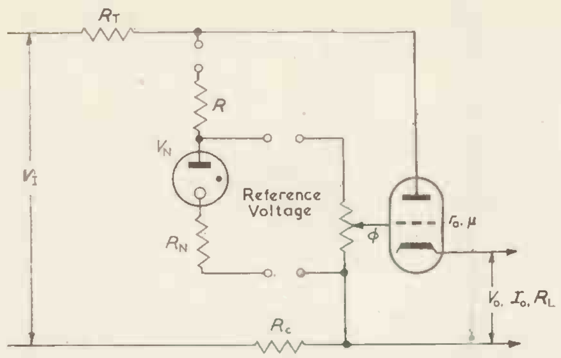


Fig. 4. Stabilized cathode-follower power supply, d.c. equivalent circuit

Typical values for the components of this circuit:

R_T 300Ω	V_N 390V
R 7 500Ω	g_m 12×10^{-3} mA/V
R_N 850Ω	r_a 1 250Ω
R_c 300Ω	μ 15

Typical values for the constants of equations (5) and (6) are given beneath Fig. 4. These assume a maximum output voltage of 400V needing a reference voltage consisting of a chain of three neon tubes. Z_o is then 180Ω for $\phi = 1{\cdot}0$ and 120Ω for $\phi = 0{\cdot}1$. The percentage stabilization ratio for $\phi = 1{\cdot}0$ is 5·9 times and for $\phi = 0{\cdot}1$ is 2·2 times.

A Practical Circuit

A power supply incorporating the features of both the series stabilizer and the cathode-follower type power pack has been built for use in the Pathology Department of this hospital. A general purpose bench power supply was required having good regulation over a current range of 0 to 40mA and an output voltage variable from approximately 0 to 400V.

The circuit used is shown in Fig. 5 and its equivalent circuit is that of Fig. 4. The choice of circuit values needs little comment except perhaps that of the series valves V_1 and V_2. Equations (5) and (6) show that a large value is required for both the μ (for improved voltage stabilization) and the g_m (for the lower output impedance) of the series valve, and some compromise must be reached. In addition, a valve with a low d.c. impedance is desirable so that minimal anode voltage is required to pass maximum load current at zero grid volts, thus enabling the secondary voltage of the mains transformer to be kept to a minimum. Also the valve must be able to dissipate the full output of the transformer as would be necessary when operating with a nearly shorted output at maximum load current.

The obvious first choice for V_1 and V_2 is a power triode or a triode-connected power tetrode or pentode. These valves normally have a low μ (<10) and in many cases considerable anode voltage is required to provide 40mA anode current at zero grid volts. For example, power tetrode type 6BW6, triode connected needs an anode voltage of 110V for a cathode current of 40mA at zero volts bias, at which conditions μ is 5·3 and g_m 2·8mA/V. A larger μ would be desirable and could be obtained from a valve such as type 12AT7, but in this case the d.c. impedance of the valve is excessive and the anode dissipation a little low. The compromise was resolved using the double triode type 12BH7. Two of these are employed with the two halves of each valve connected in parallel. The pertinent characteristics of these three valves are given in Table 1.

The transformer and chokes are standard components and the power pack is housed in a small "Imhof" case. A 3½in meter on the front panel can be switched to read either 400V or 40mA full scale. A 5V and 6·3V a.c. supply is also provided on the panel.

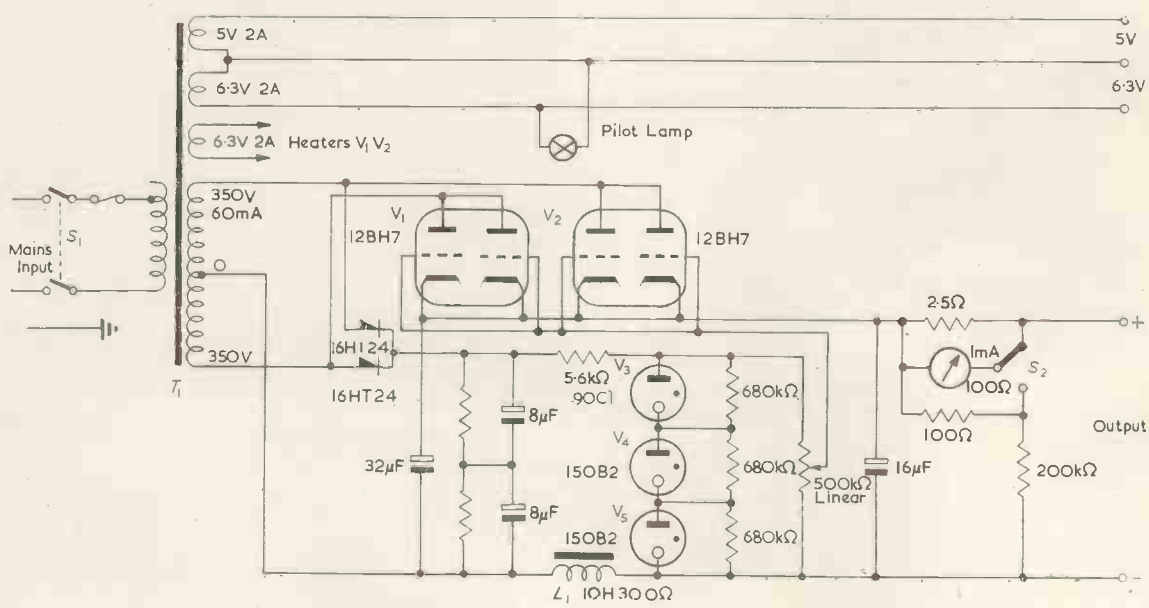
TABLE 1

Pertinent constants for three valve types when used as cathode-follower rectifier

		6BW6 (g_2 to anode)	12AT7 (halves in parallel)	12BH7 (halves in parallel)
Anode voltage	for i_a = 40mA at g_1 = 0 volts	110V	*180V	70V
μ		5·3	*70	16·6
gm		2·8mA/V	*18mA/V	12·2mA/V
Total anode dissipation . .		12·5W	5W	7W
Total maximum anode current		75mA	40mA	40mA

* *These values are outside the anode dissipation of the valve but in a full-wave rectifier circuit two valves are used and thus the total dissipation over half a cycle can be doubled.*

Fig. 5. Stabilized cathode-follower power pack, full circuit diagram



Performance of the Circuit

The load characteristics for various zero current voltages are shown in Fig. 6. The average output impedance over the middle current range of these curves is approximately 200Ω, with a higher value at smaller currents due to decrease of valve mutual conductance and also at larger currents, due to grid current limitations. This value compares well with the calculated ones which use values for the valve constants derived from a steady and not alternating anode voltage.

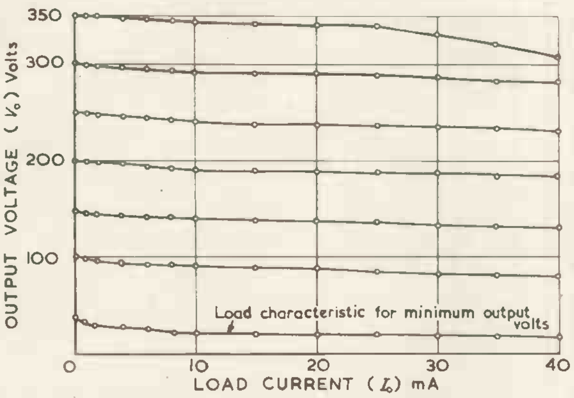


Fig. 6. Load characteristics, stabilized cathode-follower power supply (Fig. 5), at various initial (no load) voltages

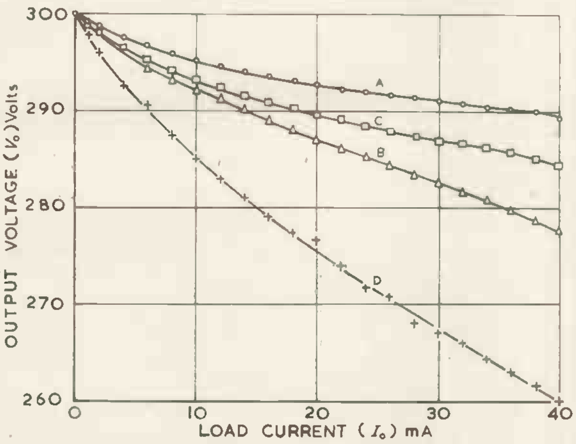


Fig. 7. Load characteristics for the circuit of Fig. 5 for various reference voltage arrangements, at an initial (no load) voltage of 300V

Curve A: *reference voltage supplied by battery instead of neon tubes*
Curve B: *reference voltage supplied by battery, with smoothing choke moved to output side of reference voltage*
Curve C: *reference voltage supplied by neon tubes as in Fig. 5*
Curve D: *no reference voltage (neons in Fig. 5 removed)*

Fig. 7 shows the effect of altering the type and position of reference voltage supply upon the load characteristics. The best regulation is obtained when the neon chain is replaced by battery A. The effect of placing the smoothing choke in the more conventional position (after the reference voltage) is shown by curve B; the output impedance increases from 150 to 470Ω. The characteristics of the practical circuit of Fig. 5 (curve C) is seen to lie between curves A and B. If an unstabilized source is used for the control voltage, the load characteristic worsens very considerably (curve D).

The effective stabilization of the circuit at 300V, no load, is shown in Fig. 8. It shows, as a pen recorder trace, the effect of a 5V change in the a.c. input voltage in the circuit of Fig. 5; (a) with the neon tubes removed, and (b) as run normally with the neons in circuit. The percentage stabilization ratio is approximately 4 times. The calculated value from the values of Fig. 4 is 5·6 times. Run for a period of 24h at 300V, the variation of output voltage was within one per cent.

An Inherent Instability

If the circuit of Fig. 3 is re-drawn as a feedback circuit (Fig. 9) it can be seen that oscillation will occur if the transfer characteristic of the passive four-terminal network has a value greater than the inverse gain of the cathode-follower. At resonance:

$$\omega = \omega_0 = \sqrt{\left(\frac{R_L + R_c}{LCR_L}\right)}$$

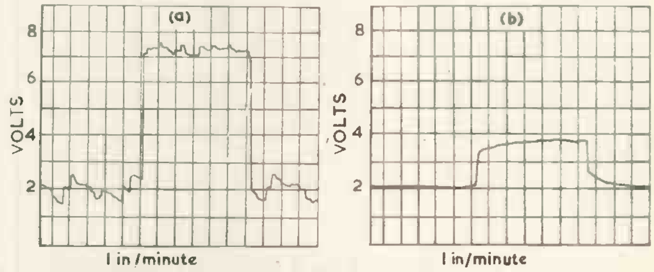


Fig. 8. The effect of a 5V change in input voltage on the output of the circuit of Fig. 5

(a) Final circuit, neons removed
(b) Final circuit, neons replaced

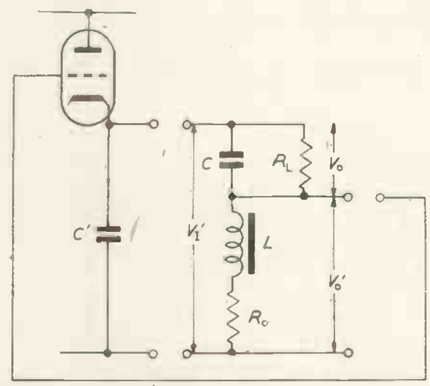


Fig. 9. Stabilized cathode-follower power supply, a.c. equivalent circuit

The value of the transfer characteristic is then:

$$|V_o'/V_I'|_{\omega=\omega_0} = \sqrt{\left(1 + \frac{LCR_L^3}{(R_L + R_c)(R_LR_cC + L)^2}\right)}$$

This expression must be smaller than $1/G$, where G is the cathode-follower gain. At small values of load resistance (R_L), the expression tends to unity. As $R_L \to \infty$, it becomes:

$$|V_o'/V_L'|_{\omega=\omega_0} = \sqrt{[1 + (L/CR_c^2)]}$$

Therefore $|V_o'/V_I'|_{\omega=\omega_0}$ can be reduced towards unity by increasing C or decreasing L provided that in the latter case R_c is not decreased or the smoothing adversely affected. If the power pack is stable on no-load it would seem at first sight to be stable over all its operating range. Unfortunately, the gain of the cathode-follower varies with R_L being considerably reduced at small values of load current. Thus, although stable at no load conditions, a small range of instability may exist at middle current values. This remaining oscillation can be removed by decreasing the effective load on the cathode-follower by increasing the value of C' (Fig. 9).

Assessment of Cathode-Follower Power Supplies

As previously suggested, the main advantage of the cathode-follower type power supply is the ease with which the output voltage can be varied. When the properties of the series type stabilizer are added to it, its performance compares very favourably with its nearest fixed-voltage counterpart, the simple neon tube stabilized power pack. Both percentage stabilization ratio and output impedance are better and the cost is very little altered, the normal rectifier being replaced by the triode valves and a small metal rectifier added. The capacitor and transformer ratings may, in fact, be considerably lower than in other stabilizer circuits having a comparable output.

The cathode-follower stabilized power pack, when designed for current ranges up to about 60mA and output voltages up to 250V is a very flexible and useful circuit. The voltage range can be considerably extended if a longer chain of reference-voltage neon tubes is not thought undesirable, or if a suitable external source of reference-voltage is available. It does, however, seem particularly suited to lower voltage applications. A power supply unit embodying these principles and which is particularly suitable for electrophoresis applications is being manufactured by Messrs. A. Gallenkamp and Co. Ltd, Sun Street, London, E.C.2.

Acknowledgments

The author is indebted to Dr. C. W. Wilson, in whose department this work was carried out and to Mr. N. H. Pierce, for his constructional help.

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A Television Line Selector Unit

By P. L. Mothersole*, Grad.I.E.E.

This article describes a line selector unit, designed to enable a normal triggered oscilloscope to be used as a television line waveform monitor.

The difficulties of obtaining a jitter-free display are discussed, and methods described to overcome them.

IN television experimental work it is frequently desirable to display a single line of video signal on an oscilloscope. With a normal repetitive oscilloscope display all lines are superimposed, and it is impossible to observe phenomena connected with one particular line.

The separate display of a single line, or a group of lines, can be achieved by using a triggered oscilloscope with the trigger having a variable delay from some fixed point in either the complete picture cycle, or the frame cycle.

If one particular line is to be displayed then the p.r.f. of the trigger pulse must be the same as the picture repetition frequency, e.g. 25c/s for the 405- and 625-line systems.

When investigating waveforms associated with a frame time-base and its synchronizing circuits it is often desirable to superimpose the odd and even frame waveforms. The p.r.f. of the trigger pulse is then required to be 50c/s.

This article describes a unit designed to enable any normal triggered oscilloscope to display a selected line waveform. The p.r.f. of the output trigger pulse can be 25 or 50c/s.

Delay Circuit

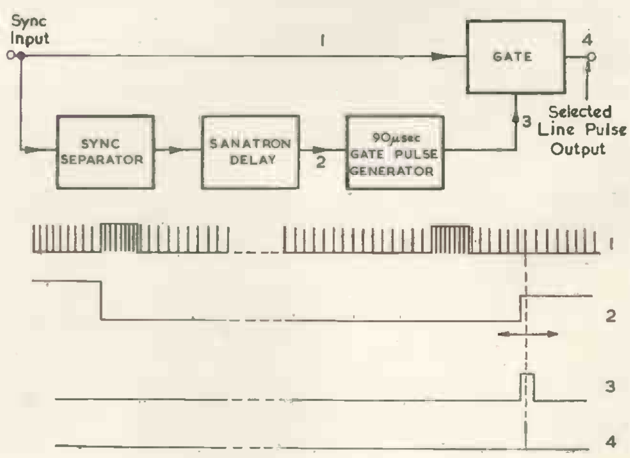
The delay circuit must be operated by every other frame synchronizing pulse to produce an output p.r.f. of 25c/s and be free from jitter. A normal trigger delay circuit (e.g. sanatron) is satisfactory for delays up to about 30 times the final sweep time, i.e. delay 3msec, sweep time 100μsec. A line selector must, however, provide a delay of over 0·02sec with a sweep time of 100μsec. When sweep expansion is used to examine part of a line the final display time may be 10μsec or less.

* *Mullard Research Laboratories.*

To obtain a jitter-free display of a particular line waveform, the line synchronizing pulse may be used to trigger the oscilloscope. A conventional delay generator can be used to operate a gate circuit to which is applied the normal synchronizing waveform. Single line synchronizing pulses are passed through the gate.

A selector system operating on this principle is shown in Fig. 1. The frame synchronizing pulse is separated from the input waveform and used to trigger a sanatron delay circuit. The delayed edge triggers a gate pulse generator which opens a gate for the duration of one line, in this case 90μsec approximately. The appropriate line synchronizing pulse passes through the gate and may be used to trigger the oscilloscope. Any jitter associated with the delay generator and gate circuit cannot produce jitter

Fig. 1. Selection of a synchronizing pulse using a gate circuit



on the oscilloscope trace since this is triggered by the line pulse of the displayed line.

An alternative method is to use the line synchronizing pulses to trigger back the delay generator. This saves the complexity of a gate pulse generator and coincidence gate. A convenient method is to add the differentiated line synchronizing pulses to the linear rundown waveform of a Miller circuit, triggered by the frame synchronizing pulses. This principle is shown in Fig. 2. The combined waveform is d.c. coupled to a cathode coupled Schmitt trigger circuit. The trigger point of this circuit is determined by its d.c. level and will be initiated by the leading edge of the line synchronizing pulse. The negative edge

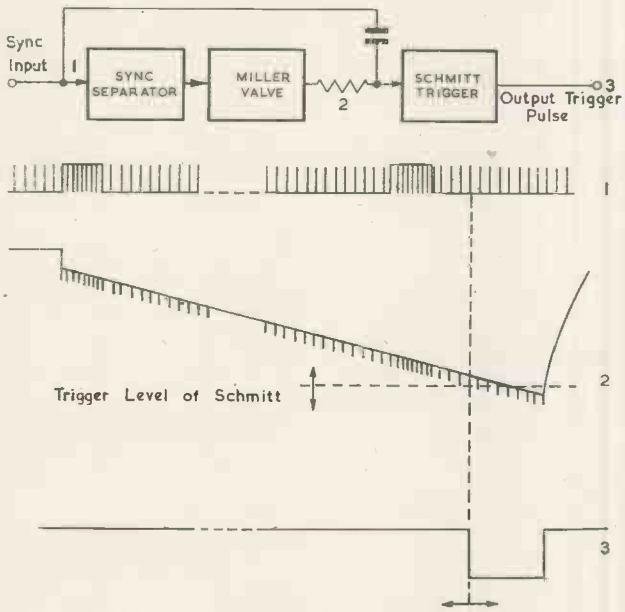


Fig. 2. Selection of a synchronizing pulse using a trigger circuit

of the output pulse may therefore be used to trigger the oscilloscope.

The above systems are satisfactory only for the display of waveforms at a p.r.f. of 25c/s. A p.r.f. of 50c/s is often required for investigating faulty interlacing in frame circuits. The frame pulse cannot be used to initiate the delay since it is often the frame pulse and the preceding lines that are under investigation. Most laboratory pattern generators can provide a 50c/s square or sine wave output locked to the frame pulse, and this waveform may be used to initiate a delay circuit. Since a trigger pulse derived from the line pulses cannot be used, a stepped delay is only possible if the twice line frequency (20·25kc/s) pulses are used.

In practice the oscilloscope trace duration will be about 600μsec, to cover the complete frame synchronizing waveform. The maximum ratio of delay to sweep time is therefore about 30 and a conventional delay circuit can be used.

Complete Instrument Requirements

Input

The instrument must be able to handle signals of either polarity, at a high or low impedance level, and provide an 80Ω termination if required. A synchronizing separator should be incorporated, to enable it to operate from a composite video waveform. The input circuits must handle signal voltages between 0·5 and 50V. This may be covered in two ranges, 0·5 to 5V, and 5 to 50V. The input impedance on this latter range must be high to enable the instrument to operate from the cathode of a c.r.t. in a television receiver.

Facilities must be provided to operate the delay circuit as a conventional trigger delay with an output p.r.f. of 50c/s when a 50c/s input is available. This input may be a sine wave or square wave with an amplitude between 1 and 10V. Provision should also be provided to derive this input from the mains input.

Delay Range and P.R.F.

When the output p.r.f. is 25c/s the stepped delay must cover a complete frame period (0·02sec). The delay must be switched to select trigger pulses from odd or even frames.

With a p.r.f. of 50c/s, the continuous delay should cover half a frame period (0·01sec), the phase being switched to cover the complete frame period. With a p.r.f. of 50c/s the output jitter, expressed as a percentage of non-interlace, must be less than 5 per cent.

Output Pulse

Two output pulses should be provided, one to trigger the oscilloscope, the other for application to the grid of a c.r.t. in a video monitor to mark the selected line. A pulse amplitude of 30V and a duration corresponding to half a line width (50μsec approximately) has been found satisfactory.

Complete Circuit Description

A block diagram of the instrument is shown in Fig. 3, and the circuit diagram in Fig. 4. The mode of operation of the unit is controlled by the system switch, S_1, shown in the video position (line selector p.r.f. 25c/s).

The low impedance input terminal is coupled through a capacitor to the first triode (V_{1a}) grid. An 80Ω resistor may be switched across this input to terminate the cable if required (S_5). The high impedance input terminal is connected to a compensated resistive divider providing about 10 : 1 attenuation of applied signals. The triode V_{1a} operates as a phase-splitter, with equal anode and cathode loads. The phase of input to the cathode compensated pentode video amplifier, V_{1b}, is selected by the phase switch, S_2.

The amplified negative-going video signal is a.c. coupled to the pentode limiter V_{2a} and d.c. restored by grid current. Since the video component is beyond cut-off, negative synchronizing pulses only are produced at the anode. This waveform is integrated by R_{17} and C_9 and d.c. coupled to the triode limiter V_{2b}. This valve is held in grid current by the d.c. coupling but cut-off by the integrated frame pulse. The positive pulse at the anode is differentiated and coupled to the pulse shaper V_{3a}. This valve is held cut-off by the positive cathode potential, its anode being at h.t. potential.

The short suppressor grid base pentode, V_5, is connected as a Miller integrator. The control grid is held in grid current, but the anode current is cut-off by the d.c. connexion from the screen to suppressor grid ($R_{34}R_{35}$) and the positive cathode potential. The anode potential is therefore close to the h.t. and the gate diode V_{4a} almost conducting.

The negative frame pulse from V_{3a} anode initiates the anode rundown, the rate of fall being determined by the grid aiming potential (VR_1) and the time-constant of the circuit R_{28}, R_{29} and C_{14}. The Miller valve is reset when the anode bottoms by the transitron action due to the screen suppressor connexion, C_{15}, R_{34}. The grid time-constant is adjusted for a 25c/s p.r.f. (VR_1), the complete cycle just occupying a double frame period (0·04sec).

To change from odd to even frames, a single run-down may be lengthened to miss one trigger pulse. This is con-

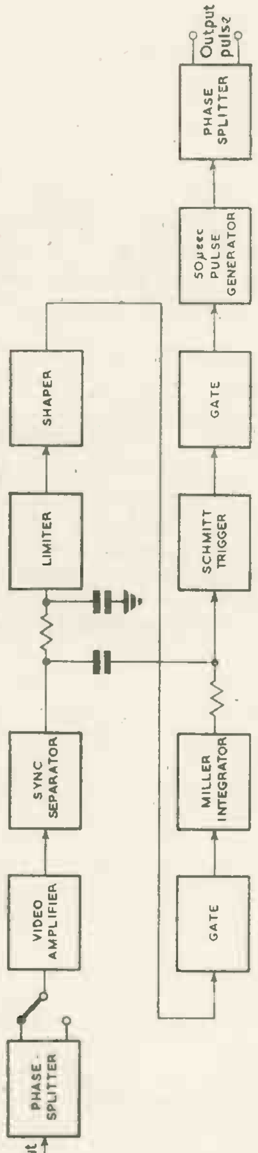


Fig. 3. Block diagram of the complete instrument

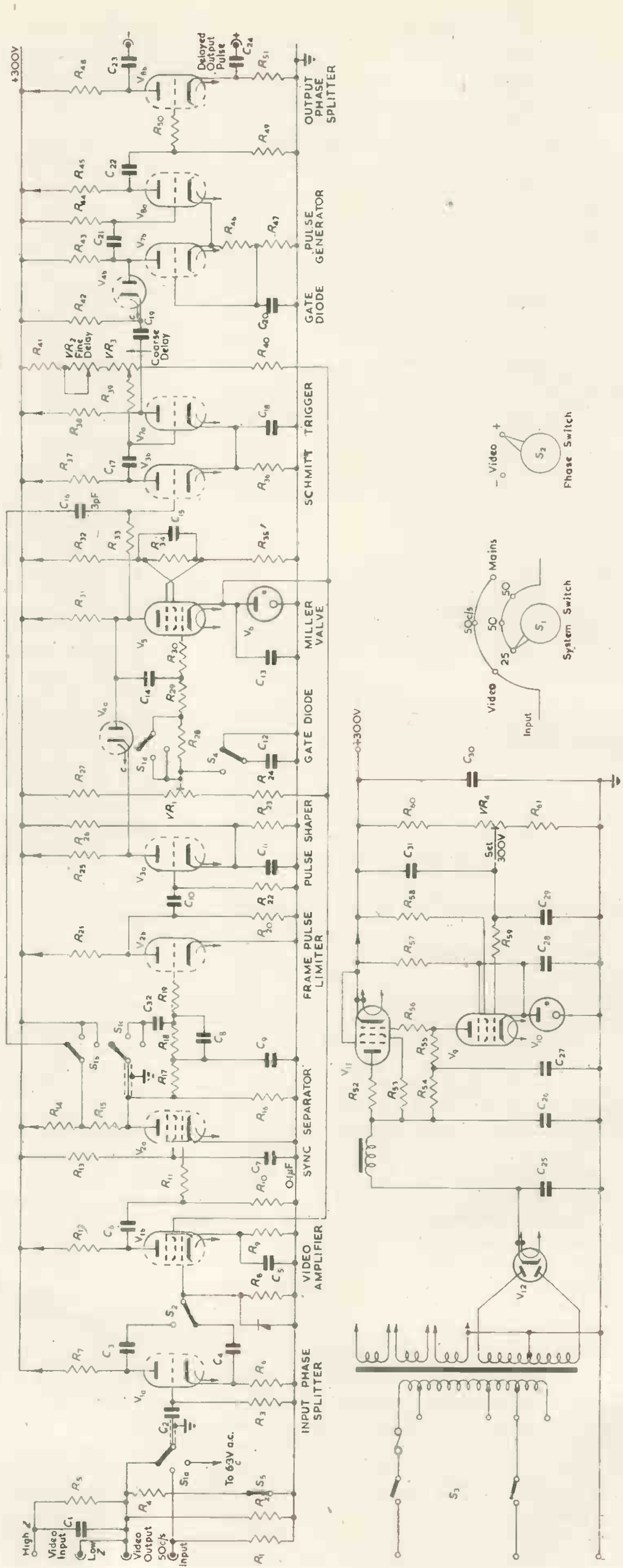


Fig. 4. Circuit diagram of the complete instrument

veniently done by momentarily lowering the aiming potential. The capacitor C_{12} and the press-button switch S_4 enable this to be done.

The Miller valve's anode is d.c.-coupled via R_{33} to the control grid of a Schmitt trigger circuit, $V_{3b}V_{7a}$. Also applied to this grid are the negative line synchronizing pulses differentiated by C_{16}. The trigger level of the Schmitt is determined by the d.c. level of V_{7a} control grid. This potential is varied by the delay controls VR_2, VR_3. At the start of the run down V_{3b} is conducting, V_{7a} cut-off, its anode at h.t. potential. When the circuit triggers, the negative going output is differentiated and coupled via the gate diode V_{4b} to the monstable pair, $V_{7b}V_{8a}$. In the stable state, V_{7b} is cut off by the cathode current of V_{8a}. The time-constant of the circuit ($C_{21}R_{44}$) is arranged to produce a 50μsec positive pulse at V_{8a} anode.

This pulse is coupled to the output valve, V_{8b}, which operates as a phase-splitter with equal anode and cathode loads.

When the system switch, S_1, is switched to the 50c/s position, the 50c/s sine or square wave input signal is applied to the input phase splitter. The video amplifier and pentode limiter operate as before. The integrator circuit is by-passed by C_{32}, and the Miller circuit is initiated by the negative-going edge. The Miller valve's grid circuit time-constant is shortened to enable the circuit to reset in one frame period, the output p.r.f. being 50c/s. The Schmitt trigger circuit is triggered by the Miller output only. The delayed output pulse being determined as before by the d.c. level of the circuit.

The total continuous delay in the 50c/s position corresponds to a half frame period, (0·01sec). Full coverage may be obtained by using the input phase switch to produce a 180° phase change in the square wave triggering the Miller circuit.

An additional position is provided on the system switch to enable the 50c/s trigger pulse to be derived from the mains input. Due to the 'spongy' mains lock in pattern generators this last position is not suitable for very expanded oscilloscope displays.

The circuits are designed to operate from a +300V supply, the consumption being 50mA. The power supply is conventional, using a single series valve, V_{11}, a control amplifier, V_9, and reference tube, V_{10}.

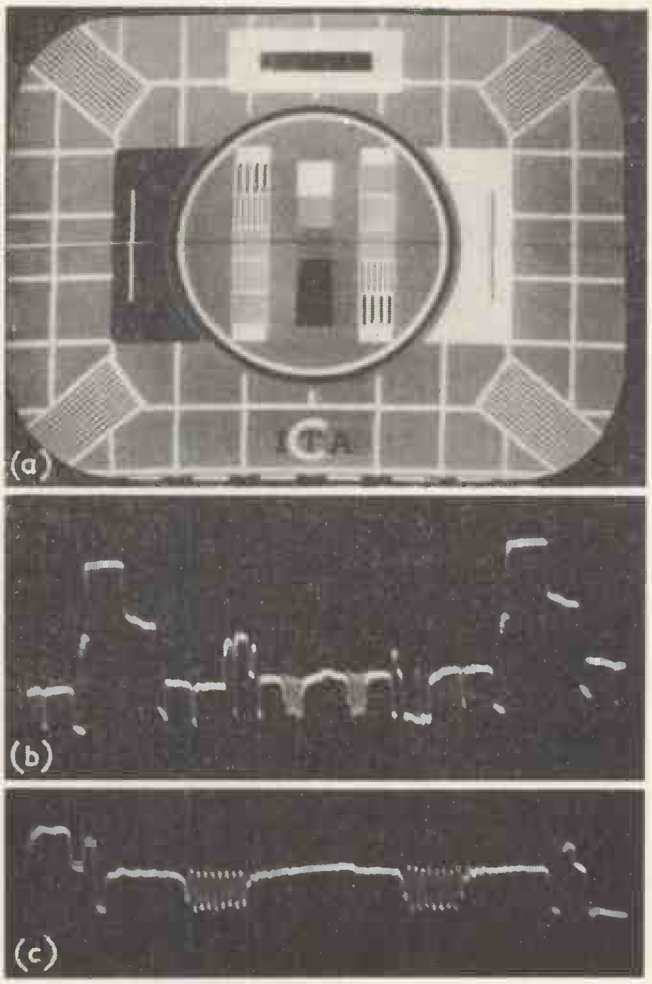
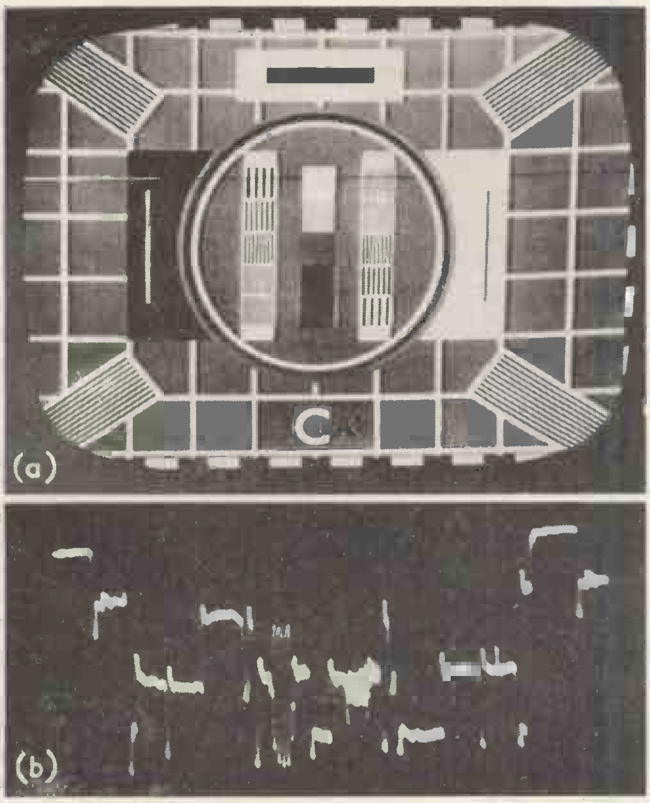


Fig. 6(a). Test card 'C' with marker pulse, (b) corresponding video waveform, (c) expanded portion of video waveform

Fig. 5(a). Test card 'C' with marker pulse, (b) corresponding video waveform



Results and Conclusions

The instrument described has had many months of continuous use on both 405-line and 625-line systems. It enables test signals to be examined in detail and photographs of test card 'C' with the marker pulse together with the corresponding video waveforms are shown, Figs. 5 and 6. From the video waveforms, the overshoot and high and low frequency response can be accurately measured.

Instruments providing the facilities described in this article are believed to be indispensable in any television development or research laboratory. It is also felt that an instrument of the type described would be very useful in television service work.

Acknowledgments

The author wishes to thank the Manager of the Mullard Research Laboratories and the Directors of Mullard Limited for permission to publish this article.

The line selector described in this article is marketed by Mullard Equipment Limited having the type number L196.

Microwave Measuring Devices

By H. H. Klinger*, Dipl.—Ing.

The article reviews the principles and characteristics of microwave measuring devices for radio link systems and in particular deals with devices required for developing testing, installing and monitoring link systems, e.g. measuring oscillators, frequency meters, impedance measuring devices, calibrated attenuators, wattmeter multipliers, and low-pass filters.

VERY high frequencies have attained considerable importance due to the development of the technique of radio link transmission. Suitable measuring devices are required for these frequencies and their applications. Their essential characteristics are the use of coaxial transmission lines, waveguides, coaxial circuits and cavity resonators as measuring circuits. In the following, an account is given of measuring devices for radio link transmission systems in the decimetre and centimetre range which have been developed by Siemens & Halske A. G. for this purpose.

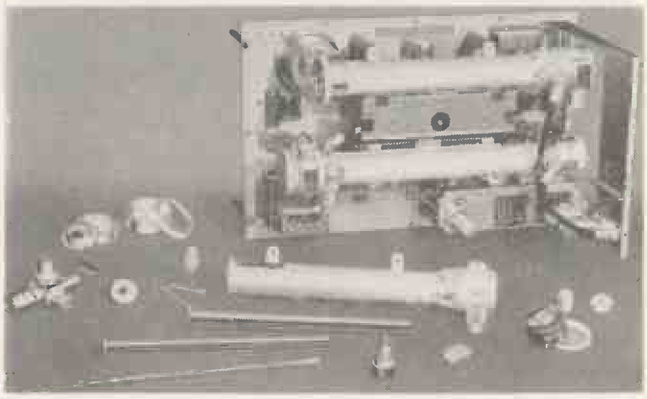
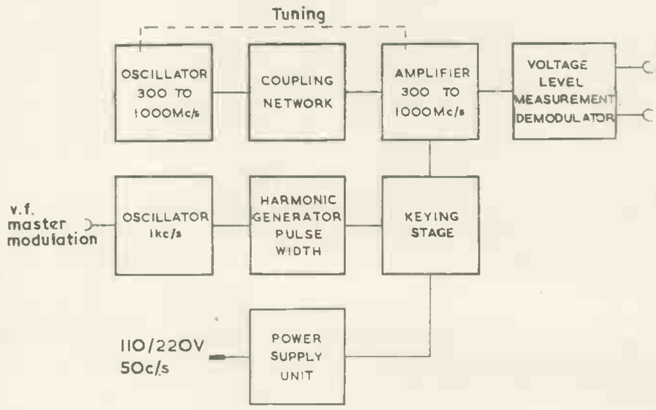


Fig. 1. Two-stage measuring oscillator 3W58 for frequencies 300 to 1 000Mc/s (Above). Outside view of assembled oscillator (Below). Inside view of assembled oscillator

Measuring Oscillators

Suitable measuring oscillators are of primary importance for carrying out exact measurements at high frequencies. They must possess a high stability of frequency and amplitude and be highly insensitive to fluctuations in the supply voltage and to mechanical vibrations. In addition, frequency and output must be finely and reproduceably adjustable over a wide range.

The two-stage measuring oscillator type 3W58 (Fig. 1) was developed for the range of 300 to 1 000Mc/s. It has a disk triode type 2C39A operating as an oscillator valve in a grounded-grid circuit. The resonant circuits are coaxial transmission lines adjusted by short-circuit pistons. The high stability of frequency of this device, its fine and reproduceable frequency adjustment to 10^{-5} and an additional fine adjustment—range of variation up to



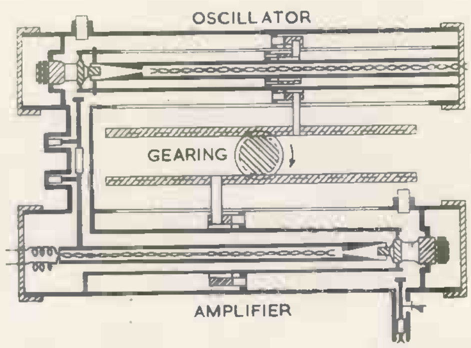


Fig. 2. Basic diagram and construction of measuring oscillator 3W58

$\pm 3 \times 10^{-4}$—are of advantage, particularly for node measurement in measuring circuits, for measurements on steep filter curves and for the determination of the high figure of merit of cavity circuits. The oscillator works practically non-reactively, owing to its two-stage design (oscillator and buffer-stage) so that the transmission frequency changes but slightly even with high output and variations in load. Simple manipulation is accomplished by single button control notwithstanding the multi-stage design of the oscillator. Fig. 2 shows the basic diagram of the test oscillator which may be self- or externally-modulated. Pulses with a p.r.f. of 1 000p/s are produced in the modulation element for self-modulation. The pulse ratio may be adjusted from 1 : 1 to 1 : 10.

The measuring oscillator type 3W59 for the range of 1 500 to 2 700Mc/s has a disk triode type 2C40 in a grounded-grid circuit with coaxial tuners. This device is suitable as current source for impedance matching measurements on aerials, transmitter and receiver cables.

* *Siemens & Halske A. G., Erlangen.*

as well as for measurements on microwave components and sensitivity measurements on receivers.

The test oscillator type 3W513 was developed for the range of 2 400 to 4 500Mc/s. A reflex klystron is used as an oscillator. A coaxial transmission line with a capacitive short-circuit piston is used as resonant circuit. The reflector voltage of the klystron is varied simultaneously with the tuning so that the correct supply voltage is used on the klystron independent of frequency. A 50c/s voltage is used for wobbling the frequency of the output voltage.

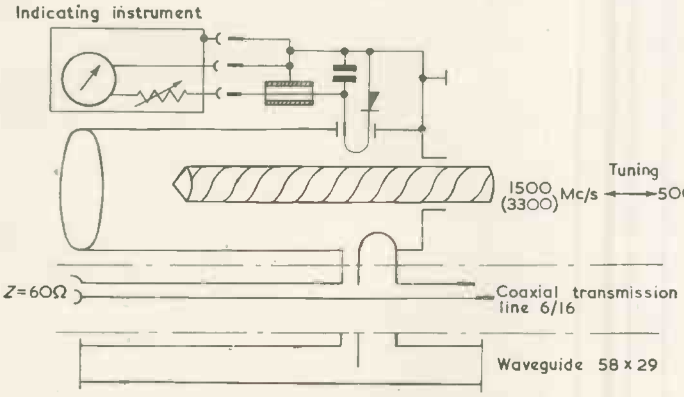


Fig. 3. Diagram of frequency meter 3F120/121 for the range 1 500 to 5 000Mc/s

The maximum high frequency power is about 100mW. An accurately adjustable waveguide attenuator with a range of 120dB is installed for using the device as a receiver test oscillator. The oscillator can be self- or externally-modulated with rectangular pulses.

The measuring oscillator type 3W515 for the frequency range of 4 400 to 9 100Mc/s is similar in design.

Frequency Meters

Frequency measurements up to about 2 000Mc/s may be attained by using coaxial strip tuners arranged in a ring as resonant circuits, in which case an accuracy of measurement of 0·5 to 1 per cent is obtained. A greater accuracy of measurement corresponding to 10^{-4} is obtained with frequency meters in which tunable coaxial quarter-wavelength lines serve as resonant circuits, one end being short-circuited while the other end is open. Figs. 3 and 4 show the circuit diagram and construction of the frequency meter type 3F120/121 for the range of 1 500 to 5 000Mc/s (type 3F120 with coaxial through-section and type 3F121 with waveguide through-section) respectively. The through-section and the crystal rectifier, arranged in the measuring head of the frequency meter, are loosely coupled to the resonant line near the short-circuited end. The high accuracy of adjustment and measurement of this device is mainly achieved by a spindle drive made of invar-steel, for the tuning of the coaxial transmission line.

A still higher accuracy of measurement corresponding to an error $\pm 5 \times 10^{-5}$ can be achieved with the frequency meter type 3F112a/b in which a cylindrical cavity resonator excited in the fundamental mode is used as a measuring circuit. It is intended for service and laboratory measurements for radio link systems in the range of 2 400 to 2 700Mc/s. The frequency may be tuned within this range by an axial plunger in the cavity.

The heterodyne frequency meter type 3F113 (Fig. 5) was developed for particularly accurate measurements in the range of 950 to 5 000Mc/s. An approximate measurement may be made by a built-in resonant frequency meter. This contains two scales, one for the direct reading of the frequency with an error of 0·5 per cent and the second one giving the order of the harmonic required for the fine measurement of frequency. For the exact measurement the 10th to 50th harmonic of a highly stable local oscillator (frequency range 95 to 105Mc/s) is compared with the frequency to be measured. The resonant circuit of the highly stable local oscillator for the fine measurement consists of a low-loss temperature compensated coaxial circuit of high capacitance across which is connected a small variable capacitor with a linear frequency scale. A precision drive enables the frequency to be read to an accuracy of 5×10^{-6}. The frequency of the local oscillator may be standardized against a built-in crystal-controlled multi-frequency calibrator with an absolute error of 5×10^{-6}. This calibrator has scale divisions of 1 or 0·2Mc/s. The frequency of the local oscillator may be so shifted that the frequency scale coincides absolutely with the crystal standard at the multi-frequency calibrator points.

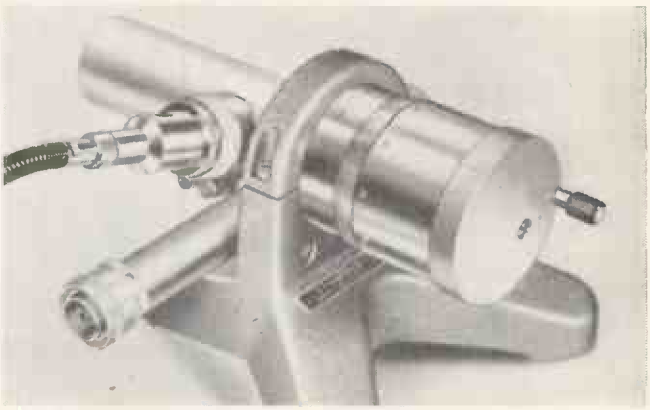
Wattmeters

High frequency power in the range up to about 3 000Mc/s may be measured with high accuracy with the thermal wattmeter type 3U81. The r.f. power is dissipated as thermal power in a resistive termination closely matching the characteristic impedance of the coaxial line so as to produce a temperature difference proportional to the r.f. power. The temperature of the resistor is measured by the aid of two copper windings arranged on the outside of the conductor, the difference in the resistance of these windings being indicated by a bridge with amplifier attached (Fig. 6). The change in resistance is proportional to the high frequency power to be measured so that the instrument may be calibrated directly in milliwatts. The terminal resistance changes its value so slightly that the wattmeter may also be used as a voltmeter with a resistance of 60Ω. A power of 16·7mW corresponds to a voltage of 1V. The calibration of the instrument is done with direct current. The minimum readable power is about 2mW at a measuring range of 200 to 500mW (full scale). Large high frequency powers, up to 10 or 25W, may be measured by using wattmeter multipliers (type 3U82 or 3U83) (Fig. 7).

Impedance Meters

The slotted coaxial transmission line is the standard device for determining impedances in the decimetre range (300 to 3 000Mc/s). Fig. 8 shows the coaxial measuring transmission line type 3R221 for the range of 600 to 6 000Mc/s. The v.s.w.r. along the line is measured by

Fig. 4. Frequency meter 3F120 for 1 600 to 5 000Mc/s



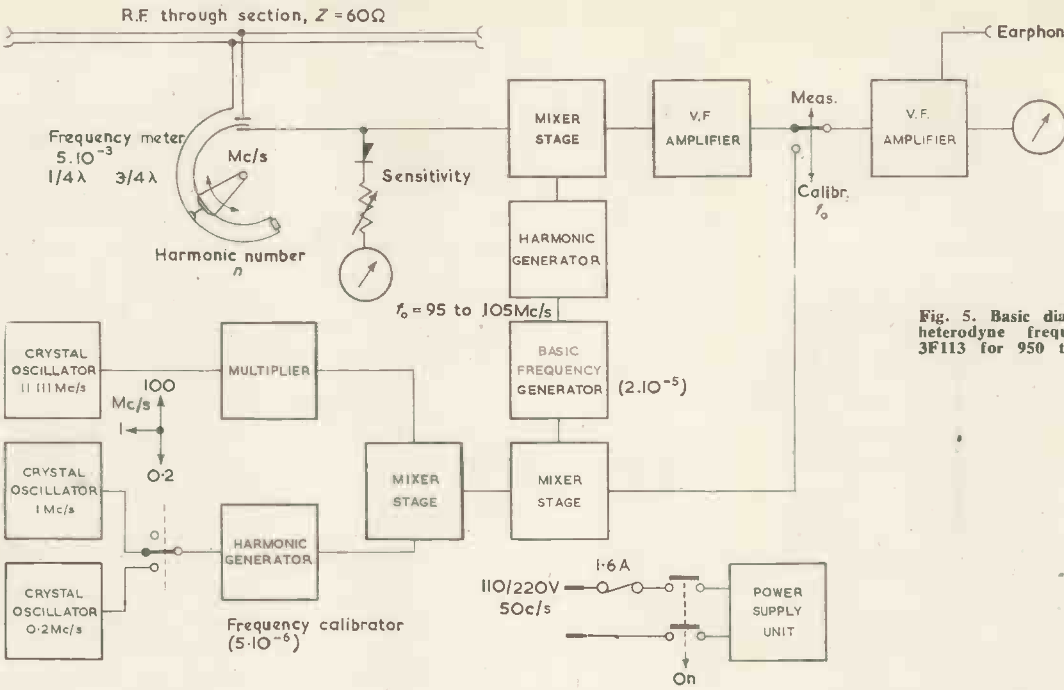


Fig. 5. Basic diagram of the heterodyne frequency meter 3F113 for 950 to 5 000Mc/s

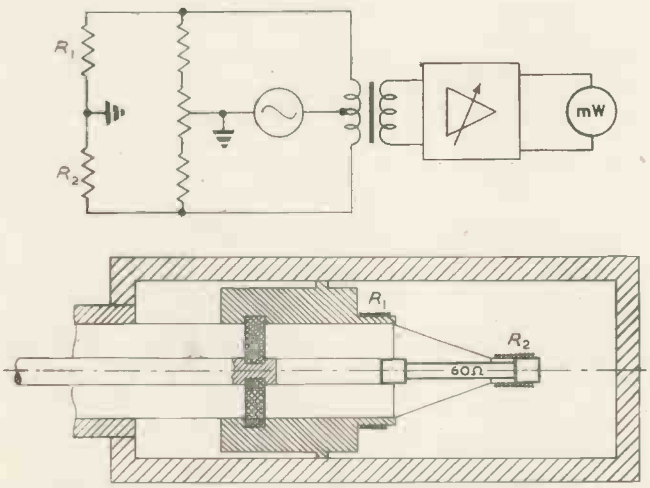


Fig. 6. Basic diagram of the thermal wattmeter 3U81 for the range of 0 to 3 000Mc/s and cross-section of the measuring head

means of a crystal rectifier probe with adjustable depth inserted into a slot in the line. The position of the probe can be read accurately to 0·01mm on a drum fitted with a vernier. An indicating amplifier is necessary for precision measurements requiring considerable decoupling between the oscillator and measuring circuit. A d.c. voltage amplifier, e.g. type Rel 3U13, is used for the purpose in measurements with unmodulated high frequency voltage, and a level meter, e.g. Rel 3D311, with modulated high frequency voltage.

Waveguide measuring circuits prove to be of increasing importance in the centimetre range (3 000 to 30 000Mc/s) when the wavelength of the electromagnetic oscillations begins to be less than the circumference of the outer tube of the coaxial conductor. The waveguide measuring circuit type Rel 3R224 (Fig. 9) was developed for the range of 2 600 to 12 400Mc/s. This measuring circuit possesses interchangeable waveguides with five different cross-

Fig. 7. Thermal wattmeter with terminal resistor (left) and wattmeter multiplier, with blower for cooling (right)

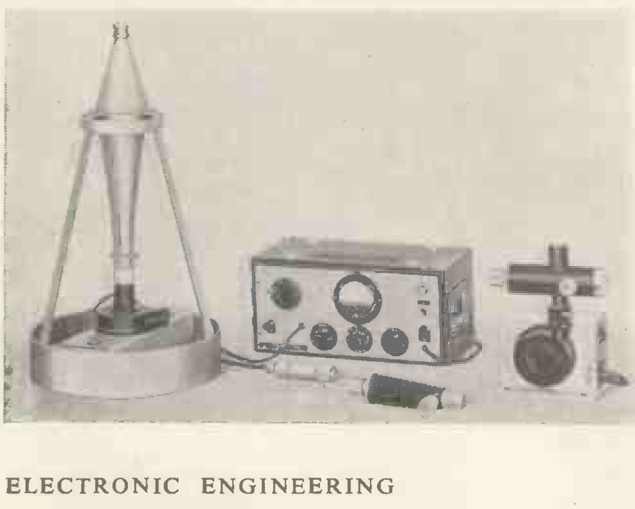
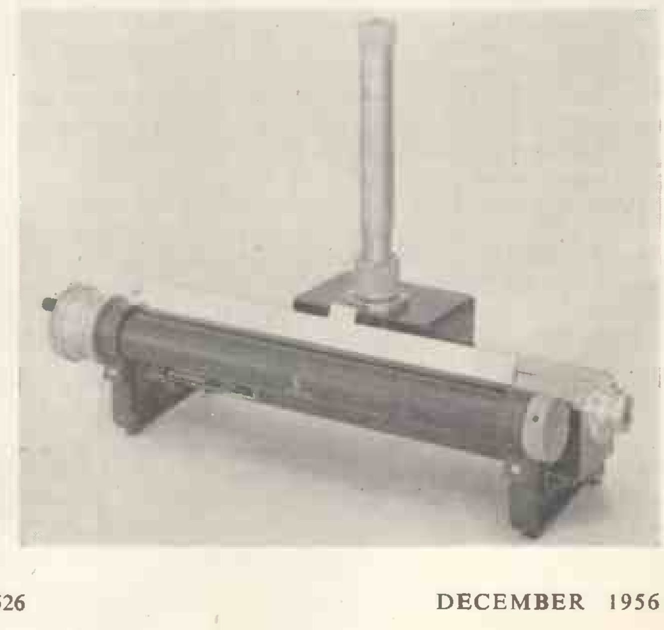


Fig. 8. Coaxial measuring circuit Rel 3R221 for the range of 600 to 6 000Mc/s



sections adapted to the different frequency ranges and fitted with slits into which the probe dips. The field along the measuring section may be explored with the probe by means of a carriage with a friction drive. Suitable indicating amplifiers as mentioned in the preceding examples are necessary for precision measurements. The location of the probe on the measuring circuit can be read with an accuracy of 0·01mm on a ground glass by means of a vernier and a special optical system.

Fig. 9. Waveguide measuring circuit Rel 3R224 for the range of 2 600 to 12 400Mc/s with interchangeable waveguides

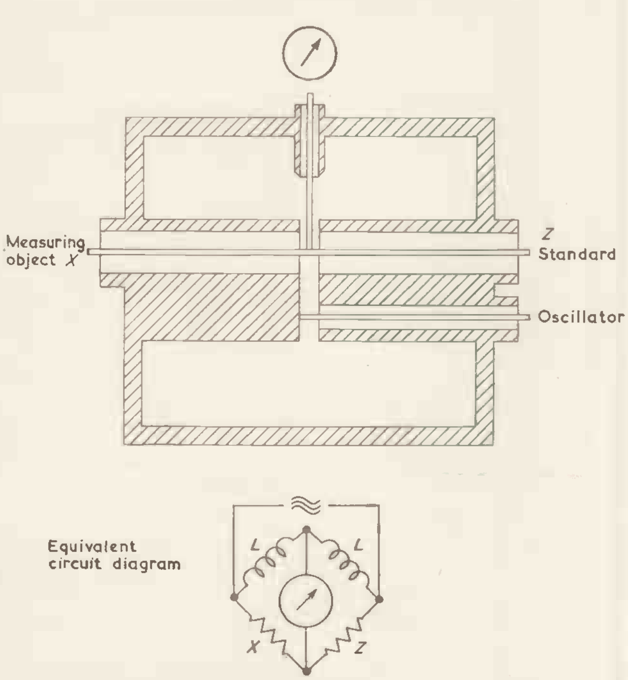


Fig. 10. Basic diagram of the reflection factor meter 3R29 for 50 to 1 000Mc/s and of the reflection factor meter 3R225 for 300 to 3 000Mc/s

Rapid determination of the amount of resistance matching between an oscillator and the characteristic impedance of supply cables and aerials must frequently be carried out. 'Reflection coefficient meters' have been developed for this purpose, the performance of which corresponds in principle to the balance attenuation method well known in low frequency engineering (Fig. 10). A coaxial circuit to which the high frequency voltage is applied coaxially in the centre replaces the differential transformer usual at low frequencies. The bridge output voltage is taken from the junction 'standard-measuring specimen' in the symmetry plane of the coaxial circuit and rectified by a built-in rectifier. The ratio between bridge output voltage and bridge input voltage is a measure of the reflection coefficient. In order to enable very small values of the reflection coefficient to be measured the device is fitted with a chopper relay which converts the d.c. voltage on the rectifier of the bridge into an alternating voltage at 50c/s. This a.c. is fed through a two-stage stabilized amplifier to the indicating instrument which is calibrated directly in values of the reflection factor for constant bridge input voltages. The reflection coefficient meter type 3R29 is for frequencies of 50 to 1 000Mc/s. The reflection coefficient meter type 3R225 is a corresponding instrument for the range of 300 to 3 000Mc/s. The special advantage of these indicators is that the accuracy of measuring increases with decreasing values of the reflection coefficient.

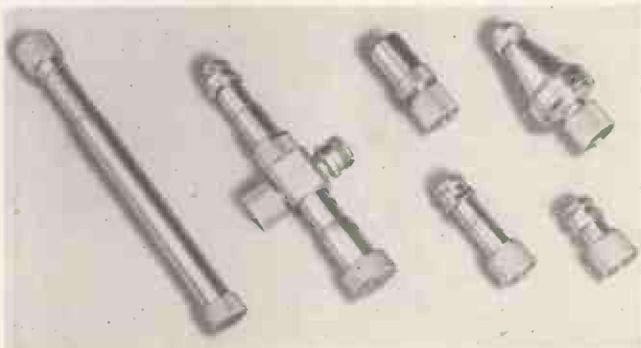


Fig. 11. Accessories for coaxial measurement: short-circuit stub, 60Ω capacitive divider, 20dB attenuator, 60Ω terminating resistor, 10dB attenuator

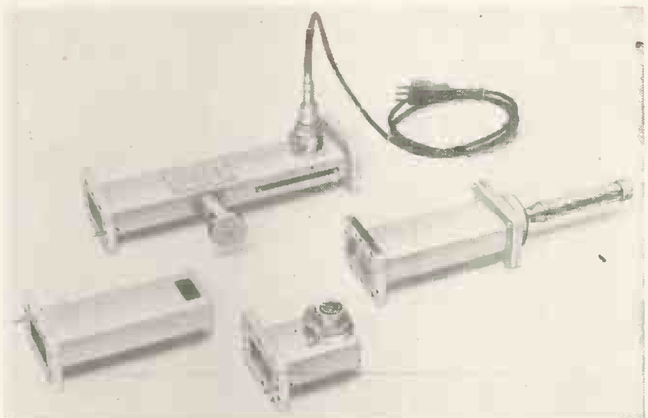


Fig. 12. Accessories for waveguide measurement: terminating resistor, attenuator, coupler waveguide-coaxial line, short-circuiting stub

Accessories

Accessories such as calibrating circuits, potentiometers, terminal elements, low-pass filters and attenuating elements are required for installing measuring sets. These accessories are made either as coaxial or as waveguide elements, depending on the technique basically applied. The accessories for the coaxial technique comprise, apart from the usual connecting lines and plugs and sockets, 60Ω terminal resistors, coaxial ohmic and capacitive potentiometers, variable attenuation elements, short-circuit lines and transit voltage testers for frequencies up to 5 000Mc/s (Fig. 11). As accessories for the waveguide technique (rectangular waveguides with an inside cross-section of 58 × 29mm) a terminal resistor for a load of 2W, variable attenuation elements with an attenuation range from 0·3 to 40dB, a reactive waveguide with short-circuiting stub and a coupler waveguide-coaxial line for the frequency range from 3 300 to 5 000Mc/s have been developed (Fig. 12).

The Design of Cold-Cathode Valve Circuits

(Part 3)

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

Gating Circuits Using Diodes

The previous sections have almost entirely been confined to triode circuits and the reader may wonder why diode circuits were not discussed first, since the valve itself is a more elementary device. This has been done deliberately, however, because the diode has its own properties that make it complementary to the triode, but with its own special applications. A common failing of engineers just commencing to design cold-cathode valve circuits is to regard the diode as inherently inferior to the triode, which they attempt to use for all purposes. This approach results in circuits that are not only expensive but clumsy, because arrangements always have to be made to extinguish triodes.

Basic circuits for OR and AND gates using cold-cathode diodes are shown in Fig. 26(a) and (b). These circuits can use small inexpensive diodes such as the 2N1 and NT2. In

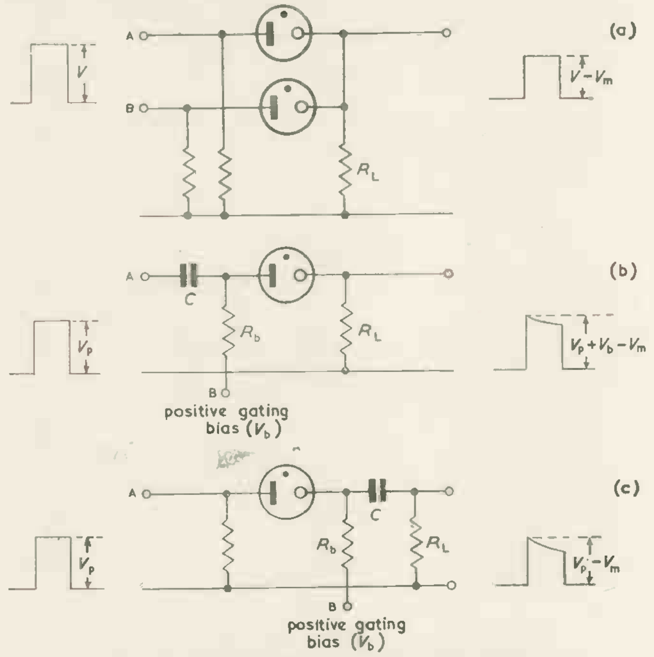


Fig. 26. Gating circuits using cold-cathode diodes
(a) A OR B *gate*
(b) A AND B *gate*
(c) A AND NOT B *gate*

Fig. 26(a), a voltage V (which exceeds the striking voltage V_s) applied to either input terminal will cause a valve to fire. The voltage which then appears at the output is $(V - V_m)$. If this is less than V_s, the other valve cannot fire and the input terminals remain isolated from each other. The AND gate shown in Fig. 26(b) operates on the pulse-plus-bias principle. The signal voltage V applied to each input terminal should be less than V_s, so that neither input alone can fire the valve. When a pulse is applied from terminal A after the bias from terminal B, the voltage applied to the valve is $2V$. The valve then fires and a voltage $(2V - V_m)$ appears at the output. For the circuit to operate as described, one obviously requires:

$$V < V_s < 2V$$

When gating circuits were considered previously, it was shown that OR and AND circuits could be simply designed using triode valves, but it was very difficult to give a NOT condition. The diode, however, can be used for a NOT condition as simply as for an AND condition; this is shown in Fig. 26(c). If a positive V_p (greater than V_s) is applied to input terminal A the valve will fire and a voltage $(V_p - V_m)$ is delivered to the output terminal as a pulse via the capacitor. However, if a positive bias voltage V_b is applied to terminal B the p.d. developed across the diode will be too small to fire it and no voltage appears at the output terminal.

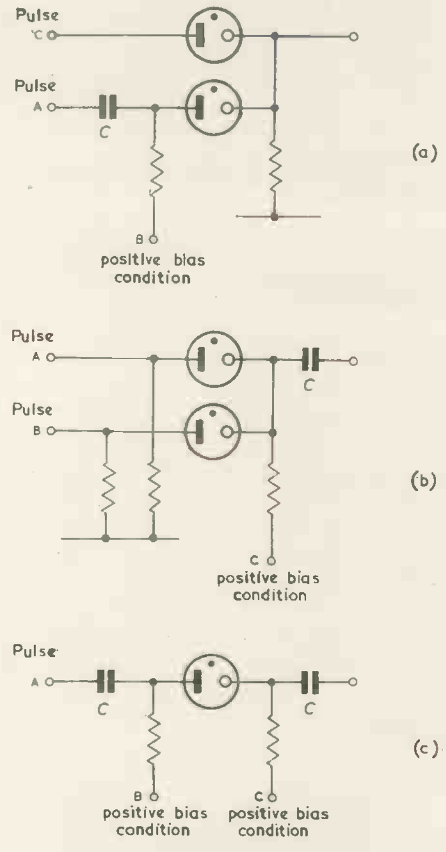


Fig. 27. More complex gating circuits
(a) A AND B OR C
(b) A OR B AND NOT C
(c) A AND B AND NOT C

To operate as described, the conditions applied to the circuit must comply with the following:

$$V_b < V_s < V_p$$

and

$$(V_p - V_b) < V_s$$

The circuits shown in Figs. 26(a), (b) and (c) can be combined to perform more complicated logical functions. Some examples are shown in Fig. 27. In all these circuits, the diodes extinguish automatically when the input voltages are removed.

* Siemens Brothers & Co. Ltd.

Diode Trigger and Counter Circuits

The cold-cathode diode can be used as a trigger circuit because its strike voltage V_s is greater than its maintain voltage V_m. Thus, if a signal causes the valve to fire by making the p.d. across it more than V_s, the valve will still conduct after the signal is removed provided that the p.d. across it still exceeds V_m. For this action to be reliable and to provide a useful output voltage, the difference between V_s and V_m should be large; diodes such as the XC14 and 2N3 are suitable and are sometimes called 'difference diodes'. The chief difficulty encountered when using a diode in trigger circuits is to provide isolation between the input and output terminals, because these have to be connected to the same pair of electrodes. This difficulty does not arise when a triode is used, because the input signal is applied to the trigger-cathode gap and the output taken from the anode-cathode gap. Most cold-cathode trigger circuits and counters have hitherto been designed using triodes, but the recent introduction of difference diodes has led to the development of practical trigger circuits using diodes.

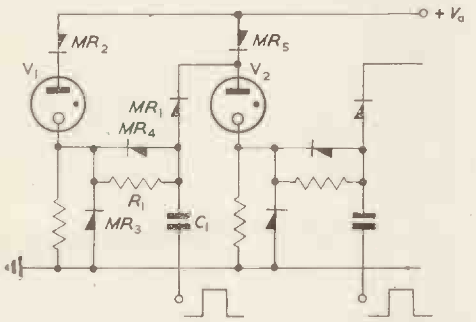


Fig. 28. Trigger circuit stages using 'difference' diodes

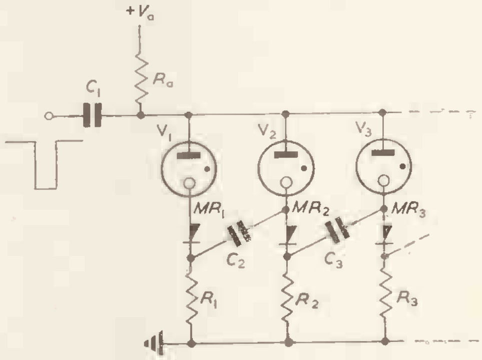


Fig. 29. Chain counter using diodes

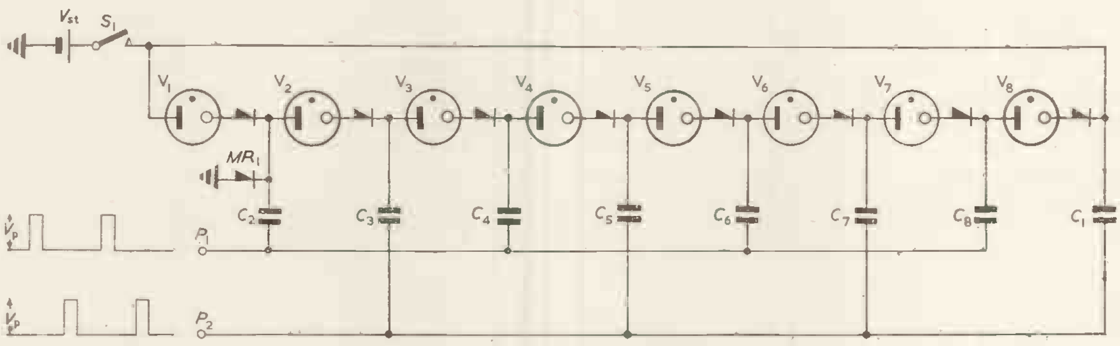


Fig. 30. Pattern register using diodes

One way of obtaining the necessary isolation between input and output circuits is to use rectifiers. Two stages of a trigger circuit which make use of this technique[19] are shown in Fig. 28. If, initially, the diode V_1 is 'on' and V_2 is 'off', the cathode of V_1 is providing a positive output voltage ($V_a - V_m$) which charges capacitor C_1 via R_1. When a positive pulse is applied via C_1, rectifier MR_4 conducts, the cathode potential of V_1 is raised and this valve extinguishes.

The pulse applied via C_1 also makes MR_1 conduct and raises the anode potential of V_2 above the h.t. voltage V_a; this biases-off rectifier MR_5 and fires the valve. The pulse has thus transferred the conducting condition from V_1 to V_2. The output voltage from V_2 can be applied to another similar stage or can be connected back to the anode of V_1. This circuit is reliable and versatile. It can be used as a standard building block to construct a variety of more complex circuits, such as counters and pattern stores.

When such circuits as counters are to be designed, it is sometimes possible to design a simpler circuit by considering the counter as a whole rather than as a number of interconnected trigger circuits. A very simple counter circuit using diodes[20] is shown in Fig. 29. If, initially, diode V_1 is 'on' and the remainder are 'off', the cathode of V_1 only is at a positive potential. Rectifier MR_1 is conducting and the capacitor C_2 charges up to the cathode potential of V_1. A negative pulse applied via capacitor C_1 extinguishes V_1 and its cathode potential falls to earth. The capacitor C_2 retains its charge, so the cathode of V_2 becomes negative and rectifier MR_2 is biased off. When the pulse disappears, the anode voltage of the valves rises towards the h.t. voltage and the p.d. across V_2 (whose cathode is negative) becomes sufficient to fire this valve. The pulse has thus transferred the conducting condition from V_1 to V_2 and subsequent pulses will transfer it further along the chain.

A very simple pattern register using diodes[21] is shown in Fig. 30. Marking signals are stored as charges on capacitors and are transferred from one capacitor to the next when the diode between them is fired by an applied pulse. Two pulse supply lines are required for stepping the store; alternate pulses appear on each line. Assume that, initially, capacitor C_1 is charged by momentarily closing switch S_1 but all the other capacitors are uncharged. Application of pulse P_2 raises the anode potential of V_1 sufficiently for the diode to fire and a current flows which discharges C_1 and charges C_2. The next pulse (P_1) is applied to C_2 and causes V_2 to fire and transfer charge from C_2 to C_3. Meanwhile capacitor C_1 can be recharged by closing S_1 again to insert another marking in the store.

Each pulse P_1 causes charge to be transferred from C_2 to C_3, from C_4 to C_5 and so on; each pulse P_2 causes charge to be transferred from C_1 to C_2, from C_3 to C_4, from C_5 to C_6, and so on. These pulses occur alternately and cause the pattern of stored charges to circulate round the ring.

If the current which flows during a pulse causes charge to transfer from capacitor C_n to C_{n+1}, the reduction in voltage across C_n equals the increase in voltage across C_{n+1} because the capacitances are equal. Thus, if the p.d. across C_n falls from V_n to V_n', while that across C_{n+1} rises from zero to V_{n+1}:

$$V_{n+1} = V_n - V_n' \quad \ldots\ldots\ldots\ldots\ldots \quad (8)$$

It is assumed that, during this process, the p.d. across the diode remains at its maintain voltage V_{am} until the valve extinguishes. If the voltage of the applied pulse is V_p, then at the time at which the valve extinguishes its anode voltage is $V_p + V_n'$ and its cathode voltage is V_{n+1}.

$$\therefore V_{am} = V_p + V_n' - V_{n+1} \quad \ldots\ldots\ldots\ldots \quad (9)$$

Solving equations (8) and (9) gives:

$$V_{n+1} = \tfrac{1}{2}(V_n + V_p - V_{am}) \quad \ldots\ldots\ldots \quad (10)$$

$$V_n' = \tfrac{1}{2}(V_n - V_p + V_{am}) \quad \ldots\ldots\ldots \quad (11)$$

It can easily be seen that if $V_n = V_p - V_{am}$, then:

$$V_{n+1} = V_n \text{ and } V_n' = 0.$$

Thus, if the pulse height is correctly chosen, all the p.d. across one capacitor is transferred to the next and the stored charge remains unaltered as it progresses along the chain.

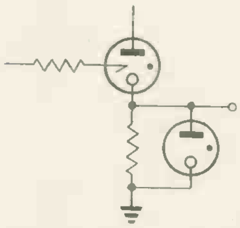


Fig. 31. Use of diode for clamping

In practice, the valves extinguish before their current has fallen to zero and the valves have slightly different values of V_{am}, not all the capacitances are equal and some charge is lost due to leakage resistance. It might be thought therefore that repeated operations of the circuit would cause the stored voltage to decay until it was finally insufficient to fire the diodes. In fact, the circuit has a compensating action.

If, for example, one valve has a high value of V_{am}, the voltage transferred to the subsequent capacitor is low, but a positive voltage is left on the previous capacitor. This supplements the p.d. produced next time the stage is operated so that repeated circulations cause the p.d. transferred to the subsequent capacitor to build up towards the correct value $V_p - V_{am}$. If a capacitor leaks, its stored voltage falls below the correct value and a negative voltage is left on the capacitor after the stage has operated. As the store rotates, the negative potential becomes distributed over all the unmarked stages and equilibrium is reached when leakage of negative charge from the unmarked capacitors equals the leakage of positive charge from the marked capacitors. The stored pattern thus continues to rotate, but the mean d.c. potential of the valves drifts away from earth. This is undesirable in practice and is prevented by the clamping rectifier MR_1 which restores the anode potential of V_2 to earth on every revolution of the store. Clamping one stage is also effective in restricting variations in the mean potentials of all the other stages. This circuit operates reliably up to at least 1 000 steps/sec.

Miscellaneous Diode Applications

Cold-cathode diodes can be used for clamping the potentials of various points in circuits, for example as shown in Fig. 31. When the triode is 'on', its cathode potential is high and the diode is conducting. The circuit thus provides a low output impedance.

Another useful application for diodes occurs when pulses for operating cold-cathode valves have to be transmitted over lengths of multi-conductor cable. Unless the conductors are screened, capacitance coupling between them causes considerable crosstalk because of the high impedance of the terminating circuits. Interference picked up on one lead due to pulses on other leads can be prevented from disturbing the circuits connected to it by means of a series diode as shown in Fig. 32. Provided that disturbing voltages are less than the striking voltage V_s, the valve will not fire and no voltage appears at the output. The correct signal voltage (V_k) will exceed V_s and so cause a voltage ($V_k - V_m$) to appear at the output.

Diodes can also be used, when required, for reducing or increasing the amplitudes of pulses used in cold-cathode valve circuits. A series diode, as shown in Fig. 32, can be used for reducing the size of pulses. Thus, if the input pulse is 110V and V_m is 60V, the output pulse height is 50V. A series diode is often preferable to a resistance potentiometer because of its much lower output impedance. The amplitude of pulses can be increased by using the circuit of Fig. 26(b), but with a diode having a large difference between V_s and V_m. If a fixed bias V_b, greater

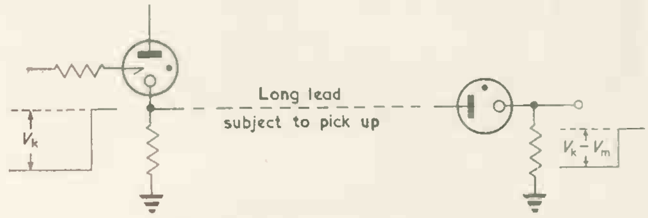


Fig. 32. Use of series limiting diode

than V_m, but less than V_s is applied to terminal B and a pulse is applied to terminal A, a pulse of larger amplitude can be obtained from the output terminal. Thus, if V_s = 180V, V_m = 80V, V_b = 150V and a 40V pulse is applied to the input terminal, then a pulse of height 110V will appear at the output terminal. Two methods are available for extinguishing the valve at the end of the pulse. The first method requires the time-constant C_cR_b to be small, so that the output voltage decays an appreciable amount during the pulse. This causes a negative overshoot at the end of the pulse, which reduces the anode voltage below V_b and extinguishes the valve. This method is only applicable when the pulse can sag sufficiently for its overshoot to make the p.d. across the valve less than V_m. When this cannot be done, as in the example given, another method is available. The second method is to shunt the load resistor with a capacitor as described in Part 1. The capacitor is charged by the pulse so that, when it ends, the cathode of the diode is at a positive voltage, the p.d. across the diode is less than V_m and the valve is extinguished. If the input pulses can be superimposed on a suitable d.c. potential, then the circuit can be directly coupled and R_b and C_c eliminated.

Multi-Cathode Valve Circuits

The circuits described so far have used simple diodes and triodes, but no survey of cold-cathode valve techniques would be complete without mentioning the use of multi-electrode cold-cathode counter valves[22,23,24,25,26]. These valves have a single central anode and a number of cathodes arranged in a ring with guide electrodes between each pair of cathodes. By applying suitable negative pulses to the guide electrodes, the glow can be transferred from each cathode to the next, thus producing a counting action. A counter circuit using a multi-cathode valve can operate faster than one using separate triodes. However, in order to operate a multi-cathode valve at high speed a thermionic valve must be used to drive it. When a cold-cathode valve is used for the driving circuit, the driving valve used limits

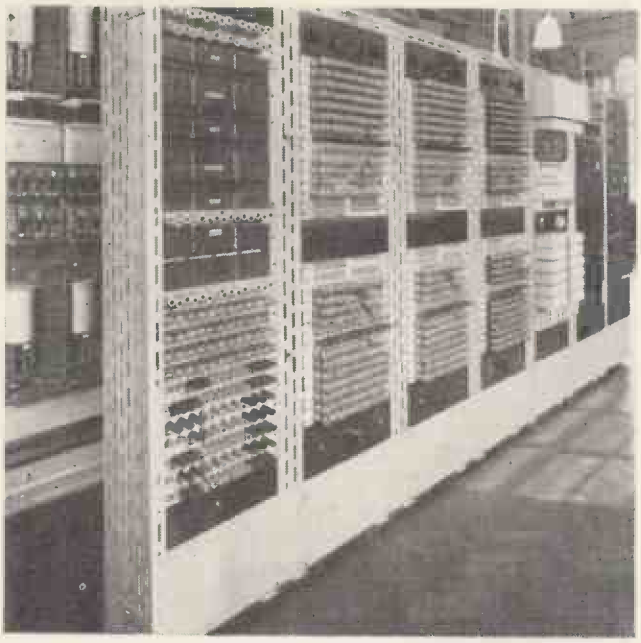


Fig. 33. Experimental directors in Richmond telephone exchange

both the speed of operation and the output voltage obtainable. The output voltage obtainable from the cathodes of the counter valve in theory cannot exceed the driving voltage applied to its guides, and in practice cannot approach this figure, so it is often difficult to obtain sufficient output voltage to operate cold-cathode valves reliably.

Multi-cathode valves have found their greatest application in straightforward scaling circuits[27,28]. However, they have been used successfully in circuits performing more complicated operations[29,30]. Because of the limitations described above, the multi-electrode valves appear to be more compatible with thermionic valve circuits than with cold-cathode circuits. For this reason, and because the fixed number of cathodes results in a certain inflexibility unsuited to sophisticated arrangement, most complicated cold-cathode valve digital circuits have been designed to use simple diodes and triodes.

Equipment Using Cold-Cathode Valves

Cold-cathode valves have been used in a number of simple circuit applications for about 20 years[31], but only during the past 10 years have equipments been designed using large numbers of them in complicated digital circuits. An obvious application for cold-cathode valve circuits is to common-control apparatus in telephone exchange systems. This type of equipment has to perform a large number of operations in a short time; when electro-mechanical switches are used they wear rapidly and require frequent maintenance adjustment. An example of such equipment is the director which is used in the telephone exchanges of some large cities like London and Manchester. The director translates the digits dialled by a telephone subscriber to represent an exchange name (e.g. CEN, BRI, etc.) into another set of digits which determine the route over which the call is to be set up. The British Post Office designed some experimental electronic directors[32] which were installed in Richmond (London) exchange in December 1951 which have been in full public service since May 1952.

The equipment installed at Richmond, which includes six electronic directors and their common translator, is shown in Fig. 33. The field trial has proved the reliability of cold-cathode valve circuits and has demonstrated the practicability of using this type of equipment to give a satisfactory 24-hour service with a lower fault-rate than on the best electro-mechanical equipment[33]. Many other applications of cold-cathode valves to telephony are being developed in various laboratories[17,34,35].

In addition to applications in telephony, cold-cathode valves have been used successfully for computing purposes[29] and in industrial equipment[27]. Another important application has been to telemetering for the control of electric power supply systems[36,37].

During recent years, improvements have been made in cold-cathode valves themselves, in the design of circuits using them and in suitable mechanical mounting methods for incorporating them in equipment. The last point is not unimportant, as can be seen by comparing Fig. 34 with Fig. 33. The directors shown in Fig. 33 are, of course, experimental models, but they use large (octal base) cold-cathode valves, about 300 of which are mounted on each rack. Fig. 34 shows a recently designed telephone operator's keysender which uses subminiature cold-cathode valves. Over 100 valves, in addition to two relays, are mounted on a telephone-type mounting plate of the size which normally accommodates 24 relays.

Acknowledgments

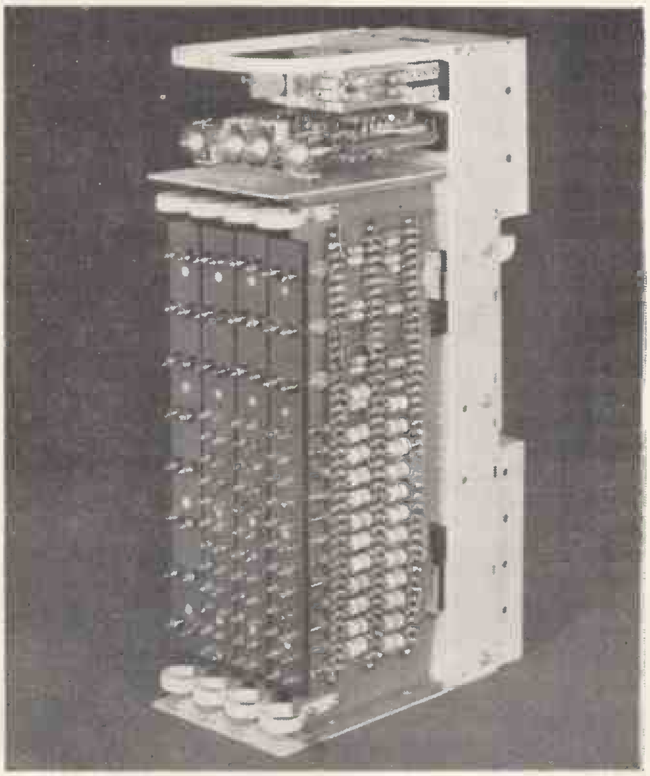
Acknowledgment is made to Siemens Brothers & Co. Limited for permission to publish this article. Fig. 33 is included by kind permission of the Engineer-in-Chief of the G.P.O.

Appendix

Pulse Supplies

The use of pulsed-anode supplies for extinguishing cold-cathode valves and for co-ordinating the operations of different stages in complicated digital circuits was discussed in Part 1. Where several pulse supplies are required it is usually convenient to derive them from a common master

Fig. 34. Telephone operator's keysender using sub-miniature cold-cathode valves mounted on a 23-way relay plate



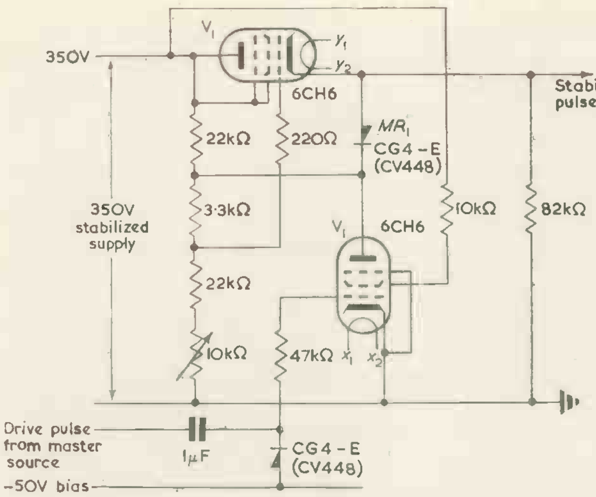


Fig. 35. Pulse supply circuit using a series stabilizing valve

pulse generator, such as a relaxation oscillator driving a cyclic counter whose stages deliver pulses at the appropriate times, which drives pulse supply circuits arranged to deliver the pulses at the required output voltage and power. Since the pulses constitute the anode supply voltage for cold-cathode valves, their voltages must be adequately controlled and it is usual for the pulse supply circuits to take the form of series-or shunt-stabilizer power supplies which are switched on and off by the master pulse generator.

Fig. 35 shows a pulse supply circuit which uses a series stabilizing valve V_1. This valve derives its reference potential from the d.c. power supply via a potential divider whose ratio can be adjusted to give the required output voltage. When a positive input pulse causes valve V_2 to conduct, the reference potential is reduced below the bottoming potential of V_2. The output voltage from the cathode of V_1 is thus less than the maintain potential of the cold-cathode valves it supplies and these are extinguished. But for the rectifier MR_1, the stray capacitance of the load would prevent the cathode of V_1 falling immediately the reference potential was removed, with the result that the valve would be cut off and the output voltage decay only slowly. However, when valve V_2 is turned on, its fall in anode potential causes MR_1 to conduct and this rapidly discharges the stray capacitance.

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Remote Control of Oil Pumps

A remote control device, operated from a tanker moored at the end of submarine cargo loading lines $3\frac{1}{2}$ miles out at sea, which is capable of stopping the cargo loading pumps on shore, has recently been installed at a cost of £10 000 at Lutong Refinery in Sarawak, British Borneo. This equipment, now in service, was developed jointly by the Shell Petroleum Company Limited and the General Electric Company Limited.

The shallowness of the water prevents tankers from coming close to the shore, and the four loading moorings are situated $3\frac{1}{2}$ miles out. They are connected to the shore by underwater pipelines. As a tanker approaches the moorings, one of the seven available portable transmitter/receiver units is taken aboard. In addition to providing a radiotelephone service with the shore terminal, this portable set can transmit on a pump control channel.

Loudhailing facilities are incorporated and a remote control unit enables the equipment to be operated up to a distance of 300ft from the main instrument. The radio telephone channel transmits on 80Mc/s and the pump control on 80·2Mc/s. The radio telephone shore station transmits on 86Mc/s and selective calling enables this channel to be operated on the party line principle.

To stop the pumps in an emergency, the pressing of a control button on the operating panel of the portable set overrides the telephone channel and brings the pump control channel into operation. On reaching the shore pumphouse the signal is fed to the particular engines pumping to the tanker on which the button was pressed. Electrically operated valves are then activated and stop the engines.

The signal transmitted on the pump control channel consists of one of four tones, each mooring being allocated a specific tone, which is selected by a four-position switch on the portable unit. The receiving equipment, which is located near the shore pump house is therefore able to determine from which mooring the emergency signal has been received. A signal is then passed to the pumphouse control panel, where a system of selector switches and indicator lights feeds it to the particular engines pumping to the tanker on which the button was pressed. The control panel also increases the power of the signal to enable it to work the solenoid-operated valves, which, mounted one on each engine, are the actual means of stopping the pumps.

The pumphouse receiving equipment comprises two complete receivers working simultaneously so that in the event of one developing a fault an emergency signal can be handled by the other, and at the same time a warning light indicates which receiver is defective.

The control panel, built by Pool & Partners of Marple, Cheshire, to Shell specification, enables the various pumps and moorings to be correlated. It also provides a means of testing the equipment and a light-and-horn indication to the pumpman that an emergency stoppage has occurred.

The solenoid-operated control valves were supplied by Electro-Hydraulics Limited of Warrington, and are the first valves of this type and size to be made intrinsically safe.

Efficiency-Diode Line Scanning Circuits

A Simplified Design Method

By K. G. Beauchamp*, A.M.Brit.I.R.E.

A rapid method of circuit and transformer design is given in which a series of 'abacs' is used to evaluate the design constants.

The method also facilitates the prediction of new design figures, should any of the initial constants be altered, and the effects on performance obtained.

THE theory of efficiency-diode scanning circuits has been previously described in some detail in several excellent papers, notably those of Friend[1], Schade[2] and Jones[3]. The latter gives a very useful design procedure applicable to modern wide-angle scanning circuits and the following treatment is based largely on this.

The design of an efficiency-diode circuit is often commenced on the basis of several known constants, viz: h.t. supply potential V_{ht} (volts), cathode-ray tube accelerating potential E (volts), type and size of tube chosen etc., all of which form part of the overall design specification for the television receiver.

It is the object of this article to show how, with the aid of a series of abacs, the design commencing from these initial design constants, may be facilitated and the effect of any changes in design constants on circuit performance quickly predicted.

Scanning Coil Sensitivity

A suitable commencing point for the design lies in consideration of the scanning coil chosen. For a given scanning angle, θ (degrees) and e.h.t., E (volts), the peak-to-peak scanning current required I_y (amps) is given approximately by the relationship:

$$\sin\theta = \frac{I_y \sqrt{L_y}\,.\,l}{kd\sqrt{E}} \quad \ldots\ldots\ldots\ldots\ldots \text{(1)}$$

where

k = a constant

d = mean diameter of scanning coils (similar units)

l = length of coils, measured along an axis parallel to the c.r.t. beam (similar units)

L_y = scanning coil inductance (henries).

For a given ampere-turns in the coil the sensitivity is governed largely by the length of the scanning coil l. This in turn is dependent on the scanning angle θ, due to the necessity of avoiding neck-shadowing or cut-off of the c.r.t. beam.

It can be shown from geometrical considerations that the maximum length of coil l possible in order to avoid this is:

$$l_{max} = D/\tan(\theta/2) \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(2)}$$

where D = internal diameter of the c.r.t. neck at the Z-Z line, i.e. at the point of junction between the c.r.t. neck and bulb.

For $\theta = 65°$, and D = 36cm, (the conditions met with in most modern 14in and 17in tubes) l_{max} = 51·5mm.

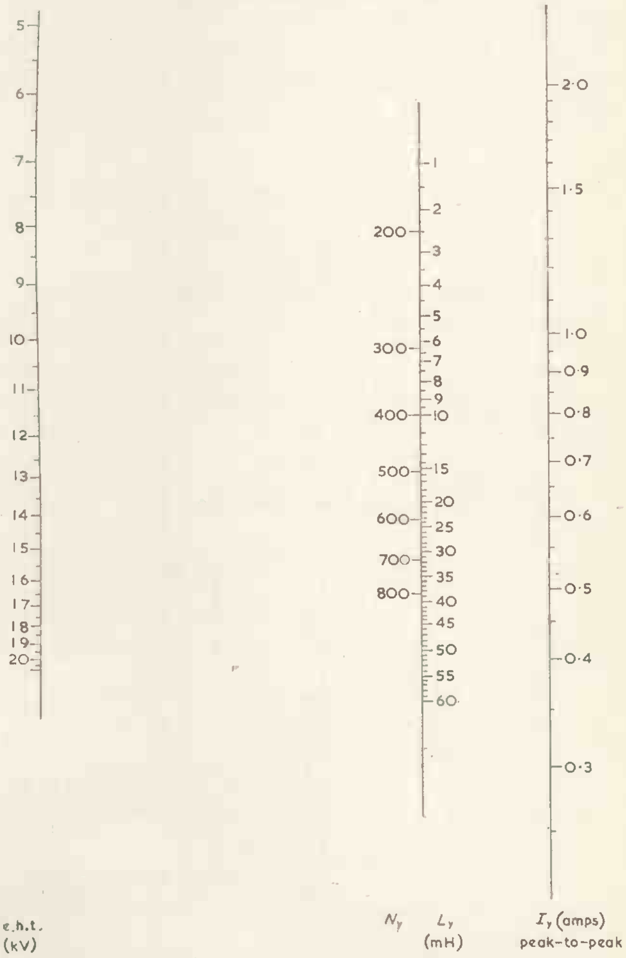
Some increase on this figure is possible if the ends of the coils are flared to fit further up the tube neck, and in the design data that follows a figure of 53mm is assumed for l_{max} in order to take this effect into account.

The assumption is also made that saddle-type scanning coils are used. Where castellated cores are used the current figures arrived at may be reduced by some 5 per cent to allow for the greater sensitivity obtained with this form of construction.

The abac of Fig. 1 has been prepared from empirical data and shows the relationship between scanning coil inductance L_y (or total turns N_y), the peak-to-peak scanning current I_y, and the e.h.t. E, for the conditions described above.

This abac can be used for either series or parallel connexion of the two line coils, but in the latter case the current I_y refers to the total current supplied to the two coils in parallel, the inductance figure also referring to the parallel connexion.

Fig. 1. Relationship between scanning coil inductance, scanning current and e.h.t. applied to the c.r.t.



* *A.E.I. Research Laboratories, Aldermaston, formerly with the G.E.C. Television Laboratory, Coventry.*

Transformer Design

It has been shown that the performance of the line-scanning transformer is largely controlled by the coupling factor k obtained between scanning coil winding $L_{1\text{-}2}$ and pentode winding $L_{1\text{-}4}$ (Fig. 2).

Jones[3] has shown that for maximum energy transfer the relationship between L_y and $L_{1\text{-}2}$ is dependent on k as given by:

$$L_{1\text{-}2}/L_y = \frac{k^2}{\sqrt{(1-k^2)}} \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

However, in order to avoid leakage reactance resonances in the completed design, it is advisable to depart from the optimum value of $L_{1\text{-}2}$ and this may be reduced by a factor α, thus:

$$\alpha = \frac{L_{1\text{-}2}\,(\text{actual})}{L_{1\text{-}2}\,(\text{optimum})} \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

Using modern core materials little improvement in efficiency is obtained by making α greater than about 0·8.

Fig. 3 incorporates the relationships given in equations (3) and (4) and enables $L_{1\text{-}2}$ to be evaluated for a given value of L_y, k and α.

The peak voltage developed across winding $L_{1\text{-}2}$ during the scanning period is given as:

$$V_y = L_y \,.\, (dI_y/dt) \text{ volts} \quad \ldots\ldots\ldots\ldots \quad (5)$$

where t = scanning period (sec).

Due to the inclusion of coils for controlling picture width, a correction factor p will be required to the volt-amperes V_yI_y developed across the transformer secondary winding.

A figure of 12 per cent may be taken as adequate to allow for valve deterioration in service[4].

Also, as $L_{1\text{-}2}$ has been reduced from its optimum value, an increase in energy is required to maintain the required potential across the scanning coils. This can also be assessed as a correction to the VA required, giving:

$$\text{VA} = p\eta\, I_yV_y \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (6)$$

for the actual volt-amperes supplied to the system, where:

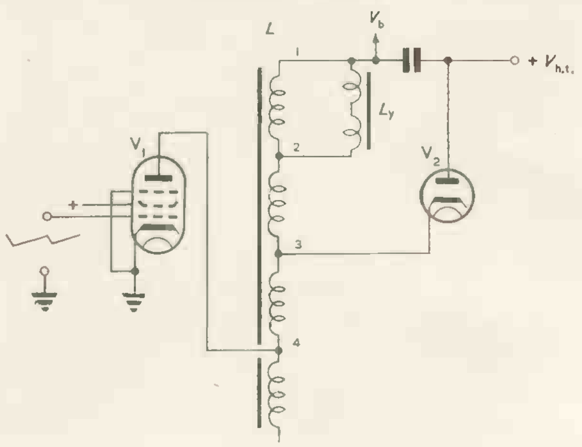
$$\eta = \frac{\text{VA supplied to the system}}{\text{VA in } L_y} \quad \ldots\ldots\ldots\ldots \quad (7)$$

An expression for η has been previously given[3] as:

$$\eta = \frac{2-k^2}{k^2} + \left[\frac{(1+\alpha)^2}{\alpha}\right] \cdot \frac{\sqrt{(1-k^2)}}{k^2} \quad \ldots\ldots \quad (8)$$

This is of the form $\eta = a + cb$, where a and b are functions of k and c is a function of α.

Fig. 2. Simplified circuit of an efficiency-diode scanning circuit, showing transformer connexions



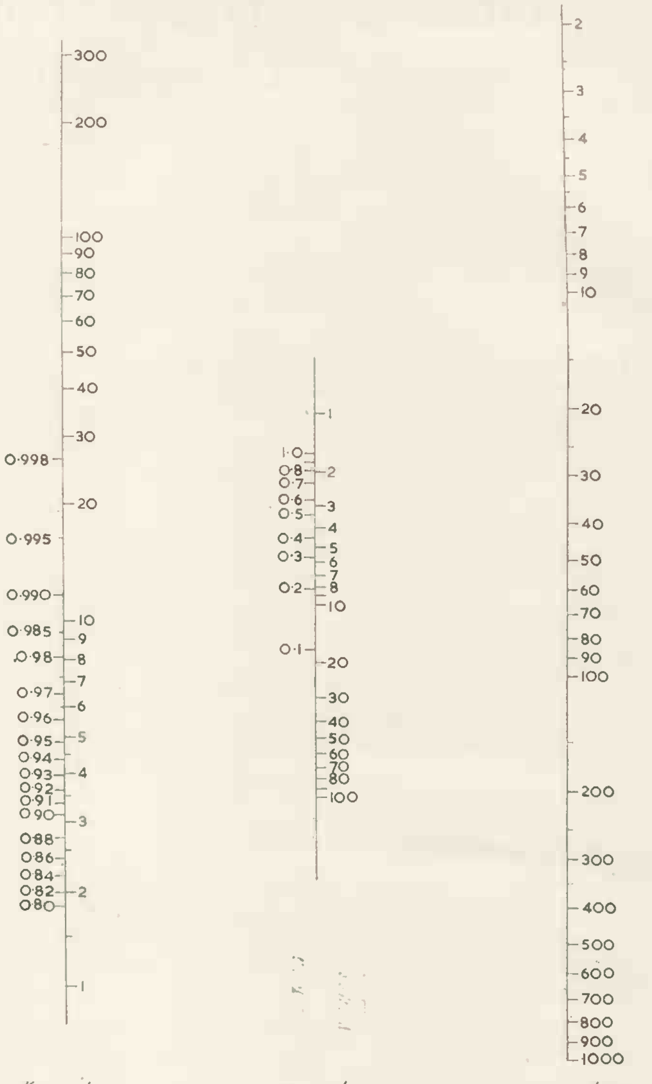


Fig. 3. Relationship between scanning coil inductance and optimum transformer secondary inductance

The expression is not directly amenable to abac form, but by separating the two constituents of η, this factor may be determined by the addition of a and cb which may be found from the abac of Fig. 4 in terms of k and α.

Making certain assumptions regarding boosted h.t. potential V_b and minimum potential for V_1 anode, $V_{a(min)}$, then the peak current in winding $L_{1\text{-}4}$ is given by:

$$I_p = \frac{\text{VA}}{V_b - V_{a(min)}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

With resonant-return systems V_1 and V_2 conduct alternately and the division of this peak current between the two valves is determined by the Q-factor of the coupling transformer. In practice, the current taken by the pentode V_1 usually lies between 0·5 and 0·6 of I_p.

Assuming the waveform of V_1 anode current to be that of a truncated sawtooth, the r.m.s. value is given by the relationship:

$$I_{a(r.m.s.)} = I_{a(peak)} \,.\, \sqrt{(d/3T)} \quad \ldots\ldots\ldots \quad (10)$$

where T = period of one cycle

d = period of pentode conduction.

For the case of $d = 0{\cdot}55T$, the r.m.s. current taken by V_1 will be approximately 0·43 of the peak value.

The transformer ratios can now be evaluated. These are $N_{1\cdot3}/N_{1\cdot2}$, matching scanning coils to the boosting diode impedance and $N_{1\cdot4}/N_{1\cdot3}$, effecting a match between the two valves.

The value of these will be dependent largely on the mode of operation desired, but for most purposes a mode will be required where the diode does not quite cut off at the end of the scanning period, while the pentode is operated below the 'knee' of its I_a/V_a characteristic.

Thus the root of the diode can be chosen to be immediately below the peak pentode current $I_{a(p)}$. The anode potential at this point can be ascertained from the valve characteristics and designated $V_{a(min)}$.

Then:

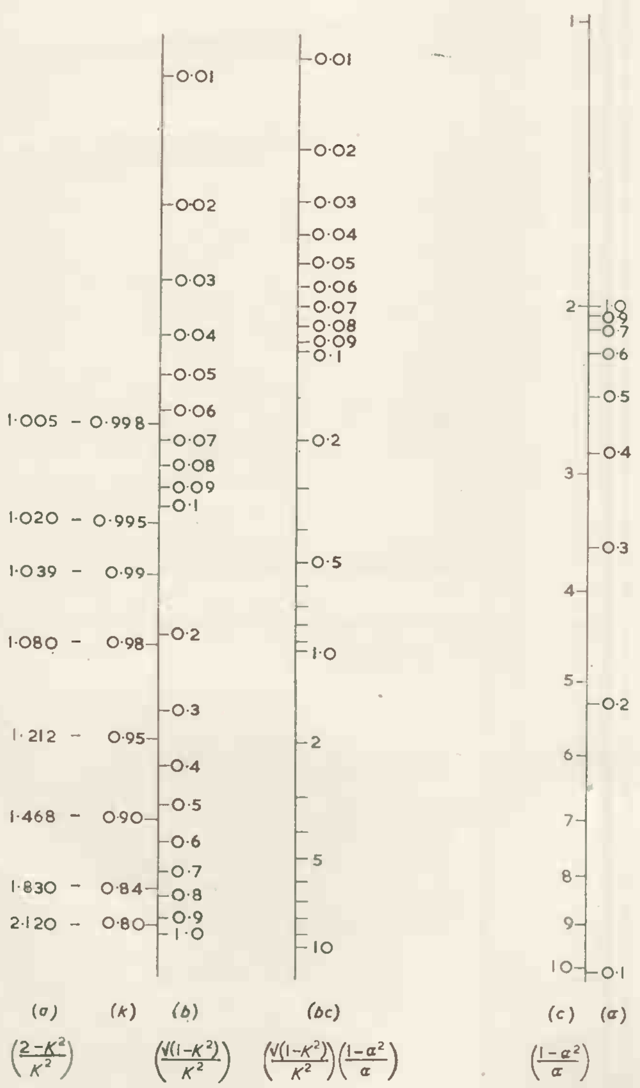
$$N_{1\cdot4}/N_{1\cdot3} = \frac{V_b - V_{a(min)}}{V_b - V_{ht}} \quad \text{.......... (11)}$$

and the diode ratio:

$$N_{1\cdot3}/N_{1\cdot2} = \frac{V_b - V_{ht}}{V_y} \quad \text{.......... (12)}$$

Realization of the second assumption mentioned above,

Fig. 4. Derivation of VA correction factor η in terms of coupling factor k and α



i.e. boost potential, is dependent on the mode of operation (controlled by $N_{1\cdot4}/N_{1\cdot3}$) efficiency of the circuit (Q-value), and any current drain from the boosted h.t. supply to other parts of the receiver (e.g., frame scanning circuit). Previous design experience is desirable here in order to make an intelligent estimate of expected boost potential.

An example will now be given, to illustrate the application of the method outlined above to a typical scanning circuit design.

Let the following design constants be applicable:

$$\theta = 65°, \; L_y = 10\text{mH}, \; E = 13\text{kV},$$
$$V_b = 450\text{V}, \; V_{ht} = 200\text{V}, \; V_{a(min)} = 60\text{V}.$$

These are typical of those found in, say, a 17in a.c./d.c. domestic television receiver.

(a) Scanning Coil Design

From Fig. 1, $N_y = 400$ turns, i.e. 200 turns per coil, using a ferrite yoke, series connexion and saddle-type construction.

Also from Fig. 1, $I_y = 0{\cdot}85$A peak-to-peak.

A suitable size of wire at a current density of 3 000A/in² is 0·0148in diameter (28 s.w.g.)

(d) Transformer Design and Operating Conditions

k can be obtained from a test transformer winding or previous known designs using the same core and method of construction.

Let $k = 0{\cdot}998$, then from Fig. 3, $L_{1\cdot2(optimum)} = 156$mH.

At these values of coupling factor there is very little loss of efficiency using low values of α, so let $\alpha = 0{\cdot}5$, giving $L_{1\cdot2} = 78$mH.

From equation (5), $V_y = 86$V pk-pk for $L_y = 10$mH and $I_y = 0{\cdot}85$A.

With $\alpha = 0{\cdot}5$ and $k = 0{\cdot}998$,

$$\eta = 1{\cdot}005 + 0{\cdot}17 = 1{\cdot}175 \text{ (from Fig. 4).}$$

The VA required are:

$$\text{VA} = 1{\cdot}175 \times 1{\cdot}12 \times 0{\cdot}85 \times 86 = 106, \text{ from equation (6).}$$

Taking $V_{a(min)}$ as 60V from the valve characteristics, then:

$$I_p = (106/390) = 272\text{mA}$$

Thus:

$$I_a = 0{\cdot}55 \times 236 = 150\text{mA}$$

the r.m.s. value being 150 × 0·43 = 64mA.

Transformer ratios are obtained from equations (11) and (12).

$$N_{1\cdot4}/N_{1\cdot3} = \frac{450 - 60}{450 - 200} = 1{\cdot}56$$

$$N_{1\cdot3}/N_{1\cdot2} = \frac{450 - 200}{86} = 2{\cdot}9$$

Finally, the transformer is wound with taps on either side of those calculated and the turns ratio adjusted to obtain optimum working conditions.

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Resistance-Capacitance Tuned Amplifiers Using Negative Feedback

By D. J. O'Connor*, B.Sc.

The transfer function of an LC tuned amplifier is compared with that obtained for a feedback amplifier using particular types of RC feedback networks. A design procedure is developed which permits the design of circuits having predicted performance.

A CONVENIENT starting point in the design of tuned amplifiers is the response of a single valve amplifier having a parallel *LCR* network as its anode load. Fig. 1 shows the current generator equivalent of such an amplifier for which the gain function is

$$G = -\frac{pLg_m}{1 + p(L/R) + p^2LC} \qquad (1)$$

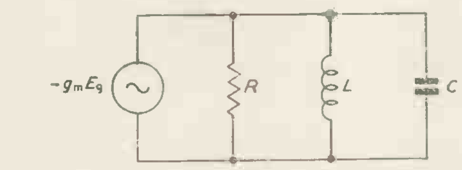


Fig. 1. Current generator equivalent circuit of LC tuned amplifier

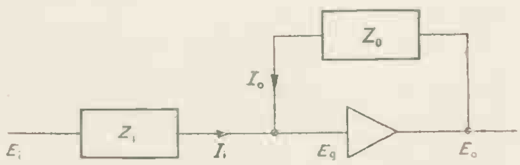


Fig. 2. Block diagram of feedback amplifier

We can identify three parameters for the function

$$\left.\begin{aligned} &\text{Peak frequency} & \omega_0 &= [1/\sqrt{(LC)}] \\ &\text{Peak gain} & G_0 &= g_mR \\ & & Q_0 &= (R/\omega_0L) \end{aligned}\right\} \qquad (2)$$

Equation (1) is of the form

$$G = -\frac{pc}{1 + pa + p^2b^2} \qquad (3)$$

and one may write

$$\left.\begin{aligned} &\text{Peak frequency} & \omega_0 &= (1/b) \\ &\text{Peak gain} & G_0 &= (c/a) \\ & & Q_0 &= (b/a) \end{aligned}\right\} \qquad (4)$$

In order to see how a function of the general form (3) may be generated, consider the circuit shown in Fig. 2. At this stage loading effects of the input and output networks on each other and on the amplifier will be ignored. If the gain of the amplifier is large and there is a large degree of feedback then the voltage E_g at the input grid is small compared with both E_i and E_o and in fact tends to zero. Further, if the amplifier input impedance is high the current to the input grid tends to zero. It follows that $I_i = -I_o$ and therefore

$$(E_o/E_i) = (Z_oI_o/Z_iI_i) = -(Z_o/Z_i) \qquad (5)$$

where Z_i and Z_o are the transfer impedance functions of the input and output networks as shown in Fig. 2. The transfer impedance function of a network is defined as the ratio of the input voltage to the output short-circuit current. Fig. 3 gives the transfer impedance functions of the networks which are considered in this article. In all cases the network of Fig. 3(a) is used as the input network.

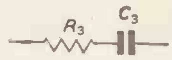
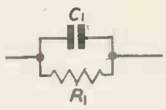
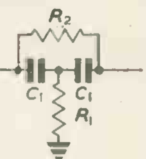
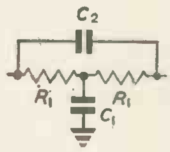
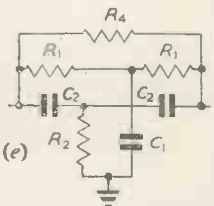
NETWORK	TRANSFER IMPEDANCE FUNCTION	RELATIONS
(*a*) 	$(1/pB)\,(1+pT_3)$	$B=C_3$ $T_3=R_3C_3$
(*b*) 	$\frac{A}{1+pT_1}$	$A=R_1$ $T_1=R_1C_1$
(*c*) 	$\frac{A(1+pT_1)}{1+pT_1+p^2T_1T_2}$	$A=R_2$ $T_1=2R_1C_1$ $T_2=\frac{R_2C_1}{2}$
(*d*) 	$\frac{A(1+pT_2)}{1+pT_1+p^2T_1T_2}$	$A=2R_1$ $T_1=2R_1C_2$ $T_2=\frac{R_1C_1}{2}$
(*e*) 	$\frac{\frac{2R_1R_4}{2R_1+R_4}\,(1+PT_1)}{1+\frac{2R_1T_1}{p^2R_1+R_4}+\frac{2R_4T_1T_2}{p^2R_1+R_4}}$	$T_1=\frac{R_1C_1}{2}$ $=2R_2C_2$ $T_2=R_1C_2$

Fig. 3. Transfer impedance functions of networks used

Case 1. Fig. 3(b) Used as Output Network

$$Z_i = (1/pB)\,(1 + pT_3) \qquad (6)$$

$$Z_o = \frac{A}{1 + pT_1} \qquad (7)$$

$$G = -(Z_o/Z_i) = \frac{-pAB}{1 + p(T_1 + T_3) + p^2T_1T_3} \qquad (8)$$

For this network combination

$$Q_0 = \frac{\sqrt{(T_1T_3)}}{T_1 + T_3} \qquad (9)$$

which has a maximum value of 0·5 when $T_1 = T_3$.

Case 2. Fig. 3(c) Used as Output Network

$$Z_i = (1/pB)\,(1 + pT_3)$$

$$Z_o = \frac{A(1 + pT_1)}{1 + pT_1 + p^2T_1T_2} \qquad (10)$$

* *The General Electric Company Limited, England.*

$$G = -\frac{pAB(1 + pT_1)}{(1 + pT_3)(1 + pT_1 + p^2T_1T_2)} \quad \ldots\ldots (11)$$

If $T_1 = T_3$, then

$$G = -\frac{pAB}{1 + pT_1 + p^2T_1T_2} \quad \ldots\ldots (12)$$

This is of the form of (2) so we may write

$$\omega_o^2 = (1/T_1T_2) = \frac{1}{R_1R_2C_1^2} \quad \ldots\ldots (13)$$

$$G_o = \frac{R_2C_3}{2R_1C_1} = (R_2/R_3) \quad \ldots\ldots (14)$$

$$Q_o = \sqrt{(T_2/T_1)} = \tfrac{1}{2}\sqrt{(R_2/R_1)} \quad \ldots\ldots (15)$$

Suppose now the value of R_2 is fixed. Then R_1 will be fixed by the required Q_o and R_3 will be fixed by the required G_o. C_1 can be selected to give the required value of ω_o and C_3 is selected so that the relation $T_1 = T_3$ is satisfied. Thus components can be selected to give any desired values of ω_o, G_o and Q_o.

Case 3. Fig. 3(d) Used as Output Network

$$Z_i = (1/pB)(1 + pT_3)$$

$$Z_o = \frac{A(1 + pT_2)}{1 + pT_1 + p^2T_1T_2} \quad \ldots\ldots (16)$$

It will be found that provided $T_2 = T_3$, then

$$\omega_o^2 = \frac{1}{R_1^2C_1C_2} \quad \ldots\ldots (17)$$

$$G_o = (C_3/C_2) \quad \ldots\ldots (18)$$

$$Q_o = \tfrac{1}{2}\sqrt{(C_1/C_2)} \quad \ldots\ldots (19)$$

Again, components can be chosen to give any desired values of ω_o, G_o and Q_o.

Case 4. Fig. 3(e) Used as Output Network

$$Z_i = (1/pB)(1 + pT_3)$$

$$Z_o = \frac{\dfrac{2R_1R_4}{2R_1 + R_4}(1 + pT_1)}{1 + p\dfrac{2R_1T_1}{2R_1 + R_4} + p^2\dfrac{R_4T_1T_2}{2R_1 + R_4}} \quad \ldots\ldots (20)$$

Making $T_1 = T_3$, it will be found that

$$\omega_o^2 = (1/T_1T_2) \cdot \frac{2R_1 + R_4}{R_4} \quad \ldots\ldots (21)$$

$$G_o = (R_4/R_3) \quad \ldots\ldots (22)$$

$$Q_o = \tfrac{1}{2}\sqrt{\left(\frac{R_4}{2R_1R_2}(2R_1 + R_4)\right)} \quad \ldots\ldots (23)$$

Again components can be selected to give any desired value of these parameters.

It appears that in three of the cases considered components can be selected to give any desired response. In practice some of the assumptions that have been made may not be valid. As a first departure from the previous assumptions consider the case when the gain M of the amplifier is not large compared with the required overall gain. E_g will be regarded as small in comparison with E_o, but comparable with E_i. Then from Fig. 2 (since Z_i is a two-terminal network),

$$E_i = E_g + Z_iI_i \quad \ldots\ldots (24)$$

$$E_o = ME_g = Z_oI_o = -Z_oI_i \quad \ldots\ldots (25)$$

$$\therefore E_i = -(E_o/M) + Z_iI_i = -(E_o/M) - (E_o/Z_o) = -E_o[(1/M) + (Z_i/Z_o)] \quad \ldots (26)$$

$$\therefore G' = (E_o/E_i) = -\left(\frac{M}{1 + (Z_i/Z_o)M}\right) \quad \ldots\ldots (27)$$

For the networks considered, Z_i/Z_o has the form

$$\frac{1 + pa + p^2b^2}{pc}$$

$$\therefore G' = -\frac{pcM}{M + p(c + aM) + p^2Mb^2} = -\frac{pc}{1 + p[(c/M) + a] + p^2b^2} \quad \ldots\ldots (28)$$

which is of the same general form as equation (3). The peak frequency is unchanged, but now

$$G_o' = \frac{M}{1 + (a/c)M} = \frac{G_oM}{G_o + M} \quad \ldots\ldots (29)$$

$$Q_o' = \frac{b}{(c/M) + a} = \frac{(b/a)M}{(c/a) + M} = \frac{Q_oM}{G_o + M} \quad \ldots\ldots (30)$$

In equations (29) and (30), G_o and Q_o indicate the functions of resistance and capacitance which have been derived earlier for the various network configurations. They are retained as convenient design parameters.

The analysis becomes a little more complex when it is desired to take into account the loading effect of the networks on each other. It will now be assumed only that the amplifier has high input impedance, low output impedance and $M \gg 1$. 'High' and 'low' are, of course, relative terms and must be considered in conjunction with the network impedances. It is shown in Appendix 1 that under these conditions the gain function of the amplifier is

$$G' = \frac{-M}{1 + (Z_i/Z_s) + (Z_i/Z_o)M} \quad \ldots\ldots (31)$$

where Z_s is the impedance of each shunt arm of a π-network, equivalent to the output network. It is also shown in the Appendix that for output networks of the forms of Fig. 3(c), (d) and (e), the term Z_i/Z_s is independent of frequency. One can therefore rewrite equation (27) as

$$G' = \frac{-M'}{1 + (Z_i/Z_o)M'} \quad \ldots\ldots (32)$$

where $M' = \dfrac{M}{1 + (Z_i/Z_s)}$

It has been seen already that this form of gain function is of the same general form as equation (3) and it may be noted that the peak frequency is unchanged.

Using the values of Z_i/Z_s derived in the Appendix it may easily be verified that using either of the bridged-T networks the peak gain is given by

$$G_o' = \frac{G_oM}{G_o + 2Q_o^2 + M} \quad \ldots\ldots (33)$$

and Q is given by

$$Q_o' = \frac{Q_oM}{G_o + 2Q_o^2 + M} \quad \ldots\ldots (34)$$

Considering equation (34) it is seen that Q_o' falls steadily as G_o is increased, but has a maximum value as Q_o is varied. This maximum occurs when $2Q_o^2 = G_o + M$ and is

$$Q_{o'(\max)} = \frac{M}{2\sqrt{(G_o + M)}} \quad \ldots\ldots (35)$$

The corresponding value of peak gain is

$$G_o' = \frac{G_oM}{2(G_o + M)} \quad \ldots\ldots (36)$$

Eliminating G_o between equations (35) and (36),

$$Q_o'_{(\max)} = \tfrac{1}{2}\sqrt{\left(\frac{M - G_o'}{2}\right)} \qquad (37)$$

Thus, using the bridged-T output networks for a given value of M, if G_o' is fixed then there is a maximum value of Q' that can be attained.

When the twin-T output network is used it is found, using the value of Z_i/Z_s derived in the Appendix, that the peak gain is given by

$$G_o' = \frac{G_o M}{G_o + \frac{2(n+1)}{\sqrt{(n)}} Q_o + M} \qquad (38)$$

and Q is given by

$$Q_o' = \frac{Q_o M}{G_o + \frac{2(n+1)}{\sqrt{(n)}} Q_o + M} \qquad (39)$$

It is seen that Q_o' decreases as G_o is increased, but as Q_o increases $Q_o' \rightarrow \frac{M\sqrt{n}}{2(n+1)}$. This has a maximum value for $n = 1$, so that for this network

$$Q_o'_{(\max)} = (M/4) \qquad (40)$$

Design Procedure

In designing an amplifier, peak frequency, peak gain and Q will be specified and the value of M for the amplifier to be used will be known. Since none of the circuits considered can give $Q' > (M/4)$ it can be seen immediately if the amplifier is capable of meeting the requirements on this parameter. From equation (37), it can be ascertained if the specified value of Q_o' can be achieved with a bridged-T network. If a bridged-T network is to be used, then eliminating G_o between equations (33) and (34),

$$2Q_o^2 - \left(\frac{M - G_o'}{Q_o'}\right) Q_o + M = 0 \qquad (41)$$

$$Q_o = \frac{M - G_o' \pm \sqrt{[(M - G_o')^2 - 8MQ_o'^2]}}{4Q_o^2} \qquad (42)$$

Also from equations (33) and (34),

$$G_o = (G_o'/Q_o') \cdot Q_o \qquad (43)$$

The network components can now be specified to give these values of G_o and Q_o.

If the twin-T network is to be used, then putting $n = 1$ and solving for G_o and Q_o between equation (38) and (39),

$$G_o = \frac{MG'_o}{M - G_o' - 4Q_o'} \qquad (44)$$

$$Q_o = Q_o'(G_o/G_o') = \frac{MQ_o'}{M - G_o' - 4Q_o'} \qquad (45)$$

and again the network components can be specified to give these values of G_o and Q_o.

It is not possible to solve explicity for the value of each component until one component has been specified arbitrarily. The designer will endeavour to specify this component so that the conditions assumed in the analysis are satisfied.

Example

It is decided to design an amplifier having a Q of 8·25, a peak gain of 40 and a peak frequency of 100c/s. The amplifier to be used has a gain of 630 and an output resistance of about 1kΩ. The low output resistance makes it comparatively easy to ensure that the output network does not load the amplifier. From equation (40) it can be seen that the specified Q can be obtaintd with a twin-T circuit and from equation (37) it is seen that a bridged-T network can be used. The latter should be used, since it is economical in components.

From equation (42),

$$Q_o = \frac{590 - \sqrt{(590^2 - 8 \times 630 \times 68)}}{33}$$

$$= 15{\cdot}6$$

$$G_o = 40 \times (15{\cdot}6/8{\cdot}25) = 76.$$

Taking the circuit of Fig. 3(c) as the output network to be used, then, from equation (15),

$$(R_2/R_1) \simeq 1\,000.$$

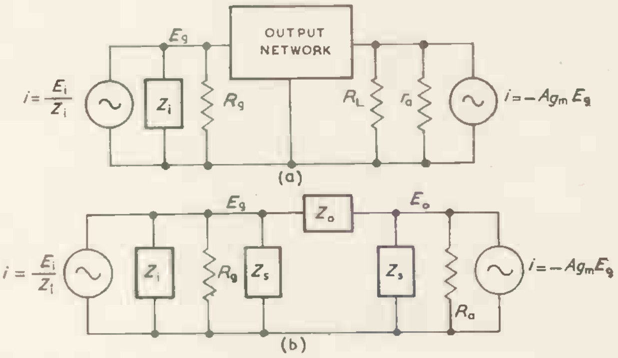


Fig. 4. Current generator equivalent circuit of feedback amplifier

At this stage one of the components can be specified, say R_1. Putting $R_1 = 1\text{k}\Omega$, then $R_2 = 1\text{M}\Omega$.

From equation (14), $R_3 = (10^6/76) = 13{\cdot}2\text{k}\Omega$

From equation (13), $C_1 = \frac{1}{2\pi \cdot 100\ \sqrt{(10^9)}} = 0{\cdot}05\mu\text{F}$

The only component left to specify is the capacitor of the input network which is specified by the relationship

$$T_1 = T_3$$

i.e., $R_3C_3 = 2R_1C_1$

$$C_3 = \frac{2{\cdot}10^3 \times 5 \times 10^{-8}}{1{\cdot}32 \cdot 10^4} = {\cdot}0076\mu\text{F}.$$

In a practical amplifier based on these calculations the actual component values used were:

$$R_1 = 980\Omega,\ R_2 = 980\text{k}\Omega,\ C_1 = 0{\cdot}052\mu\text{F},\ R_3 = 13\text{k}\Omega,$$
$$C_3 = {\cdot}0077\mu\text{F}.$$

These components should give a peak frequency 99c/s, $Q_o' = 8{\cdot}25$, $G_o' = 39{\cdot}5$. Measured values were $f_o = 98$c/s, $Q_o' = 8{\cdot}15$, $G_o' = 38{\cdot}2$.

Appendix

If a multi-stage amplifier is considered, in which the number of stages is such that there is a phase reversal between input grid and output terminals, an equivalent circuit can be drawn in the form of Fig. 4(a). The gain of the amplifier up to the grid of the output valve is A; the terms g_m and r_a refer to the output valve. The three terminal output network can be transformed to an equivalent π-network and the equivalent circuit can be redrawn as in Fig. 4(b), where Z_o is the transfer impedance of the output network. Since all the output networks considered are symmetrical, the two shunt arms of the equivalent π-networks are equal. R_a is the parallel combination of load resistance R_L and anode resistance r_a of the output valve. Putting $Y_i = (1/Z_i)$, $Y_g = (1/R_g)$, etc.,

$$E_g(Y_i+Y_g+Y_s+Y_o) - E_oY_o = E_iY_i \quad \ldots\ldots\ldots\ldots (46)$$

$$-E_gY_o \qquad + E_o(Y_s+Y_o+Y_a) = -Ag_mE_g \quad \ldots\ldots\ldots\ldots (47a)$$

i.e. $$E_g(Ag_m - Y_o) \qquad + E_o(Y_s+Y_o+Y_a) = 0 \quad \ldots (47b)$$

Then

$$E_o = \frac{\begin{vmatrix} Y_i+Y_g+Y_s+Y_o, & E_iY_i \\ Ag_m - Y_o & , \quad 0 \end{vmatrix}}{\begin{vmatrix} Y_i+Y_g+Y_s+Y_o, & -Y_o \\ Ag_m - Y_o & , \quad Y_s+Y_o+Y_a \end{vmatrix}}$$

$$= \frac{E_iY_i(Y_o - Ag_m)}{(Y_i+Y_g+Y_s+Y_o)(Y_s+Y_o+Y_a)+Y_o(Ag_m-Y_o)} \quad \ldots\ldots\ldots\ldots (48)$$

$$G' = \frac{E_o}{E_i} = \frac{Y_i(Y_o - Ag_m)}{(Y_i+Y_g+Y_s+Y_o)(Y_s+Y_o+Y_a)+Y_o(Ag_m-Y_o)} \quad \ldots\ldots\ldots\ldots (49)$$

This equation gives the complete gain function of the amplifier. Assume

$$Ag_m \gg |Y_o|$$
$$Y_a \gg |Y_o + Y_s|$$
$$Y_g \to 0$$
$$M = Ag_mR_a \gg 1$$

The validity of these assumptions can usually be ensured by suitable choice of components.

$$G = \frac{-Y_iAg_m}{(Y_i+Y_s+Y_o)\,Y_a+Y_og_m} = \frac{-Y_iM}{Y_i+Y_s+Y_o+Y_oM}$$

$$= \frac{-M}{1+(Y_s/Y_i)+(Y_o/Y_i)\,(M+1)}$$

$$= \frac{-M}{1+(Z_i/Z_s)+(Z_i/Z_o)\,M} \quad \ldots (50)$$

If the network of Fig. 3(c) is used as output network, then using the star-delta transformation it follows immediately that

$$Z_s = \frac{1+pT_1}{pC_1}$$

Thus when $T_1 = T_3$,

$$Z_i/Z_s = (C_1/C_3) = (R_3/2R_1) = \tfrac{1}{2}\,(R_3/R_2)\,.\,(R_2/R_1) = (4Q_o^2/2G_o) = (2Q_o^2/G_o)$$

This term is independent of frequency.

Similarly, when the circuit of Fig. 3(d) is used as output network,

$$Z_i/Z_s = (C_1/2C_3) = (2Q_o^2/G_o)$$

which is again independent of frequency.

When the circuit of Fig. 3(e) is used as output network,

$$Y_s = \frac{pC_2}{1+pT_1} + \frac{pC_1}{2(1+pT_1)} = \frac{pC_2 + \frac{1}{2}pC_1}{1+pT_1}$$

Thus when $T_1 = T_3$,

$$Z_i/Z_s = \frac{C_2 + \frac{1}{2}C_1}{C_3}$$

Putting

$$(R_1/2R_2) = n = (2C_2/C_1)$$

Then

$$Z_i/Z_s = \frac{(C_1/2)\,(n+1)}{C_3} = (R_3/R_1)\,(n+1)$$

which is independent of frequency.

For this circuit it has been shown that

$$G_o = (R_4/R_3)$$

$$Q_o = \tfrac{1}{2}\sqrt{\left[\frac{R_4}{2R_1R_2}(2R_1+R_4)\right]}$$

$$\therefore\ 4Q_o^2 = \frac{R_4(2R_1+R_4)}{2R_1R_2} = \frac{nR_4(2R_1+R_4)}{R_1^2} = \frac{nG_oR_3(2R_1+G_oR_3)}{R_1^2}$$
$$= 2nG_o\,(R_3/R_1) + nG_o^2\,(R_3/R_1)^2$$

i.e. $$(R_3/R_1)^2 + (R_3/R_1)\,.\,(2/G_o) - \frac{4Q_o^2}{nG_o^2} = 0.$$

$$\therefore\ R_3/R_1 = \frac{\sqrt{[(1+4Q_o^2)/n - 1]}}{G_o}$$

$$Z_a/Z_s = \left(\frac{n+1}{G_o}\right)\sqrt{\left[\frac{1+4Q_o^2}{n}\right]} - 1$$

If $Q_o \gg 1$

$$Z_i/Z_s = \frac{2Q_o\,(n+1)}{\sqrt{n}\,.\,G_o}$$

Acknowledgment

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Research into Tropospheric Scatter

In common with other countries, intensive research into the possibilities of tropospheric scatter is being carried out in the United Kingdom. Earlier in the year it was stated that the radar stations of N.A.T.O. from the Arctic to the Mediterranean would be co-ordinated by the use of such techniques.

Marconi's have now brought into operation a tropospheric scatter link between Great Bromley, Essex and Sutton Bank near Thirsk, Yorkshire, a distance of about 200 miles.

Field strengths are recorded continuously over this link; various types of modulation are available so that studies can be made of multipath effects, the effect of aircraft and other relevant problems. Preliminary tests have already been made over a 400 mile circuit to Aberdeen, and it is hoped later to establish a link over this distance.

Plans are currently in hand to set up 10kW transmitters and associated receivers at Newcastle and in the London area, with the object of being able to operate up to 36 simultaneous telephone channels or a television link between these points. The equipment is being designed and constructed for tests and demonstrations only.

The present transmitter at Great Bromley, is housed in a Services-type trailer. Operating on a frequency of 858Mc/s, the transmitter has an output of 500W and is capable of being frequency-modulated by a multi-channel signal or amplitude-modulated by a television waveform, or by pulses. The actual radiator consists of a 30ft diameter dish, the lower edge of which is mounted about 30ft above ground level; this is energized by a feed horn mounted on the transmitter trailer.

The radio beam can be steered in the horizontal plane by positioning the transmitter trailer on an arc of railway track which is laid in front of the radiating dish. Adjustment of the beam in the vertical plane is also possible since the feed horn is mounted on a small boom, which can be raised or lowered as necessary.

The receiver is contained in a mobile laboratory. In order to obtain the maximum mobility, the receiving aerial, which is mounted on a trailer, is only 10ft in diameter.

The link has now been in continuous operation for seven months and results from this and the original research are such as to indicate a sound commercial future for communication systems making use of the tropospheric scatter principle.

For the future, the ultimate limits of transmission will depend largely on the transmitter power available. With a power of 100kW it might be possible to achieve telegraph transmission up to distances of 700 miles. The ultimate operational limits are not yet established, but it would seem that there is a possibility of an eventual transatlantic television service by the tropospheric scatter system. This, however, would entail a series of 'hops', as the maximum range for reliable television transmission by any one link is only about 250 miles. A much nearer possibility is a direct European television link, for example from Holland to Britain.

A High Speed Oscillograph Cathode-Ray Tube for the Direct Recording of High Current Transients

By R. Feinberg*, Dr.Ing., M.Sc.

A high speed sealed-off high voltage cathode-ray tube is described which has straight acceleration, magnetic focusing, a voltage time-base deflector, and a current signal deflector consisting of a single-turn coil. As an example of the performance of the tube an oscillogram is shown of a current pulse having a peak value of about 1650A, a duration of the order of 15μsec and in the steepest part a rate of change of about 250A/μsec.

An oscillogram of a transient current of high peak intensity is commonly obtained with a voltage signal deflector oscillograph cathode-ray tube in combination with a suitable shunt and, if necessary, an amplifier of sufficient

Fig. 1. Prototype of the high current transient oscillograph cathode-ray tube with an experimental low sensitivity signal deflector coil

bandwidth. This method of current oscillography thus utilizes a voltage proportional to the current to be measured. The precautions to be taken when applying the method in cases of a transient duration of the order of microseconds have been pointed out, for example, by P. R. Howard[1].

There is another method of oscillographing a high current transient which does not appear to be generally used. The method employs a current transient oscillograph cathode-ray tube for direct signal deflexion. A high speed tube of this type is described in this article, and an example of its application is given.

Design of the Current Transient Oscillograph Tube

Fig. 1 shows a prototype of the tube, Ferranti 06/10PM, together with an experimental deflector coil. The coil is shown mounted around the neck of the tube.

The glass envelope of the tube is of conventional design, having a cone with a flat face of nominally 150mm diameter. The luminescent screen which is aluminium backed, is made of silver activated zinc sulphide phosphor

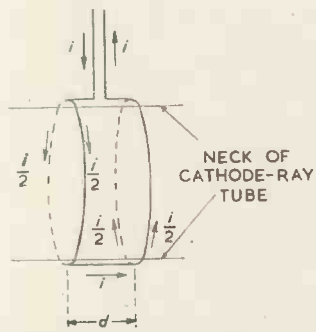

Fig. 2. Principle of design of the current deflector coil indicating the flow of current

which, at the present state of phosphor development, is considered to be the most suitable for the photographic recording of high speed traces. The electron gun is of the triode type with straight acceleration and magnetic final focusing, and is designed to operate at a maximum accelerator voltage of about 25kV.

The tube has two deflector systems of different designs. The X-deflector is of the electrostatic type for voltage deflexion, and the Y-system is of the magnetic type for current deflexion.

The X-deflector consists of a pair of plates of conventional design, with a deflexion sensitivity of about $500/V_a$mm/V, where V_a denotes the accelerator voltage. The capacitance of the deflector plates combined with that of the lead is about 1·5pF. The deflector loop from the terminals has an inductance of about 0·9μH. Thus the natural resonant frequency of the deflector system is about 400Mc/s.

The Y-deflector is constructed as a single-turn coil consisting of two symmetrical sections screwed together to

* Ferranti, Limited

form a compact unit placed closely around the neck of the tube, see Fig. 1. The principle of coil design is illustrated in Fig. 2, which also indicates the flow of current in the various parts of the coil. By adjusting the spacing d, see Fig. 2, between the two ring parts of the coil, the sensitivity of electron beam deflexion is set, a narrow spacing, as in Fig. 1, giving a low value of deflector sensitivity, but a wider spacing resulting in an increased sensitivity. The inductance of the coil varies in proportion with the magnitude of d. Its value for the coil shown in Fig. 1 where $d = 1$mm and whose ring parts are made of brass 3mm × 4·5mm, was measured at 180Mc/s as about 0·02μH.

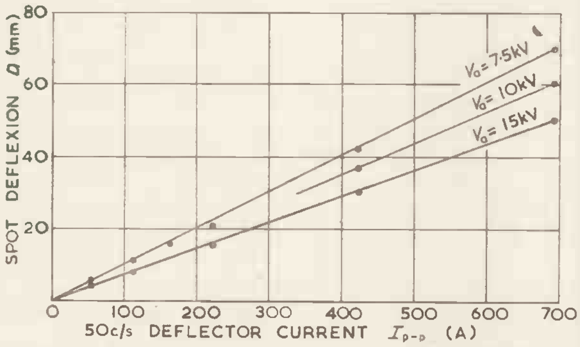

Fig. 3. Deflexion characteristic measured at 50c/s operation

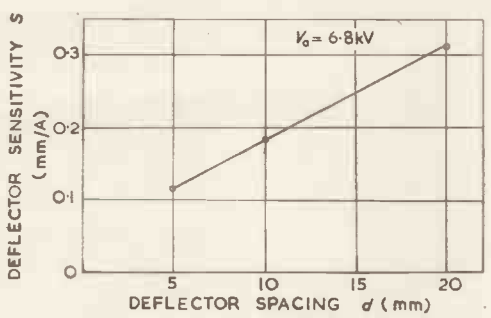

Fig. 4. Deflector sensitivity characteristic measured at 50c/s operation

Performance of the Oscillograph Tube

Fig. 3 gives the spot deflexion D obtained for various values of accelerator voltage V_a and measured as a function of the peak-to-peak value I_{p-p} of a 50c/s current passing through an experimental deflector coil similar to the one shown in Fig. 1. The deflector current was limited to a maximum of 250A r.m.s. because of the output limit of the current transformer employed to supply it; that means the maximum of I_{p-p} was 705A. The curves indicate linearity of D and that D, for a given value of deflector current, is inversely proportional to $\sqrt{V_a}$, in accordance with theory.

In Fig. 4 is shown the deflector sensitivity S measured with a 50c/s deflector current at $V_a = 6·8$kV using a set of coils as in Fig. 1, but with different values of spacing d (see Fig. 2). Within the range of measurement, i.e. $d =$ 5mm to 20mm, the value of S varies in linear relationship with d.

The deflexion Dmm, caused by a current iA, is:

$$D = Si$$

with S, in the range of measurements of Figs. 3 and 4 with the deflector coils used, being represented by the empirical expression:

$$S = a(d_o + d)/\sqrt{V_a}$$

where S is expressed in millimetres per ampere, a and d_o are dimensionless constants with numerical values dependent on the design of the deflector coil and its distance from the screen of the tube, d is in millimetres and V_a is in kilovolts. For the coils used for Fig. 4, $a = 0·034$ and $d_o = 4$.

The oscillogram, Fig. 5, was obtained with the tube and deflector coil shown in Fig. 1. It depicts the current pulse discharging a 10μF capacitor at 1·8kV through a Ferranti type AD30 cold-cathode gas-filled arc conduction valve. The pulse duration, measured at one-fifth of peak current, is about 17μsec and the peak current is 1650A. In order to obtain the trace within the useful screen area of the oscillograph tube it was necessary to reduce the effective deflector sensitivity of the coil; this was achieved by shunting the coil with a simple small wire loop.

In the steepest part of the current pulse, the rate of change of current through the deflector coil and shunt combination is about 250A/μsec. Assuming the coil-shunt combination to have only inductance, the value being about 0·01μH, the theoretical value of voltage drop across it at the steepest part of the current pulse should thus be about 3V. A low value of voltage drop across the coil-shunt combination was confirmed in an oscillogram.

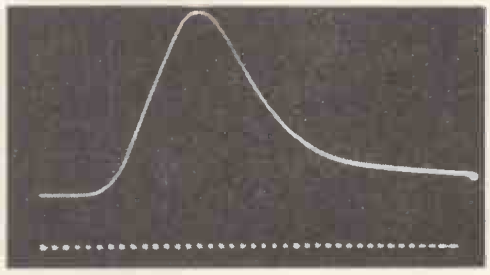
Fig. 5. Oscillogram of a current pulse with a peak of 1650A, the time calibration dots being spaced at 1μsec intervals

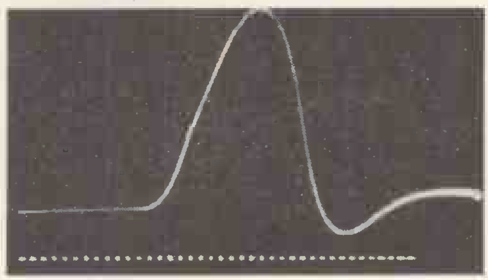
Fig. 6. Record of the distorted trace of a current pulse with a peak of 3300A, the time calibration dots being spaced at 1μsec intervals

The effect of insufficient screening is demonstrated in the oscillogram Fig. 6, which shows the current pulse through the AD30 valve from the 10μF capacitor at 2·5kV. The pulse duration is about the same as in Fig. 5, but the peak current is about 3300A (the shunt to the deflector coil consisted of a 6cm length of wire connected to the deflector coil leads). By more elaborate screening of the oscillograph tube, the distortion of the trace could have been eliminated and a curve similar to Fig. 5 obtained.

Acknowledgments

The prototype oscillograph cathode-ray tube described was developed and constructed in the Physical Laboratory of Ferranti Ltd. Mr. M. E. Roberts measured the tube performance and obtained the oscillograms.

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25 YEARS OF PROGRESS IN MICROWAVE LINKS

A survey in retrospect of the development of microwave or super-high-frequency techniques in telecommunications.

THE world's first introduction to what was then known as 'Micro-Ray' communication took place just over a quarter of a century ago, on 3 March 1931, when Standard Telephones and Cables Limited and their associates, Le Matériel Téléphonique of Paris demonstrated telephone and teleprinter links between Dover and Calais.

The two stations at Dover and Calais were in all essentials identical. That at Dover comprised a transmitter and a receiver and terminal equipment of normal design for connecting them together, so as to give facilities for two-way communication.

The outgoing signals were applied to what was designated a 'micro-radion' tube, in which the high frequency oscillations were generated. A short transmission line connected the 'micro-radion' tube to the radiating system or doublet, which was about 2cm in length. The amplitude of this h.f. current along the doublet at any instant was substantially the same. The doublet, situated at the focus of a paraboloidal reflector some 3m in diameter, served to concentrate the radiated waves into a 3° beam directed towards the distant receiver.

In order further to increase the efficiency of the system by the prevention of radiation other than in the required direction, a hemispherical reflector was located at the opposite side of the doublet to the paraboloidal reflector, with the doublet at its centre.

The radius of the hemispherical reflector was so chosen that when the reflected radiations reached the focus again they were in phase with those being radiated at that instant. The appropriate radius depended upon the wavelength, the relation being that it should be substantially a multiple of half wave lengths. The radius was large enough to ensure that the reflector had satisfactory electro-optical properties, but not so large as to intercept unduly the radiations reflected forward from the paraboloidal reflector.

It was estimated that the gain due to the paraboloidal reflectors on one channel was of the order of 46dB, to which the hemispherical reflector added another 6dB.

For the purpose of measuring the high frequency output at the transmitter, an aperture was provided in the centre of the paraboloidal reflector through which part of the radiation passed. By making the diameter of the aperture slightly smaller than that of the hemispherical reflector, no loss of radiated power resulted. The radiations passing through the aperture fell upon the measuring instrument employed, which took the form of a wavemeter calibrated for, and normally set to, the transmitted frequency. It comprised a small receiving antenna in which the induced e.m.f. was used to act upon a thermocouple junction. The readings of the associated galvanometer were an indication of the radiated power, while the distance between antenna and metal screen, being adjustable, also enabled wavelength measurements to be made. In the demonstration the wavelength used was 17·6cm (1705Mc/s), while the radiated power was of the order of 0·5W.

The receiver was a counterpart of the transmitter, except that no high frequency measuring device was provided. That is to say, it comprised a doublet connected by a transmission line to the 'micro-radion' tube, where detection took place. Paraboloidal and spherical mirrors exactly similar to those of the transmitter were also provided for concentrating the received waves upon this doublet.

To avoid coupling, the receiver was situated about 80 yards from the transmitter at each terminal and arranged to be in its electro-optical shadow, adequate allowance being made for diffraction. The same wavelength was used both for sending and receiving.

Though a certain number of experimenters had already succeeded in generating and utilizing oscillations of such wavelengths, nothing beyond what was described as laboratory investigations had up to then resulted. The enormous advance in technique shown by the 1931 demonstration, both as to distance covered and results obtained, indicated that the range of wavelengths between 10 and 100cm (3 000 to 300Mc/s) was ready for commercial use. The two-way radio telephone circuit was noteworthy for the quality of the speech received. Not only was it well up to the standard of a high-quality metallic circuit, but it showed little signs of being affected by fading.

Compared with radiations of the more usual wavelengths, the 'micro-rays' presented many striking features. Their extremely short wavelength permitted the use of electro-optical devices, such as reflectors or refractors, in addition to diminutive antennæ systems. It was shown that a further similarity between these radiations and light was that it was necessary to have virtual optical visibility between transmitter and receiver, or at least that conductive obstacles of too great length should not be interposed.

The success of the 1931 demonstration was widely hailed at the time, the equipment used being so advanced in principle and performance that it still resembles closely some of the latest types used in the same field of application today. An interesting forecast of future development was made in a note issued by S.T.C. in 1931, of which the following is an extract:

> "The frequency band available will allow of a very large number of permanent and continuous channels between the same places without mutual interference, while the directional properties and comparatively short range of the waves will make possible the use of the same frequencies over other routes. A further very important use will be for television, the development of which is hampered at the present time by the very large frequency band required for satisfactory definition of the object transmitted. It should now be possible to allocate as wide a band as is necessary for television without causing any ether congestion. It is easy to imagine the establishment of national 'micro-ray' networks for use in conjunction with television apparatus."

The development of the early experimental system to meet the needs of commercial communications was rapid, and early in 1934 the world's first commercial 'micro-ray' radio service between the civil airports at Lympne in Kent and St. Inglevert in France was opened for service by Sir Philip Sassoon, the then Under-Secretary of State for Air. This telephone and telegraph link was operated jointly by the British and French Air Ministries to speed up transmission of essential traffic messages between the two countries, and it remained in regular operation until after the outbreak of war.

The Lympne station occupied a commanding position overlooking the Channel, within optical range of the corresponding station at St. Inglevert, 35 miles away. The

Terminal tower and antenna at Lympne in the s.h.f. link with St. Inglevert, France.

wavelengths used were slightly staggered, a wavelength of 17cm (1 760Mc/s) being employed for transmission and one of 17·5cm (1 715Mc/s) for reception. Duplex working took place simultaneously by teleprinter and telephone. The teleprinter had been used on land line commercial telegraph services for some years, but had hitherto been used on w.t. for experimental purposes only, owing to the difficulty of obtaining a satisfactory radio signal. The 'micro-ray' system overcame this handicap and, in service, messages were transmitted at a speed of 60 to 70 words/min.

Research, development and operational experience over a period of nearly two decades resulted in prototype models of the portable s.h.f. radio link equipment which began extensive field trials with the BBC in 1948. This led to the use of equipment of this type for an outside television broadcast in 1950, and also in that year, the first Franco-British television relay, with an s.h.f. link spanning the Channel. Since this time portable s.h.f. links have been extensively used by the BBC and other television authorities for outside broadcast work, the most notable applications being the television relay of the Coronation in 1953 to France, Belgium, the Netherlands and Germany, and within the Eurovision network.

A typical portable s.h.f. link, currently produced by S.T.C. comprises a light-weight equipment forming a complete transmitter and receiver operating in selected bands in the frequency range 4 400 to 4 750Mc/s. It can rapidly be set up on a temporary or permanent basis for connecting an outside broadcast television camera with the permanent television network.

The transmitter has a velocity-modulated coaxial-line valve delivering a power of 250mW into a four-foot diameter paraboloid antenna. A similar valve is used as the beating oscillator in the receiver. By using a directly-modulated coaxial-line oscillator to feed the antenna direct all of the apparatus which is operating at s.h.f. has been made compact, being mounted as an integral part of the antenna assembly. This eliminates the need for extensive waveguide runs to the horn feed and permits the use of flexible cables between the antennæ and the control units of the transmitter and receiver. The antenna may therefore be easily mounted on a roof-top, on a mobile tower, or in any such elevated position as may be made necessary by the terrain traversed. Control units may also be set up under cover at ground level or beside the carrier-telephone equipment in the terminal or repeater station.

The receiver output of one link can be connected to the transmitter input of another link to form a repeater, enabling a limited number of links, each of which normally spans about 30 to 40 miles, to be connected in tandem. It also permits a deviation of route, around obstacles which would otherwise obstruct the line-of-sight path of a single link.

The performance which can be achieved on any particular project depends upon the length of route, the spacing between repeaters and the type of country over which the system operates. When a link is to form part of a telephone network, the objectives of which are C.C.I.F. requirements for international land line circuits, up to 60 telephone channels may be operated over a distance of some 300 miles. More channels could be obtained under emergency conditions with shorter routes and closer repeater spacing.

The system is frequency modulated with a deviation of 4Mc/s, but an increase in frequency deviation up to 8Mc/s accompanied by some loss of linearity is available for use under difficult transmission conditions. The intermediate frequency of the equipment is 30Mc/s and the i.f. amplifiers have a bandwidth of 20Mc/s.

Portable s.h.f. equipment of this type, based on early 'micro-ray' experience and field service with the BBC and other broadcasting administrations, has also been influenced by concurrent design, manufacture and installation of large-scale projects such as the Manchester to Kirk o'Shotts permanent television transmission links produced and installed for the General Post Office.

The longest link in the G.P.O.'s permanent television transmission network, the Manchester to Kirk o'Shotts s.h.f. f.m. radio system, was opened for public service in 1952 when the BBC inaugurated regular low-power transmission from Kirk o'Shotts. This television service has since been extended to Aberdeen, similar equipment providing the link for the section Dundee-Aberdeen.

The total length of the Manchester-Kirk o'Shotts route is 250 miles with seven repeater stations spaced at an average distance of some 30 miles. Equipment at terminal and repeater stations affords one uni-directional channel

Mobile television unit at Ticknock Hill, near Dublin.

from Manchester to Kirk o'Shotts simultaneously with a second similar channel in the reverse direction. Each channel is designed to handle the 3Mc/s bandwidth of a 405-line definition, 50 frames/sec, double-interlaced television programme in one direction of transmission. Full remote control and supervisory facilities are provided so that each terminal station is able to control the operation of the equipment at all points along the route to the other terminal.

To ensure a high degree of reliability, all transmission equipment with the exception of antennæ and main waveguide feeds is provided in duplicate, the change-over from one set of equipment to the other being fully automatic in the event of breakdown. In the same way, automatic standby sources of power supply cater for any breakdown in local mains supplies.

The intermediate two-way repeater stations are designed to work unattended and, in the case of that at Pontop Pike, the equipment provides for a feed to a local television transmitter without interference to through-routed signals in either direction.

Much experience in the development of these systems has been gained during the past two or three years from operating under test conditions over difficult propagation paths a multi-channel system in South Africa with 120 telephone channels. This experience had led to the development of an s.h.f. radio system which provides seven both-way radio channels, each of which is designed to be capable of handling up to 600 telephone channels or video signals for 405 to 525 or 625-line television providing either monochrome or colour pictures.

Since those early days of 1931, the experience of a quarter of a century in the development and application of s.h.f. techniques has shown that, with reliable service as the criterion, a properly-engineered s.h.f. radio system can take its place with any other telecommunication system. In fact, it will generally be found that the choice between a 'wire' or s.h.f. system rests not with performance, which in practice is comparable, but is influenced largely by political and geographical factors.

Parallel with the successful development of the s.h.f. system there has grown a tremendous demand for radio channels for multi-circuit telephony and television. Modern s.h.f. radio systems have played an outstanding part in meeting the demand and also contributed to the relief of the congested radio spectrum.

S.H.F. systems are under development which are designed to meet performance requirements to international standards for 600 telephone channels or appropriate television channels.

Improvement of the layout and performance of portable s.h.f. radio systems for outside broadcast links and short-term usage, is constantly under review and the design of equipment for the 7000Mc/s range is an important part of the development programme.

Thus that first venture in s.h.f. radio transmission 25 years ago has more than fulfilled its promise, and holds at least equal hope for the future.

Temperature Stability of Transistor Amplifiers

By G. Stuart-Monteith*, B.Sc., A.M.I.I.A.

An analysis is made of the most general form of transistor d.c. amplifier circuit, and a 'figure-of-merit' is proposed, in terms of which the current and voltage stability factors and the voltage gain can be expressed. The argument is extended to the stability of a multi-stage d.c. coupled amplifier.

ONE of the principal difficulties encountered in the design of transistor amplifiers, is due to the variation of the transistor characteristics with ambient temperature. Shea[1,2] has shown how the circuit design may compensate for these variations and he introduced a stability factor S, defined as $\partial I_c/\partial I_{co}$ which can be stated in terms of the transistor current gain α, and the circuit parameters. Hurley[3] has extended the argument and, since I_{co} may not vary with temperature in the same manner for all transistors, he introduced a level factor, the product of which and the stability factor gives a more correct forecast of the amplifier temperature stability.

Slightly different methods of approach have been used by other workers[4,5], but in each case the expressions derived are considered too complex for a rapid evaluation to be made of the stability of the circuit, and no relationship has been given between the voltage gain and the stability of the stage.

The Stability Factors

In Fig. 1 is shown the most general form of transistor d.c. circuit. It includes both d.c. feedback between the collector and the base and the input voltage, which may be d.c. and/or a.c. When the biasing conditions are considered, then, denoting d.c. currents and voltage by capital letters, it is shown in Appendix 1 that:

The collector current stability factor $S_c \equiv (\partial I_c/\partial I_{co}) = \dfrac{1}{1-\alpha\xi}$

The emitter current stability factor $S_e \equiv (\partial I_e/\partial I_{co}) = \dfrac{\xi}{1-\alpha\xi}$

The collector voltage stability factor $S_v \equiv (\partial V_c/\partial I_{co})$

$$= -\frac{1}{1-\alpha\xi}\cdot\frac{R_f}{R_f+R_L}(\xi R_e + R_L)$$

The output voltage stability factor $S_o \equiv (\partial V_o/\partial I_{co})$

$$= -\frac{1}{1-\alpha\xi}\,\frac{R_f R_L}{R_f+R_L}(1-\xi(R_e/R_f))$$

where the factor ξ, defined as the 'figure of merit', is given by:

$$\xi \equiv \frac{R_f}{R_f+R_L}\cdot\frac{R_{bb}}{R_e+R_{bb}} \text{ hence } 0<\xi<1.$$

The resistance R_{bb} is the total resistance in the base and comprises the parallel network formed by R_s, R_r, and $(R_f + R_L)$. In most practical cases $(R_f + R_L)$ will be much greater than R_s and R_r and in consequence can be neglected.

For perfect stability the stability factors should be zero, but this ideal figure cannot be approached, except for the

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emitter current stability factor, and hence it follows that the grounded-collector configuration can achieve a very high stability; unfortunately, the lack of gain of this configuration must rule it out for most applications.

The smaller the value of the figure of merit ξ, the better will the stability become, that is R_e must be of the order of R_{bb}, and the feedback resistance R_f must not be so large that the load R_L is negligible in comparison; it will be seen, however, that even a value of R_f ten times the load R_L can improve the stability.

Voltage Gain

A direct consequence of the insertion of the stabilizing network is the loss of gain, and in Appendix 2, the voltage gain G_v for the grounded emitter configuration at a con-

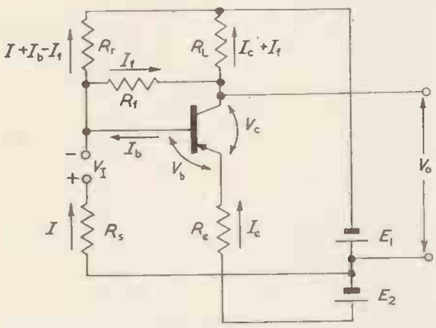

Fig. 1. General form of transistor d.c. circuit

stant value of I_{co}, that is at constant temperature, is shown to be:

$$G_{v(\text{temp/const})} \equiv (\partial V_o/\partial V_I)_{I_{co}=\text{const}}$$

$$= -(R_L/R_s) \cdot \frac{\alpha\xi}{1-\alpha\xi}[1-(1/\alpha)(R_e/R_f)]$$

If d.c. feedback is not present, that is R_f is infinity, then the above formula becomes:

$$G_v = -(R_L/R_s) \cdot \frac{\alpha\xi}{1-\alpha\xi}$$

It can be shown that the maximum gain obtainable from a grounded-emitter stage is given approximately by:

$$G_{v(\max)} \simeq -(R_L/R_s)\frac{\alpha}{1-\alpha}$$

Therefore it may be said that the effect of the stabilizing circuit is to reduce the current gain α by the factor ξ; it is for this reason that ξ has been termed the 'figure of merit'.

In the case of an a.c. circuit, the above formulæ may be modified by the presence of by-pass and/or blocking capacitors. For example R_e may be by-passed, when α will only be modified by the factor:

$$\frac{R_f}{R_f + R_L}$$

the value of ξ, however, remaining unchanged in the stability factor expressions.

It is not proposed to discuss the effect of the current and voltage feedback resistances R_e and R_f in this article, but if voltage feedback is required to extend the frequency range, the above expressions indicate that a d.c. path is preferred, since in addition to providing feedback it also improves the stability of the stage.

The Level Factors and Stability Products

The level factors have been defined as follows[3]:

Emitter current level factor $\equiv L_e \equiv (I_{co}/I_e) \times 100$
Collector current level factor $\equiv L_c \equiv (I_{co}/I_c) \times 100$
Collector voltage level factor $\equiv M_c \equiv (I_{co}/V_c) \times 100$
Output voltage level factor $\equiv M_o \equiv (I_{co}/V_o) \times 100$

The products of these factors with their respective stability factors give the stability products P in terms of percentages, which, as they take into account the operating conditions of a particular transistor, will give a quantitative evaluation of the temperature stability of the transistor amplifier.

For example, the operating conditions of a transistor having been determined, and knowing the variation of I_{co} with temperature, then the value of the stability product, for a given collector voltage can be found, so as to ensure that the transistor will not bias itself off over the desired temperature range.

Multi-Stage Amplifier

In multi-stage d.c. coupled amplifiers, a problem that faces the designer is the cumulative effect of the drift in each stage. From the above formulæ this may be readily determined, since:

$$dV_o = (\partial V_o/\partial V_I)\,\delta V_I + (\partial V_o/\partial I_{co})\,\delta I_{co}$$

then for two stages:

$$dV_o'' = \frac{\partial V_o''}{\partial V_I''}\left\{\frac{\partial V_o'}{\partial V_I'}\delta V_I + \frac{\partial V_o'}{\partial I_{co}}\delta I_{co}'\right\} + \frac{\partial V_o''}{\partial I_{co}''}\delta I_{co}''$$

where the superscripts denote stages 1 and 2. For identical stages this resolves to:

$$dV_o'' = \left(\frac{\partial V_o'}{\partial V_I'}\right)^2 \delta V_I' + \frac{\partial V_o'}{\partial I_{co}'}\left(1+\frac{\partial V_o'}{\partial V_I'}\right)\delta I_{co}'$$

Assuming the input voltage remains constant, we obtain the output voltage stability factor of a two-stage amplifier:

$$\frac{\partial V_o''}{\partial I_{co}} = \frac{\partial V_o'}{\partial I_{co}'}\left(1+\frac{\partial V_o'}{\partial V_I'}\right) \equiv S_o\,(1+G_v)$$

that is the drift to be expected in the final stage will be approximately equal to the product of the output voltage stability factor and the voltage gain of one stage. For an amplifier comprising n similar stages,

$$(\partial V_o^n/\partial I_{co}) = S_o\left\{\frac{(\partial V_o/\partial V_I)^n - 1}{(\partial V_o/\partial V_I) - 1}\right\} \simeq S_o\,(\partial V_o/\partial V_I)^{n-1}$$

since $(\partial V_o/\partial V_I) \gg 1$.

Acknowledgments

This theoretical analysis was carried out at the Admiralty Signal and Radar Establishment, and the writer is indebted to the Admiralty for permission to publish this article.

Appendix 1

Stability of Transistor Amplifiers

In Fig. 1 is shown the general form of transistor circuit with d.c. feedback. The stability may then be predicted if the following assumptions are made:

(a) The current gain α is constant.

(b) The input diode voltage drop V_b is constant.

(c) $I_c = I_{co} + \alpha I_e$ i.e., $I_e = \dfrac{I_c - I_{co}}{\alpha}$ .... (1)

(d) $I_b = I_e - I_c$
i.e., $I_b = I_e(1-\alpha) - I_{co} = (1/\alpha)[I_c(1-\alpha) - I_{co}]$ .. (2)

Denoting d.c. currents and voltages by capital letters, the loop equations are:

$$O = I_fR_f + (I_c + I_f) R_L - (I - I_f + I_b)R_r \tag{3}$$

$$E_1 = IR_s + V_1 + (I - I_f + I_b)R_r \tag{4}$$

$$E_1 + E_2 = I_eR_e + V_b + (I - I_f + I_b)R_r \tag{5}$$

$$E_1 + E_2 = I_eR_e + V_c + (I_c + I_f)R_r \tag{6}$$

From equation (3),

$$I_f(R_f + R_L + R_r) - IR_r + I_cR_L - I_bR_r = 0 \tag{7}$$

From equation (4),

$$I(R_s + R_r) - I_fR_r + I_bR_r = (E_1 - V_I) \equiv E_1' \tag{8}$$

$$\text{i.e., } I = \frac{E_1' + I_fR_r - I_bR_r}{R_s + R_r} \tag{9}$$

Substituting in equation (7),

$$I_f(R_f+R_L+R_r) - \frac{R_r}{R_s + R_r}[E_1' + I_fR_r - I_bR_r] - I_bR_r + I_cR_L = 0 \tag{10}$$

$$\text{i.e., } I_f\left[R_f+R_L+\frac{R_sR_r}{R_s + R_r}\right] - I_b\cdot\frac{R_sR_r}{R_s + R_r} + I_cR_L - \frac{R_sR_r}{R_s + R_r}\cdot(E_1'/R_s) = 0 \tag{11}$$

Denoting the parallel combination of R_s and R_r, i.e., $R_sR_r/(R_s+R_r)$, by R_b we have:

$$I_f = [(R_b/R_s)E_1' + R_bI_b - R_LI_c]/[R_f+R_L+R_b] \tag{12}$$

Substituting equation (4) in equation (5),

$$E_1+E_2=I_eR_e+V_b+\left[\frac{E_1'+I_fR_r-I_bR_r}{R_s+R_r} - I_f+I_b\right]R_r \tag{13}$$

i.e.,

$$E_1 + E_2 - V_b - (R_b/R_s)E_1' = I_eR_e - I_fR_b + I_bR_b \tag{14}$$

$$= I_eR_e+I_bR_b - \frac{R_b}{R_f+R_L+R_b}[(R_b/R_s)E_1'+R_bI_b-R_LI_c] \tag{15}$$

$$\therefore E_1+E_2-V_b-E_1'\left[\frac{R_f+R_L}{R_f+R_L+R_b}\right](R_b/R_s) = I_eR_e+I_bR_b\left[\frac{R_f+R_L}{R_f+R_L+R_b}\right] + I_cR_L\left[\frac{R_b}{R_f+R_L+R_b}\right] \tag{16}$$

Denoting the parallel combination of R_b and (R_f+R_L) i.e., $\frac{R_b(R_f+R_L)}{R_f+R_L+R_b}$ by R_{bb} and replacing E_1' by (E_1-V_I) (equation (8)), equation (16) may be written

$$E_1[1-(R_{bb}/R_s)]+E_2-V_b+V_I(R_{bb}/R_s) = I_eR_e+I_bR_{bb}+I_c\frac{R_LR_{bb}}{R_f+R_L} \tag{17}$$

Substituting for I_e and I_b from equations (1) and (2),

$$E_1[1-(R_{bb}/R_s)]+E_2-V_b+V_I(R_{bb}/R_s)=(1/\alpha)\left\{I_c\left(R_e+(1-\alpha)R_{bb}+\alpha\frac{R_LR_{bb}}{R_f+R_L}\right)-I_{co}(R_e+R_{bb})\right\} \tag{18}$$

$$= (1/\alpha)\left\{I_c\left(R_e + R_{bb} - \alpha\frac{R_fR_{bb}}{R_f+R_L}\right) - I_{co}(R_e+R_{bb})\right\} \tag{19}$$

The collector current stability factor is defined as

$$S_c \equiv (\partial I_c/\partial I_{co}) = \frac{R_e+R_{bb}}{R_e+R_{bb}-\alpha[R_fR_{bb}/(R_f+R_L)]} \tag{20}$$

making assumptions (a) and (b),

$$\text{i.e., } S_c = \frac{1}{1-\alpha[R_f/(R_f+R_L)]\,[R_{bb}/(R_e+R_{bb})]} \equiv \frac{1}{1-\alpha\xi} \tag{21}$$

Differentiating equation (1) we obtain the emitter current stability factor, defined as

$$S_e \equiv \frac{\partial I_e}{\partial I_{co}} = (1/\alpha)\left[\frac{\partial I_c}{\partial I_{co}} - 1\right] = (1/\alpha)\left[\frac{1}{1-\alpha\xi} - 1\right] \tag{22}$$

$$\text{i.e. } S_e = \frac{\xi}{1-\alpha\xi} \tag{23}$$

Substituting equation (12) in equation (6), we obtain

$$E_1+E_2=I_eR_e+V_c+\left\{I_c+\frac{(R_b/R_s)E_1'+R_bI_b-R_LI_c}{R_f+R_L+R_b}\right\}R_L \tag{24}$$

$$\text{i.e., } E_1+E_2-\frac{R_LR_b}{R_s(R_f+R_L+R_b)}E_1' - V_c = I_eR_e+I_c\left\{1-\frac{R_L}{R_f+R_L+R_b}\right\}R_L + I_b\frac{R_bR_L}{R_f+R_L+R_b} \tag{25}$$

Substituting for I_b from equation (2) and E_1' from equation (8),

$$E_I\left[1-\frac{R_LR_b}{R_s(R_f+R_L+R_b)}\right] + E_2 + \frac{R_LR_b}{R_s(R_f+R_L+R_b)}V_I - V_c = I_e\left[R_e+\frac{R_bR_L}{(R_f+R_L+R_b)}\right] + I_c\left[\frac{(R_f+R_b)R_L - R_bR_L}{R_f+R_L+R_b}\right] \tag{26}$$

Differentiating, we obtain

$$-\frac{\partial V_c}{\partial I_{co}} = \frac{\partial I_e}{\partial I_{co}}\left\{R_e+\frac{R_bR_L}{R_f+R_L+R_b}\right\} + \frac{\partial I_c}{\partial I_{co}}\left\{\frac{R_fR_L}{R_f+R_L+R_b}\right\} \tag{27}$$

Substituting equations (21) and (23) for $\partial I_c/\partial I_{co}$ and $\partial I_e/\partial I_{co}$ respectively,

$$-\frac{\partial V_c}{\partial I_{co}} = \frac{1}{1-\alpha\xi}\left\{\xi R_e + \xi\frac{R_bR_L}{R_f+R_L+R_b} + \frac{R_fR_L}{R_f+R_L+R_b}\right\} \tag{28}$$

$$= \frac{1}{1-\alpha\xi}\left[\xi R_e + \xi\frac{R_LR_{bb}}{R_f+R_L} + \frac{R_fR_LR_{bb}}{(R_f+R_L)R_b}\right] \tag{29}$$

$$-(\partial V_c/\partial I_{co}) = \frac{\xi}{1-\alpha\xi}\left[\frac{R_e(R_f+R_L)+R_LR_{bb}}{R_f+R_L} + \frac{R_fR_LR_{bb}}{(R_f+R_L)R_b}\cdot\frac{R_f+R_L}{R_f}\cdot\frac{R_{bb}+R_e}{R_{bb}}\right] \tag{30}$$

$$= \frac{\xi}{1-\alpha\xi}\left[\frac{R_e R_f}{R_f+R_L} \frac{(R_{bb}+R_e)R_L}{R_f+R_L} \frac{R_L(R_{bb}+R_e)}{R_b}\right]. \quad (31)$$

$$= \frac{\xi}{1-\alpha\xi}\left[\frac{R_e R_f}{R_f+R_L} + \frac{R_L(R_{bb}+R_e)}{R_{bb}}\right] \ldots\ldots \quad (32)$$

Hence

$$-(\partial V_c/\partial I_{co}) = \frac{\xi}{1-\alpha\xi}\,\frac{R_f}{R_f+R_L}\left[R_e + R_L \frac{R_{bb}+R_e}{R_{bb}}\,\frac{R_f+R_L}{R_f}\right] \ldots\ldots \quad (33)$$

$$= \frac{1}{1-\alpha\xi}\,\frac{R_f}{R_f+R_L}\,(\xi R_e + R_L) \ldots\ldots \quad (34)$$

By definition the collector voltage stability factor is

$$S_V \equiv (\partial V_c/\partial I_{co}) = -\frac{1}{1-\alpha\xi}\,\frac{R_f}{R_f+R_L}\,(\xi R_e + R_L) \ldots \quad (35)$$

The d.c. output voltage, i.e. the voltage between the collector and earth is given by

$$V_o = I_e R_e + V_c - E_2 \ldots\ldots \quad (36)$$

$$\therefore\ (\partial V_o/\partial I_{co}) = (\partial I_e/\partial I_{co})\,.\,R_e + (\partial V_c/\partial I_{co}) \equiv S_e R_e + S_V \ldots \quad (37)$$

Substituting for S_e and S_V from equations (23) and (35) we obtain the output voltage stability factor,

$$S_o = (\partial V_o/\partial I_o) = \frac{\xi}{1-\alpha\xi}\,.\,R_e - \frac{1}{1-\alpha\xi}\,\frac{R_f}{R_f+R_L}\,(\xi R_e + R_L) \ldots\ldots \quad (38)$$

$$= \frac{\xi R_e}{1-\alpha\xi}\,\frac{R_L}{R_f+R_L} - \frac{1}{1-\alpha\xi}\,\frac{R_f R_L}{R_f+R_L} \ldots \quad (39)$$

$$\text{i.e., } S_o = -\frac{1}{1-\alpha\xi}\,\frac{R_f R_L}{R_f+R_L}\,[1 - (\xi R_e/R_f)] \ldots\ldots \quad (40)$$

Appendix 2

Voltage Gain of a Transistor Amplifier

Considering the general form of transistor circuit shown in Fig. 1, we have (Appendix 1, equation (19))

$$E_1\,[1 - (R_{bb}/R_s)] + E_2 - V_b + V_I\,(R_{bb}/R_s)$$

$$= (1/\alpha)\left\{I_c\left(R_e + R_{bb} - \alpha\,\frac{R_f R_{bb}}{R_f+R_L}\right) - I_{co}(R_e+R_{bb})\right\} \ldots \quad (41)$$

then as I_{co} only varies with temperature and not with input volts, and again assuming V_b is constant,

$$R_{bb}/R_s = (1/\alpha)\left(R_e + R_{bb} - \alpha\,\frac{R_f R_{bb}}{R_f+R_L}\right)\partial I_c/\partial V_I \ldots\ldots \quad (42)$$

$$\text{i.e., } (\partial I_c/\partial V_I)_{Ico} = \alpha(R_{bb}/R_s)\,\frac{R_f + R_L}{(R_e+R_{bb})(R_f+R_L) - \alpha R_f R_{bb}} \ldots\ldots \quad (43)$$

Also from equation (1), Appendix 1,

$$(\partial I_e/\partial V_I)_{Ico} = (1/\alpha)\,\partial I_c/\partial V_I)_{Ico} \ldots\ldots \quad (44)$$

Substituting equations in Appendix 1, (2), (8), (12) and (36) in equation (6), we obtain

$$E_1\left\{1 - \frac{R_L}{R_f+R_L+R_b}\,.\,(R_b/R_s)\right\} - V_o = \frac{R_L}{R_f+R_L+R_b}$$

$$\left\{I_c(R_f+R_L+R_b) + (I_e - I_c)R_b - I_c R_L - (R_b/R_s)\,V_I\right\} \ldots \quad (45)$$

Then differentiating,

$$-\frac{\partial V_o}{\partial V_I} = \frac{R_L}{R_f+R_L+R_b}\left\{R_f\,\frac{\partial I_c}{\partial V_I} + R_b\,\frac{\partial I_e}{\partial V_I} - (R_b/R_s)\right\} \ldots\ldots \quad (46)$$

Substituting equations (43) and (44) above,

$$-\frac{\partial V_o}{\partial V_I} = (R_L/R_s)\,.\,\frac{R_{bb}}{R_f+R_L+R_b}$$

$$\left\{\frac{(R_f+R_L)(\alpha R_f + R_b)}{(R_e+R_{bb})(R_f+R_L) - \alpha R_f R_{bb}} - (R_b/R_{bb})\right\} \ldots \quad (47)$$

$$= (R_L/R_s)\,\frac{R_{bb}}{R_s+R_L+R_b}\left\{\frac{(\alpha R_f + R_b)}{(R_e+R_{bb})(1-\alpha\xi)} - (R_b/R_{bb})\right\}. \quad (48)$$

where

$$\xi \equiv \frac{R_f R_{bb}}{(R_f+R_L)(R_e+R_{bb})}$$

Hence

$$-\frac{\partial V_o}{\partial V_I} = (R_L/R_s)\,.\,\frac{1}{R_f+R_L+R_b}\left\{\frac{\alpha R_f + R_b}{1-\xi\alpha}\,.\,\xi\,.\,\frac{R_f+R_L}{R_f} - R_b\right\} \ldots\ldots \quad (49)$$

$$= (R_L/R_s)\,.\,\frac{1}{R_f+R_L+R_b}\left\{\frac{\xi(\alpha R_f + R_b)(R_f+R_L) - R_b R_f(1-\alpha\xi)}{R_f(1-\alpha\xi)}\right\} \ldots\ldots \quad (50)$$

$$= (R_L/R_s)\,.\,\frac{1}{1-\alpha\xi}\left\{\frac{\alpha\xi(R_f+R_L+R_b)R_f + R_b(\xi R_f + \xi R_L - R_f)}{R_f(R_f+R_L+R_b)}\right\} \ldots\ldots \quad (51)$$

$$= (R_L/R_s)\,.\,\frac{\alpha\xi}{1-\alpha\xi}\left\{1 + \frac{(1/\alpha\xi)\,(R_b/R_f)\,[\xi(R_L+R_L) - R_f]}{R_f+R_L+R_b}\right\} \ldots\ldots \quad (52)$$

$$= (R_L/R_s)\,.\,\frac{\alpha\xi}{1-\alpha\xi}\left\{1 + (1/\alpha)\,.\,(R_b/R_f)\,.\,\frac{(R_f+R_L) - \frac{(R_f+R_L)(R_e+R_{bb})}{R_{bb}}}{R_f+R_L+R_b}\right\} \ldots\ldots \quad (53)$$

$$= (R_L/R_s)\,.\,\frac{\alpha\xi}{1-\alpha\xi}\left\{1 + (1/\alpha)\,.\,(R_{bb}/R_f)\,.\left(1 - \frac{R_e+R_{bb}}{R_{bb}}\right)\right\} \ldots\ldots \quad (54)$$

$$= (R_L/R_s)\,.\,\frac{\alpha\xi}{1-\alpha\xi}\left\{1 - (1/\alpha)\,(R_e/R_f)\right\} \ldots\ldots \quad (55)$$

That is, at constant temperature,

$$G_v \equiv (\partial V_c/\partial I_I)_{Ico} = -(R_L/R_s)\,.\,\frac{\alpha\xi}{1-\alpha\xi}\,[1 - (1/\alpha)\,(R_e/R_f)] \ldots\ldots \quad (56)$$

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Waveguide Surface Finish and Attenuation

(Part 2)

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

Polished Waveguide Surfaces

The surface finish of drawn waveguides of copper, brass, silver or aluminium can be improved by careful electrolytic or chemical polishing under closely-controlled conditions.

The underlying principle of electropolishing is selective dissolution under the influence of current, making the surface progressively smoother and more brilliant. The actual mechanism is still obscure, but several theories have been put forward[24-27].

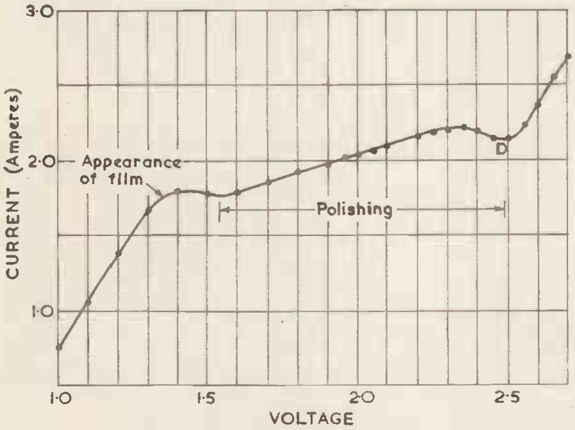


Fig. 15. Voltage-current characteristic of brass in orthophosphoric acid electrolyte

Fig. 16. Electropolished precision-drawn copper waveguide surface from 1in × ½in guide

Section perpendicular to waveguide axis
Magnification: 940
Copper waveguide in lower part of photograph

Chemical polishing is successful, provided the correct solution composition and operating conditions are found. Theories of levelling produced by this method have been discussed by Pinner[28].

The authors have electropolished specimens of precision-drawn copper and brass waveguide in orthophosphoric-acid baths (two parts acid and one part water). The optimum voltage in such a bath is very critical, higher and lower voltages cause pitting and etching respectively. The optimum voltage can be deduced from the voltage-current characteristic of the electrolyte. The best polish is obtained at a voltage just below that which causes gassing at the anode, as indicated by a sharp increase in current density for a small change in voltage (Fig. 15).

Electropolishing removes all microscopic roughness, there being no evidence of a microprofile even when the magnification is 4 000. No 'draw' lines are present although there are small pits on electropolished-copper surfaces (Fig. 16) probably due to copper-oxide inclusions. Large undulations remain on electropolished brass surfaces (Fig. 17), but there are no small pits.

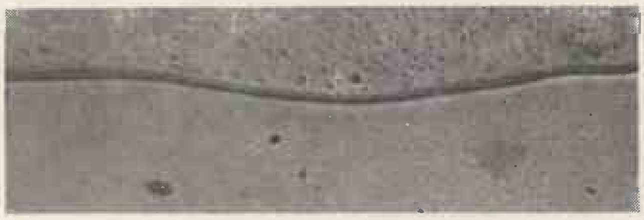


Fig. 17. Electropolished precision-drawn brass waveguide surface from 1in × ½in guide

Section from long side parallel to waveguide axis
Magnification: 940
Brass in lower part of photograph

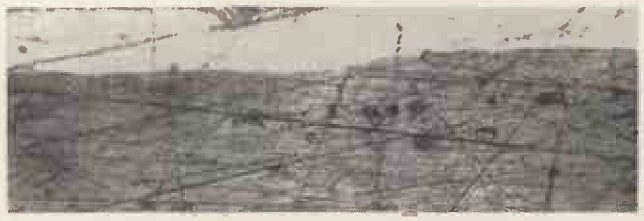
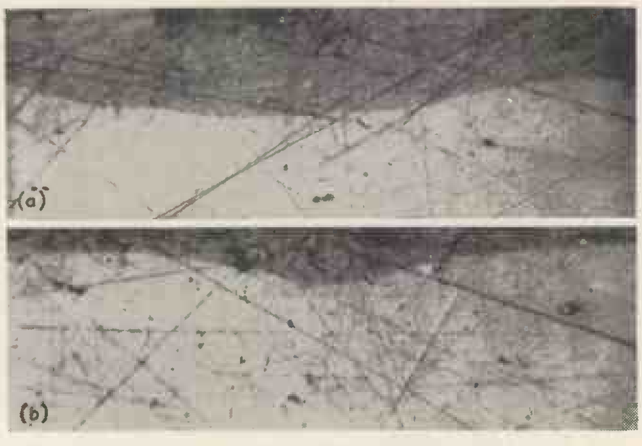


Fig. 18. Chemically polished copper waveguide surface from 1in × ½in guide

Section from long side parallel to waveguide axis
Magnification: 940
Copper in lower part of photograph



Figs. 19(a) and (b). Chemically polished brass waveguide surface from 1in × ½in guide

Sections perpendicular to waveguide axis
Magnification: 940
Brass in lower part of photographs

Drawn-aluminium waveguides have also been electropolished and anodized successfully by the authors[20] and have been compared with cast, sprayed, chemically-polished and drawn aluminium components.

Precision-drawn copper and brass waveguides have been chemically polished in phosphoric-acid solutions. The surfaces of such copper tubes are very smooth with no acute 'draw' lines (Fig. 18). Fig. 19(a) shows that for

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brass too there has been a microscopic polishing action, but brass surfaces show gentle undulations with a few sharp changes in level and there are some shallow hollows on both long and short sides particularly near the corners (Fig. 19(b)). Although these can be smoothed out by chemical polishing a good deal of metal would have to be removed.

Both electropolishing and chemical polishing seem to be more effective on copper guides because of their greater original roughness although they can be usefully employed on brass guides.

Electroformed Waveguides

The electroforming[29-34] process consists of electroplating a suitable mandrel with a heavy deposit. After machining the outside of the deposit to give the required wall thickness, the mandrel is withdrawn. Thus, all precision machining is done on the outside of a mandrel, whereas to produce the same finish and accuracy on the inside of the component by normal machining techniques would be nearly impossible. An interesting paper on the electroforming of components and instruments for millimetre wavelengths has been published by Harvey[32].

The technique is useful for manufacturing pieces of irregular shape. Pins and plates intended to project inside the waveguide can be incorporated in the mandrel. It is also possible to plate layers of different metals on the mandrel, e.g. copper on the inside of a guide for good conductivity and a chromium deposit on the outside for durability.

For all the electroformed guides except one examined by the authors the copper was deposited by means of the periodic-reverse-current plating process[34,35] which permits the use of high current densities. Due to the levelling action of this type of plating[36], heavy deposits are not subject to nodules. The remaining electroformed guide inspected during the present investigations was produced in an acid-copper bath.

Mandrels are of either the permanent or expendable type. The permanent type can be withdrawn without damage to either itself or the electroformed component. It is, therefore, used for electroforming samples with a uniform cross-section or having a unidirectional taper. Multiple permanent mandrels can also be used for complex sections. Permanent mandrels are usually made of stainless steel which is inert in most plating solutions and can be machined to fine tolerances. Separation of the mandrel from the finished piece is possible because the natural oxide film responsible for the metal's passivity prevents adhesion. For long uniform sections stainless-steel mandrels are made undersize and then plated with tin to the required dimensions. After electroforming the tin is melted out, facilitating easy withdrawal of the mandrel. Any remaining tin is removed by etching in hydrochloric acid for several hours. Invar, covar or glass made conducting by coating with silver by chemical reduction or evaporation, may be utilized for long untapered mandrels where differential expansion facilitates simple withdrawal.

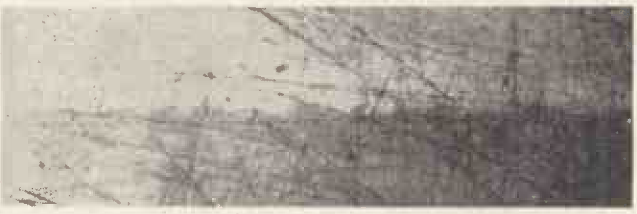


Fig. 20. Electroformed copper waveguide surface from 1in × ½in guide. Expendable fusible alloy mandrel. Copper sulphate bath

Section perpendicular to waveguide axis
Magnification: 940
Copper in lower part of photograph

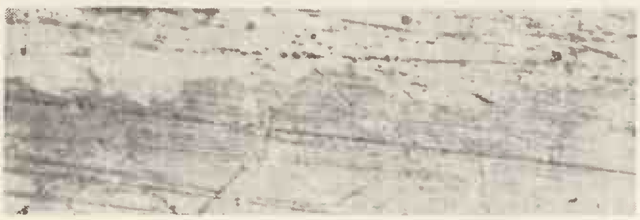
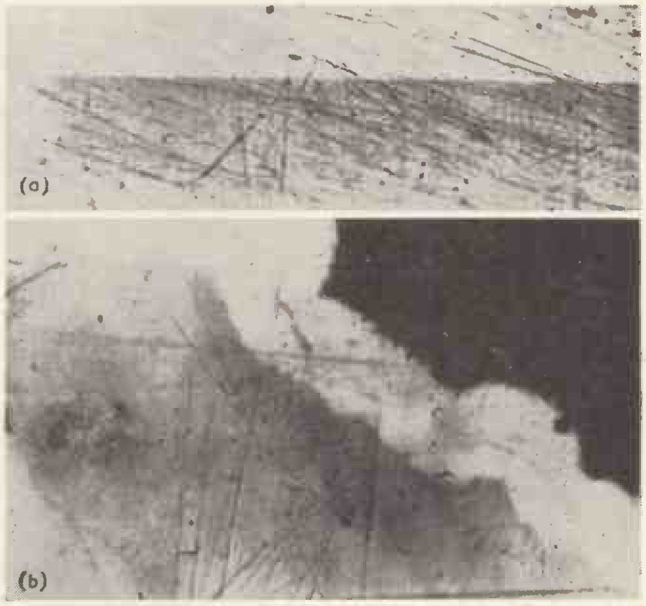


Fig. 21. Electroformed copper waveguide surface from 1in × 1in guide. Expendable fusible alloy mandrel. Copper cyanide bath

Section from long side parallel to waveguide axis
Magnification: 940
Copper in lower part of photograph



Figs. 22(a) and (b). Electroformed copper waveguide surfaces from 1in × ½in guide. Composite stainless steel mandrel. Copper cyanide bath

Sections perpendicular to waveguide axis
Magnification: 940
Copper in lower part of photographs

Expendable mandrels are used when the mandrel must change form to permit its withdrawal from the work and can be subdivided into fusible and soluble types. Fusible mandrels are generally made of low melting-point alloys or waxes the latter being sprayed with a conducting coating of metal. This type does not give a very good internal finish and it is difficult to hold precise tolerances. Soluble mandrels can be made of aluminium, soluble in caustic soda, zinc, soluble in hydrochloric acid, or ordinary rolled steel which dissolves slowly in hydrochloric acid. These are useful where better finishes and tolerances are required than those which are obtainable with fusible mandrels, but where permanent mandrels cannot be withdrawn. They are also used in preference to fusible mandrels when heat might cause distortion of the electroformed shell.

The surface finish of electroformed guides depends on the type of mandrel employed[33] and the method of separation of the guide and mandrel. A transverse section of a bend electroformed on an expendable mandrel of tin-lead-bismuth alloy in an acid-copper bath was found to have a very poor internal surface which is quite undular and microscopically

Fig. 23. Electroformed copper waveguide surface from 1in × ½in guide. Permanent stainless steel mandrel. Copper sulphate bath

Section perpendicular to waveguide axis
Magnification: 940
Copper in lower part of the photograph

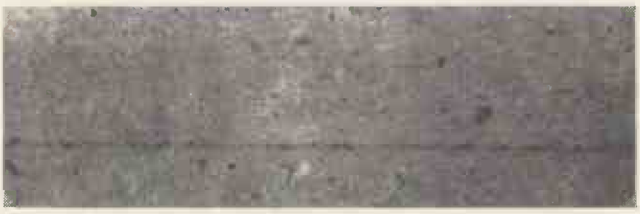


Fig. 24. Electroformed copper waveguide surface from 1in × ½in guide. (Later sample). Permanent stainless steel mandrel. Copper sulphate bath

Section perpendicular to waveguide axis
Magnification: 940
Copper waveguide in lower part of photograph

rough in places (Fig. 20). The roughness of a section parallel to the guide axis is of the same order as that of a transverse section. Another component electroformed on a similar mandrel in a cyanide bath had a much worse internal topography, the microscopic roughness being much more general (Fig. 21). The difference in internal roughness is attributed to differences in mandrel finish and methods of separating the mandrel and component.

The effect of various types of stainless-steel mandrel on the internal roughness was then determined. A sample was electroformed in a cyanide bath and had a composite rectangular mandrel fabricated from stainless steel plates with fusible-alloy plugs at the corners to hold them together. The mandrel was removed by melting the alloy at the corners. A second sample was electroformed in an acid bath on a permanent stainless-steel mandrel, the mandrel being withdrawn by applying a separating force from a screw mechanism. A third sample was produced on a similar type of mandrel which had a better finish. Withdrawal of the mandrel was assisted by heating.

Transverse sections from the first sample show that the surface is extremely flat and smooth over the majority of long and short sides (Fig. 22(a)). There are large cavities at the corners, however, due to the plugs (Fig. 22(b)). On the second sample there are numerous scratch marks due to the withdrawal process (Fig. 23). There is a complete absence of scratches on the third sample due to the withdrawal technique and the resulting finish is better than that of any other type of waveguide (Fig. 24).

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An Artificial Heart and Lung Machine

A British machine which makes possible direct access to the interior of the heart during cardiac surgery is now in production. The machine takes over the function of the patient's heart and lungs after the heart has been stopped by a chemical injection.

The idea was conceived by Dr. D. G. Melrose at the Post-Graduate Medical School of London. The second prototype model of the machine was built with the backing of the Nuffield Foundation and proved to be effective for clinical use. New Electronic Products Ltd designed and manufactured the production model, assisted by Tricorn Designs Ltd.

The whole machine is housed in one trolley measuring approximately 3ft × 3ft × 2ft. The main units are the oxygenator, which takes over the function of the lungs, the pumps which maintain the artificial circulation and an electronic control unit.

The pumps are driven by two 1/4 h.p. motors which in the event of the power failure can be replaced instantaneously by a manually-operated crank handle. Blood passing through the venous pump is passed to the oxygenator, a rotating cylinder set at slight angle to the horizontal having many internal projections, over which the blood flows, providing a total re-oxygenating area of $1 \cdot 3m^2$.

The control unit contains the switches for the electric motors and the electronic units for measuring continuously the pressures within the hydraulic circuit. Should the pressures exceed a chosen limit the machine is switched off by means of a pressure-sensitive diaphragm moving a magnetic core between the coils of an inductance bridge.

The bridge is fed from an audio frequency oscillator, any out-of-balance voltage being amplified sufficiently to operate a phase-sensitive demodulator which in turn drives a milliammeter calibrated in pressure and a trip relay.

An additional control circuit ensures that the volume of blood in the machine is constant within narrow limits under varying conditions. A brass disk at the lower end of the oxygenator is separated from the pool of blood by a 'Perspex' disk 1mm thick, forming a capacitor, changes in mean level changing the capacitance.

This in turn frequency modulates an oscillator and any deviation from the pre-set frequency acts as a proportional controller to the two servo motors driving the back-stops of the pumps. Direction of rotation of the servo motor is such that if the level of blood within the oxygenator decreases, delivery of the venous pump and arterial pump are increased and decreased respectively.

The servo loop is stabilized by positive and integral feedback networks to ensure utmost control with maximum response. In practice, the particular circuit adopted controls the total volume of blood of approximately two litres within 50ml.

Short News Items

A Conference of the Physical Society on Crystal Dynamics will be held at Queen Mary College, University of London, on Monday and Tuesday, 7-8 January 1957, in the Department of Physics. The meeting is open to non-members and enquiries should be addressed to the Physical Society, 1 Lowther Gardens, Prince Consort Road, London, S.W.7, marked for the attention of Miss Miles.

The College of Science and Technology announce the following courses. A course of nine lectures on High Quality Sound Reproduction commencing on 7 January; a course of seven lectures on Acceptance Testing of Electrical Equipment for Power Supply Systems commencing on 21 January, and a two-day course of lectures on Patents for Engineers from 8-9 March. Full details and enrolment forms may be obtained from the Registrar, College of Science and Technology, Manchester 1.

The Twelfth Annual Electronics Exhibition and Convention of the Northern Division of the Institution of Electronics will be held at the College of Science and Technology, Manchester, from 11-13 July and 15-17 July 1957. This exhibition will incorporate a Scientific and Industrial Research Section, a Manufacturers' Section and a programme of lectures and film shows on electronic topics. Further particulars may be obtained from the Honorary Exhibition Organizer, Mr. W. Birtwistle, 78 Shaw Road, Rochdale, Lancs.

The Fourteenth Annual Exhibition organized by the Radio and Electronic Component Manufacturers' Federation will be held at Grosvenor House and Park Lane House, Park Lane, London, W.1, from Monday, 8 April, to Thursday, 11 April, 1957. Formerly known as the Radio Component Show, the title has now been changed to Radio and Electronic Component Show to be more descriptive of its scope, covering as it does components for the radio, television, telecommunications and electronic industries in their widest aspects. For overseas visitors and other special guests a preview will be held from 10 a.m. to 2 p.m. on the first day, 8 April. Admission will be by ticket obtainable from the Secretary, R.E.C.M.F., 21 Tothill Street, Westminster, London, S.W.1.

Fielden Electronics Ltd have recently released details of the Tekto J.3 Level Controller and the Telstor L.F.3 Continuous Level Indicator. Five years ago this company introduced original instruments based on the principle of electrical capacitance. These years of experience, development and application have resulted in the Tekto J.3 and the Telstor L.F.3.

Isotope Developments Ltd are acting as consultants to Landis & Gyr A.G. for the instrumentation of the new high power Swiss 20MW materials testing heavy water moderated reactor which is being designed and built by Reaktor A.G. They are responsible for the recommendation and supply of the full nucleonic instrumentation, which includes the equipment for the channels for controlling and safeguarding the reactor and its surroundings.

Pye Marine Ltd have received an order to equip fifteen Russian trawlers, being built at Brooke Marine Ltd, Lowestoft, with their latest electronic Fishfinder. The Pye Fishfinder is the latest advance in fish detection, employing a cathode ray tube indicator which gives a visual indication of fish shoals under the vessel.

Hallam, Sleigh & Cheston Ltd announce that their London area Sales and Technical Office has been transferred to Oldfield Road, Maidenhead, Berks. Telephone Maidenhead 4311.

Square D Ltd have opened new offices at 72a Queen's College Chambers, Paradise Street, Birmingham. Mr. A. S. Paice, Area Manager for the Midlands, is in charge of the new office.

E.M.I. Electronics Ltd have been commissioned to provide the telemetry equipment for the series of experimental high altitude test vehicles (rockets) which will be used in the upper atmosphere rocket research programme to be undertaken by the Gassiot Committee of the Royal Society during the forthcoming International Geophysical Year (1957-58).

Leo Computers Ltd, the J. Lyons & Co subsidiary engaged in the manufacture and sale of the automatic office known as LEO, announce that all publicity enquiries should be directed to Mr. J. Morrish, Voice and Vision Ltd, 107 Baker Street, London, W.1, and not to the Lyons Publicity Office.

The Research Department of Fry's Metal Foundries have introduced a new product called Alcho-Re Soldering Fluid type S.64. This fluid allows an excess temperature of 50-100°C above the carbon point of resin fluxes. Soldering is thus completed without flux carbonization.

Remington Rand Ltd have announced the formation of an Electronics Division. Mr. C. W. Elliott is head of the new division.

Cossor Communications Co Ltd has been formed, as a subsidiary of A. C. Cossor Ltd, to extend the telecommunications interests of the Cossor Group. Mr. H. Chisholm, Joint Managing Director of A. C. Cossor Ltd, is Chairman of the new company and Mr. T. S. Heftman General Manager and Technical Director. The Sales Division is at Cossor House, Highbury Grove, London, N.5.

Racal Engineering Ltd announce that they have now appointed Farnell Instruments Ltd of Hereford House, North Court, Vicar Lane, Leeds 2, sole distributors for their range of digital counters and frequency meters in the north of England area. Farnell Instruments are fully qualified to discuss any special application of these equipments and their technical staff is being trained in the servicing of these Racal products.

The British Sound Recording Association, in connexion with its twenty-first birthday celebrations, will hold a technical convention and exhibition of sound recording, reproducing and allied electroacoustic equipment at the Waldorf Hotel, London, W.C.2, from 20-22 September 1957.

Notes from — NORTH AMERICA

U.H.F. Klystron Transmitter

The 225 to 400Mc/s Levinthal Model PC33 transmitter employs a unique form of high-level high-efficiency amplitude modulation in the audio range up to 20kc/s, the system being capable of 90 per cent modulation on a 10kW carrier with overall harmonic distortion of about 4 per cent and an r.f. efficiency of up to 40 per cent. The klystron used is the Eimac X590E, which incorporates a modulating anode. The system is also capable of up to 20kW c.w. operation and can be used for f.m. or f.s.k. by modulating the r.f. drive accordingly. The equipment consists of a beam power-supply unit, a modulator unit, a heat-exchanger unit, and an r.f. unit. The beam power supply is rated for 30kV at 2A d.c., with less than 0·04 per cent ripple. The modulator unit includes a low-level audio amplifier and a high-level 1kW anode dissipation modulation tetrode, a 0 to 15kV bias supply, a d.c. filament supply for the klystron, a d.c. focus-electrode supply, five focusing-magnet supplies rated for 150V, 4A each and a performance monitoring system. The heat exchanger is rated for 50kW at 115°F ambient and provides up to 30 gal/min at 60 lb/in². The r.f. unit consists of the X590E klystron, focus coils, tuning boxes, air blowers, dummy load and input and output directional couplers.

Aircraft Research Department

The McDonnell Aircraft Corporation, St. Louis, Missouri, has recently established a research department which will be seeking the new knowledge being called for with increasing urgency for the advanced fighters, missiles and helicopters which the company is developing. The new department will be working in the fields of electronics, supersonic and hypersonic aerodynamics, thermodynamics, advanced structures, certain aspects of nuclear physics and studies of new weapon systems and will be headed by Dr. Albert E. Lombard, Jr.

V.L.F. Symposium

The National Bureau of Standards announces the Very Low Frequency Symposium which will be held at the Boulder Laboratories, Boulder, Colorado, from 23 to 25 January 1957. The low frequency range of the radio spectrum has not received a great deal of attention in recent years, but new applications are now coming to light. In 1919 some experiments showed it was possible to communicate with a totally submerged submarine by using low frequencies. Today there is a renewed interest because high frequencies require that the antenna be raised above the water.

Other applications concern research laboratories that are asking for more stable l.f. standard signals which provide a quicker and more accurate measurement. Also at the present time, the civil aviation organizations are experimenting with navigation systems which will depend on a relatively unvarying l.f. signal to indicate distance and direction at all times.

High Temperature Components

The penetration of a critical temperature barrier in the development of guided missiles and supersonic aircraft has been demonstrated by scientists of the General Electric Company's Research Laboratories Schenectady, with laboratory models of electronic devices and circuits operating literally 'red hot'. The achievement was described as a major step towards overcoming the inability of electronic controls to withstand the heat generated by air friction at extreme speeds.

Inside electric furnaces, new valves, capacitors, resistors, transformers, inductors, wires, printed circuit boards and an electric motor were operated at temperatures ranging from 900 to 1 500°F, metals such as titanium and special ceramics being used in the construction of the components.

Operations Research Symposium

The Institute of Radio Engineers announces an Operations Research Symposium to be held at the University Museum Lecture Hall of the University of Pennsylvania on 7 February, 1957, sponsored jointly by the Professional Group on Engineering Management of the Philadelphia Section of the I.R.E. and the Society of Industrial and Applied Mathematics. The major theme will be Mathematical Models in Management Decision Making.

Navigation and Interception Computer

The Canadian Government has placed a multi-million dollar contract for the R-Theta Navigation and Interception Computer. By being adopted as standard on the RCAF'S CF-100 twin-jet interceptor planes, the equipment may be adopted by at least eight other fighting services of allied countries outside Canada, including NATO.

The equipment has been ordered by the British,Canadian, and United States armies, for evaluation as an aid in tank warfare and also by the Swedish Air Force.

The system can be used as an interception computer on a moving target, of value in defence, particularly in view of its immunity to jamming.

The system is continuously being improved, the present version being transistorized and of very small weight and volume. The R-Theta is claimed to be the world's only transistorized navigational computer.

Radio Fire-Fighting Network

A radio communications network, serving three New England states, has been installed and is in operation.

Eventually, 46 communities in Massachusetts, New Hampshire and Vermont will be equipped with transmitters and receivers which, under a system of simultaneous monitoring, make it possible for fire vehicles in each participating community to hear not only their local messages but also those of the three-state mutual aid control centre at Greenfield, Mass.

The monitoring system was installed by the General Electric Company and as a result of the new radio link, rural communities within the three-state hookup are receiving quick fire department assistance of the type usually found only in large metropolitan areas.

Currently, as many as 35 pieces of equipment may be dispatched immediately to any town needing help.

A much larger number will be possible when the system is expanded shortly to include 57 mobile units and 17 base stations.

Instrument Company's Canadian Representation

The merger of Baird Associates Inc. and Atomic Instrument Company, both of Cambridge, Mass., under the name of Baird Associates—Atomic Instrument Company, was recently announced. The new company will be represented in Canada by the Canadian Marconi Company, which has been acting as Canadian representative for the Atomic Instrument Company for some time.

Baird Associates, Inc., manufacturer of spectrochemical instrumentation and equipment for infra-red spectrophotometry for the past twenty years, has pioneered the development of emission and absorption type apparatus and more recently has entered the field of transistor test equipment and transistorized devices.

The Atomic Instrument Company is known for equipment used in the detection and analysis of radioactive phenomena and high speed data-handling through specialized electronic systems and the manufacture of electronic counting and control systems for industrial applications.

Mr. Arthur Kingsnorth, Supervisor of Instrumentation for the Commercial Products Division of Canadian Marconi Company, 2442 Trenton Avenue, Montreal, 16, Quebec, will be handling all inquiries regarding the sale of Baird Associates—Atomic Instrument Co. equipment.

X-Band Calibrated Load

Color Television Inc., of San Carlos, California, has announced Model 128A X-Band calibrated load. This is an adjustable mismatch load, which is calibrated at frequencies of 8·6, 9·0 and 9·5kMc/s with a v.s.w.r. of 1·10, 1·16, 1·4 and 1·8 for each frequency. The accuracy is ±1 per cent of the calibrated value. The load is produced with a UG-39/U waveguide fitting, modified with two locating guide pins and the overall length is 8¾in.

LETTERS TO THE EDITOR

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

Function Generation

Dear Sir,—Several articles have appeared in your journal which use the same basic technique of function generation although the general possibilities of the method have not been pointed out. Reference may be made to I. G. Baxter "A Square Root Law Circuit" (March 1954, page 97), and to P. N. Nikiforuk "A Technique for Non Linear Function Generation" (March 1955, page 118).

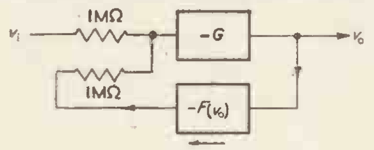


Fig. 1

Consider the basic circuit shown in Fig. 1. The resistors are summing junction inputs to a high gain d.c. computing amplifier (G). The block (F) represents a given function generator.

We have

$$[V_i - F(v_o)] = V_o / -G$$

Since G may be of order 10^8,

$$V_i = F(v_o)$$

If Φ is the inverse function of F,

$$\Phi(v_i) = V_o$$

The following table shows the possibilities of the method:

F	Φ
$V_o dt$	(dv_i/dt)
$\sin v_o$	$\sin^{-1} v_i$
$\exp(v_o)$	$\log v_i$
$(V_o)^n$	$(V_i)^{1/n}$

We find that the method has been used for taking inverse log by C. J. Savant and R. C. Howard (Trans. Inst. Radio Engs., Electronic Computers, September 1954).

If there is an external input (y) to block F and if F is a multiplier then $V_o (= V_i/Y)$

Yours faithfully,

G. Lapierre, J. Dolan, J. Kydd,
Computing Devices of Canada Ltd.,
Ottawa.

A Photo-transistor Trigger Circuit*

Dear Sir,—The circuit shown in Fig 1. constitutes a transistor version of the Eccles-Jordan circuit in which one of the transistors is photo-sensitive.

When the circuit is in the condition X_1 cut off and X_2 conducting it will change its state if the photo-transistor is exposed to a suitable amplitude (either a pulse or a slow rise in level) of thermal or visible radiation. Both the sets of component values shown were chosen so that the circuit would not be triggered by the fluctuations in the ambient daylight illumination to be found in a shaded portion of the laboratory. The circuit responded to a candle at a range of one foot and to a cigarette or small soldering iron passed near to it. The speed of response in the latter two cases indicated that triggering was due to the incident radiant flux and not to an increase in the temperature of the photo-transistor.

Heating the circuit slowly, in an oven, showed that it will also be triggered by an increase in ambient temperature. With the component values shown unbracketed the circuit triggered when the oven temperature was 26°C. With the component values shown in brackets the corresponding temperature was 37°C.

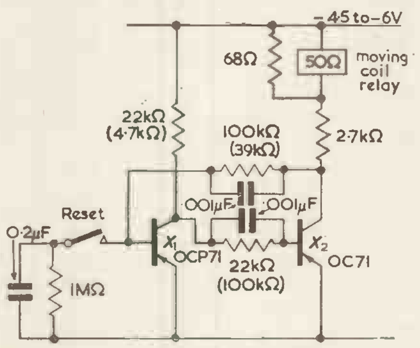


Fig. 1. The trigger circuit

Yours faithfully,

J. H. McGuire,
D.S.I.R.,
Fire Research Station,
Boreham Wood, Herts.

* U.K. Patent Application No. 27492/56.

L.F. High-Pass Filters

Dear Sir,—I have found Mr. Crombie's article (June, 1956) very useful and interesting and am grateful to him for the labour which went into it and to you for publishing the result.

The working points for the valves seem however, to be on the non-linear parts of the characteristics for V_1 and the second section of V_2 while grid-current bias is hardly to be recommended for linear working—even of a cathode-follower—but is used for the first section of V_1. The valves are not specified but I take them both to be 6SL7's. Another mild criticism that may be made is that V_1 is a very imperfect 'see-saw' with the values given.

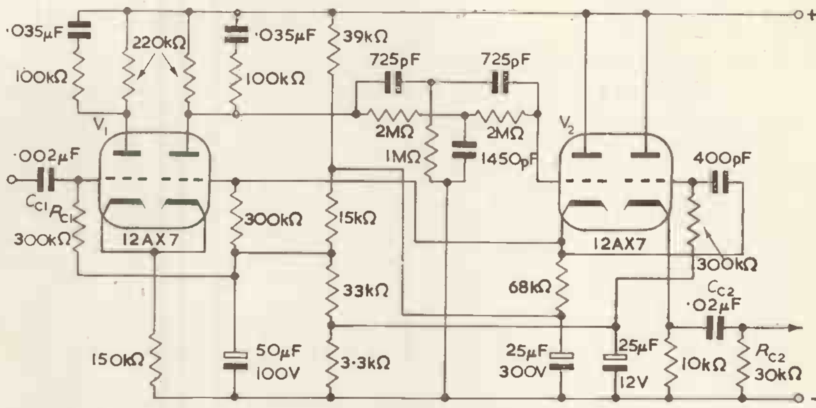


Fig. A. $V_1 = V_2 = 12AX7$. It is convenient to duplicate the high-pass section of the "α" network rather than use a high impedance RC feed (necessary if the network section is not to be upset) of turnover ω_o for the feed to the output cathode follower

I have worked out values for a pair of 12AX7's which should be linear up to a peak input of 1V. The circuit is shown in Fig. A. It may be necessary to explain to those unfamiliar with the long-tailed pair that each half must, relative to its own cathode, develop twice the required gain: the dynamic loads in the anode circuits of V_1 are adjusted for this state.

Your faithfully,

W. Grant
22 Montague Avenue,
London, S.E.4.

The author replies:

Dear Sir,—I agree with Mr. Grant's comments on the operating conditions of the valves in my circuit. His circuit should be an improvement on mine at high input levels.

However, it was not intended that the published circuit would meet all requirements, it being included merely as an illustration of the general method of design. For this reason the valve types were not specified.

The published circuit was designed as a scale model of a filter having a cut off frequency of 1c/s, and operating with a small input signal. For this reason, no attempt to provide optimum bias was made, since it was felt that phase shift difficulties might arise from inadequate bypassing. During tests it was found that at an input level of about 0·1V distortion was negligible, except at the cut-off frequency.

Yours faithfully,

D. D. Crombie
Dominion Physical Laboratory,
Lower Hutt, New Zealand.

last named article has been added as a supplement to the preceding one but formed no part of the lecture course.

While the specialist in any one of the fields dealt with here will find perhaps not much that is new to him in the article dealing with his special field there seems to be no doubt that a mutual fecundation, as indicated for example in Professor Kroebel's contribution, is made possible by arranging the lectures on the various fields side by side. The subjects dealt with lie far enough apart to avoid the danger of overlapping and repetition so frequently found in composite volumes of this kind.

The different subjects are dealt with in a clear, straightforward manner and the book as a whole is not beyond the grasp of the ordinary student. Nearly every chapter ends with a bibliography. A common subject matter index concludes this well produced book.

R. NEUMANN.

BOOK REVIEWS

Mechanical Design for Electronic Engineers

By R. H. Garner. 223 pp. 40 figs. Demy 8vo. George Newnes Ltd. 1956. Price 25s.

THIS is an extremely useful book for the electronic engineer wishing to extend his grasp of the mechanical aspect of his profession. It has chapters on, among other things, Racks, Chassis, Serviceability, Sheet Metalwork, Finishes, Printing, Potting, Soldering, and Codes of Practice. The volume is well printed on good paper, the text is well illustrated with diagrams and photographs, and printing errors and typographical inconsistencies are infrequent; there must be few technologists for whom it is not worth the 25s. asked for it. But it is questionable whether, with the exception of the excellent introductory chapter, it is actually about design.

The value of the endeavours of any individual depends on how much he knows, and what he can do with what he knows. In terms of mechanical engineering this means on the one hand a sound knowledge of materials, processes and components, their advantages and drawbacks, and on the other the ability to select appropriate materials, processes and components in the light of particular requirements and combine them to produce a good piece of apparatus. It is this latter which constitutes the act of designing. Knowledge and ability are distinct and complementary; neither is any use without the other, and no amount of knowledge will make a good designer if he lacks creative imagination. Leaving aside the question whether such imagination can ever be *taught*, it is certainly true to say that it can be trained. Thus what one looks for in a primer on design is a number of operational procedures, ranging in formality from mathematical expressions, through rules of thumb, to "tips". What one finds in Mr. Garner's book are lists of possible systems, generally supported by useful comment, and a kind of distillate from manufacturers' catalogues. It will certainly increase one's knowledge, but it is rather a work of reference than a textbook on design.

In the chapter on ventilation, for example, Mr. Garner enumerates the various systems for keeping apparatus cool, but the information is advanced with insufficient reference to the range of application of each technique. The engineer who has to get a completed electrical design into a cabinet of prescribed dimensions perhaps hopes to rely on convective cooling and not be obliged to use a blower. How large, then, should this louvre be? How much will it help if holes are punched in the chassis around that beam tetrode? Mr. Garner tells us that the ports should be "adequate". Even the roughest rule of thumb would be helpful here. If it be objected that it is simpler to experiment and find out, then well and good, but this is not design.

The author's preface states that "if every possible aspect of mechanical design were to be described fully, an extremely large volume would be required, and for economic reasons such a book would fail to reach those for whom it would be of greatest use . . . nevertheless it is hoped that the information and ideas encompassed in the present work will prove of very real value to all who wish to extend their knowledge of this important subject". It is the reviewer's opinion that this book contains most valuable information, but is disappointingly deficient in ideas.

P. E. K. DONALDSON

Impulstechnik

Edited by F. Winckel. 346 pp., 242 figs. Large Octavo. Springer-Verlag, Berlin-Göttingen-Heidelberg. 1956. Price DM 37·50.

THIS book is a collection of ten lectures given at a university extension course recently held at the Technical University of Berlin-Charlottenburg in collaboration with the Elektrotechnische Verein Berlin. It deals with new problems and fields of application of the impulse technique, in particular with methods and circuits for the generation, modulation, synchronization and transformation of simple and complex impulses. The treatment is partly mathematical, partly more physical and technological. From the list of titles it will be seen what a wide field has been covered and that not only technical problems have been dealt with but also physiological subjects like human hearing and generally the impulse transmission by nerves.

The single subjects are: Impulse analysis by F. A. Fischer; Application of information theory to impulse problems by W. Meyer-Eppler; Impulse technique used as measuring method in physics by W. Kroebel; Problems of the multiple utilization of communication channels by pulse modulation by H. Holzwarth; The impulse technique in radio navigation by E. Kramar; The impulse method in ionospheric research by W. Dieminger; Impulse problems in electronic computers by A. Speiser; Optical impulse technique by P. K. Hermann; Impulse transmission in the nervous system by H. Fack; Hearing treated on the basis of information theory by H. Fack. This

Reliability Factor for Ground Electronic Equipments

Editor-in-chief: Keith Henney. 243 pp. 152 figs. Demy 4to. McGraw Hill Publishing Co. Ltd. New York and London. 1956. Price 56s. 6d.

THE reliability of electronic equipments has become of increasing importance since the last war, during which it was found that designs and constructions of military equipment on commercial lines could not withstand the arduous service conditions which they were called upon to experience. The increasing complexity of electronic equipments has emphasized the problems and, since the war, considerable effort has been put into improving the reliability of military equipment by the United States of America and the United Kingdom. This is the first book to be published giving detailed information on the problem of poor reliability of electronic equipments.

The book was originally prepared by the McGraw Hill Book Company for the Rome Air Development Center, Griffiss Air Force Base, Rome, N.Y., under a U.S. Government contract and a great deal of the information in the book was provided by the General Engineering Laboratory of the Rome Air Development Center. The McGraw Hill Book Company was then given the right to print and sell the book commercially.

The book is divided into twelve chapters and a fair summary of its contents can be seen from the chapter headings—Reliability Concepts, Causes of Unreliability, Systems Aspects, Mathematical Approach to Reliability, Electrical and Electronic Factors, Mechanical and Environmental Factors, Human Engineering, Component Parts, Interference, Automatic Production, Equipment Publications and Maintenance.

It covers useful fundamentals of statistical methods of analysis, electrical and mechanical engineering principles, the human engineering side of the problem, some broad principles governing the selection and use of valves and components and some comments and data on training and service manuals and maintenance.

While the book is written primarily for designers of military equipments, it is useful also to industrial equipment designers, for as equipments become more complex the need for reliability in industrial and commercial equipments becomes more essential. The book is excellently wrtten and faces up extremely well to the task of collecting and sorting the mass of information and data which exists both in America and in this country in a somewhat intangible form. The result is a reasonably clear exposition of the problems involved and of present methods of overcoming them.

There are one or two points which might have been stressed, such as the use of wired-in valves in improving reliability—this is more prevalent in this country than in the United States of America; also that the situation on components is constantly changing—components fitted in equipment may be several years old, making any reliability analysis difficult.

In Chapter IV, where the mathematical approach to reliability is discussed, it might be useful to comment that in order to obtain an accurate scientific analysis in the true sense it would be necessary to include failures from *all* models of a particular type of equipment which have been tested under the *same* environmental conditions. As this is not often practical, approximations only can be given.

The chapter on human engineering is particularly valuable and contains a great deal of useful information not generally available in concise form. Considerable space is devoted to valves and their operation, as it is stated that valves are, at the moment, the single largest cause of unreliability in the United States equipments.

It is obvious that a great deal of thought and care has been taken in the preparation of this book and very little detail affecting reliability has been omitted. Even the effect of lowest contract bidding has been covered.

It is interesting that British work in this field is mentioned in places throughout the book.

The book can, therefore, be thoroughly recommended to those who are concerned with building reliable equipments whether for military use under arduous environmental conditions or for industrial use where a high standard of reliability is required.

G. W. A. DUMMER

BBC Handbook 1957

288 pp. Pott 8vo. The British Broadcasting Corporation. 1956. Price 5s.

THIS handbook is a basic reference book for all matters connected with the BBC, including its constitution, the Charter, and the historical background. It gives facts about the nation-wide audiences running up to well over ten million people for individual sound and television programmes. It deals with the need for development in the Middle East transmissions, and surveys the vast amount of material broadcast in 44 languages in the External Services, which amounts to more hours of broadcasting than all the domestic sound services combined.

Proceedings of The 1956 Electronic Components Symposium

240 pp. 120 figs. Demy 8vo. Engineering Publishers, New York. 1956. Price $5 paper cover, $8·25 cloth cover.

THESE are the authorized proceedings of the 1956 Electronic Components Symposium sponsored by the American Institute of Electrical Engineers, the Institute of Radio Engineers, the Radio Electronics - Television Manufacturers Association and the West Coast Electronic Manufacturers Association, with participation by agencies of the Department of Defence and the National Bureau of Standards.

Forty-three different papers cover a variety of subjects such as materials progress, electron tubes, solid state devices, passive components, instrumentation, measurements, etc.

ELECTRONIC EQUIPMENT

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

MAGNETIC MATERIALS

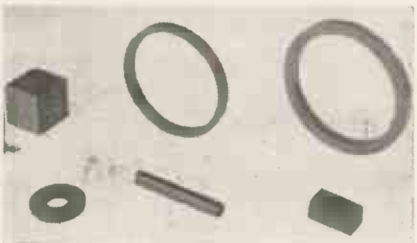
(*Illustrated below*)

The Plessey Co. Ltd, Ilford, Essex.

THE range of 'Caslox' magnetic materials, manufactured for use in television receiver focusing magnets, has been extended by the introduction of a new type known as 'Caslox' Grade 3.

Basically, this material, like earlier 'Caslox' grades is a non-metallic magnetic ferrite for use in geometrical configurations where metallic magnets are unsuitable. It has a high stability which results from a high value of coercivity and a permeability only slightly greater than unity.

The new Grade 3 material has been developed as an alternative to existing grades, for application where high standards of dimensional accuracy are of greater importance than maximum magnetic properties. This material is light in weight, can be easily machined with carbide-tipped tools and consists of a powdered magnetic ferrite which is bonded into the shapes desired by the addition of a small quantity of resinous binder. This technique enables tolerances of ±·005in to be achieved and also permits the moulding-in of metallic inserts if required.



When 'Caslox' is used in scanning coils and ion traps, the magnets can be very compact, with high magneto-motive force and very low leakage flux.

The existing 'Caslox' Grade 2 will continue to be employed in cases where maximum magnetic properties are of paramount importance.

VIBRATION GENERATOR

(*Illustrated above right*)

Goodmans Industries Ltd, Axiom Works, Lancelot Road, Wembley, Middlesex.

MODEL V.G.108 has been especially developed for vibration testing of heavy components for aircraft, guided weapons and electronic applications. Continuous dynamic forces up to ±3 000lb are available at the mounting table, with a maximum stroke of ±0·35in.

The 16½in diameter mounting table is cast in a special magnesium alloy for strength and lightness. A combination of T-slots will permit bolt fixings at points on the table.

A central shaft is employed to maintain a high radial stiffness, but this does not transmit the driving force, which is

developed immediately under the periphery of the table.

This method eliminates the mechanical disadvantages associated with a spindle drive and table attachment when testing bulky objects.

A single large casting of a special steel forms the main part of the magnet system which provides the annular magnetic gap for the moving coil.

A separate supply panel houses the rectifier set and smoothing circuit to convert the three-phase mains supply to d.c. for exciting the electromagnet field supply and a separate blower motor can be remotely installed with the air supply ducted into the vibrator.

Test loads up to 100lb can be attached to the mounting table and excited up to 20*g* but much larger loads can be applied at correspondingly lower accelerations.

The moving coil has a low impedance and the effective inductance has been reduced to negligible proportions, the change in impedance with frequency being reduced to a minimum.

The vibrator installed with a 10kVA amplifier is capable of producing dynamic forces up to 4 000lb for short duration. Continuous forces of 2 500lb can be generated at up to 500c/s and over 2 000lb at 1kc/s.

POWER TRANSISTORS

Distributed by Tellux Ltd, West Mall Works, 27/39 Rabbit Row, London, W.8.

TYPE OD604 power transistor was developed by Telefunken to enable the final stages of l.f. amplifiers to handle bigger outputs and to increase the d.c. power in d.c. convertors.

With a peak current of 1·4A, a residual voltage of less than about 0·8V is required at the collector. For this peak current a control voltage of 0·7V is sufficient. It is thus possible with a push-pull circuit including two transistors in class B operation to obtain a useful output of approximately 3·5W with a battery voltage of 6V. In a d.c. convertor with a push-pull oscillator circuit, approximately 8 to 10W of d.c. power can be achieved.

The transistor is so constructed that its thermal internal resistance, i.e. the thermal resistance between collector barrier and base, is small. By means of cooling fins or a good thermal contact between transistor and chassis, the temperature in the barrier can be limited to 75°C with a d.c. load of 1·3W.

INSTRUMENT CATHODE-RAY TUBE

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

THE 4GP range of 3½in instrument tubes replaces the existing 3½in E4412 series, four varieties of screen being available, with persistences ranging from 1msec to 20sec. A further screen, suitable for radar applications, is to be introduced at a later date although the tube can be made to special order with any of the majority of other standard screens.

The deflector plate sensitivity of the new tube does not vary by more than 2 per cent for deflexions up to 75 per cent of the useful scan. Improved spot centring ensures that the undeflected spot will fall within a radius of 5mm concentric with the tube face. The deflexion axes are orthogonal to within 1°. Other changes from the E4412 series are single stage post-deflexion acceleration, reduced inter-electrode capacitance, a flat-plate glass screen and a 6·3V heater.

STROBOSCOPIC EQUIPMENT

(*Illustrated above*)

Dawe Instruments Ltd. 99 Uxbridge Road, London, W.5

THE electronic flash tubes normally used in stroboscopes all emit light over a comparatively large area. Lens systems and reflectors, however, require a point-source of light for maximum efficiency and Type 1202C/1 point-source stroboscopic lamp unit has been produced for use with Type 1202C Strobotorch. Using a Mazda FA5 lamp, the unit provides a point-source of stroboscopic light over the range 120 to 10 000 flashes/min. The design of the unit permits the use of either a reflector or lens system for projection and microscopy.

The output of the unit is within the range 0·1 to 2·35 joules, depending on the flash-rate. The normal lamp is rendered inoperative, but its speed control is used to control the flash rate of the point-source lamp. The independent mounting of the new lamp unit enables it to be used with optical systems of any required focal length, since the mounting provides free access from three sides.

STABILIZED POWER UNIT

(*Illustrated above*)

Racal Engineering Ltd, Western Road, Bracknell, Berkshire

POWER supply unit type PU.156 has been designed so that it can be built into experimental or production equipments. Its dimensions are such that it occupies the minimum depth when mounted behind a standard 19in front panel, thereby allowing space for the circuits which it will be used to power.

In addition to being offered as a bare chassis, the PU.156 can be supplied assembled behind a standard 19in front panel complete with output terminals, metering and the necessary controls.

The unit employs a conventional series stabilizer circuit. Two 5V4G valves in parallel are arranged for full-wave rectification, and the cathodes connected to two A1834 double triodes in parallel, acting as the series valve.

The single stage control circuit uses an EF91 which obtains its reference voltage from a 150C2 neon stabilizer. As a safety precaution the anode of the EF91 is connected to earth through five CC3L neons so that the potential on the grids of the series valves remains at a fixed safe value even if the control valve fails.

The positive h.t. output may be varied between 200 and 300V by operation of the potentiometer connected to the grid of the control valve.

Two metal rectifiers connected in a full-wave system provide the −150V supply. The 150C stabilizer holds the voltage consistant to within 0·04 per cent for loads of 0 to 15mA.

An additional transformer supplies 8·5A at 6·3V a.c. unstabilized.

CRYSTAL CALIBRATOR

(*Illustrated below*)

Advance Components Ltd, Roebuck Road, Hainault, Essex

CALIBRATOR type 74 is designed to facilitate the setting of signal sources to multiples of 5Mc/s or 1Mc/s over the frequency range 1 to 250Mc/s and can be used at higher frequencies with reduced sensitivity.

The instrument comprises a 5Mc/s crystal oscillator which beats with the unknown signal in the grid circuit of a pentode valve. This valve functions as a detector and optional 1Mc/s oscillator. The resulting beat frequency between the unknown signal and the nearest 5Mc/s (or 1Mc/s) multiple is fed to an audio amplifier. The amplified beat note is then fed to the speaker (or phones) and to a rectifier circuit. The resultant d.c. output is applied to the magic-eye tuning indicator, which is of the twin-sensitivity type, permitting the use of a wider range of acceptable input voltages without recourse to adjustment of the gain control. At zero beat the audio amplifier response falls to zero, the eye opening when exact resonance is established.

The instrument may also be used for receiver calibration, but this application will be greatly dependent on receiver characteristics, e.g. input impedance, sensitivity, signal-to-noise ratio, etc. A small fraction of the 5Mc/s signal appears at the input socket, together with a weak signal at 1Mc/s. Thus if the r.f. input socket is connected to the input terminals of a sensitive receiver with the b.f.o. switched on, beats will be heard when the receiver is tuned to multiples of 1Mc/s or 5Mc/s.

The r.f. input is 30mV or less for indicator saturation and the aural and visual indication accuracies are better than ±2 parts in 10 000 ±500c/s and 15c/s respectively.

UNIVERSAL SPECTRUM ANALYSER

(*Illustrated above right*)

Winston Electronics, Ltd, Shepperton, Middlesex

A UNIVERSAL instrument for the precision measurement of the spectra of pulsed signals in the S, X and L bands has been produced which also acts as a sensitive receiver of the frequencies covered by these bands. The requisite spectra are displayed on a 5in long-persistence c.r.t.

The equipment consists of an electronic rack and a microwave component section at X-band frequencies. The microwave component is mounted on the front panel of the electronic rack. At S-band, the r.f. components are mounted in a second rack, both racks being fitted into a single case.

The analyser is designed for the viewing of signals of pulse length between 0·2 and 2·5μsec at repetition frequencies from 200 to 3 000 p/s.

The sweep speeds are 2, 5, 10 and 25c/s, the intermediate frequency 20Mc/s and the i.f. bandwidth 50kc/s.

Both sections use a mechanically tuned local klystron oscillator which can be frequency-modulated over a range varying between 40 and 10Mc/s.

The sawtooth modulating voltage applied to the reflector of the klystron is derived from a Miller transitron oscillator which also produces a horizontal deflexion on the cathode-ray tube.

Since the deflecting voltage applied to the X-plate is also used to frequency-modulate the klystron, the horizontal deflexion is proportional to frequency,

while the vertical deflexion is proportional to the vertical amplitude.

Determination of the spectrum width and the centre frequency are obtained by means of a wavemeter pip with a pip mark oscillator for S-band measurements at 1Mc/s intervals.

The wavemeter operates in the TE_{011} mode and has a Q-factor of 15 000. The accuracy of absolute frequency measurements is ±4Mc/s or ±1·5Mc/s and of relative measurements 250 or 150kc/s. according to the bend.

ULTRASONIC TRANSDUCERS

(*Illustrated below*)

Technical Ceramics Ltd, Towcester, Northants

THESE piezo-electric transducers, manufactured for use in ultrasonic generators, are fabricated in ceramic, made principally from barium titanate in a polycrystalline form, and are impervious to moisture and unaffected by temperature. Ceramic transducers are rugged and able to handle high powers, their overall characteristics, such as coupling coefficient, dielectric constant, temperature stability, and sensitivity being generally superior to those found in other transducer materials. Transducers can be fabricated to any desired shape or size and, in addition to simple rings, plates, disks, or cylinders, transducer shapes may be combined into dual elements, mosaic arrays, series or parallel systems and other arrangements.

Electrodes, consisting of a special silver fused to the ceramic at high temperature, normally cover the whole of the two major surfaces of the transducers, but specific areas can be left bare if required. The transducers can be supplied with any type of lead or, alternatively, tinned areas can be provided so that leads can be fitted by the user.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 12 December. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: Principles of the Light Amplifier and Allied Devices.
By: T. B. Tomlinson.

South Wales Section

Date: 5 December. Time: 6.30 p.m.
Held at: Cardiff College of Technology and Commerce, Cathays Park, Cardiff.
Lecture: Voltage Stabilization.
By: F. A. Benson.

North Western Section

Date: 6 December. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester 1.
Lecture: Electronic Automation Applied to the Wind Tunnel.
By: D. S. Wilde.

West Midlands Section

Date: 12 December. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: Design of an Experimental Colour Television Receiver.
By: H. A. Fairhurst.

North Eastern Section

Date: 12 December. Time: 6 p.m.
Held at: Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Lecture: The Design and Application of Quartz Crystals.
By: R. A. Spears.

Merseyside Section

Date: 13 December. Time: 7 p.m.
Held at: The Council Room, Chamber of Commerce, 1 Old Hall Street, Liverpool 3.
Lecture: An Automatic System for Electronic Component Assembly.
By: K. M. McKee.

Scottish Section

Date: 13 December. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders. 39 Elmbank Crescent, Glasgow.
Lecture: The Design and Manufacture of Modern Capacitors.
By: J. H. Cozens.

South Midlands Section

Date: 14 December. Time: 7 p.m.
Held at: North Gloucestershire Technical College, Cheltenham.
Lecture: Radio Astronomy.
By: R. L. Adgie.

THE BRITISH KINEMATOGRAPH SOCIETY

Date: 13 December. Time: 7.30 p.m.
Held at: The Lecture Hall of the Royal Society of Arts, John Adam Street, London, W.C.2.
Lecture: A Magnetic Tape Recording System for Colour Television Signals.
By: H. R. L. Lamont.

THE INSTITUTE OF PHYSICS

Date: 12 December. Time: 6.30 p.m.
Held at: The Institute's House, 47 Belgrave Square, London, S.W.1.
Lecture: The Preparation and Testing of Diffraction Gratings.
By: L. A. Sayce.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Radio and Measurement Sections

Date: 3 December.
Informal evening on Electronics and Automation—The Use of Nucleonic Devices.
Talk by: Denis Taylor.

Measurement and Control Section

Date: 4 December.
Lecture: The Absolute Calibration of Voltage Transformers.
By: W. K. Clothier and L. Medina.
Date: 18 December.
Discussion on Breakdown in Dielectrics.
Opened by: C. G. Garton and J. H. Mason.

Ordinary Meeting

Date: 6 December.
Lecture: The Measurement of Earth-loop Resistance.
By: G. F. Tagg.

Informal Meeting

Date: 10 December.
Discussion on Unsolved Problems Arising from Automation.
Opened by: G. L. E. Metz.

Radio and Telecommunication Section

Date: 12 December.
Lecture: The BBC Sound Broadcasting Service on Very High Frequencies.
By: E. W. Hayes and H. Page.

Supply Section

Date: 19 December.
Lecture: Choice of Insulation and Surge Protection of Overhead Transmission Lines of 33kV and above.
By: A. Morris Thomas and D. F. Oakeshott.

Cambridge Radio and Telecommunication Group

Date: 4 December. Time: 6 p.m.
Held at: The Cambridgeshire Technical College, Collier Road, Cambridge.
Informal evening on Ultrasonics in Industry.
Talk by C. F. Brocklesby.

East Midland Centre

Date: 14 December. Time: 6.30 p.m.
Held at: Leicester College of Technology.
Lecture: A Transatlantic Telephone Cable.
By: M. J. Kelly, Sir Gordon Radley, G. W. Gilman and R. J. Halsey. (Joint meeting with the Leicester Association of Engineers.)

East Anglian Sub-Centre

Date: 10 December. Time: 7.30 p.m.
Held at: The Assembly House, Norwich.
Lecture: A. C. Traction.
By: A. Mandl.

Mersey and North Wales Centre

Date: 10 December. Time: 6.30 p.m.
Held at: The Town Hall, Chester.
Lecture: The Control of Nuclear Reactors.
By: R. J. Cox and J. Walker.

North-Eastern Radio and Measurements Group

Date: 3 December. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Lecture: Ship Stabilization: Automatic Controls, Computed and in Practice.
By: J. Bell.

North Midland Centre

Date: 4 December. Time: 6.30 p.m.
Held at: The Central Electricity Authority, Yorkshire Division, 1 Whitehall Road, Leeds.
Informal evening on Electronics and Automation.
Talk by H. A. Thomas.

North-Western Radio and Telecommunication Group

Date: 5 December. Time: 6.45 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lectures: A Point-Contact Transistor Sealing Circuit with 0.4 μsec resolution and
A Junction Transistor Scaling Circuit with 2 μsec resolution.
By: G. B. B. Chaplin.

North Lancashire Sub-Centre

Date: 12 December. Time: 7.15 p.m.
Held at: The North Western Electricity Board Demonstration Theatre, Darwen Street, Blackburn.
Discussion on Coordination of Overload Capacity.
Opened by: H. G. Bell.

North Scotland Sub-Centre

Date: 13 December. Time: 7 p.m.
Held at: The Electrical Engineering Department, Queen's College, Dundee.
Lecture: TRIDAC A Large Analogue Computing Machine.
By: F. R. J. Spearman, J. J. Gait, A. V. Hemingway and R. W. Hynes.

Rugby Sub-Centre

Date: 11 December. Time: 6.30 p.m.
Held at: The Rugby College of Technology and Arts.
Lecture: The Application of Digital Computers.
By: R. L. Grimsdale.

Southern Centre

Date: 5 December. Time: 6.30 p.m.
Held at: The College of Technology Extension, Anglesea Road, Portsmouth.
Lecture: Power Station Auxiliary Plant.
By: G. F. Kennedy and F. J. Hutchinson.

Western Centre

Date: 10 December. Time: 6 p.m.
Held at: The Lecture Hall, Engineering Laboratories, University Walk, Bristol.
Lecture: The Potentialities of Railway Electrification at the Standard Frequency.
By: E. L. E. Wheatcroft and H. H. C. Barton.
(Joint meeting with the Supply and Utilization Groups).

South-Western Sub-Centre

Date: 4 December. Time: 3 p.m.
Held at: The Electric Hall, Union Street, Torquay.
Lecture: Digital Computers.
By: A. C. D. Haley.

West Wales (Swansea) Sub-Centre

Date: 13 December. Time: 6 p.m.
Held at: The Conference Room, South Wales Electricity Board Showrooms, The Kingsway, Swansea.
Lecture: Germanium and Silicon Power Rectifiers.
By: T. H. Kinman, G. A. Carrick, R. G. Hibberd and A. J. Blundell.

Hatfield District

Date: 11 December. Time: 7 p.m.
Held at: The Hatfield Technical College.
Lecture: A Transatlantic Telephone Cable.
By: M. J. Kelly, Sir Gordon Radley, G. W. Gilman and R. J. Halsey.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 11 December. Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Lecture: The Independent Television Network—Its Operation and Maintenance.
By: C. E. Clinch and J. B. Sewter.

THE TELEVISION SOCIETY

Date: 7 December. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: 90° Scanning.
By: R. H. C. Morgan and K. E. Martin.

PUBLICATIONS RECEIVED

ANNUAL REPORT OF THE BRITISH STANDARDS INSTITUTION 1955-56. The number of subscribing members—commercial firms, trade associations and professional bodies—reached the record figure of 8 650, over 4 per cent higher than in the previous year. The first section of the report includes references to the progress made in the study of fuel efficiency, the completion of the first British Standard concerned with smoke abatement, and preliminary steps taken in the preparation of standards required for the development of nuclear energy. A page of statistics records that there were nearly 4 000 meetings of technical committees whose work resulted in the publication of more than 200 new standards. The British Standards Institution, British Standards House, 2 Park Street, London, W.1.

CERTIFICATES AND AWARDS is a booklet which gives details of nearly 80 operating certificates and awards issued by member societies of the International Amateur Radio Union and certain commercial journals and is reputed to be the first of its kind published in the United Kingdom. The Radio Society of Great Britain, New Ruskin House, Little Russell Street, London, W.C.1. Price 2s. 6d., by post 2s. 10d. For U.S.A. and Canada 50 cents, post free.

RADIO HANDBOOK, 14th edition, has recently been published. Chapters are included on Introduction to Radio, d.c. circuits, a.c. Circuits, Semiconductors, the oscilloscope, television and broadcast interference, etc. In all there are thirty-two chapters. Editors and Engineers, Ltd, Summerland, California, U.S.A. Price \$7·50.

TIME-SAVING NETWORK CALCULATIONS by H. Stockman gives a set of general rules for network calculations, and describes the use of Thevenin's theorem, the potentiometer method, and other techniques for the transient and steady states. Two output immittance theorems, are formulated. The Symbolic Method and Laplace Transform are used in many of the solved problems, and 280 selected literature references are included. SER Co, 543 Lexington Street, Waltham, Mass, U.S.A. Price \$1·75.