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Commentary

THE Physical Society's 43rd exhibition was held from 19 to 22 January and a description of a few of the exhibits appears elsewhere in this issue. This exhibition is now largely one of commercially produced equipment and, while the almost complete passing of the private exhibitor and the lessening of the 'academic air' the exhibition once had may be regretted, there is no denying that it is a first class display of scientific equipment. Furthermore it has the great advantage over many other exhibitions that the majority of the stands are staffed by technical personnel who are more interested in discussing the techniques employed in their equipment than in its commercial exploitation.

One of the outstanding overall impressions of the exhibition, and one which has grown steadily year by year over the last decade or so, is the extremely large part that electronic engineering, in one form or another, is playing in almost every branch of science. Indeed, at this year's exhibition the number of stands on which there were no items of 'electronic interest' were very much in the minority.

Next in number to the 'electronic' exhibits came those connected with radioactivity and these are of special significance this year since it is the birth centenary of Pierre Curie.

Pierre Curie was born in Paris on 15 May 1859, he married the young Polish chemist Manya Sklodovska in July 1895 and was run over by a dray and killed on 19 April 1906. A life span of barely forty-seven years—but how much he accomplished and saw accomplished.

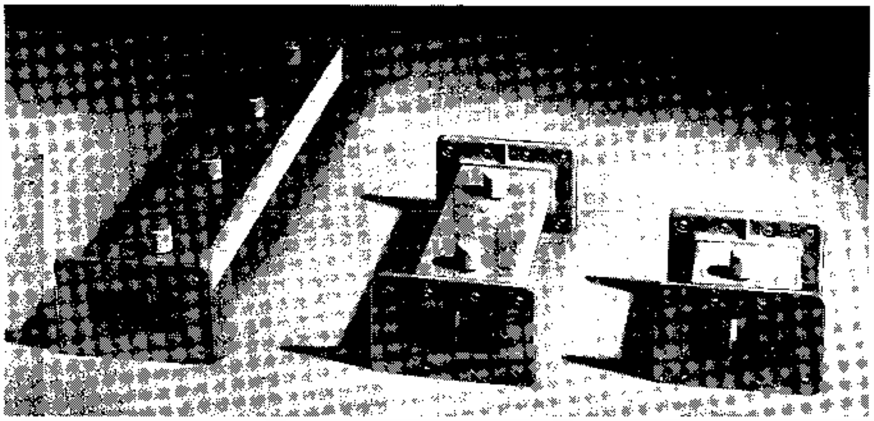
Nowadays one tends to regard Pierre Curie as merely the husband and assistant of Mme. Marie Curie and his own genius is rather hidden by the glamour that attaches to his illustrious wife. Frequently the vital part he played in the early discoveries and elucidation of the phenomena of radioactivity is overlooked, and so too are his other important contributions to scientific knowledge. Indeed, he was a physicist of no mean attainments and reputation before he met his wife; he had already made a considerable study of the behaviour of crystals in a magnetic field and, in association with his brother Jacques, had discovered the piezoelectric effect. In addition he had studied the variation in the magnetic properties of materials with changes of temperature and had discovered that at a certain temperature these properties change abruptly—what is today known as the 'Curie Point'. Rather strangely, at this time he was held in considerably more respect abroad than in his own country and one of his greatest admirers was Lord Kelvin who, in 1893, had written to Curie about his 'magnificent experimental discovery of piezoelectric quartz' and asked when it might be convenient to visit him in his laboratory: and this at a time when Curie had no more than a minor post as a lecturer in practical physics and the facilities of a very poor laboratory in the Paris Municipal School of Physics and Chemistry.

That he occupied so lowly a position was, no doubt, partly due to his highly original personality. He was a quiet reflective man of an extremely uncompromising nature and these inborn characteristics had been strengthened by his unconventional childhood. Instead of sending him to school his father had him carefully educated at home and this had helped to develop his natural tendency for profound original thinking. He had graduated at sixteen and took his Master's degree at eighteen, but, unfortunately, his unconventional training had prevented him from learning the habits and traditions of other scientists and made it very difficult for him to fit in with them. On two occasions he applied for a chair at the Sorbonne and was rejected, he also failed to gain election to the Academy of Sciences.

However, by the time of his death his genius had overcome whatever handicaps he may have had: a chair had been created for him at the Sorbonne; he had shared in a Nobel Prize; he had been elected to the Academy of Sciences; he had been invited to lecture at the Royal Institution in London and the Royal Society had awarded him and his wife the Davy medal. He was, in fact, regarded the world over as one of the greatest scientists of his day.

How much of the credit for the research into radioactivity belongs to Pierre Curie and how much to Marie Curie it is difficult to say. A great deal of the work was carried out as a joint effort and even when they were pursuing their own lines of investigation they discussed their work with one another to their mutual benefit. It is, however, certain that Pierre's part was not inconsiderable and that he was largely responsible for the researches into the physical properties of the radioactive elements and for the very fine instrumentation used in many of their joint experiments. He may, without doubt, be regarded as one of the greatest pioneers in what has become one of our most important branches of science.

Waveguide Switches and Branching Networks



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

Waveguide switches of the 'barrel' type and those actuated by shutters, gas discharge tube switches, and ferrite switches are described and discussed. A comparison of the properties of each type is made.

The properties of the branching tee in waveguide, and several of its applications are discussed in detail. A multiple branching network using hybrids is described.

(Voir page 120 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 127)

THE need frequently arises in microwave systems to switch a signal from one waveguide to another. The features of such a waveguide switch vary considerably with the application in which it is to be used, and the purpose of this article is to describe a number of waveguide switches in current use, indicate their properties, and show how these properties fit them for particular applications.

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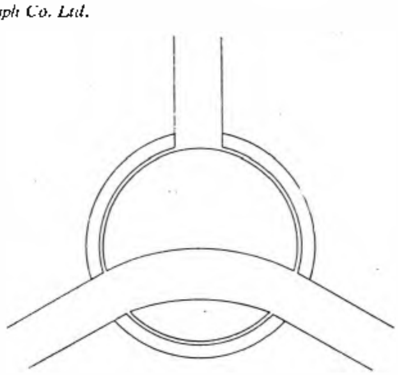
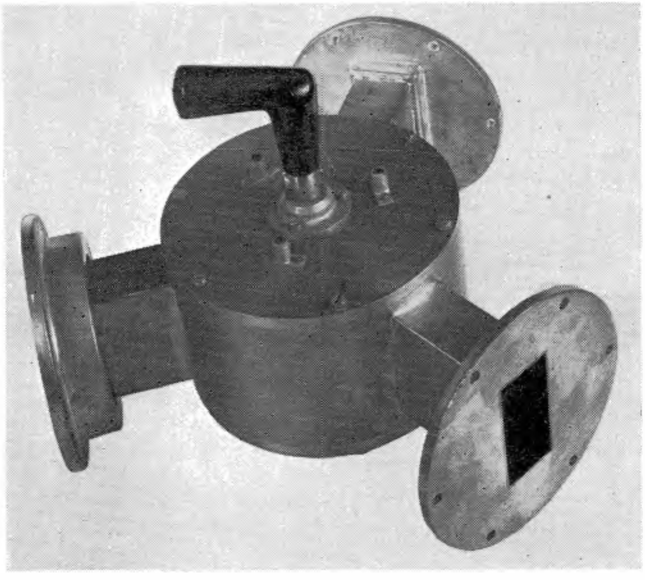


Fig. 1. Waveguide barrel switch (E-plane 3 000Mc/s)



The 'Barrel' Switch

This type of switch consists of an outer cylinder into whose walls are fixed a number of waveguides arranged radially, and an inner rotating cylinder containing channels of waveguide cross-section such that in certain positions of the cylinder two or more of the external waveguides are joined together. Fig. 1(a) makes this operation clear. This switch can be made with 3 or 4 or even more external waveguides. The channel in the inner cylinder can be a radius bend, in either the E-plane or the H-plane (as Fig. 1(a)), or an abrupt corner with suitable matching (as Fig. 2) in either plane. The required geometrical layout of the finished switch, and ease of manufacture will determine the design to be adopted. The E-plane radius type switch offers the best compromise for most purposes, and Fig. 1(b) is a photograph of a switch of this type which was designed to change over alternative transmitters into a single aerial system. In a practical switch, the r.f. match and power handling capability of the switch is considerably improved by the use of 'chokes' in the outside of the switch rotor.

Consider the design of a switch for operation in waveguide size 11 at 4 000Mc/s; in the sectional view (Fig. 3) A-B is the nominal inside dimension of the waveguide (1·122in) and B-C, C-D and D-E are each equal to a quarter of a waveguide wavelength. Since the gap X is very small (of the order ·008 to 0·010in), and the gap Y is considerably greater (mean value about $\frac{3}{8}$in) B-C and D-E are low impedance sections and C-D is a high impedance section. There is therefore an effective short-circuit between the rotor and stator of the switch at the edges of the aperture. A small discontinuity in the narrow wall of the waveguide perpendicular to the axis of the waveguide will not materially disturb the electromagnetic field, and therefore no choking is used here. As the illustration shows, the choke dimensions in this design of switch determine the overall size of

The above photograph shows 4 000Mc/s band-pass filters.

the switch, and also the centre line radius of curvature of the channel cut in the rotor. In this example this is 2·635in.

The performance of a switch of this type is extremely good. The impedance match is excellent over a very wide band (v.s.w.r. better than 0·98 over 20 per cent band), the isolation between adjacent waveguide outlets is high (exceeding 50dB over the same band) and the power-handling capacity is good. For example the switch in the photograph Fig. 1(b), in waveguide size 10 at 3 000Mc/s, withstands a peak power of at least 2MW without breakdown. Furthermore, the design is such that the whole switch can be simply sealed and can form part of a pressurized waveguide system. Disadvantages of this type of switch

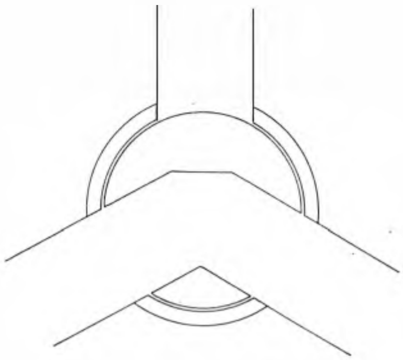


Fig. 2. Waveguide 'barrel' switch (H-plane 10 000Mc/s)

are firstly that it presents a large mismatch during the switch-over period, and secondly, since it must be located reasonably accurately in its operating position and possesses a fairly high inertia, the operating time is relatively high—say up to ½sec.

'Shutter' Type Switches

There is a number of waveguide switches with moving parts, in which the switching element is a moving shutter acting as a waveguide short-circuit. In the majority of cases, the shutter is associated with a choke flange on the live side of the switch, so that effective continuity is maintained when the shutter is not in position, and effective short-circuit when the shutter is across the waveguide, without the necessity for actual physical contact. A simple shutter type switch consists of a waveguide 'T' junction (in the example considered, an H-plane 'T') with two arms A and B (Fig. 4). When a shutter is closed at an appropriate position in arm A, all power entering the upright of the 'T' is passed to B, and a good impedance match is presented. If now the shutter in A is opened and B is closed, the power passes to B. This is a reasonably quick acting device, with a fairly good match and good isolation; it suffers from the disadvantage of breaking down at moderate powers (around 1MW at 3 000Mc/s) and having a bandwidth not more than 3 per cent for v.s.w.r. to 0·9.

Waveguide switches with considerably enhanced bandwidth can be made using the short slot hybrid[1]. In Fig. 5(a) the input to the switch (A) passes to a hybrid, in which the power is split into two equal components in phase quadrature; when the shutter is not in position, the two components recombine and the original input signal is fed to

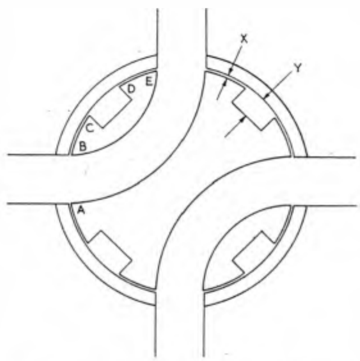


Fig. 3. Waveguide 'barrel' switch (E-plane 4 000Mc/s)

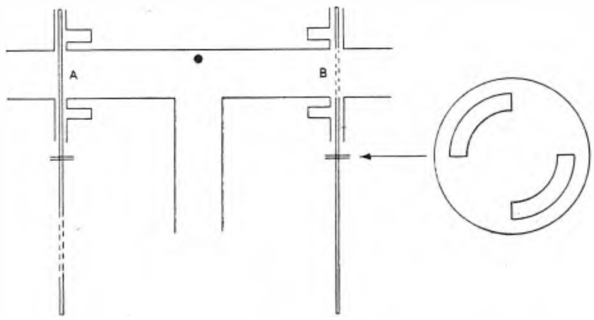


Fig. 4. H-plane 'T' and shutter

(B); when the shutter closes both components are reflected and recombine at (C). Teeter[2] describes the extension of this simple unit to a multiple arrangement of four such switches and shutters in cascade, enabling the input signal to be directed to any one of five outlets by closing the appropriate shutter.

This type of switch has a bandwidth of the order of 10 per cent to 0·9 v.s.w.r., a relatively quick action, but is mismatched at the instant of changeover. Another arrangement of hybrids with a simple switching action is shown in Fig. 5(b); a signal entering at A is divided into two equal amplitude, phase quadrature components at C and D. These will arrive at E and F and be recombined; if the path lengths C to E and D to F (via hybrid H_3 and reflected at K, L and recombined at F) differ by an integral number of wavelengths, all the power will be combined at J. (At the same time power entering the system at B will emerge at G). KM and LN are equal to a quarter of a wavelength, and consequently when the shutter is opened, the path from D to F is lengthened by an effective half of a wavelength, and the signal entering at A is switched from J to G and the signal entering at B switched from G to J. As in the previous arrangement, the speed of action of the switch depends to some extent on the design of the shutter, but in a repetitively switched arrangement the switching time can be made to be about one-tenth of the period. The switch is matched over a band of about 8 per cent, and is capable of handling moderate powers. Isolation of the order of 35dB can be achieved over a band of about 6 per cent. One function of this kind of switch is the comparison of the amplitude of two signals arriving from the same source in equal phase

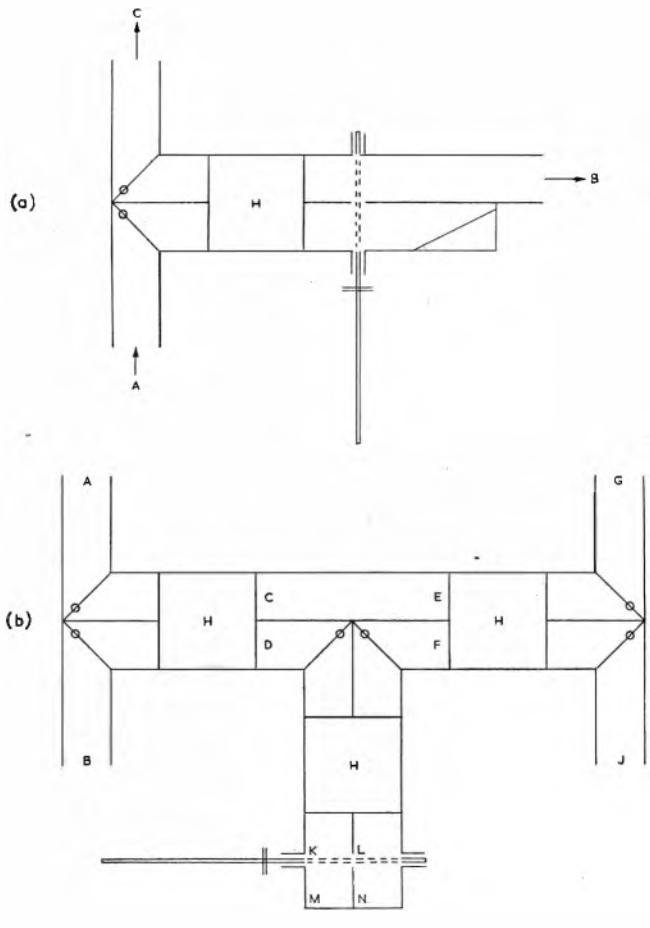


Fig. 5(a). Hybrid and shutter
(b). Three hybrids and shutter

at A and B (or the comparison in phase of signals of equal amplitude arriving at A and B). The oscillograms shown in Fig. 6 give the waveform to be expected at G for different relative levels of A and B. This gives an indication of the speed of action and the sensitivity of such a system. A typical application of this type of switching arrangement is in an automatic following radar system. Signals A and B originate from separate feeds to the same aerial such that the two beams diverge slightly either side of the aerial axis; when the aerial is on target, the signals A and B are equal, and the output at G or J is a steady level. If A and B are not equal, however, the output at G is a square wave, which can be made to operate a servomechanism, driving the aerial back on to target. The phase of the square wave determines whether the correction is 'left' or 'right', as the waveforms show.

Gas Discharge Switches

There is a group of waveguide switches operated by gas discharge tubes. The photograph and sketch, Fig. 7(a) and 7(b) show a simple on-off switch comprising a short slot hybrid, gas discharge tube and twin waveguide loads, which has been used most satisfactorily to 'pulse' a microwave signal. The gas discharge tube is matched to the waveguide in the 'cold' condition. The tube itself is made from precision glass tube containing argon at 10mm of mercury pressure, but alternative fillings may be used for special applications. It is mounted in a close-fitting brass tube

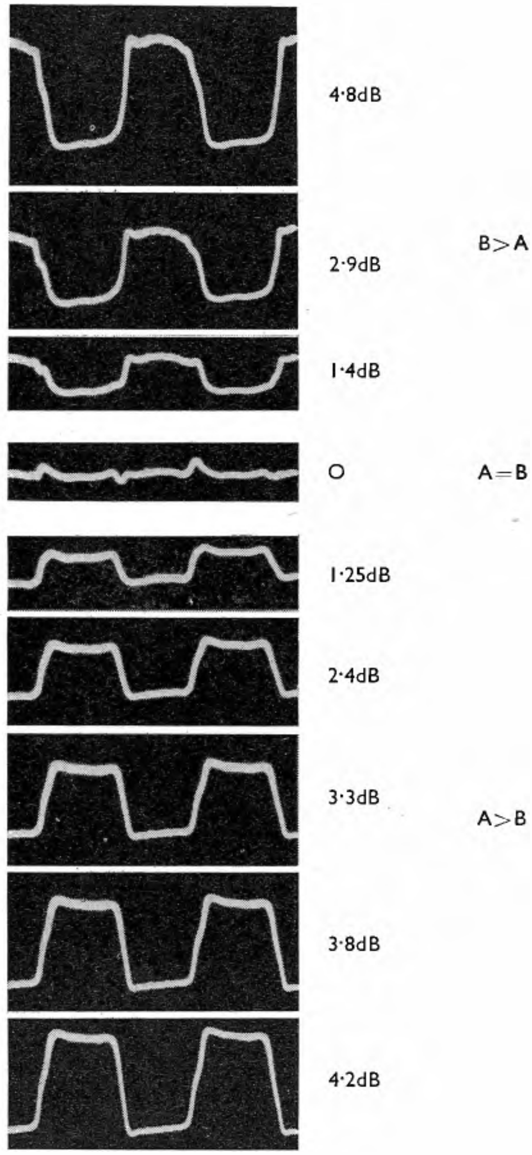


Fig. 6. Oscillograms of hybrid switch shown in Fig. 5(b)

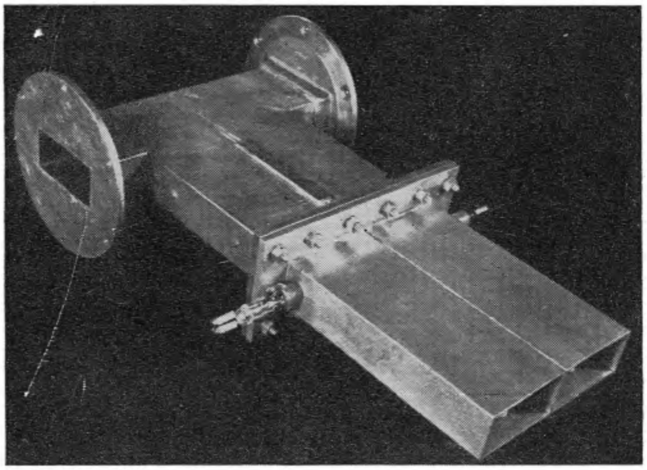
across both waveguides in the H-plane. The gap between the mounting tube and the upper and lower waveguide walls is filled with a brass plate; the brass mounting tube is slotted to give a match at the operating frequency with a suitable Q.

In Fig. 7(a), power entering at A is divided in the hybrid, and when the tube is cold, will be dissipated in the loads. When the tube strikes, however, the power is reflected, recombines in the hybrid and is transmitted at B. This device is somewhat narrow band—2 per cent for an isolation of

33dB, the forward loss is less than 1dB and the match is better than 0·95. The switch is capable of rapid action, and the only limitation is the striking and recovery of the gas filling. With suitable filling, switching times of less than a microsecond are easily achieved.

Ferrite Switches

There is a fairly large group of waveguide switches which depend for their action on the Faraday rotation of the plane of polarization of an electromagnetic wave by a ferrite in a circular waveguide. Since this type of device forms a separate subject of its own, only the briefest description



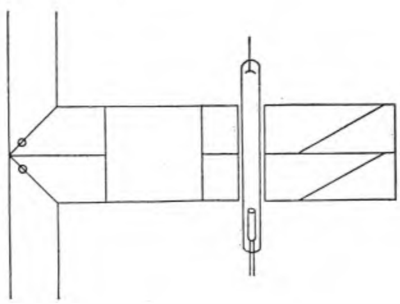


Fig. 7. Gas discharge tube and hybrid

will be given in the present article. A linearly polarized wave is propagated in a circular waveguide; it can be coupled to either of two rectangular waveguide outputs, according to its plane of polarization, which is itself determined by the external field applied to a ferrite located in the circular waveguide.

The device is for low power applications, has a switching rate dependent on the rate of change of applied field (of the order of a few microseconds) and is relatively wideband. The drawback of this switch is that for a loss of 1dB or less and isolation of say 35dB, the setting of the applied field is extremely critical. An alternative arrangement of ferrite switch described by Sullivan[3] does not suffer from this disadvantage. The ferrite is located in a cylindrical cavity, and switching times of 0·05μsec are claimed with an isolation of 30 to 35dB for a wide range of applied field.

Table 1 compares the properties of the waveguide switches described above.

Branching Networks

In communication systems at s.h.f. and u.h.f. there are a number of applications for 'frequency sensitive switches', i.e. networks which can discriminate between two or more bands of frequencies, and 'sort them out' into individual branches of the network with a high degree of isolation. The fundamental unit of such a network is the waveguide filter; the photograph on page 64 shows typical 1, 2 and 4 cavity filters operating in the 4 000Mc/s band, and Fig. 8 shows the attenuation and v.s.w.r. characteristic of the 4 cavity filter.

The Branching Tee

The simplest branching network is a 'T' junction in waveguide associated with two filters F_1 and F_2 tuned to bands of frequencies centred on f_1 and f_2. It is found that an H-plane 'T' is easier to match and is somewhat broader

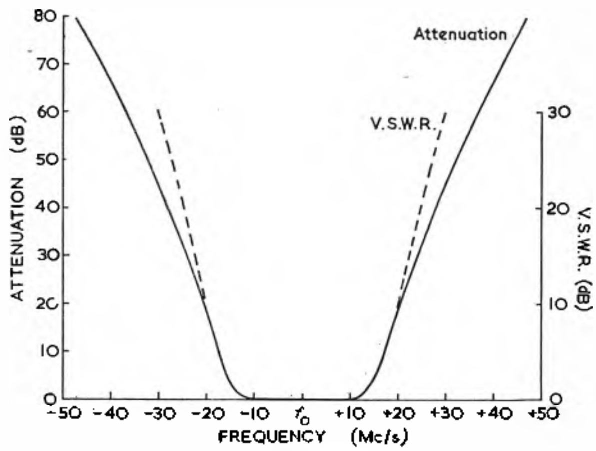


Fig. 8. Characteristics of 4-cavity waveguide filter (4 000Mc/s)

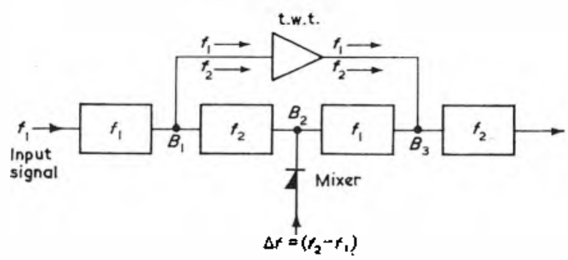


Fig. 9. Use of branching networks in reflex repeater

TABLE 1

TYPE OF SWITCH	POWER	BAND-WIDTH (per cent)	ISOLA-TION (dB)	SPEED OF ACTION
Barrel switch	High power	20	>50	Slow
Shutter and H-plane 'T'	Moderate	3	~40	Moderate (order of 1μsec)
Shutter and hybrid	Moderate	10	~40	Moderate (order of 1μsec)
Discharge tube and hybrid	Moderate	2	33	Fast <1μsec
Faraday rotation (ferrite)	Moderate or low	Approx. 5	Approx. 35	<10μsec
Cavity and ferrite	Moderate	Not known but essentially narrow band	30–35	Very fast (0·05μsec is claimed)

band than an E-plane 'T'. The position of the short-circuit plane of filter F_1 measured at frequency f_2 is located such that the 'T' is matched at f_2 and all power at this frequency passes through F_2, and vice-versa for frequency f_1. An obvious application of this principle is the sharing of a single aerial by two radio frequency channels. Typical performance of a branching 'T' is a v.s.w.r. of 0·95 over a 22Mc/s band at each frequency f_1 and f_2 with a maximum insertion loss of 0·5dB and isolation of the unwanted frequency in excess of 65dB.

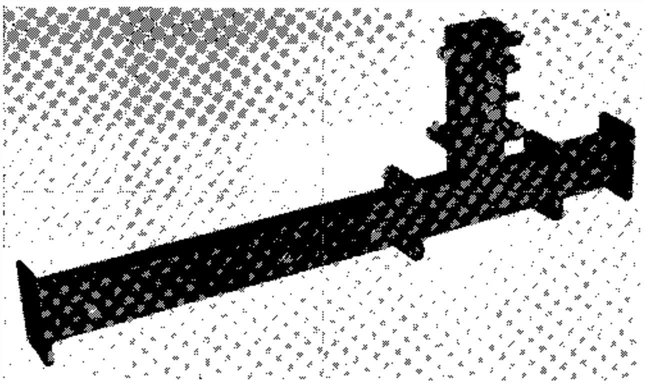


Fig. 10. Use of branching network on local oscillator feed

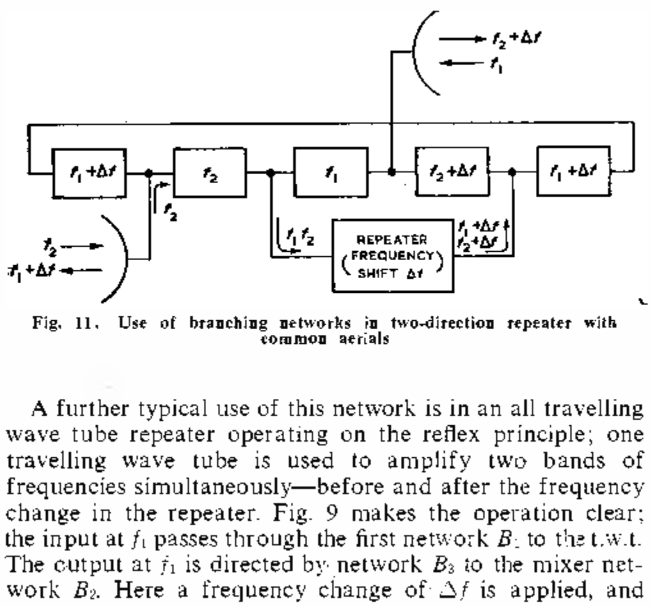


Fig. 11. Use of branching networks in two-direction repeater with common aerials

A further typical use of this network is in an all travelling wave tube repeater operating on the reflex principle; one travelling wave tube is used to amplify two bands of frequencies simultaneously—before and after the frequency change in the repeater. Fig. 9 makes the operation clear; the input at f_1 passes through the first network B_1 to the t.w.t. The output at f_1 is directed by network B_3 to the mixer network B_2. Here a frequency change of Δf is applied, and the new frequency f_2 is passed back via B_1 to the t.w.t. Since the t.w.t. is a very wideband amplifier, f_2 is also amplified, and passed by B_3 to the output. The principle of using a branching network in a mixer system has also been applied to a microwave receiver at 4 000Mc/s. In one arm of the 'T' is a four cavity band-pass filter tuned to the signal frequency, and in the other arm a single cavity high-Q filter tuned to the local oscillator frequency, and in the upright of the 'T', the mixer/head amplifier assembly. In this way only a small local oscillator power is required, and yet a very high isolation is maintained between the local oscillator and signal channels. Fig. 10 is a photograph of this arrangement.

Another system in which branching networks facilitate a compact r.f. arrangement is the use of a single channel repeater for the amplification of signals passing in opposite directions but sharing common aerials. In the schematic (Fig. 11), five filters arranged in four branching networks enable signals incoming at frequency f_2 to be amplified and retransmitted at $f_2 + \Delta f$, and signals incoming from the other direction at f_1 to be amplified and retransmitted at $f_1 + \Delta f$. In a practical system for optimum operation there must be a difference in level between the two signals passing through the repeater.

Multiple Branching Network

In more complicated communication systems, it is necessary to carry several adjacent radio frequency channels in the same aerial and feeder. The composite input signal

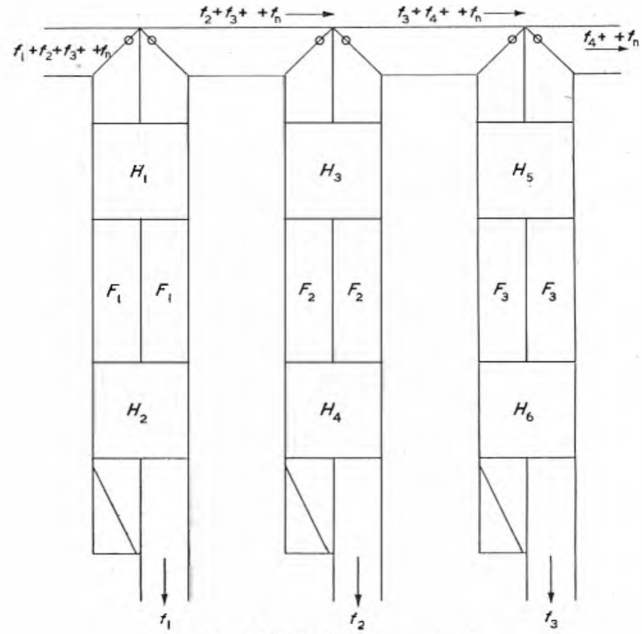


Fig. 11. Multiple branching network

containing bands of frequencies centred on $f_1, f_2 \ldots f_n$ are all passed into a hybrid H_1; two arms of the hybrid contain identical filters tuned to f_1, and two equal quadrature components of the signal f_1 are transmitted through the filters and recombined in a second hybrid H_2. At the fourth arm of H_1, the frequencies $f_2, f_3 \ldots f_n$ are transmitted.

A second pair of hybrids and filters is used to select f_2 and so on, until only f_n is left.

Short slot hybrids and waveguide filters make a particularly convenient mechanical arrangement, as can be seen from Fig. 12.

The insertion loss at any one frequency is small, since the power is shared between two filters; the isolation is high, since the filters can be made with steep rejection characteristics; the match of the system depends upon the care with which the hybrids H_1, H_3, H_5 etc. are designed and manufactured, and will obviously be somewhat worse for f_n than for f_1, but an overall figure of 0·9 v.s.w.r. can be achieved with care.

Acknowledgments

Acknowledgment is made to Marconi's Wireless Telegraph Co. Ltd for permission to publish this article. Fig. 1(b) is reproduced by the permission of Metropolitan Vickers Electrical Co. Ltd.

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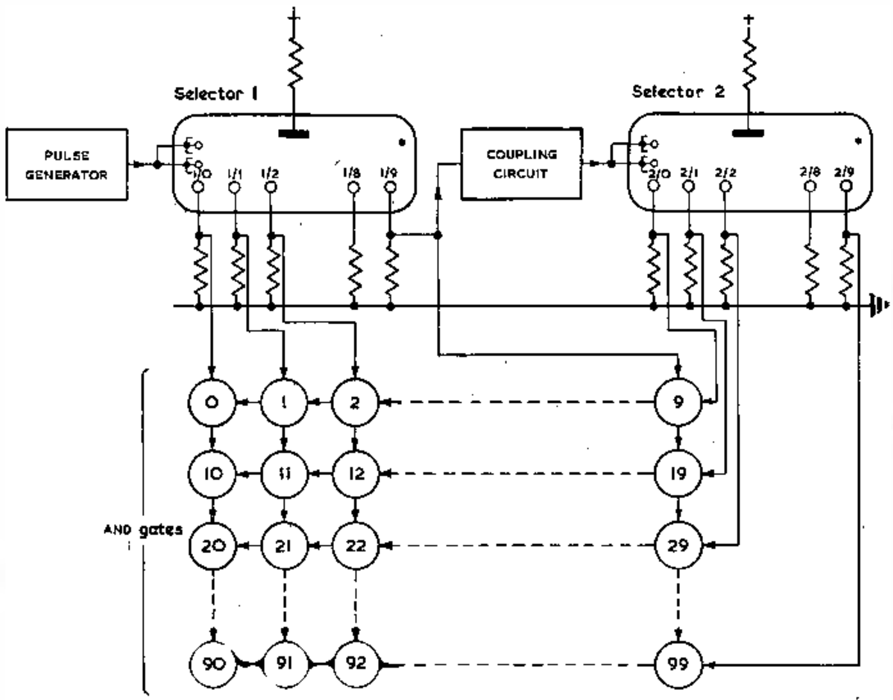
The Use of Dekatrons for Pulse Distribution

By G. H. Stearman*, B.Sc.(Eng.), A.M.I.E.E.

Cold-cathode selector tubes are frequently used where it is necessary to have a source capable of generating pulses on several output lines and in a fixed time sequence. The number of output lines is, however, limited to twelve and to increase this number it is necessary to employ special methods. In this article one well-known solution (the diode matrix method) is mentioned and two others are suggested, both of which may be advantageous in certain circumstances.

(Voir page 120 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 127)

IN certain problems involving the sequential control of a number of pulse-operated circuits, it is convenient to have available a source capable of generating pulses on several output lines and in a fixed time sequence. If the rate of generation need not be high the cold-cathode selector tube is very suitable for this purpose. The number of output lines is, however, limited to twelve (Type GS12C) and to increase this number it is necessary to adopt special methods. This note mentions one well-known solution and suggests two others which may not be widely known.



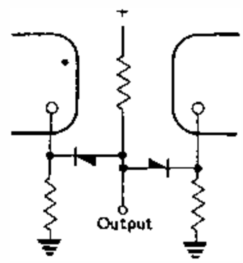


Fig. 1 (left). Matrix method

Fig. 2 (above). Typical AND gate for matrix method

Diode Matrix Method

In this method, two or more selectors are employed and the outputs combined in coincidence gates (Fig. 1). The first selector drives the second in the usual manner so that if x outputs from the first are combined with y from the second, then xy output lines are available.

Using 10-way selector tubes (GS10C) up to 100 lines may be used, a number usually more than sufficient, and the number of diodes is 200. If the full number of outputs is not required, then for repeated operation it becomes necessary to reset both selectors to their starting point when the appropriate number of pulses has been distributed. For example, if 20 lines are in use a pulse from the 20th line is used to operate a reset circuit returning both selectors to zero.

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Fig. 2 shows a typical coincidence or AND gate from which it follows that the number of diodes necessary is always twice the number of output lines. For 20 output lines, each cathode of the first selector carries two diodes and of the second selector, ten; if there is any particular disadvantage of this type of circuit it lies in the difficulty of designing these coincidence gates to give outputs adequate to operate further circuits without the introduction of buffer stages. In addition, the back resistance of the diodes may not be high, particularly if they are germanium types, and where ten are connected to one cathode, excessive loading may occur.

The two alternative methods to be mentioned avoid these particular difficulties and in some circumstances effect an economy in components.

Glow-Extinguishing Method

The principle is illustrated in Fig. 3.

Two Dekatron selectors are driven by common guide pulses and share a common anode resistor. One only of the selectors can glow at any one time; suppose that selector No. 1 contains the glow and that this invests cathode No. 1/0. Under the influence of the guide pulses the glow progresses to cathode 1/9, but as it leaves this cathode the consequent negative-going potential is used to trigger a monostable multivibrator; this generates a large negative pulse which is applied to cathode No. 2/0 of selector No. 2. This pulse is large enough to cause the glow to transfer from selector 1 to selector 2, arriving on cathode 2/0. The diode D_2 serves to isolate the output line of 2/0 from the transfer pulse.

The glow now progresses to cathode 2/9 where an exactly similar process occurs and selector 1 is again brought into operation.

This method has been tried and, with certain reservations, found satisfactory. The circuits were not developed to a state in which they could safely be handed on, so it is intended only to point out certain factors which became evident during development. These are:—

(1) The negative-going transfer pulse must be not less than 200V in amplitude or less than 8msec in length if a satisfactory transfer of the glow is to be achieved. A longer pulse may be smaller in amplitude and a larger pulse shorter in time. In other words it is the energy content which is important.

These conditions set a rather severe limit to the maximum pulse repetition rate, particularly as any attempt to reduce the length of the pulse by increasing its amplitude will eventually result in the tube ratings being exceeded, with the consequence of possible damage.

(2) There is no output possible from cathodes 1/0 or 2/0 until the corresponding transfer pulses have fully decayed and the glow is firmly established on one of these cathodes. In order that the glow shall dwell on them for a reasonable period it is therefore essential that the transfer pulse shall be appreciably shorter than the interval between guide pulses. If it is not, the glow will probably progress to cathode 1/1 or 2/1 during the transfer pulse.

(3) The reliable operation of this method depends to a certain extent on ambient light conditions. In darkness, some selector tubes require far greater energy in the transfer pulse to cause them to glow, so for reliability with randomly selected tubes a fairly high level of illumination is essential.

(4) A useful feature of the method is the simplicity with which the number of output lines may be altered. All that is necessary is the switching of the inputs and outputs of the transfer pulse generators to appropriate cathodes.

(5) The number of selectors may certainly be increased to three and possibly to more. The limit will probably be set by the accumulation of stray capacitance across the common anode resistor. A circuit employing three selectors and three transfer-pulse generators was built and found to be satisfactory.

Quiescent Cathode Method

The circuit of Fig. 4 illustrates this method. The two selectors are again driven by common guide pulses but have separate anode resistors so that both tubes glow. V_3 and V_4 comprise a bistable multivibrator circuit with cathode-follower outputs which are either slightly positive to ground or very much negative, depending upon the state of the bistable circuit.

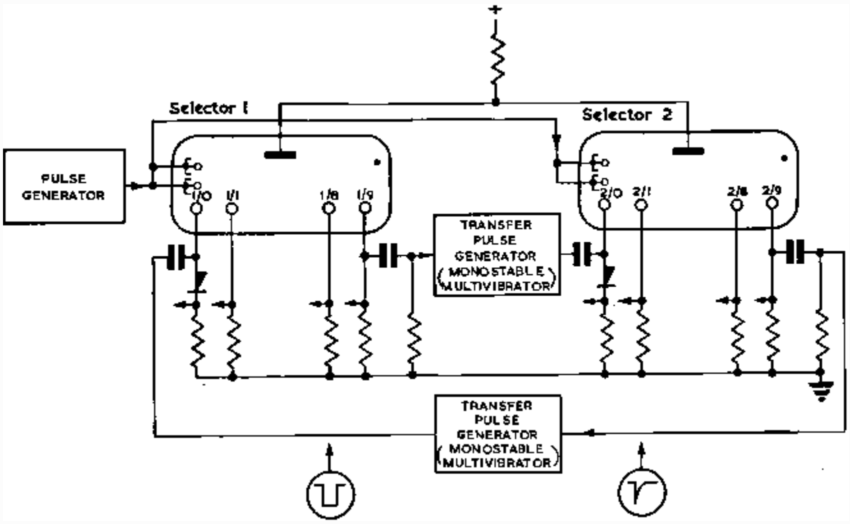

Fig. 3. Glow-extinguishing method

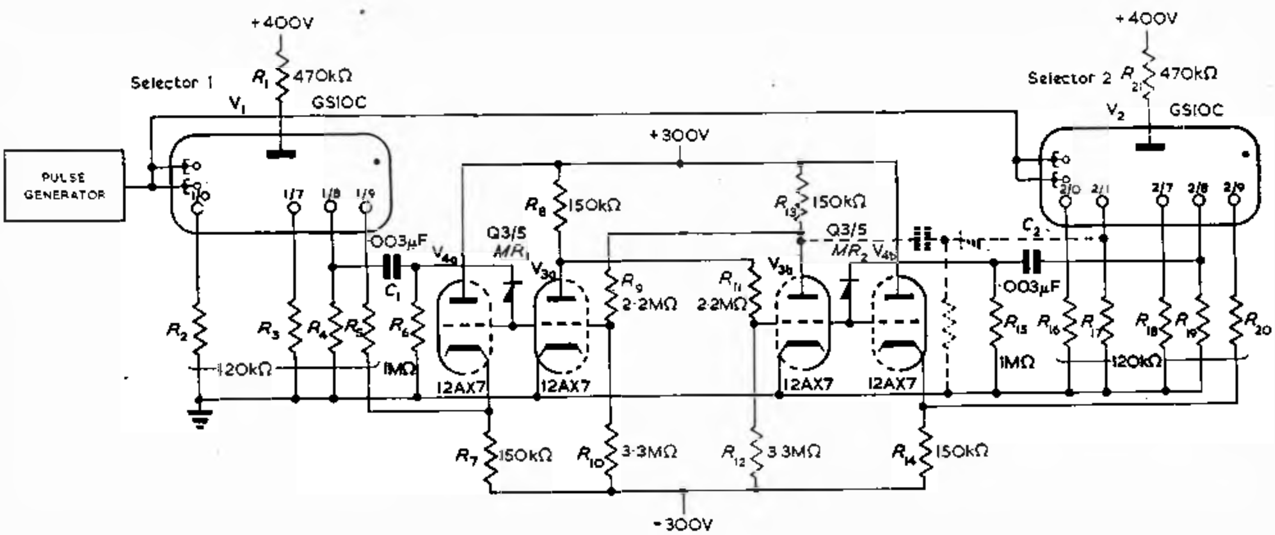

Fig. 4. Quiescent cathode method

Suppose that V_{3b} is conducting. Then the grid of V_{3a} will be at a negative potential and so also will be the lower end of the cathode resistor R_5. This negative potential suffices to lock the glow in selector 1 on cathode 1/9. The corresponding cathode of selector 2 is, however, not biased, so the glow in tube 2 is free to circulate in the usual manner.

When this glow is leaving cathode 2/8, a negative voltage pulse is generated to trigger the bistable circuit into its other stable state. Whereupon the glow in selector 2 is locked on cathode 2/9 and that in selector 1 becomes free to circulate. The reverse process occurs when this glow is leaving cathode 1/8, when the bistable circuit is restored to its original state.

Thus cathodes 1/9 and 2/9 may be referred to as quiescent and clearly no output pulses may be taken from occurs at V_{3b} anode and, after differentiation, is applied to cathode 2/1 so that the glow in selector 2 jumps almost instantaneously to cathode 2/1.

At the finish of this short 'reset' pulse guide pulse N, which is still operating, transfers the glow to cathode 2/2. Finally guide pulse ($N + 1$) gives an output from cathode 2/2. The 'skip' pulse could alternatively be taken from the cathode of V_{4a}.

This scheme is possible because the requirements for resetting a glow to any cathode in a selector already ionized, are not nearly so stringent as those mentioned earlier for the initiation of a glow in an extinguished tube.

Extension to more than 18 outputs

In theory, at any rate, any number of selectors may be connected in this fashion provided the bistable circuit is

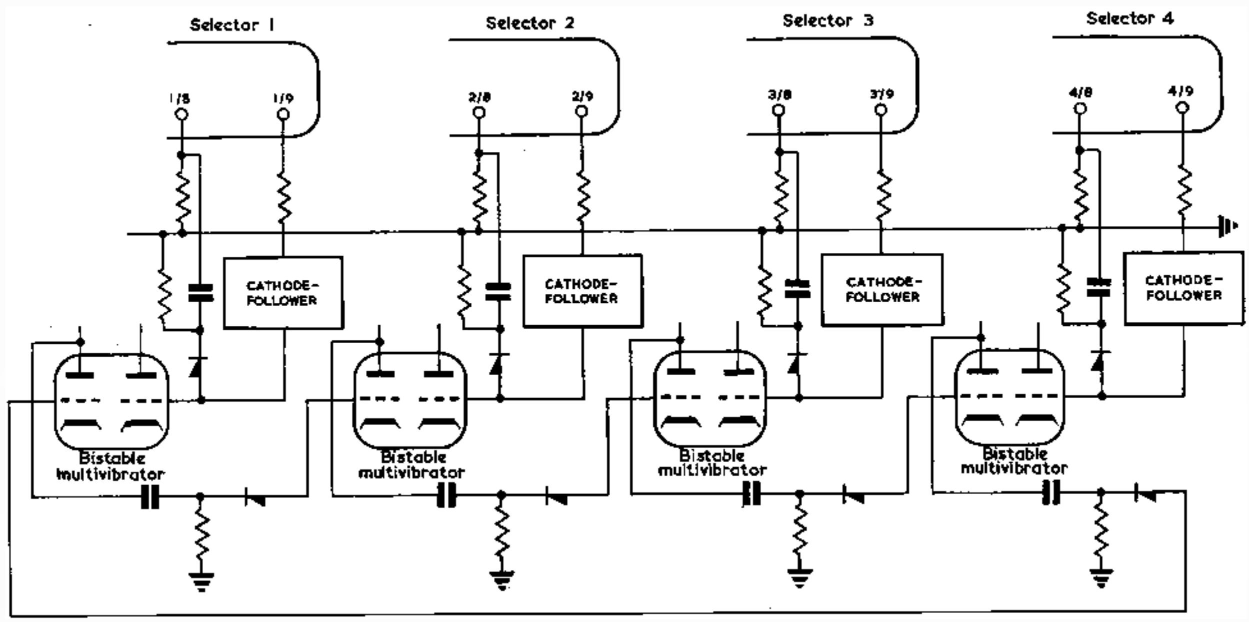

Fig. 5. Quiescent cathode method with ring counter for four selectors

them. Hence only eighteen lines are available from two ten-way selectors. There is no loss of sequential continuity at the transfer, however; for example, if guide pulse N removes the glow from cathode 1/8, then, since the glow on cathode 2/9 is released instantaneously, the same guide pulse is able to transfer this glow to cathode 2/0, from which it will be removed by guide pulse ($N + 1$). Thus the output pulses from cathode 1/8 and 2/0 are sequential in synchronism with the guide pulses.

Using less than 18 outputs

One disadvantage of the scheme is that to use a number of output lines less than eighteen it becomes necessary to devise some means to cause the glows to skip the unwanted outputs. Although this has not actually been tried, a fairly simple means, employing no more valves, suggests itself. This is indicated by the broken lines in Fig. 4.

Suppose the glow in selector 2 to be locked on cathode 2/9 and that in selector 1 to be investing cathode 1/8. Assume also that it is required to take outputs from cathodes in the sequence 1/8, 2/2, 2/3, etc., skipping 2/0 and 2/1. At guide pulse N the glow in selector 1 leaves cathode 1/8, causing the release of the glow in selector 2 from cathode 2/9. At the same time a large negative step

replaced by a ring counter having the desired number of stable states. Fig. 5 shows a possible arrangement using a conventional ring counter employing bistable stages. In a ring counter all but one of the stages are in the same state at any time. The one odd stage would correspond to the one selector whose glow is free to circulate.

An obvious further development is to replace the conventional ring-counter by another Dekatron selector, although the details of the circuits might be more difficult to arrange.

Clearly a point is soon reached where the scheme becomes uneconomical in valves and components and where a return to the matrix method is indicated, even if the latter does demand buffer stages to follow the diode coincidence gates.

Conclusion

No special advantages are claimed for the two schemes proposed, but readers may well find that they suit some particular applications better than the coincidence method.

The quiescent-cathode scheme has been found particularly satisfactory and reliable in a two-selector circuit, and as may be seen from Fig. 4, the associated circuits are rather simple.

A Regenerative Modulator Frequency Divider using Transistors

By F. Butler, B.Sc., M.I.E.E., M.Brit.I.R.E.

A description is given of a standard frequency generator consisting of a 100kc/s crystal controlled oscillator followed by two frequency dividers giving outputs through buffer amplifiers at 10kc/s and 1kc/s. The frequency dividers are of the regenerative modulator type, characterized by the fact that failure of the drive voltage results in the total cessation of output with no possibility of free-running unlocked operation.

Transistors are employed throughout; there are no preset or adjustable controls and the sinusoidal output voltages are of a convenient level for direct use without the need for supplementary apparatus. A blocking oscillator divider circuit is also described. It provides a range of frequencies, locked to the 1kc/s source, down to 25c/s.

(Voir page 120 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 127)

A PRIMARY frequency standard normally consists of a precision crystal controlled oscillator followed by a number of binary or decimal frequency divider stages, one of which drives a synchronous clock. Provision is made for comparing the indicated clock time with astronomical time. In principle, this procedure permits the measurement of frequencies in absolute terms.

A secondary standard is basically similar in layout but contains no special facilities for time comparison. It relies on the inherent stability of the driving oscillator frequency. For field and mobile use a unit of small size and low power consumption is required. It has proved possible to construct such an equipment using standard components with normal germanium transistors to give an acceptable performance in respect of battery voltage and ambient temperature changes, although considerably more development work might be required to evolve a design suitable for engineering production in quantity.

Frequency Dividers

Some standard frequency equipments employ triggered multivibrator dividers in tandem decimal stages ($\div 10$) or in rather more reliable arrangements in which the division by ten is achieved in successive stages of division by two and five. By proper design, extremely dependable operation can be secured when thermionic valves are used but it is more difficult to design transistor circuits of equal reliability. Moreover, the triggered multivibrator suffers from the defect that it will run free at an uncontrolled rate if the drive source fails.

An alternative possibility is to use blocking oscillators, triggered at some selected multiple of the output frequency. Division by ten is practicable in a single stage but once again loss of drive results in uncontrolled free-running operation. Both multivibrators and blocking oscillators should be run from well regulated power supplies.

The performance of a blocking oscillator divider may be improved by using a 'staircase' waveform generator to produce the locking trigger signal. Fig. 1 shows the essentials of the so-called diode pump circuit sometimes used to generate this special waveform[1]. As shown the voltage steps are of varying amplitude but there are several known ways of producing a linear staircase and so to achieve the most reliable triggering action. Further reference will be made to this possibility.

Continuing the discussion of the basic types of frequency divider, a third scheme involves the use of binary scaling circuits of the kind widely used in computers. Binary dividers are extremely reliable and are well adapted for use with transistors at moderate frequencies. For many purposes the scale of two is inconvenient and although it is possible to adapt a chain of these to operate decimally, great care is necessary to achieve complete reliability. Besides this the circuit employs a large number of components.

When operating at relatively low counting rates, cold-cathode tubes have advantages, especially when used in conjunction with thermionic valves. They are less convenient for use with transistors, although at least one reliable circuit has been developed[2].

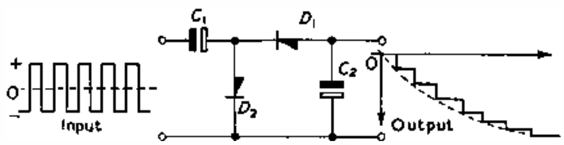

Fig. 1. Diode integrator

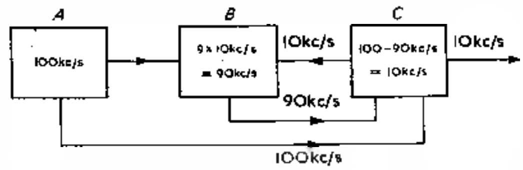

Fig. 2. Principle of regenerative modulator divider

For one reason or another, none of the dividers so far described is entirely suitable for use in transistorized apparatus although the only objection to the binary circuit is that it is expensive in components. It remains then to consider one more possibility, the regenerative modulator circuit.

Regenerative Modulator Dividers

The mode of operation of such devices may be understood by reference to the schematic arrangement shown in Fig. 2. Unit *A* is the primary 100kc/s oscillator which supplies a mixer-amplifier unit *B*, tuned to 90kc/s. The combined output from *A* at 100kc/s and *B* at 90kc/s is applied to the mixer-amplifier *C* of which the output circuit is tuned to the difference frequency 100-90 = 10kc/s. The 10kc/s output from *C* is fed to *B* where its ninth harmonic is selected to give the desired 90kc/s output on which the rest of the action depends. The ninth harmonic is of very low level and it may prove more convenient to generate the required output by supplying the mixer *B* with 100kc/s from the crystal oscillator so that a 90kc/s difference frequency is produced by mixing with the 10kc/s fundamental.

More generally, let F be the primary drive frequency, n the division ratio and f the desired output frequency so that $f = F/n$. A little consideration will show that the harmonic amplifier-mixer unit B must have an output stage tuned to the frequency $F - f$ or to $F - F/n$, i.e. to $[n - 1)/n]F$.

At first sight it is difficult to understand how the circuit action starts, although easy to see how it is sustained. Initially there is only a single input of 100kc/s from A to B and C but in fact the circuit noise is sufficient to produce a low level 90kc/s output from B which rapidly builds up as in the case of an ordinary self oscillator. Switching transients may also play a part in starting the action.

The description so far given is over simplified. It has been stated that the desired 10kc/s output is obtained by mixing 100kc/s and 90kc/s signals and selecting the difference frequency. But both high frequency sources contain harmonics and, still taking difference frequencies, it will be clear that the fundamental output at 10kc/s will be accompanied by a range of harmonics. Besides this, sum frequencies must be taken into account as they will also be present in the mixer output. However, the undesired frequencies can be partially suppressed by tuned output circuits.

The mixer stage B is supplied with 100kc/s from the crystal oscillator and 10kc/s from the mixer-amplifier C. The output is again a whole series of sum and difference frequencies and can be regarded as a 100kc/s carrier with sideband components spaced at 10kc/s intervals. There will also be harmonics of all these frequencies. Clearly there will be a multiplicity of settings of the tuned output circuit which will result in the final production of the desired 10kc/s signal. This circumstance can give rise to difficulties and ambiguities when setting up the divider circuits. The point will be taken up later, noting meanwhile that, since the whole operation depends on the continued presence of the 100kc/s drive, failure of this source will completely stop the divider action.

Practical Circuit of Regenerative Divider

Although the design of transistor multivibrators has received a good deal of attention in the literature there is very little published work on the transistor regenerative modulator. Only one useful reference can be given[3]. In Tsejtlin's paper the circuit shown in Fig. 3 is recommended and the performance is plotted graphically under various conditions, including different supply voltages and different off-tune settings of the LC circuits. The object of the arrangement is to accept an input signal of frequency F and to deliver an output of frequency f, the output being an integral submultiple of the input frequency such that $F = nf$. The circuit is a practical embodiment of the schematic arrangement in Fig. 2 and the action is as already described. The transistor X_1 has an output circuit tuned to the difference frequency $F - f$. Its output on this frequency is connected in series with the primary source of frequency F and applied to the emitter circuit of the mixer-amplifier transistor X_2. The tuned collector circuit T_3 selects the difference frequency f which is applied to the emitter circuit of X_1. The output transformer T_2 picks out the harmonic of order $n - 1$, the actual frequency being $(n-1)f$ or $F - f$ since $F = nf$.

In principle the mode of operation seems extremely simple but it is by no means easy to devise a satisfactory practical circuit. The reasons for this are:

(a) Because of the low input impedance of the transistor mixer-amplifiers the output transformer T_2 is loaded and mistuned by the transformer T_1 and the biasing components CR.

(b) The input impedance of a transistor is affected by the collector load impedance and is critically dependent on the working bias.

(c) Unpredictable phase shifts are encountered. These vary with the tuning of all the interstage transformers and with changes in transistor impedances.

(d) There are several settings of the tuned transformer T_2 which will produce the desired difference frequency f and the ambiguities are not easily detected by casual adjustments.

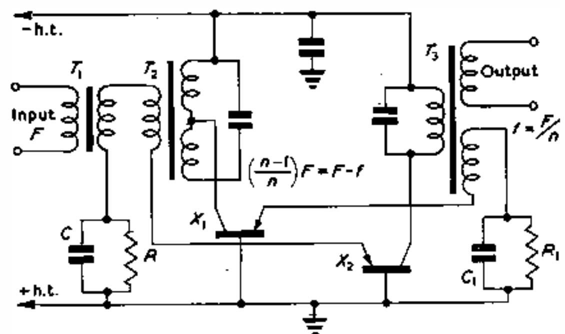

Fig. 3. Frequency divider described by Tsejtlin

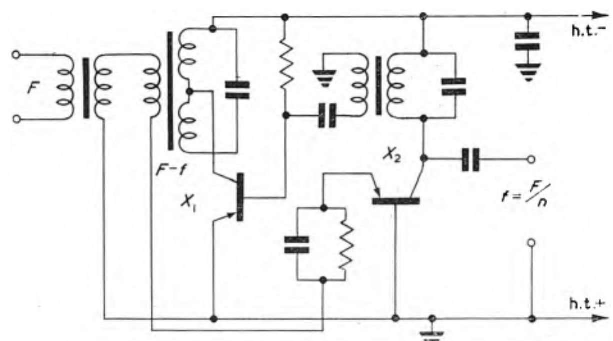

Fig. 4. Alternative circuit for frequency divider

(e) Transistor parameters are temperature dependent and somewhat sensitive to supply voltage changes.

(f) Uncontrolled self oscillation may be encountered.

This last difficulty can be avoided by proper choice and control of stage gain and phase shift of the two mixer-amplifiers. The simplest way to stabilize the performance of the equipment against temperature and supply voltage changes is to use a biased diode clamp circuit to define the peak output voltage level of the 90kc/s mixer stage.

In his paper, Tsejtlin states that the emitter drive arrangement shown in Fig. 3 proved the best in practice. In can certainly be made to work satisfactorily but several variants are possible. Using British transistors with ferrite cored transformers the arrangement shown in Fig. 4 proved rather more reliable in operation but it does not differ in essentials from the circuit of Fig. 3. The chief difference is that the first transistor is operated with the emitter earthed instead of the base. This change requires a corresponding reversal of the polarity of one of the secondary windings of the interstage transformers in order to preserve the correct phase relationship to secure regenerative operation.

Three Stage Divider Circuit

Fig. 5 shows the complete circuit diagram of a three stage divider giving outputs through buffer amplifiers at 100, 10

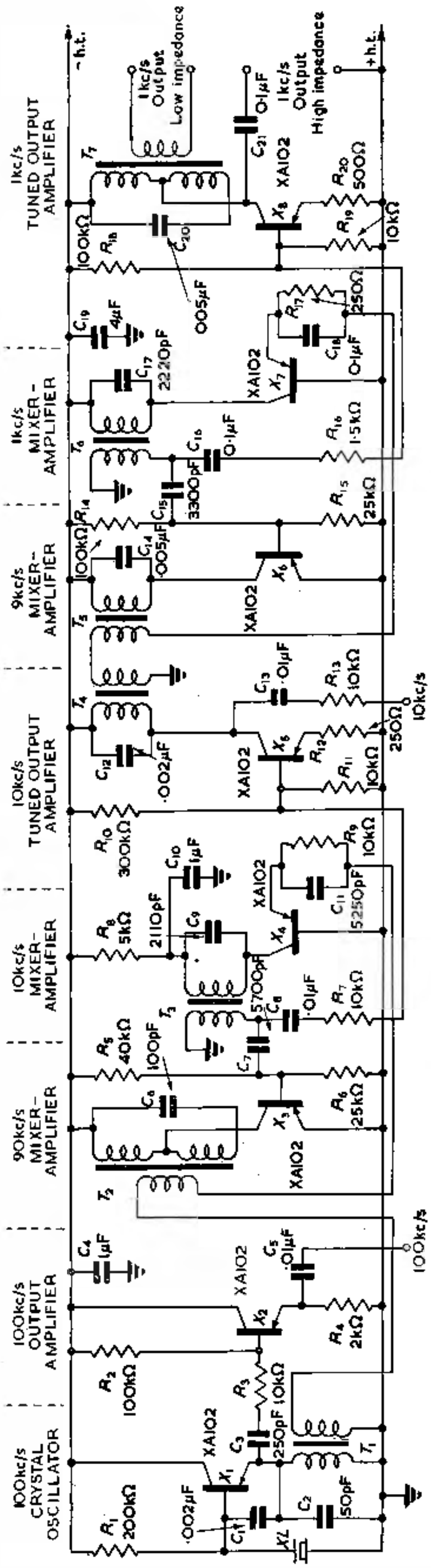

Fig. 5. Complete circuit diagram of three stage regenerative modulator frequency divider

T_1 to T_6 Ferrite grade 25, T_6 Ferrite grade 35, T_7 Nickel iron laminations.

C_1 may require selection in range 50 to 1000pF.

NOTE.—Where component values are indicated to several significant figures measured values are denoted.

	TUNED WINDING		COUPLED WINDING	
	INDUCTANCE	NO. OF TURNS	INDUCTANCE	NO. OF TURNS
T_1	40mH	400	700µH	50
T_2	4·16mH (Total)	65+65 c.t.	42µH	13
T_3	95·7mH	370	3·32mH	100
T_4	100mH	385	3·4mH	100
T_5	100mH	385	3·4mH	100
T_6	1045mH	—	5·55mH	—
T_7	5H (Total) c.t.	—	500mH	—

and 1kc/s. It must be accepted that in some cases a good deal of experimental work will be required to find the optimum values of components to ensure reliable operation over wide temperature ranges or with variable supply voltages. Once correct working is established and all output frequencies set to the desired values the operation is subsequently trouble free. Some characteristics of the equipment are as follows:

Output Voltages

100kc/s .... 1·7V }
10kc/s .... 2·8V } With 4·5V supply battery
1kc/s .... 1·8V }

Temperature Range .... 4 to 35°C

Supply Voltage Range ...... 2·5 to 4·5V

Energy Consumption

12mA at 4·5V

9mA at 2·5V

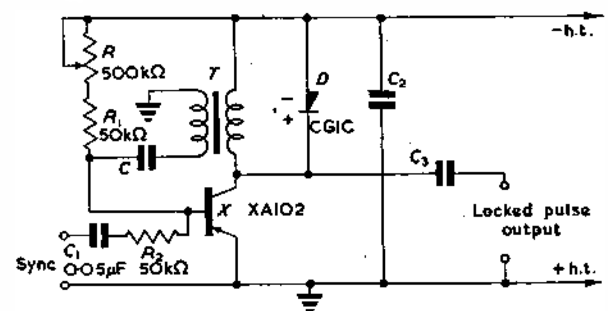

Fig. 6. Blocking oscillator

C .. 0·5µF for 25c/s, 0·01µF for 1 000c/s; C_2 .. 4·8µF; C_3 .. 0·1-0·5µF

The 100kc/s quartz crystal is an X-cut bar with gold sputtered electrodes, mounted in a glass envelope. It is not temperature controlled but was selected from a batch to produce the nominal frequency in the circuit shown at ordinary room temperature. Frequency changes due to combined voltage and temperature changes are normally within ± 20 parts in 10^6. It is probable that a much better performance could be obtained by using a GT cut plate in a series resonant mode.

Low Frequency Blocking Oscillator Dividers

Attempts have been made by the writer, without success, to produce a regenerative modulator divider giving an output at 100c/s from a 1kc/s source. However, this is readily accomplished by using a blocking oscillator. Fig. 6 gives a practical circuit. Referring to the diagram, the capacitor C is initially uncharged and the transistor is cut off. Thereafter C becomes charged through the variable resistor R and the fixed resistor R_1 until collector current begins to flow in the transistor. By means of the transformer, the base of the transistor is made more negative as a result of the changing collector current in the transformer primary winding and the collector current is cumulatively increased until the transistor is bottomed and C is charged to the maximum negative voltage.

Discharge of C through the transformer winding now initiates the reverse action, reducing the collector current and tending to produce an output voltage of opposite polarity to that of the initial pulse. It is the function of the diode D to inhibit this reverse pulse so that the output is a train of unidirectional pulses. Regular triggering is achieved by supplying a synchronizing voltage through the components C_1 and R_2. The pulse repetition frequency is primarily decided by the time-constant $C(R + R_1)$ and can be changed in coarse steps by discrete changes in C. Continuous variation is made by suitable setting of the resistance R. Triggering is reliable when the frequency of the synchronizing signal is any small multiple of the output pulse frequency.

The transformer turns ratio, inductance and winding resistance are not critical. For an output at 100c/s and a synchronizing signal of 1kc/s, typical values might be:

Primary inductance (Collector winding)	3 to 10H
Primary resistance	400 to 2 000Ω
Secondary inductance	0·5 to 2H
Secondary resistance	100 to 500Ω

The core is preferably of nickel iron but Stalloy is normally satisfactory.

Some preliminary work has been done on a variant of this circuit in which the blocking oscillator is triggered by means of a 'staircase' voltage waveform generator developed from the circuit of Fig. 1 and using a special technique[4,5] to linearize the step voltage which in the simple circuit has an exponential envelope. Fig. 7 indicates one possible arrangement for linearizing the escalator voltage and Fig. 8 shows its application to a blocking oscillator circuit in which precise triggering is required. It is hoped to publish a fuller description of this divider in the near future.

The simpler circuit of Fig. 6 is particularly useful for generating low frequency pulses from a standard 1kc/s source and can, for short periods at least, be relied upon to divide by figures as high as 40. The wide range of possible output frequencies makes it a very useful device for the calibration of low frequency oscillators at many points in the range, all frequencies having the ultimate precision of the single frequency drive source.

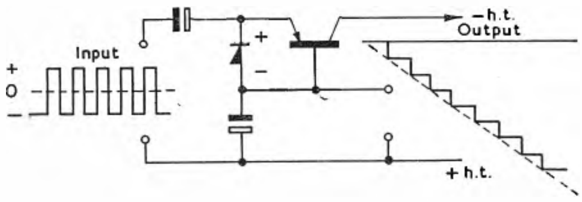

Fig. 7. Linear stepped waveform generator

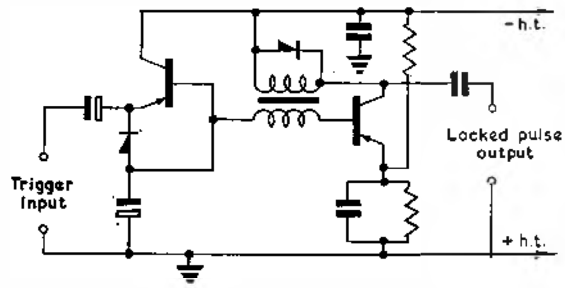

Fig. 8. Blocking oscillator triggered by escalator waveform generator

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A NEW TEACHING AID

A new aid, of American origin, for the teaching of electronics in technical colleges and universities has recently been introduced.

Known as the Erec-tronic Educational System, it is designed to replace the conventional breadboard assembly of permanently wired components used for instructional purposes.

The advantages of the breadboard are well known. The components are readily identified and the wiring is easily verified, but a student gains no experience in assembly and wiring and components cannot be changed without difficulty. Moreover, when the breadboard is used in the early stages of circuit development changes of components are usually necessary and, with frequent soldering and unsoldering, the components soon become expendable.

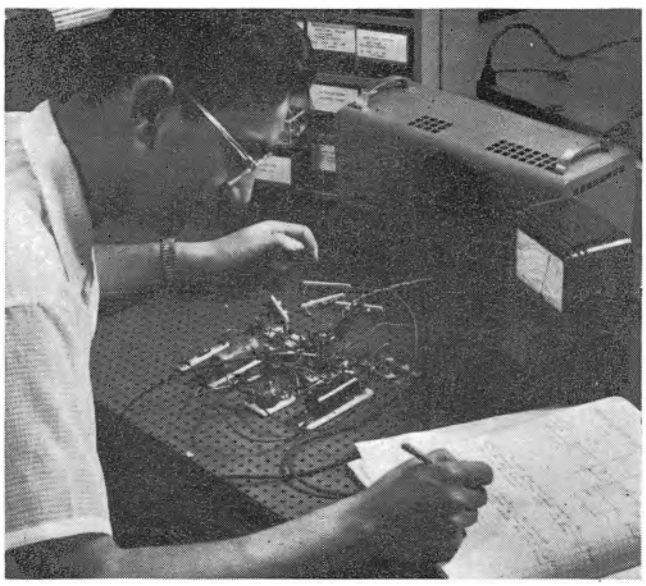
The new system consists essentially of an insulating pegboard with a large number of accurately spaced holes of standard diameter. Components such as resistors, capacitors, valveholders, etc., are mounted on polystyrene baseboards with two or more locating pins which are a push fit in the peg-board.

The components, which are connected to terminal posts on the baseboard, can thus be securely mounted and spaced according to circuit requirements.

Interconnexion between components is made by Jiffy Clip connectors.

The illustration shows the early stages of circuit development using the Erec-tronic system.

The Erec-tronic constructional system in use



Frequency Compensation for Simple Stepped Waveguide Transforming Sections

By D. Wray*, B.Sc., A.M.I.E.E.

The band-widening of stepped waveguide transforming sections is often attempted by arranging for the impedance ratios at each step to follow some prescribed (e.g.: binomial, Tschebycheff exponential) sequence. The resultant performance usually falls short of that expected because the impedance ratios are themselves frequency-sensitive, so that the desired sequence applies only at centre frequency. This article describes how the required impedance ratios can be maintained almost constant over the working band by varying both broad and narrow dimensions at each step. A substantial improvement in performance is achieved.

(Voir page 121 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 128)

A FREQUENTLY used method of 'broadbanding' components used for the change of direction or impedance in waveguide transmission is the fragmentation of the change into several discrete discontinuities and the separation of these by odd multiples of a quarter of the guide wavelength at the midband frequency[1]. The magnitudes of these discontinuities are chosen such that the reflections are as nearly as possible mutually cancelling over the required bandwidth; currently, the most highly-extolled methods of achieving this are by arranging the reflections to follow binomial[1] or Tschebycheff[2] sequences. Such methods, however, suffer from the disadvantage that, although they yield with great accuracy the ideal sequence of angles required to make up a corner, or the optimum series of impedance changes for a transforming section, these factors are themselves frequency-sensitive, so that when frequencies other than the midband frequency are being carried by the component, not only is the quarter-wavelength spacing lost between the discontinuities, but so also is the desired binomial or Tschebycheff sequence.

This effect can be combated to a certain extent in transforming sections by altering both dimensions of the rectangular cross-section at each impedance step and choosing them such that the sequence of impedance ratios is kept approximately constant over a wide frequency range.

Asymmetrically-Stepped Transforming Sections

It is often assumed that when two waveguides of different cross-sectional areas are joined with an abrupt step between them (see Fig. 1), then the impedance ratio at the step is given by:

$$Z_x/Z_y = \frac{b_x\, a_y}{b_y\, a_x}\sqrt{\left(\frac{\mu_x\, e_y}{\mu_y\, e_x}\right)}\sqrt{\left[\frac{1-(\lambda_y/2a_y)^2}{1-(\lambda_x/2a_x)^2}\right]} \quad \text{(1a)}$$

where Z is the impedance, a and b are the wide and narrow internal dimensions, μ is the permeability and e the permittivity, and λ is the unbounded wavelength in the waveguides indicated by the subscripts to the symbols. For air-filled guide this reduces to:

$$Z_x/Z_y = \frac{b_x\, a_y}{a_x\, b_y}\sqrt{\left[\frac{1-(\lambda/2a_y)^2}{1-(\lambda/2a_x)^2}\right]} \quad \text{(1b)}$$

Now this assumption is based on the steady-state equations for long waveguides, and Ragan[3] has pointed out that it does not hold precisely at step transitions where field-fringing effects are bound to modify the long-line impedance. Unfortunately, the theoretical investigation of the impedance change at an abrupt step is a complex and laborious calculation, and moreover cannot normally be carried out without employing some form of variational method yielding an answer of limited accuracy (say 5 per cent error).

The author has conducted experiments on step transitions in waveguide that show that equation (1b) can be assumed to apply, for all practical purposes, provided that the step is symmetrically disposed about a common axis of the guides on either side of the step and that the change in dimension is not greater than 10 per cent of the total dimension.

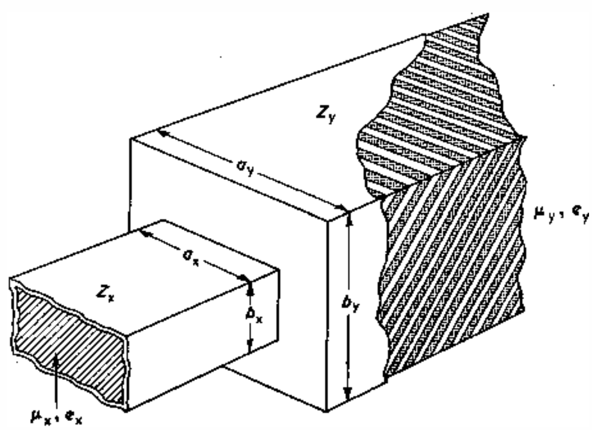


Fig. 1. Junction of dissimilar rectangular waveguides

This means that the popular method of producing transforming sections that are stepped in only one side of the waveguide is liable to error. Consider a device such as that shown in Fig. 2. It is desired that the reflection coefficients at the steps follow a sequence $1:u:v:u:1$, where $v > u > 1$, and that this is to be achieved by making a reduction in both the a and the b dimensions at the first step and then subsequently only in the b dimension at one side of the waveguide; the change in the a dimension at the first step brings it to its final value. Now since the first and last steps should introduce the smallest reflection coefficients and since changes in the a and b dimensions affect the impedance in inverse ways (see equation (1)), then the large change in a that this method of design entails must be counteracted by a large change in b to produce a small net impedance change. This is a hazardous proceeding because large steps produce impedance changes that are materially different from those predicted theoretically and, further-

* *General Post Office*

more, asymmetrical steps cause further deviation from the impedance change suggested by equation (1b). It is therefore unlikely that the combination of two large asymmetrical steps in both dimensions at the same point in the waveguide will interact so delicately as to produce the first, and smallest, reflection coefficient with any accuracy. This has been confirmed in practice: a model of a five-stepped asymmetrical transforming section having an intended binomial sequence of 1 : 4 : 6 : 4 : 1 had a worse performance than a three-stepped 1 : 2 : 1 model, although, of course, it should have been considerably better.

The asymmetrical form of construction was common in the past because it eased the fabrication of the waveguide item. Now that sprayed-metal, electro-deposition and casting methods are widely used in the production of waveguide components, this reason has lost much of its force. Indeed for many new methods, and in particular for hobbing, symmetrical components are preferable.

Binomial Matching

The method of broadband compensation to be described can be applied to any symmetrically-stepped transforming section having any desired sequence of reflection coefficients. However, to illustrate the method, the binomial matching system will be briefly discussed here so that its subsequent adaptation can be appreciated.

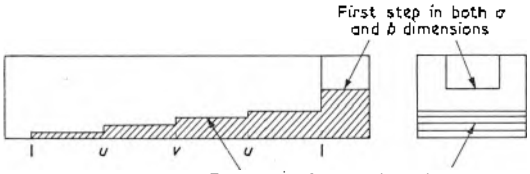


Fig. 2. Five step asymmetric transforming section

In any loss-free transmission system in which there are a number, N, of irregularities spaced equally by an electrical length θ, and having reflection coefficients ρ_1, ρ_2, $\rho_3, \ldots \rho_N$ (see Fig. 3), then the reflection coefficient of the complete system when referred to the first discontinuity is[4]:

$$R = \rho_1 + \rho_2\,e^{j2\theta} + \rho_3\,e^{j4\theta} + \ldots + \rho_N\,e^{j2\theta(N-1)} \quad \ldots\ldots (2)$$

Now, in general,

$$\cos^n\theta = \left(\frac{e^{j\theta} + e^{-j\theta}}{2}\right)^n \quad \ldots\ldots\ldots\ldots\ldots (3)$$

where n is an integer. Applying the binomial theorem to equation (3):

$$\cos^n\theta = (e^{-nj\theta}/2^n)\left\{1 + n\,e^{j2\theta} + \frac{n(n-1)}{2}\,e^{j4\theta} + \ldots + e^{j2n\theta}\right\} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

Now, if n is given a particular value, such that $N = n + 1$, and equations (2) and (4) are compared, it can be seen that R is proportional to $\cos^n\theta$, providing that $\rho_1, \rho_2, \rho_3 \ldots \rho_N$ have magnitudes that follow the sequence of ratios $1 : n : (n(n-1))/2! : (n(n-1)(n-2))/3!$ etc. The curve of $\cos^n\theta$ against θ approaches zero in the region of $\theta = \pi/2$ and becomes flatter in this neighbourhood as n is increased (see Fig. 4). If the discontinuities are arranged to be separated by a distance equal to a quarter of a wavelength at the midband frequency of the signal to be carried, then the greater the value of n the greater is the bandwidth of the signals (i.e. values of θ increasingly displaced about $\pi/2$) before any arbitrary value of reflection coefficient is exceeded. A good match is therefore provided over a wider band as the number of discontinuities is increased, providing the reflection coefficients are arranged to have magnitudes that vary in a binomial sequence.

This can be simplified in the following argument by writing:

$$r_q = kB_q \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

where r_q is the reflection coefficient at the q^{th} discontinuity, k is a constant of proportionality, and B_q is the q^{th} coefficient of a binomial expansion. If the discontinuity is a waveguide step, then the reflection coefficient depends on the waveguide impedances on either side of the step, and equation (4) can be extended to:

$$\frac{Z_{q+1} - Z_q}{Z_{q+1} + Z_q} = kB_q \quad \ldots\ldots\ldots\ldots\ldots (5)$$

where Z_q and Z_{q-1} are the waveguide impedances before and after the q^{th} step. It is therefore true to say, with only third order errors, that:

$$\ln(Z_{q+1}/Z_q) = KB_q \quad \ldots\ldots\ldots\ldots (6)$$

Fig. 3. Equally-spaced irregularities in transmission line

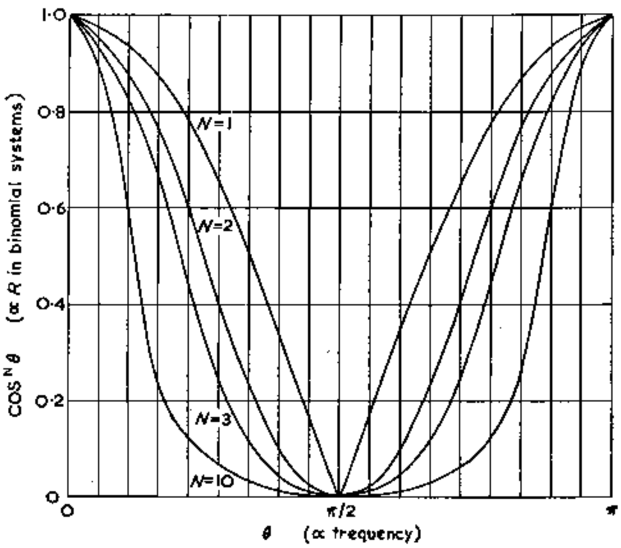


Fig. 4. Broadbanding in binomially-matched systems

where $K = 2k$, and this can be used to find the impedance ratio required at each step. The matching should improve as the number of steps is increased, so the values of B_q will depend on the maximum number of steps that can be fitted into the space available with quarter-wavelength spacing between them.

In practice, however, the performance of any item in which binomial matching is employed is rather worse than the curves in Fig. (4) would indicate. This is because the theory outlined above presumes that although the electrical length between the discontinuities is changing as the frequency changes, the binomial sequence of coefficients still remains true. Equation (1), however, shows that the impedance ratio at a step is a function of wavelength, and so unless some form of compensation is adopted, the binomial sequence only applies at the design frequency, where $\theta = \pi/2$.

Method of Compensation

In theory, there are an infinite number of a and b dimension values that combine to give any required waveguide impedance, and, since the considerations outlined previously make it desirable to alter both a and b at each step, this offers a second degree of freedom in design that can be exploited in combating the frequency dependence of the impedance ratios. This can be done by evaluating Z_x/Z_y from equation 1(b) for two frequencies near the edges of the required transmission band and working thenceforth to keep both these ratios following a binomial sequence. This leads to a pair of simultaneous equations at each step that can be solved to give a unique combination of a and b dimensions. This technique can perhaps be better understood from the following description of the design procedure.

If the transforming section is to mate two waveguides

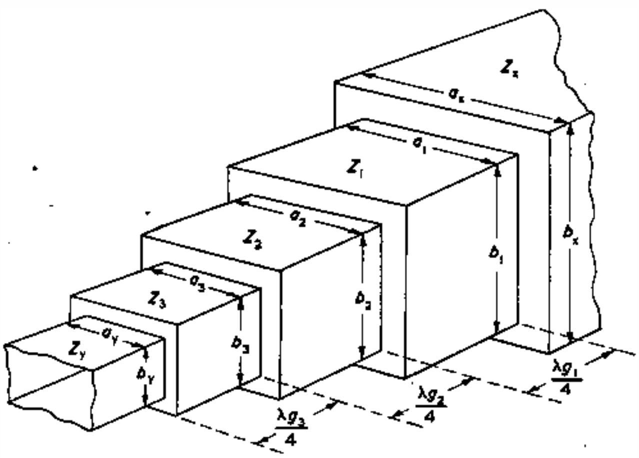


Fig. 5. Schematic of four-step transforming section

of dimensions a_x, b_x, and a_y b_y, and impedance Z_x and Z_y respectively, and the transition is to be in four steps with three intermediate impedances of Z_1, Z_2 and Z_3 (see Fig. 5), then from equation (6):

$$\left.\begin{aligned}\log(Z_1/Z_x) &= \log C\\ \log(Z_2/Z_1) &= 3\log C\\ \log(Z_3/Z_2) &= 3\log C\\ \log(Z_y/Z_3) &= \log C\end{aligned}\right\}\quad\text{(7)}$$

Hence

$$\log(Z_y/Z_x) = 8\log C \quad\text{(8)}$$

where a new constant, C, is introduced for convenience, such that $\log C = K$.

If frequencies $f_{l,u}$, near the lower and upper extremes of the required band are chosen, corresponding to unrestricted wavelengths of λ_l and λ_u then; from equations 1(b) and (8):

at f_l,

$$C_l = \text{antilog}\,\tfrac{1}{8}\,(b_y a_x/b_x a_y)\sqrt{\left[\frac{1-(\lambda_l/2a_x)^2}{1-(\lambda_l/2a_y)^2}\right]} \quad\text{(9)}$$

and at f_u,

$$C_u = \text{antilog}\,\tfrac{1}{8}\,(b_y a_x/b_x a_y)\sqrt{\left[\frac{1-(\lambda_u/2a_x)^2}{1-(\lambda_u/2a_y)^2}\right]} \quad\text{(10)}$$

C_l and C_u can then be evaluated.

The dimensions of the first step can then be calculated from these values of C from:

At f_l,

$$C_l = (b_1 a_x/a_1 b_x)\sqrt{\left[\frac{1-(\lambda_l/2a_x)^2}{1-(\lambda_l/2a_1)^2}\right]} \quad\text{(11)}$$

and at f_u,

$$C_u = (b_1 a_x/a_1 b_x)\sqrt{\left[\frac{1-(\lambda_u/2a_x)^2}{1-(\lambda_u/2a_1)^2}\right]} \quad\text{(12)}$$

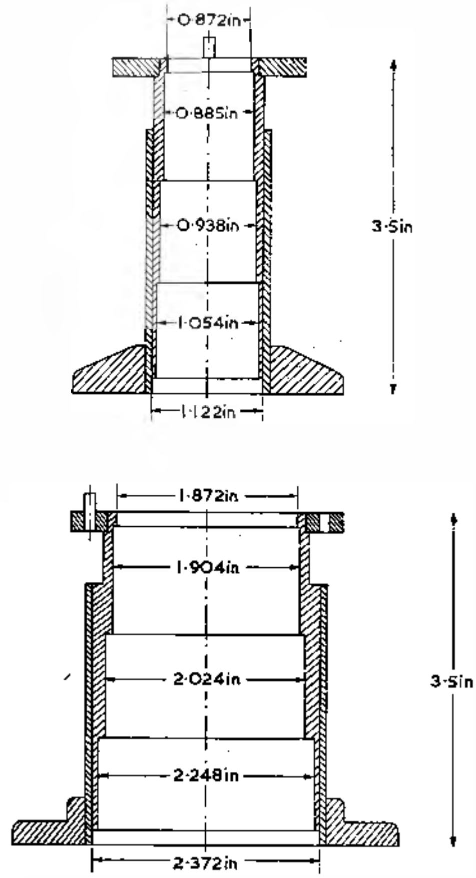


Fig. 6. Construction of simple waveguide transforming section

a_x and b_x are known dimensions, and C_l and C_u are known from equations (9) and (10). Thus equations (11) and (12) can be solved to give a_1 and b_1. Then, from equation (7):

$\log(Z_2/Z_1) = 3\log C$, thus enabling the values of Z_2/Z_1 for the two frequencies to be calculated using the appropriate values of C from equations (9) and (10). Two simultaneous equations similar to (11) and (12) can then be derived, but with a_1, b_1 as the known and a_2, b_2 the unknown dimensions. In this way, the calculations can proceed logically along the section evaluating the dimensions at each step. The accuracy of the calculation can be checked by continuing the process to the very last step, and thus ensuring that a_y, b_y are finally arrived at.

Construction and Performance

Fig. 6 shows the assembly of a four-stepped transforming section designed in the way outlined above. The transformer is intended to match to size 11 waveguide (2·372in × 1·122in) to size 12 waveguide (1·872in × 0·872in) over a band centred on 4 000Mc/s. f_l and f_u were taken as 3 800 and 4 200Mc/s respectively.

The variation of v.s.w.r. with frequency of two such transforming sections connected back-to-back is shown in Fig. 7. The v.s.w.r. of the combination is nowhere worse than 0·5dB within a ± 10 per cent range centred about the midband frequency. A slight bumping is discernible at the two design frequencies and emphasizes the analogy with conventional stagger-tuned coil transformers. Obviously the bandwidth could be extended by increasing the spacing between f_l and f_u, but this would be at the expense of the v.s.w.r. at the midband frequency.

As a comparison to the stepped transformers (which are each 3½in long), the performance of a 9in long taper section for connecting the same two waveguide dimensions is shown in the same figure.

Conclusions

By adopting a procedure that results in a performance similar to that of a double-tuned coil transformer, small stepped transforming sections can be broadbanded to give results considerably better than sections designed on a single midband frequency, and comparable in performance with much longer taper sections.

Acknowledgments

The author wishes to thank Mr. R. L. Corke for his encouragement and advice, Mr. W. R. Williams for carrying out many of the laborious measurements that this investigation entailed, and the Engineer-in-Chief of the Post Office for permission to publish this article.

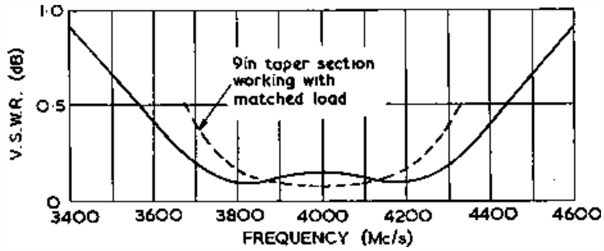


Fig. 7. Variation of v.s.w.r. with frequency of two transforming sections (as in Fig. 6) connected back to back

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Some Notes on Cathode-Ray Tubes*

Aluminized cathode-ray tubes are now well known and such tubes have the advantage that the screen is protected from ion burn while at the same time they provide a brighter picture due to the reflecting properties of the metal layer.

It is essential that the metal layer shall be very smooth and this presents a difficulty in manufacture. With screens deposited by spraying or settling, the surface formed is irregular due to the random distribution of a large number of minute crystals. To render this surface of the screen sufficiently smooth to receive the metal layer an intermediate layer, such as a film of organic material stretched over the screen surface is provided.

One usual method of forming an intermediate layer comprises introducing a quantity of an organic material such as a solution of nitro-cellulose on to the surface of a pool of water covering the screen so that the solution spreads out to form a thin film. Thereafter the water is removed, leaving the film stretched across the peaks of the phosphor particles. However, with this method extreme care has to be taken since day to day changes in temperature and general atmospheric conditions can cause changes in the viscosity and spreading power of the solution such that variations in the quality of the films produced from a single batch of such material can occur.

A variation on the above method which ensures that the conditions for the filming process are the same, tube for tube, is to form the filming material into drops each of the required amount for a single tube and to freeze these drops solid. To film a tube a frozen drop is placed on to the water pool whereby it liquifies and spreads out and covers the surface of the pool.

The use of an organic layer further complicates the manufacture of the tube since this organic material must be completely removed after the metallic layer has been evaporated thereon, otherwise the life of the tube may be impaired. Apart from the aforementioned methods of depositing screens there is a further method which consists in evaporating the phosphor material on to the screen supporting surface. Luminescent screens formed by an evaporating process appear as smooth glassy layers and metal such as aluminium can be evaporated in vacuum directly on to such screens so as to provide layers which are highly reflecting. This materially simplifies manufacture since it avoids the provision of and subsequent removal of any intermediate layer.

Reducing Reflections from Cathode-Ray Tube Screens

Reflections of ambient lighting from the screen face of a cathode-ray tube can be particularly disturbing when these reflections are towards the viewer.

These reflections can be very considerably reduced if the inner surface of the end window of the tube is rendered matt such as by frosting prior to the deposition thereon of the phosphor screen and the outer viewing surface of this end window is provided with an anti-reflection layer. The anti-reflection layer may be the same as that now commonly employed in the formation of 'bloomed' surfaces of optical components such as lenses. The combination of matt and 'bloomed' surfaces has the added advantage of not impairing the quality of the light from the screen when the glass employed for the formation of the viewing window of the cathode-ray tube is of considerable thickness.

Coloured Trace Oscillograph Tube

Multi-beam, such as double beam, oscillograph tubes are quite widely used when it is desired to have a number of simultaneous displays and quite obviously there would be considerable advantage if the displays appeared in different colours.

This could be arranged by employing the techniques which have been suggested for colour television. In one example the luminescent screen would be deposited in an orderly array of dots of different colour phosphor and co-operating therewith a perforated mask. The individual electron guns would be arranged so that their beams are guided by the mask to strike the appropriate colour phosphor.

In another example the different phosphors could be deposited in strips and have co-operating therewith a grid of parallel wires suitably biased so as to steer the beams to the appropriate phosphor.

Such a tube would provide a very clear distinction between different traces.

Improving Heater-Cathode Insulation

In expensive devices such as cathode-ray tubes additional precautions to prolong the usable life of the device is often worthwhile particularly in respect of the prevention of heater to cathode leakage or short-circuit.

This can be alleviated by inserting a very thin tube of refractory insulating material between the heater and the cathode. The heater is also coated with insulation.

Such tubes have to be of high accuracy and a simple manner of producing accurate tubes of this kind can be effected by employing a very thin tube of refractory metal such as molybdenum, which is first coated with a refractory frit and then fired to form a thin coherent coating on the inner or outer surfaces or both surfaces thereof.

* *A communication from E.M.I. Ltd.*

A Monostable Circuit Using a Transistor

By H. Stinton*

A pulse generating circuit is described which uses a simple pulse transformer and a transistor, the characteristics of which can have considerable latitude. An approximately linear increase of emitter and collector current is available, or an approximately constant amplitude voltage pulse. Low output impedance makes the device useful as a pulse forming buffer.

(Voir page 121 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 128)

THE combination of a switching element and pulse transformer, connected to give positive feedback, which forms the blocking-oscillator is well known. The circuit being considered produces a pulse of widely adjustable duration which can greatly exceed that normally associated with the pulse transformer employed.

Basic Principle

In response to the application via a transistor series switch of a voltage step-function to the series combination of a resistance and inductance there follows, subject to certain restrictions, an exponential current growth which may be treated as initially linear, and an associated potential difference across the inductance which shows a small linear

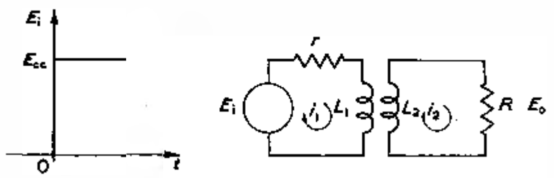


Fig. 1. Equivalent circuit

decrease over the same period. Such a potential difference may be used to generate a roughly constant secondary current which, fed back to the input of the transistor switch maintains the dynamic condition so long as the primary current rate of increase can be maintained.

The following approximate treatment indicates the behaviour of the circuit. Considering Fig. 1 and referring quantities to the primary circuit

$$E_o/n \simeq E_i \exp(-rt/L_1) \simeq E_i\,[1-(rt/L_1)]$$

assuming

(1) Transformer leakage inductance may be neglected

(2) $r \ll (R/n^2)$, where $n^2 = (L_2/L_1)$

(3) $rt \ll L_1$.

Such assumptions need not be rigorously applied, but suggest desirable circuit conditions for example to minimize output voltage droop, as may be noted from

$$dE_o/dt \simeq -(nE_i r/L_1).$$

For $t > 0$, $E_o \simeq nE_{cc}\,[1-(rt/L_1)]$

and $i_2 \simeq (nE_{cc}/R)\,[1-(rt/L_1)] \simeq nE_{cc}/R$ which is constant.

Subject to the same restrictions

$$i_1 \simeq (E_{cc}/r)\,[1-\exp(-rt/L_1)] \simeq E_{cc}/L_1$$

showing a linear increase at the rate

$$di_1/dt \simeq E_{cc}/L_1$$

which is independent of r.

Transistor Action

It is sufficient for present purposes to regard the transistor as defined by the transfer function of Fig. 2, in which I_B, V_B are the 'bottoming' values of I_c, V_c at the 'knee' of the characteristic curve of constant base current.

When connected in the circuit of Fig. 3 a pnp transistor will respond to a small negative trigger pulse, applied to its base via L_2, producing the output waveform, Fig. 4(a), associated with the sequence of transitions in Fig. 4(b).

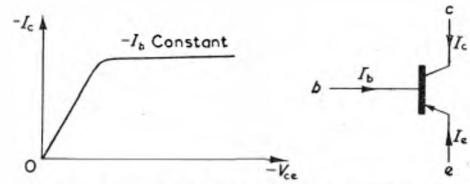


Fig. 2. Typical pnp transistor characteristic curve

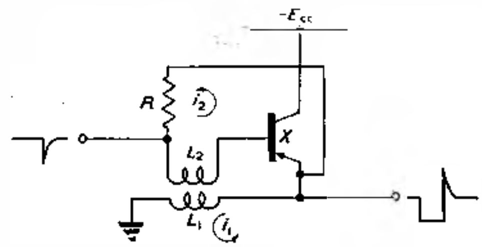


Fig. 3. Basic circuit

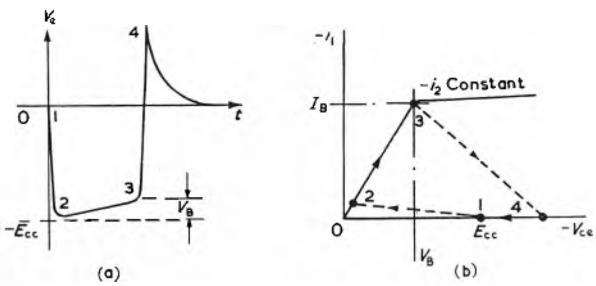


Fig. 4. (a) Emitter voltage waveform
(b) Pulse transition sequence

Circuit switching in the regions 1-2, 3-4 is dependent on transistor frequency characteristics and transformer design, however over the region 1-2 the transformer impedance is large so that the equivalent load-line intersects the appropriate characteristic curve at a correspondingly low value of emitter current.

Dynamic equilibrium is established when $i_2 = nE_{cc}/R$. Then the system is constrained to follow the curve of constant base current while the emitter current increases at a constant rate. It should be noted that during this main transition, 2-3, the transistor operates in the 'bottomed' condition in which its power dissipation is generally small.

* *Solartron Electronic Business Machines Limited*

When, at the knee of the curve, emitter current reaches the value I_B, the transistor becomes a high impedance source and the available voltage across L_1 cannot be maintained. The consequent drop of E_o and, therefore, of i_2 permits a reversal of sign of di_1/dt and the transformer voltages exhibit reverse overshoot. Provided the transformer is at least critically damped by a load, the circuit will return at the end of this overshoot to its stable quiescent condition.

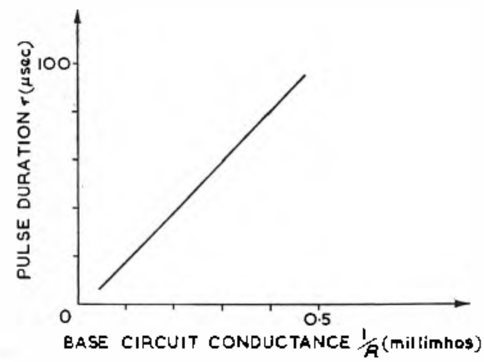


Fig. 5. Variation of pulse length with base circuit resistance
X = V6/R4, $\beta = 30$, $-E_{cc} = -4V$, $L_1 = 5mH$, $n = 1$

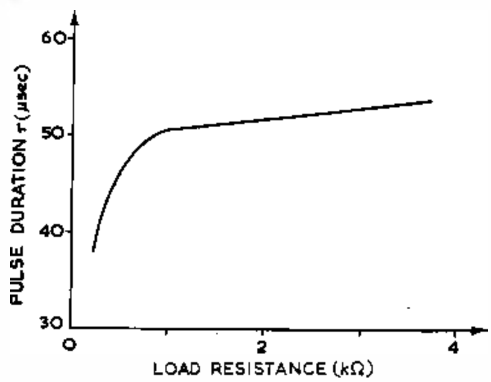


Fig. 6. Variation of pulse length with external load resistance

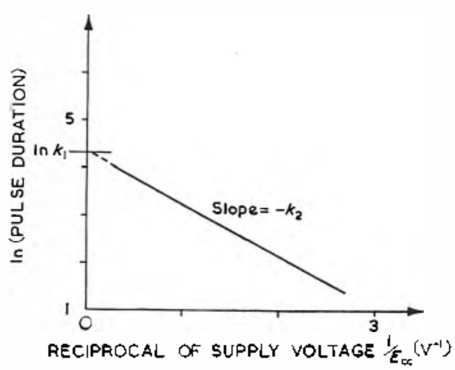


Fig. 7. Pulse length as a function of supply voltage

Pulse Duration

That the effective resistance of the transistor current source is given by the slope of its characteristic curve may be shown from

$$di_1/dV_c = -(di_1/dt)/(1/n)\; dE_o/dt \simeq 1/r$$

using the expressions derived above.

Considering again the circuit of Fig. 1

$$E_o = i_2R \simeq n\,(E_{cc} - i_1r)$$

and at the instant of pulse termination

$$i_2R \simeq n\,(E_{cc} - I_Br).$$

Let the transistor current gain $\beta = (I_B/i_2)$.

Also $V_B = I_Br$.

Hence $I_B \simeq (\beta n/R)\,(E_{cc} - V_B)$.

Now $i_1 \simeq (E_{cc}t/L_1)$

$\therefore$ Pulse duration $\tau \simeq (L_1I_B/E_{cc})$

or $\tau \simeq (\beta nL_1/R)\,[1 - (V_B/E_{cc})]$.

Some Practical Aspects

In observing variation of pulse length with supply voltage, deviation from the derived expression for τ is found unless $E_{cc} \gg V_B$. A more generally usable but entirely empirical expression is $\tau \simeq K_1 \exp(-K_2/E_{cc})$, where K_1, K_2 are experimentally determined constants. K_1 is of the order of $\beta nL_1/R$, but $K_2 \gg V_B$.

Table 1 indicates the pulses obtainable with examples of medium and low frequency transistors. In all cases the supply used was -4V, and the transformer consisted of

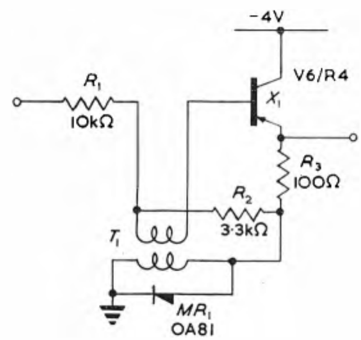


Fig. 8. Example of practical circuit

two single-layer windings, on a grade A4 Ferroxcube core, having a primary inductance L_1 = 5mH and turns ratio $n = 1$.

TABLE 1

TRANSISTOR	CUT-OFF FREQUENCY f_{co}	TYPICAL PULSE LENGTH (μsec)	APPROXIMATE RISE-TIME (μsec)
V6/R4	4Mc/s	3 to 100	½ to 1
OC76	350kc/s	10 to 300	4 to 6

Transformer overshoot peak voltage is proportional to pulse width. When wide pulses are generated an overshoot diode should be used to protect the transistor.

Some stabilization of pulse length against variation of circuit parameters, such as spread in transistor current gain, may be achieved by the inclusion of a resistor in the emitter circuit, as for example in Fig. 8.

Typical circuit properties are shown in Figs. 5 to 7. A further property worth mentioning is that the output pulse can in fact be terminated by a positive 'switch-off' pulse applied to the input.

Trigger sensitivity is usually better than $-0{\cdot}5$V peak, depending on any additional biasing arrangements, and trigger pulse width is not critical provided it exceeds 1μsec.

Acknowledgments

The author wishes to thank the Directors of the Solartron Electronic Group Limited for permission to publish this article.

The author is also indebted to Mr. J. H. Smith for his pulse transformer design, and to Mr. I. H. Mallender for experimental measurements.

A Microwave Frequency Standard

(Part 2)

By B. H. L. James*, B.Sc., and M. T. Stockford*, B.Sc.

(Voir page 57 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 61)

FREQUENCY CHECKING SYSTEM

Circular Time-Base Generators

The circular time-base generators are located on the frequency check chassis; the circuit diagram is shown in Fig. 10.

The time-bases required are at 3·7kc/s, 100kc/s, and 400kc/s; the circuits used for each are similar. A triode-pentode is used, the triode portion as an untuned amplifier feeding the pentode which has a tuned anode load. A *CR/RC* network (in which $R = n/\omega C$ at the frequency concerned, where n is the ratio of the X and Y deflexion sensitivities) is then used to obtain quadrature voltages which are selected by the check/run switch for the appropriate tube.

Instability, which might have been expected due to the common cathode in the triode and pentode sections of V_{43-45}, did not occur; the pentode decoupling is nevertheless taken to the cathode instead of to earth, to minimize this. A gain control is provided to control the circle diameter in each case.

Brightening Amplifiers

The brightening amplifiers are located on the frequency check chassis; the circuit diagram is shown in Figs. 10 and 11.

For two of the frequencies required, 100kc/s, and 350 to 450kc/s, use is made of the gain of the time-base amplifiers; for the 100kc/s the output is from the tuned circuit, and for 350 to 450kc/s from the anode of the untuned triode portion.

For the 11Mc/s amplifier (Fig. 11(a)) two stages of amplification, EF91's, are necessary; both are staggered tuned, the first to 10·9Mc/s and the second to 11·1Mc/s.

Cathode-Ray Tube Circuit

The cathode-ray tube circuit is located on the frequency check chassis; the circuit diagram is shown in Fig. 12.

The tubes used are DG4-2's with an anode potential of 750V which gives adequate focusing and brilliance for the purpose. X and Y shift controls (pre-set) are provided for each tube.

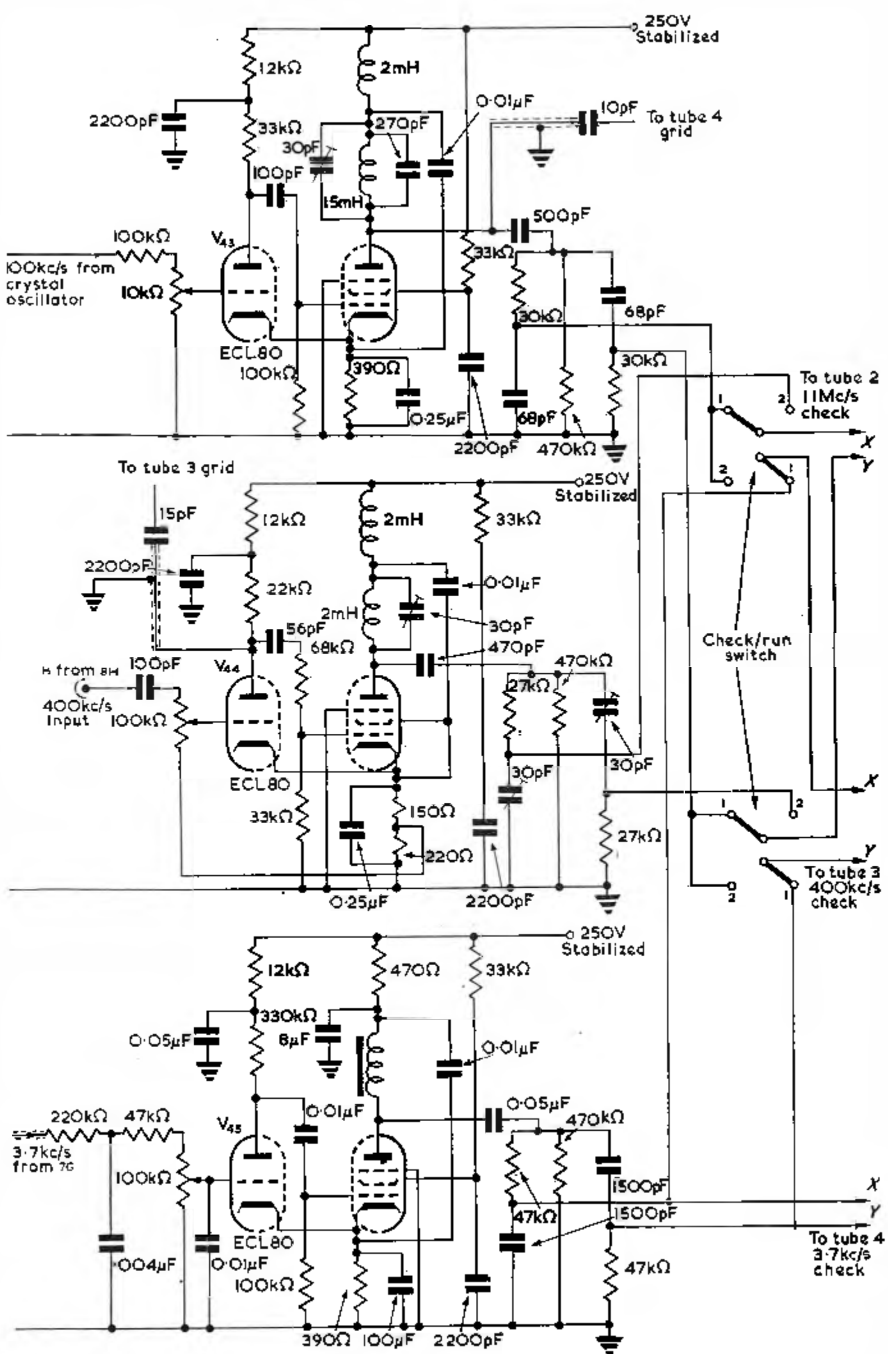
* *Mullard Ltd.*

Crystal Check Time-Base and Brightening Amplifier

The crystal check time-base and brightening amplifier are located on the frequency check chassis; the circuit diagrams are shown in Figs. 11(b) and 13 respectively.

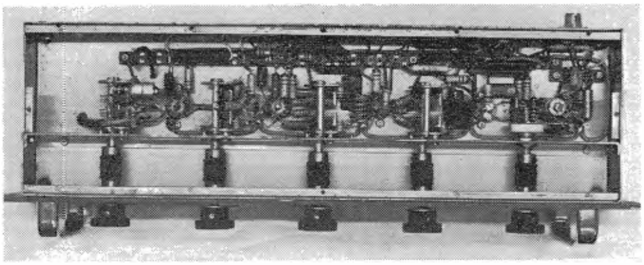
The c.r.t. displaying this check has a separate permanently connected circle generator using a phase-shift net-

Fig. 10. Circular time-base generators



work to produce the circle. The pre-set gain of the stage is variable by detuning the anode circuit.

The 200kc/s brightening amplifier V_{10} is placed on the check chassis to avoid any possibility of overall feedback in the t.r.f. receiver.



The 11-300Mc/s multiplier
(sub-chassis, with cover removed)

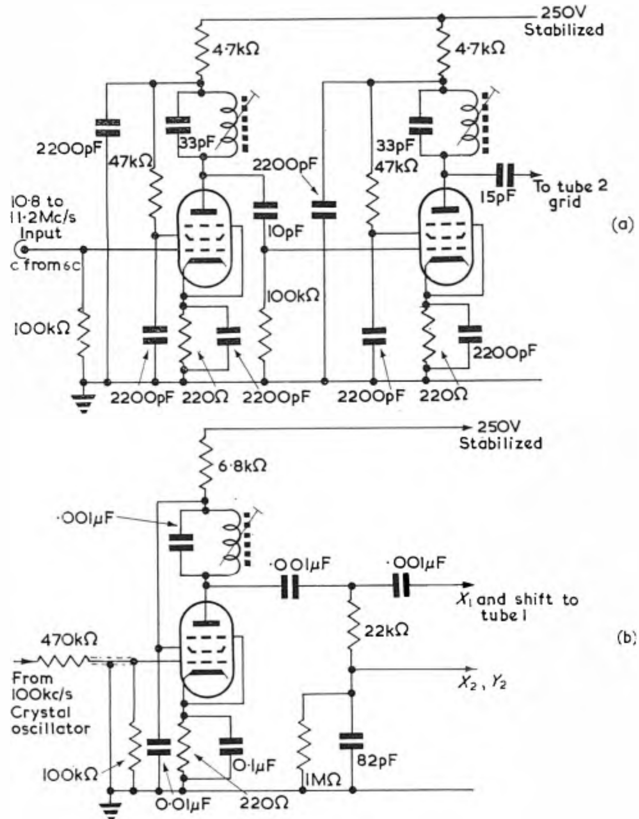


Fig. 11. 100kc/s circular time-base

Harmonic Frequency Multipliers

11Mc/s Amplifier

The 11Mc/s amplifier is shown in Fig. 14(a), V_1.

This stage has an aperiodic grid circuit giving a certain amount of grid-current bias. The anode load is a transformer with a tuned primary and untuned balanced secondary. The front panel tuning control enables this anode circuit to be tuned from 11·1 to 11·7Mc/s.

11 to 33Mc/s Tripler

The 11 to 33Mc/s tripler is shown in Fig. 14(a), V_2.

The untuned grid circuit employs a combination of grid-current and cathode bias, thus working well in a class-C condition. The anode load is tuned from the front panel by a split-stator capacitor. Two 1·5pF neutralizing capacitors (selected to be equal to C_{ag} for ECC91) are employed to remove a tendency to instability at one end of the tuning range.

33 to 100Mc/s Tripler

The 33 to 100Mc/s tripler is shown in Fig. 14(a), V_3.

The grid is capacitively coupled and the anodes are tapped one turn from the end of the anode tuned circuit to achieve a better match to the grid of the following stage.

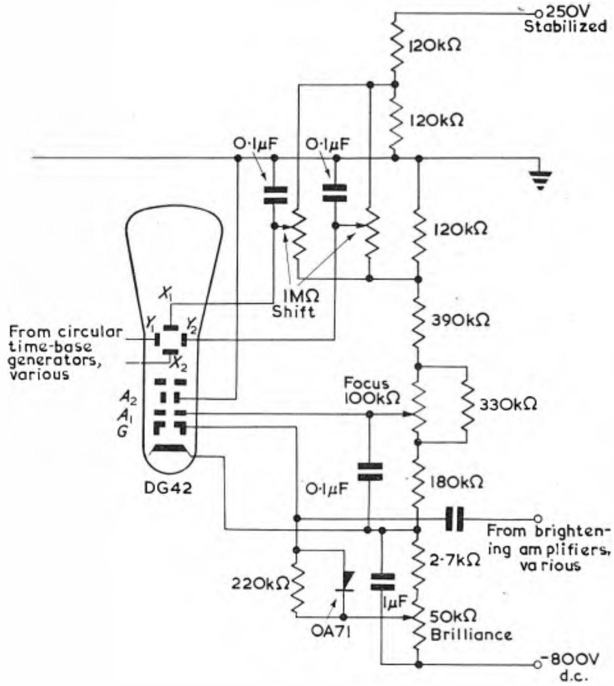
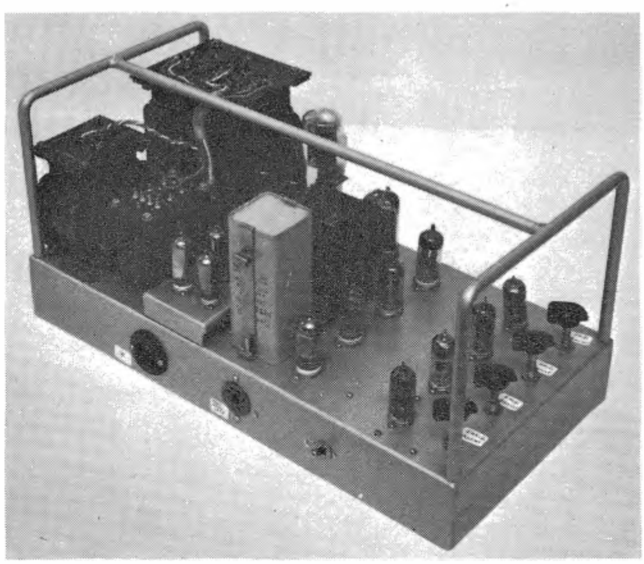
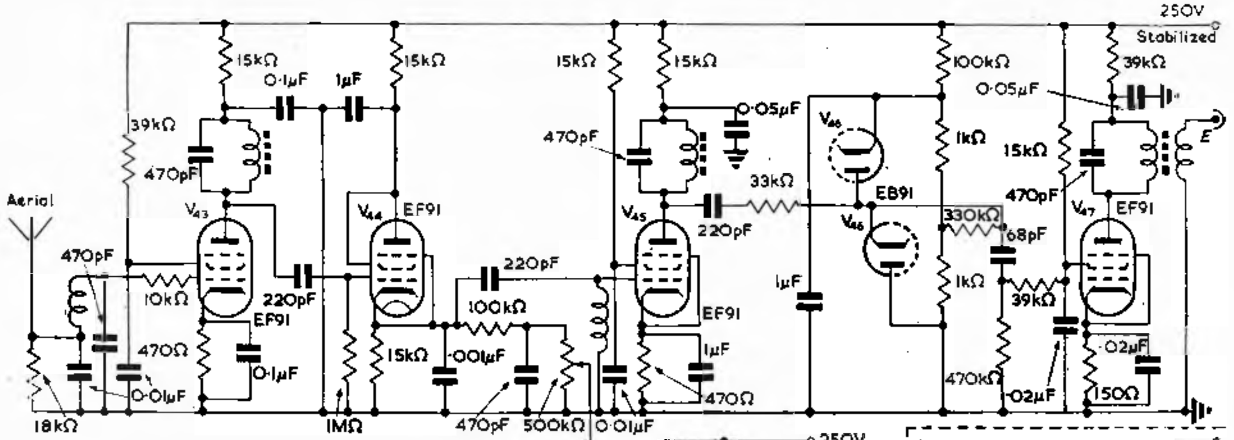


Fig. 12. Display tubes

The 100kc/s-50c/s frequency divider unit





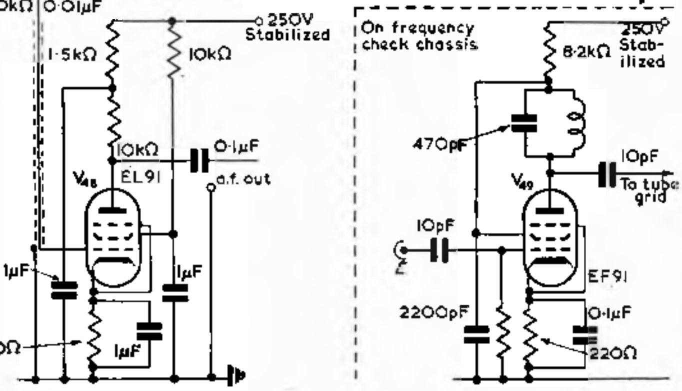


Fig. 13 (above). 200kc/s receiver

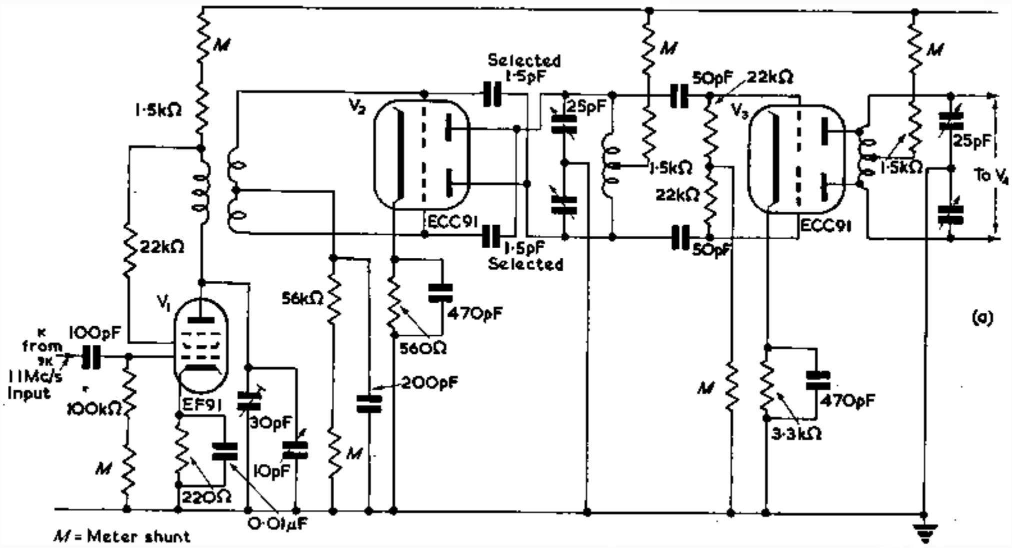
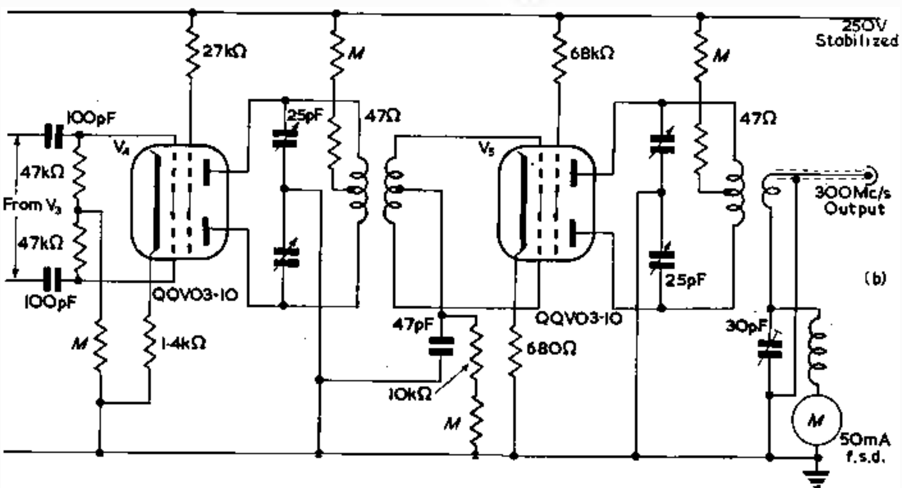


Fig. 14
(a) Frequency multiplier unit 11·4—307·8Mc/s



(b) Frequency multiplier unit 11·4—307·8Mc/s (continued)

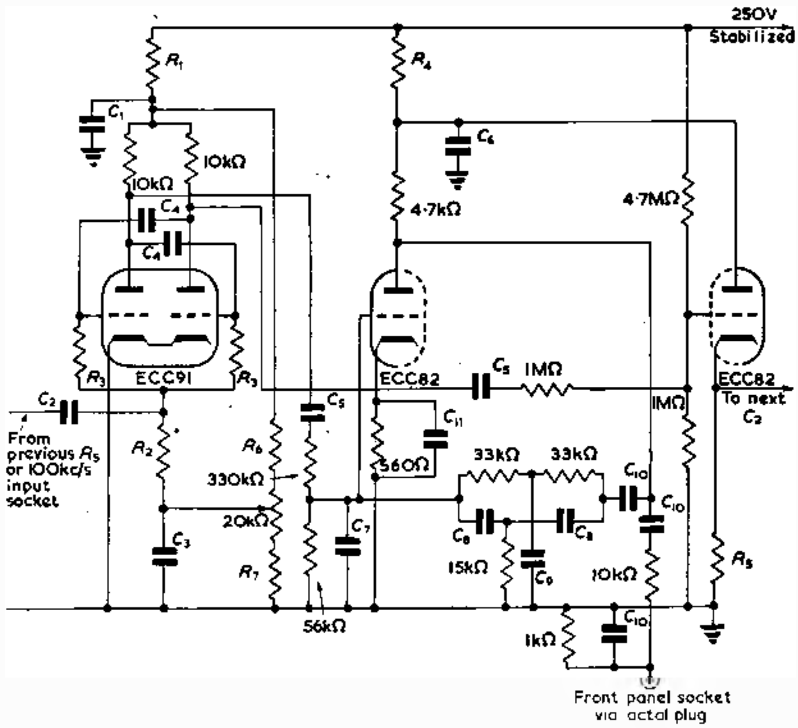


Fig. 15. Frequency divider unit

Single divider stages, 50kc/s, 5kc/s, 500c/s with selective amplifier and cathode-follower

R	50kc/s	5kc/s	500c/s	50c/s
1	8·2k Ω	6·8k Ω	4·7k Ω	4·7k Ω
2	47k Ω	22k Ω	22k Ω	22k Ω
3	100k Ω	100k Ω	100k Ω	100k Ω
4	4·7k Ω	6·8k Ω	6·8k Ω	—
5	10k Ω	33k Ω	8·2k Ω	—
6	22k Ω	68k Ω	47k Ω	47k Ω
7	180k Ω	150k Ω	150k Ω	180k Ω
C				
1	0·1μF	0.1μF	1μF	8μF
2	0·001μF	0·001μF	0·01μF	0·1μF
3	0·05μF	0·05μF	0·1μF	1μF
4	100pF	0·001μF	0·01μF	0·1μF
5	680pF	0·001μF	0·01μF	0·1μF
6	0·1μF	0·25μF	1μF	—
7	50pF	500pF	0·005μF	0·05μF
8	100pF	0·001μF	0·01μF	0·1μF
9	200pF	0·002μF	0·02μF	0·2μF
10	0·001μF	0·01μF	0·1μF	1μF
11	0·1μF	1μF	50μF	50μF

100 to 300Mc/s Tripler

The 100 to 300Mc/s tripler is shown in Fig. 14(b), V_4.

A double-triode type QQVO3-10 is employed in this stage. Internal neutralizing is incorporated in this valve probably accounting for the higher input impedance than that of the triode stages. The series resistor in the screen-grid lead is selected to make V_{g2} correct under these operating conditions.

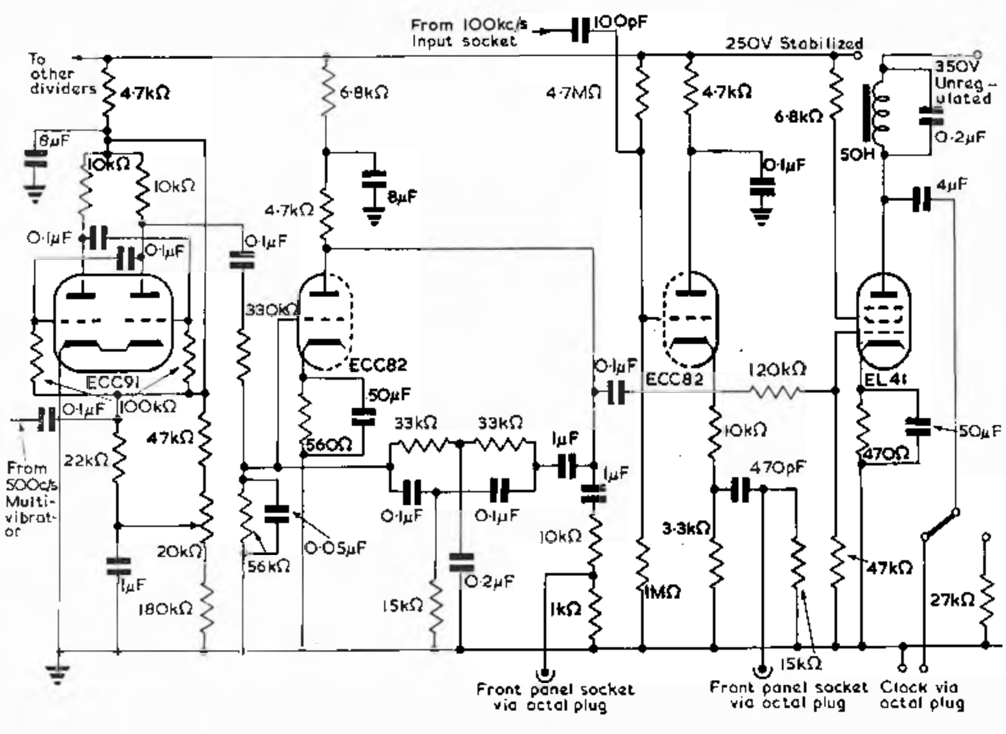
300Mc/s Amplifier

The 300Mc/s amplifier is shown in Fig. 14(b), V_5.

Due probably to an internal resonance in the QQVO3-10, this valve has a low input impedance at 300Mc/s, necessitating its being inductively coupled to the previous anode circuit. The screen resistor is again chosen to give correct voltage at this electrode. The output signal is taken from a single turn series-tuned link, coupled to the anode load, to a low impedance output connector which feeds the waveguide crystal. The rectified crystal current is by-passed through an r.f. choke and measured by a meter on the front panel. The meter may be reversed to permit crystals of differing polarity to be used.

Fig. 16. Frequency divider unit

50c/s divider, 50c/s selective amplifier, 100kc/s cathode-follower, 50c/s power amplifier



Metering

Throughout the multiplier chain small series resistors are included in the grid-earth and anode-h.t. returns to enable these currents to be monitored in each stage. A front panel uncalibrated meter may be switched across any of these, providing a check on the correct tuning of the stages.

Frequency Dividers

The frequency divider unit is illustrated in Fig. 2, 15 and 16.

Multivibrators

Frequencies in the divider chain are generated by a series of locked multivibrators separated by cathode-followers to prevent interaction between the multivibrators in the wrong direction. The free-running frequency may be adjusted by altering the d.c. level to which the grids of the multivibrators are returned, the synchronizing signal being applied symmetrically to the grids across a common resistor R_2.

Cathode-followers

The grid of one-half of an ECC82 is given a positive d.c. bias to prevent distortion of large amplitude signals, the input signal from the multivibrator anode being suitably attenuated for the next divider by the 1MΩ resistor in series with C_5.

Selective Amplifiers

To remove the harmonics from the outputs of the multivibrators each is fed to a selective amplifier consisting of one-half of an ECC82 with negative feedback through a twin-T *RC* network tuned to the appropriate fundamental. The outputs from the selective amplifiers are attenuated and taken to front panel sockets, the r.m.s. voltage at high impedance being 0·5V.

Clock Driver

This stage is a power amplifier whose input is from the 50c/s selective amplifier; the anode load is tuned to 50c/s and a self-starting synchronous clock is fed via a 4μF capacitor. When the clock is switched out, it is replaced by a load resistor to prevent excess output voltage damaging the tuned circuit.

Stabilized Power Line

The divider unit is fed from its own self-contained 250V stabilized supply so that it and the clock may be left running for long-term stability checks when the rest of the gear is not in operation. The stabilizer is of conventional design and uses EL81's as series regulator valves.

200kc/s Receiver

Except for the omission of a crystal filter the 200kc/s receiver (Fig. 13) is similar to one described in M.R.L. Report No. 163. A tuned h.f. amplifier feeds an 'infinite input impedance' detector, V_{44}, from which an a.f. output is taken to an amplifier V_{48} and thence to a front panel telephone jack. The detector is also used as a cathode-follower and a 200kc/s output is further amplified in V_{45} and limited by the double-diode V_{46}. The signal is then free from modulation except for variations corresponding to the maximum troughs which may occur, and is further amplified in V_{47}, taken by coaxial connector to V_{49} on the frequency-check chassis, and after amplification in this stage, applied to the grid of the c.r.t. to produce two bright spots on a circular time-base of 100kc/s.

Stabilized Power Supplies

For supplying the whole equipment except for the crystal oscillator unit and the divider unit, a supply giving 250V at 500mA (shared between two equal lines) is incorporated. Conventional stabilizer practice is followed, the series valves being two EL37's in parallel. A current meter and a voltmeter on the front panel can be switched to show the output of and the loading on each stabilized line.

800V at a few milliamperes for the c.r.t. monitors is obtained by full-wave voltage doubling from one of the h.t. transformers with selenium rectifiers.

Crystal Harmonic Generators, Frequency and 'Q' Measurements

Crystal Harmonic Generators

For the last stage of multiplication from the v.h.f. to the s.h.f. frequency range harmonic generation in a silicon crystal is used. Crystals suitable for this purpose are types CV2226, CS3A, etc., mounted in broadband waveguide mixer mounts for the required waveband. The maximum available power at the 30th harmonic is in the region of −60 to −70dBW and is limited by the peak inverse rating of the crystal. In the case of the CV2226 this corresponds to about 50 to 100mA of rectified v.h.f. current.

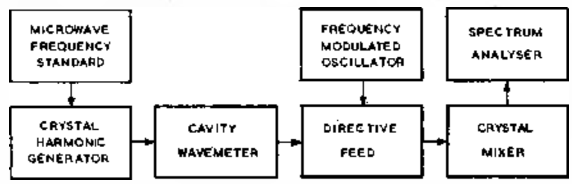


Fig. 17. Equipment for identifying harmonics

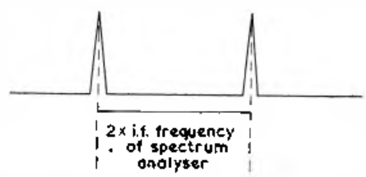


Fig. 18. Display on spectrum analyser

For operation at frequencies above 7kMc/s the 'frequency standard' gives complete coverage at intervals governed by the harmonic number and the degree of overlap of successive harmonics. It is possible to obtain complete coverage from 400Mc/s to 7kMc/s by driving the crystal multiplier from the earlier stages of the 27 × multiplier chain. Below about 3kMc/s it is convenient to work in coaxial line but it is then necessary to remove the fundamental and unwanted lower harmonics by means of filters, such as stub supports. Above 3kMc/s waveguide mixer mounts are used, the waveguide itself acting at a high-pass filter.

The 'frequency standard' was used initially only in X-band (8·5 to 10kMc/s) and the 28th to 32nd harmonics give complete coverage at intervals of not greater than 3·2Mc/s. By means of a harmonic chart the frequency may be readily determined provided that a cavity wavemeter with an accuracy of better than ±150Mc/s is available to discriminate between successive harmonics. Another suitable wavemeter for this purpose would be one of the coaxial type constructed to normal tolerances. The equipment required for identifying harmonics is illustrated in Fig. 17.

For the sake of simplicity padding attenuators have been omitted.

A display such as shown in Fig. 18 will be obtained.

The cavity wavemeter is tuned to reduce both spikes to a minimum and the approximate frequency of the 'standard' is read off the wavemeter calibration curve. The 'frequency

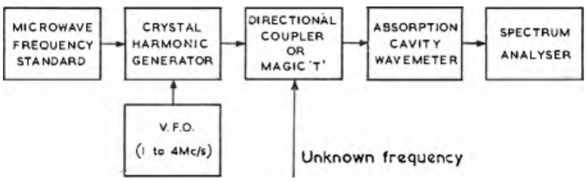


Fig. 19. Equipment for measuring unknown frequency

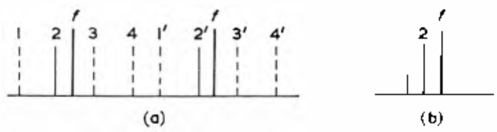


Fig. 20. Displays on spectrum analyser

standard' output frequency is read on the calibrated dials, and by reference to the harmonic chart the particular harmonic may be determined.

Use of the Frequency Standard for Microwave Measurements

In general the output level of the 'frequency standard' necessitates the use of sensitive detectors and measuring gear of the type normally incorporated in spectrum analysers. The measuring techniques described provide accurate measurement without undue complexity of equipment. For further information on these and other techniques, the bibliography should be consulted.

Measurement of an Unknown Frequency

The equipment required is shown in Fig. 19 in block form, and the measurement is carried out as follows: The spectrum analyser is tuned so that the unknown frequency is displayed and can be measured approximately with the cavity wavemeter. The 'microwave frequency standard' is then set as closely as possible to a suitable sub-harmonic of this frequency and adjusted about this setting in conjunction with the variable frequency oscillator (v.f.o.) until one sideband beats with the unknown frequency. The display will be similar to that shown in Fig. 20.

Fig. 20(a) shows the display, with the v.f.o. off, as the 'frequency standard' is adjusted in steps from 1,1', to 4,4'; these steps corresponding to $n \times 100$kc/s (where n is the harmonic in use). Fig. 20(b) shows the display as the v.f.o. is being adjusted until one sideband coincides with f. The unknown frequency is then n times the reading of the 'frequency standard' plus the v.f.o. frequency. For a fre-

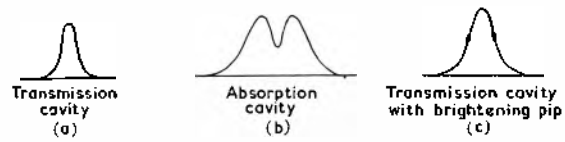


Fig. 22. Displays on spectrum analyser
(all waveforms shown with sawtooth sweep)

quency in the vicinity of 10kMc/s and assuming a v.f.o. setting of 3Mc/s the overall accuracy is 0·0003 per cent for a v.f.o. accuracy of 1 per cent.

An alternative method would be to use a tunable receiver in place of the spectrum analyser, dispensing with the v.f.o.

Measurement of Cavity Resonance Frequency and 'Q'

The equipment required is shown in Fig. 21 in block form, and the display obtained in Fig. 22.

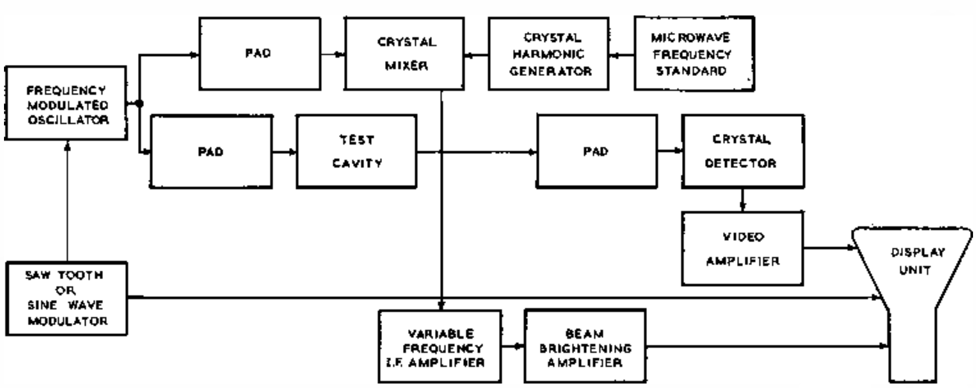
The measurement is carried out as follows. The swept oscillator is set up to cover the frequency at which calibration is required either by means of the test cavity, or initially by means of a known cavity to obtain the displays shown in Fig. 22.

The 'frequency standard' is varied until the beam-brightening spots occur symmetrically on the display pattern, using a low i.f. Since the 'standard' does not provide continuous frequency coverage the cavity has to be tuned to obtain the waveform shown in Fig. 22(c). The i.f. is then adjusted until the beam-brightening occurs at the half-power points of the cavity response curve. The cavity frequency is now read from the 'frequency standard' chart and the Q can be determined.

$$Q = \tfrac{1}{2}\,\frac{\text{Resonant Frequency}}{\text{Intermediate Frequency}}$$

The accuracy of the results is limited by the maximum gain of the video amplifier and the size of the brightening spots, the Q measurement also assuming a perfect square-law characteristic for the crystal detector. Cavity-frequency setting accuracy is high since the spot position changes rapidly with tuning at the half-power points.

Fig. 21. Equipment for measuring cavity resonance and Q



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Transistor Bias Design from Thermal Incremental Properties

By L. M. Vallese*, D.Sc.

Explicit formulae for the design of single stage bias networks are given. The use of these formulae requires the knowledge of two thermal incremental parameters, one of which is related to dI_{co}/dT and the other to the rate of change of the input current. Theoretical expressions and experimental procedures for the derivation of these properties are discussed in detail. Actual applications and experimental verifications of the theory are included.

(Voir page 121 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 128)

THE problem of the design of a transistor amplifier for prescribed thermal stability has received considerable attention in the technical literature[1,2]. In a recent paper[3] the author has discussed the analysis of the thermal behaviour of transistor circuits and has derived expressions for the thermal displacement of the operating point of a given bias network.

It is the purpose of this article to develop explicit design procedures for a number of bias network configurations, starting from a prescribed thermal displacement of the operating point. In particular, the derivation of the thermal incremental characteristics of transistors from theory and from experiment will be discussed. In the following the problem of the "spread" of the transistor parameters and of the corresponding sensitivity of the centre design formulae will not be considered.

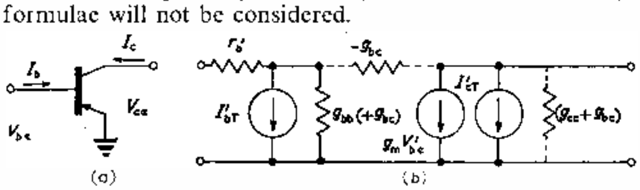

Fig. 1. Thermal incremental equivalent circuit of common emitter transistor

Theory

At a given temperature the thermal rates of change of currents and voltages are interrelated linearly by means of the 'temperature incremental equations' of the transistor[3] obtained by differentiating the static characteristics with respect to temperature. For example, for the common emitter configuration (Fig. 1(a)) these are

$$\left.\begin{aligned} I'_b &= I'_{bT} + g_{bb}V'_{b'e} + g_{bc}V'_{ce} \\ I'_c &= I'_{cT} + g_{cb}V'_{b'e} + g_{cc}V'_{ce} \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots (1)$$

where primes indicate temperature derivatives. $V'_{b'e}$ is referred to the intrinsic base, and I'_{bT}, I'_{cT} are respectively the partial derivatives of I_b and of I_c with $V_{b'e}$ = constant and V_{ce} =constant. The latter quantities may be expressed theoretically as follows:

$$\left.\begin{aligned} I'_{bT} \simeq {} & (\eta/\Lambda)(G_{ee}+G_{cc}-2G_{ce}) - (\eta/\Lambda - V_{eb'}/T) \\ & (G_{ee}-G_{ce})\exp(\Lambda V_{eb'}) - (\eta/\Lambda - V_{eb'}/T) \\ & (G_{cc} - G_{ce})\exp(\Lambda V_{cb'}) \\ I'_{cT} \simeq {} & -(\eta/\Lambda - V_{cb'}/T)\, G_{ce}\exp(\Lambda V_{eb'}) \end{aligned}\right\} \quad \ldots (2)$$

In these equations η is equal $5/2T + W_g/kT^2$, W_g is the energy gap, $\Lambda = +q/kT$ (+ for pnp, − for npn), $V_{eb'}$ $V_{cb'}$ are bias voltages, G_{ee}, G_{cc}, G_{ce} are the absolute values of the d.c. conductance coefficients. The coefficient η is typically 0·06 to 0·1 for germanium and 0·1 to 0·16 for silicon. The algebraic sign of I'_{bT} and I'_{cT} is equal to the sign of I_b and I_c respectively. Equations (1) may be represented with the equivalent circuit of Fig. 1(b).

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If the d.c. transistor characteristics are 'linearized' for the region of interest, letting (Fig. 2)

$$\left.\begin{aligned} I_b &= g_{bb}(V_{b'e} - V^o_{b'e}) + g_{bc}V_{ce} \\ I_c &= I_{ceo} + \alpha_{cb}I_b + g_d V_{ce} \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots (3)$$

where the significance of $V^o_{b'e}$ is explained in Fig. 2, and $g_d = g_{cc} - g_{bc}g_{cb}/g_{bb}$, one obtains the following additional definitions for I'_{bT} and I'_{cT} equivalent to those given in equation (2):

$$\left.\begin{aligned} I'_{bT} &\simeq \partial g_{bb}/\partial T\,(V_{b'e} - V^o_{b'e}) - g_{bb}\frac{\partial V^o_{b'e}}{\partial T} \simeq -g_{bb}\frac{\partial V^o_{b'e}}{\partial T} \\ I'_{cT} &\simeq I'_{cT}{}^* + \alpha_{cb}I'_{bT} \\ I'_{cT}{}^* &\simeq \partial/\partial T\,(I_{ceo} + g_d V_{ce}) + (\partial\alpha_{cb}/\partial T)\,I_b \end{aligned}\right\} (4)$$

In these equations $I'_{cT}{}^*$ is the partial derivative of I_c with I_b = constant, V_{ce} = constant; in addition the quantity $\partial V^o_{b'e}/\partial T$ is approximately[4] 2mV/°C. Finally in all the previous equations the intrinsic base voltage is related to the extrinsic base voltage with the equation

$$V_{be} = V_{b'e} + r_{b'}I_b \quad \ldots\ldots\ldots\ldots\ldots (5)$$

where $r_{b'}$ is the base spread resistance. From equations (3) the following thermal incremental equations (equivalent to (1)) are obtained:

$$\left.\begin{aligned} I'_b &= I'_{bT} + g_{bb}V'_{b'e} + g_{bc}V'_{ce} \\ I'_c &= I_{cT}{}^* + \alpha_{cb}I'_b + g_d V'_{ce} \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots (6)$$

The quantities I'_{bT}, I'_{cT}, $I'_{cT}{}^*$ are functions of temperature; however, only their values at room temperature, for which the bias point is specified, are needed for the design. A simple experimental procedure for this determination is described later.

It should be noted that the above thermal incremental equations are referred to the intrinsic transistor. This means that the base spread resistance $r_{b'}$ is to be considered as part of the external d.c. bias network; however, in most cases $r_{b'}$ is negligible with respect to the external base resistance.

The thermal incremental equations of the entire transistor stage are obtained combining equations (1) or (6) with the corresponding incremental equations of the associated d.c. bias network. If the bias equations are

$$\left.\begin{aligned} -I_b &= g_{11}V_{be} + g_{12}V_{ce} + \sum_i G_i E_i \\ -I_c &= g_{21}V_{be} + g_{22}V_{ce} + \sum_j G_j E_j \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots (7)$$

where E_i and E_j indicate the e.m.f.'s of the supply batteries, the corresponding incremental equations are

$$\left.\begin{aligned} -I'_b &= g_{11}V'_{be} + g_{12}V'_{ce} + \theta_1 \\ -I'_c &= g_{21}V'_{be} + g_{22}V'_{ce} + \theta_2 \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots (8)$$

where $\theta_1 = -\partial I_b/\partial T$, $\theta_2 = -\partial I_c/\partial T$ are zero if the bias network is temperature invariant.

From equations (1) (or (6)) and (8) there follow the thermal incremental equations of the entire stage in the form

$$\left.\begin{aligned}(g_{bb} + g_{11})V'_{be} + (g_{bc} + g_{12})V'_{ce} &= -I'_{bT} - \theta_1\\ (g_{cb} + g_{12})V'_{be} + (g_{cc} + g_{22})V'_{ce} &= -I'_{cT} - \theta_2\end{aligned}\right\}\ldots\ldots (9)$$

For simplification purposes in these equations $V'_{b'e}$ has been written as V'_{be}; furthermore in the following the parameters g_{bc} and g_{cc} will be neglected. Letting $\Delta g = (g_{bb} + g_{11})\, g_{22} - (g_{cb} + g_{21})g_{12}$, solving equations (9) for V'_{be}, V'_{ce}, and using also equations (1) or (6), one finds

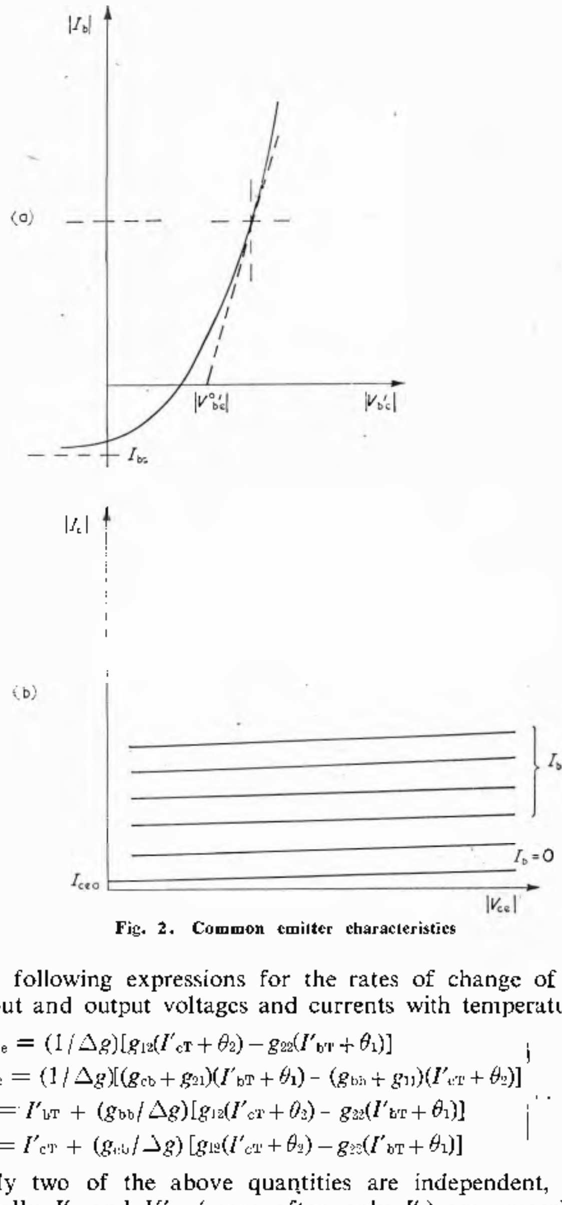

Fig. 2. Common emitter characteristics

the following expressions for the rates of change of the input and output voltages and currents with temperature:

$$\left.\begin{aligned}V'_{be} &= (1/\Delta g)[g_{12}(I'_{cT} + \theta_2) - g_{22}(I'_{bT} + \theta_1)]\\ V'_{ce} &= (1/\Delta g)[(g_{cb} + g_{21})(I'_{bT} + \theta_1) - (g_{bb} + g_{11})(I'_{cT} + \theta_2)]\\ I'_b &= I'_{bT} + (g_{bb}/\Delta g)[g_{12}(I'_{cT} + \theta_2) - g_{22}(I'_{bT} + \theta_1)]\\ I'_c &= I'_{cT} + (g_{cb}/\Delta g)\,[g_{12}(I'_{cT} + \theta_2) - g_{22}(I'_{bT} + \theta_1)]\end{aligned}\right\}\ldots (10)$$

Only two of the above quantities are independent, and usually I'_c and V'_{ce} (more often only I'_c) are prescribed for the design of the bias network. In general the design conditions are stated in terms of the maximum permissible incremental variations ΔI_c, ΔV_{ce} over a temperature range ΔT. For the computation of I'_c and V'_{ce} from the total incremental variations ΔI_c and ΔV_{ce}, it is necessary to assume the functional dependence of I_c and V_{ce} with T. If this is exponential, one has at the (junction temperature corresponding to) room temperature T_o:

$$I'_c(T_o) = (I_c(T_o)/\Delta T)\ln\,[1 + \Delta I_c/I_c(T_o)]$$

$$V'_{ce}(T_o) = (V_{ce}(T_o)/\Delta T)\ln\,[1 + \Delta V_{ce}/V_{ce}(T_o)] \quad \ldots\ldots\ldots\ldots\ldots\ldots (11)$$

In particular, using Shea's stability coefficient[1] $S_I = \Delta I_c/I_{cbo}$, $S_V = \Delta V_{ce}/I_{cbo}$, these may be written

$$I'_c(T_o) = (I_c(T_o)/\Delta T)\ln\,[1 + S_I\,I_{cbo}/I_c(T_o)]$$

$$V'_{ce}(T_o) = (V_{ce}(T_o)/\Delta T)\ln\,[1 + S_V\,I_{cbo}/V_{ce}(T_o)] \quad \ldots\ldots\ldots\ldots (12)$$

If $\Delta I_c/I_c(T_o) \ll 1$ or if the functional dependence is linear one has approximately

$$I'_c(T_o) \simeq \Delta I_c/\Delta T = S_I\,I_{cbo}(T_o)/\Delta T \quad \ldots\ldots\ldots\ldots (13)$$

A similar simplification applies for V'_{ce} (T_o).

Sometimes the values of the maximum permissible incremental changes ΔI_c and ΔV_{ce} are given for a temperature increase ΔT_1 and for a temperature decrease ΔT_2 from room temperature. In this case the rates of change, $I'_c(T_o)$ (and similarly $V'_{ce}(T_o)$) must be computed separately for the

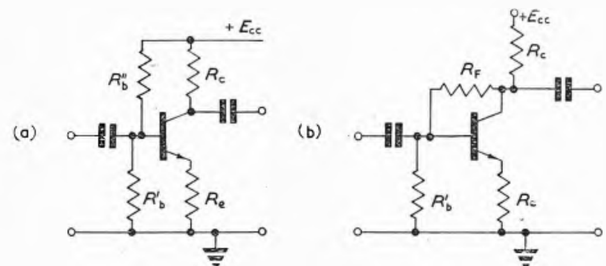

Fig. 3. Single stage bias network configurations

ΔT_1 and the ΔT_2 variations and the lower of the resulting values used for the design.

Recapitulating, the design of a bias network for prescribed total incremental variations ΔI_c and ΔV_{ce} is obtained as follows: first, the values of $I'_c(T_o)$, $V'_{ce}(T_o)$ are computed from equations (11) or (12); secondly, the bias equations (7) and the equations (10) are solved for the values of the circuit parameters.

In the following section the procedure of design is discussed in detail for two commonly used bias network configurations. Compensation by means of thermistors, junction diodes, etc., has been treated elsewhere and will not be repeated here[3]; accordingly, the quantities θ_1 and θ_2 will be equated to zero in the following.

Design of Single-Stage Temperature Invariant Bias Network

Typical one-battery bias network configurations for a single stage transistor amplifier are shown in Fig. 3. It is assumed that the desired rate of change I'_c is prescribed, together with the values of I_b, V_{be}, I_c and V_{ce} at room temperature.

For the network of Fig. 3(a), letting $R_b = R'_b \parallel R''_b$ and $E_{bb} = E_{cc}R'_b/(R'_b + R''_b) = E_{cc}R_b/R_b''$, the following three independent equations are considered

$$\left.\begin{aligned}(R_e + R_b)I_b + R_eI_c &= E_{bb} - V_{be}\\ R_eI_b + (R_e + R_c)I_c &= E_{cc} - V_{ce}\\ I'_c &= \frac{I'_{cT} + g_{bb}(R_e + R_b)I'_{cT}{}^*}{1 + g_{bb}(R_e + R_b) + g_{cb}R_e}\end{aligned}\right\}\ldots\ldots\ldots\ldots\ldots (14)$$

Solving the first two equations for R_e and R_c as a function of R_b one has

$$\left.\begin{aligned}R_e &= \frac{1}{I_c + I_b}(E_{bb} - V_{be} - R_bI_b)\\ R_c &= (1/I_c)\,(E_{cc} - V_{ce} - E_{bb} + V_{be} + R_bI_b)\end{aligned}\right\}\ldots (15)$$

Substituting these expressions into the third of equations

(14) one has

$$R_b = \frac{(E_{bb}-V_{be})[(g_{bb}+g_{cb})I'_c - g_{bb}I'_{cT}{}^*] - (I'_{cT}-I'_c)(I_c+I_b)}{g_{bb}I_cI'_{cT}{}^* - g_{bb}(I_c-\alpha_{cb}I_b)I'_c} \quad \ldots\ldots\ldots\ldots (16)$$

Since $R_b \geqslant 0$ and $|E_{bb}| > |V_{be}|$, the following limiting conditions must be satisfied

$$|E_{bb}| > \left| V_{be} + (1/g_{bb}) \frac{(I'_{cT} - I'_c)(I_c + I_b)}{(1 + \alpha_{cb})I'_c - I'_{cT}{}^*} \right| = E_{bb1} \quad \ldots\ldots\ldots\ldots (17)$$

$$|I'_c| > |I'_{cT}{}^*|/(1 + \alpha_{cb})$$

Furthermore, since $R_c \geqslant 0$, the upper limit of $|E_{bb}|$ is

$$|E_{bb2}| = \left| V_{be} + \frac{(E_{cc}-V_{ce})[I_cI'_{cT}{}^* - (I_c-\alpha_{cb}I_b)I'_c] - (I'_{cT}-I'_c)(I_c+I_b)I_b/g_{bb}}{(I'_{cT}{}^* - I'_c)(I_c+I_b)} \right| \quad \ldots\ldots (18)$$

Recapitulating, the design of a bias network of the type shown in Fig. 3(a) for prescribed I'_c is obtained as follows: First, a value of E_{bb} satisfying the limitations (17), (18) is chosen; second, the values of R_b, R_e and R_c are obtained

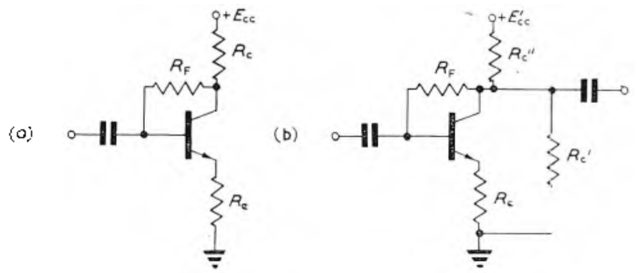

Fig. 4. Particular case of the network of Fig. 3(b)

from equations (16) and (15); third, the values of $R''_b = E_{cc}R_b/E_{bb}$ and of $R'_b = R_bR''_b/(R''_b - R_b)$ are computed.

It is immaterial which value of E_{bb} is chosen within the permitted range; however, if $E_{bb} = E_{bb1}$, R_b is zero, R_e is minimum, R_c is maximum and if $E_{bb} = E_{bb2}$, R_b and R_e are maxima and R_c is zero. Therefore, if an additional constraint is imposed on the design, such as a specified value of R_c, for example, the arbitrariness in the choice of E_{bb} is removed. The rate of change of V_{ce} is given approximately by the expression

$$V'_{ce} \simeq -I'_c(E_{cc} - V_{ce})/I_c \quad \ldots\ldots\ldots\ldots (19)$$

and therefore is not affected by the choice of E_{bb}.

In practice the value of R_c is usually prescribed. In this case R_e is obtained from equation (15), and R_b is

$$R_b = \frac{R_e[(1+\alpha_{cb})I'_c - I'_{cT}{}^*] \quad (I'_{cT} \quad I'_c)/g_{bb}}{I'_{cT}{}^* - I'_c} \quad \ldots (16')$$

Finally, computing $E_{bb} = R_e(I_c + I_b) + V_{be} + R_bI_b$, one obtains R_b' and R_b'' as before.

Consider now the design of the bias network of Fig. 3(b). This is simplified considerably if R'_b is infinite (Fig. 4); in this case, letting $G_3 = 1/(R_c + R_e)$ the following design equations are obtained:

$$\left.\begin{aligned} G_F(V_{ce} - V_{be}) &= I_b \\ G_3(E_{cc} - V_{ce}) &= I_c + I_b \\ I'_c &= \frac{g_{bb}I'_{cT}{}^*(G_3 + G_F) + G_3G_FI'_{cT}}{g_{bb}(G_3 + G_F) + (g_{cb} + G_3)G_F} \end{aligned}\right\} \ldots (20)$$

Since three design values must be determined, three variables, namely E_{cc}, G_3 and G_F, are considered. One computes G_F from the first of equations (20), G_3 from the relation

$$G_3 = G_F \frac{(1+\alpha_{cb})I'_c - I'_{cT}{}^*}{(I'_{cT}{}^* - I'_c) + G_F(I'_{cT} - I'_c)/g_{bb}} \quad \ldots\ldots (21)$$

and finally E_{cc} from the second of equations (20). In general the value of E_{cc} is not directly available. However, if the supply voltage is $|E'_{cc}| > |E_{cc}|$, the network may be realized as shown in Fig. 4(b) where $R'_c \,\|\, R''_c = R_c$ and $E'_{cc}R_c/R''_c = E_{cc}$. The only limitations associated with the design of the network are

$$\frac{|I'_{cT}{}^*|}{1 + \alpha_{cb}} \leqslant |I'_c| \leqslant \frac{|g_{bb}I'_{cT}{}^* + G_FI'_{cT}|}{g_{bb} + G_F} \quad \ldots\ldots\ldots\ldots (22)$$

where the minimum value of $|I'_c|$ corresponds to the condition $|E_{cc}| \gg |V_{ce}|$, and the maximum value of $|I'_c|$ to the condition $E_{cc} = V_{ce}$. It is noted that the design restricts only the value of $G_3 = 1/(R_c + R_e)$. Hence the designer may use any part[6] of $1/G_3$ for R_e.

If the value of $|E_{cc}|$ from the above design is larger than that of the available supply $|E'_{cc}|$, the complete network of Fig. 3(b) may be considered. The problem of the design of this network will be solved first for the case in which $R_e = 0$. Letting $G_b = 1/R'_b$, $G_c = 1/R_c$ one has:

$$\left.\begin{aligned} I_c &= G_c(E_{cc} - V_{ce}) - G_F(V_{ce} - V_{be}) \\ I_b &= G_F(V_{ce} - V_{be}) - G_bV_{be} \\ I'_c &= \frac{(g_{bb}I'_{cT}{}^* + G_bI'_{cT})(G_c + G_F) + G_cG_FI'_{cT}}{(g_{bb} + G_b)(G_c + G_F) + (g_{cb} + G_c)G_F} \end{aligned}\right\} \ldots\ldots (23)$$

In the above equations E_{cc} is prescribed and the variables are G_c, G_b and G_F; elimination of G_c and G_b between the first two equations (23) and the third one gives the following quadratic equation for the determination of G_F:

$$aG_F^2 + bG_F + c = 0 \quad \ldots\ldots\ldots\ldots (24)$$

where

$$a = \frac{E_{cc}V_{ce}}{V_{be}}(I'_{cT} \quad I'_c) - I'_cV_{ce}$$

$$b = (E_{cc}[g_{bb}(I'_{cT}{}^* - I'_c) - I_b(I'_{cT} \quad I'_c)/V_{be}] - g_{cb}(E_{cc} - V_{ce})I'_c + I_c\left((V_{ce}/V_{be})\,I'_{cT} - \frac{E_{cc}I'_c}{E_{cc} - V_{ce}}\right)$$

$$c = I_c[g_{bb}(I'_{cT}{}^* - I'_c) - I_b(I'_{cT} - I'_c)/V_{be}]$$

Using the positive solution of equation (24) one obtains the values of G_c and G_b from the relations

$$\left.\begin{aligned} G_c &= \frac{I_c + G_F(V_{ce} - V_{be})}{E_{cc} - V_{ce}} \\ G_b &= \frac{G_F(V_{ce} - V_{be}) - I_b}{V_{be}} \end{aligned}\right\} \ldots\ldots (25)$$

The case in which G_c rather than E_{cc} is prescribed is discussed in the Appendix.

Finally, the network of Fig. 3(b) with $R_e \neq 0$ may be considered when the previous design provides excessive values of G_F or of E_{cc}. In this case the design equations are written:

$$\left.\begin{aligned} & V_{be} + R_e(I_b + I_c) + R'_b(I_b - (V_{ce} - V_{be})G_F) = 0 \\ & E_{cc} - V_{ce} - R_c(I_c + G_F(V_{ce} - V_{be})) - R_e(I_c + I_b) = 0 \\ & I'_c = \frac{(R_c + R_b' + G_F\Delta R')(g_{bb}\Delta R' I'_{cT}{}^* + R_eI'_{cT}) + R'_b(R_e + G_F\Delta R')I'_{cT}}{(g_{bb}\Delta R' + R_e)(R_c + R'_b + G_F\Delta R') + (g_{cb}\Delta R' + R'_b)(R_e + G_F\Delta R')} \end{aligned}\right\} \ldots (26)$$

In the last equation $\Delta R' = R_eR_c + R_cR'_b + R'_bR_e$. The bias design may be obtained from solution of these equations. However, a much simpler procedure follows from using the one-to-one correspondence between the networks of Fig. 3(a) and 3(b). Using subscripts to identify the networks and letting $V_1 = V_{be} + R_e(I_c + I_b)$, $V_2 = V_{ce} +$

$R_e(I_c + I_b)$, one has

$$\left.\begin{aligned}(R_e)_b &= (R_e)_a\,;\ (G_F)_b = (E_{cc})_a/(R''_b)_aV_2\\ (G_c)_b &= 1/(R_c)_a - (G_F)_b\\ (E_{cc})_b &= ((E_{cc})_a/(R_c)_a - (G_F)_bV_1)/(G_c)_b\\ (G'_b)_b &= 1/(R_b)_a - (G_F)_b\end{aligned}\right\}\ \ldots\ldots\ldots \quad (27)$$

Therefore, the design may be performed on the network of Fig. 3(a) and then transferred, if desired, to that of Fig. 3(b).

Measurement of Transistor Thermal Incremental Properties

The measurement of the values of I'_{bT}, $I'_{cT}{}^*$, I'_{cT} is obtained by placing the transistor in an oven, biasing it at the desired point, and taking the values of V_{be} and of I_c as a function of the temperature, while I_b and V_{ce} are kept constant. In this case $I'_b = 0$ and therefore the following relation is derived from equation (5):

$$V'_{be} = V'_{be'} + r'_{b'}I_b \ \ldots\ldots\ldots\ldots \quad (28)$$

In equation (28) $r'_{b'}$ is the rate of change of the base spread resistance with temperature. The latter is positive[5] in the ranges of temperature of interest, and for this reason $V'_{b'e}$ (which has the sign $-V_{b'e}$) has sign opposite to that of $r'_{b'}I_b$. However, except for very large values of I_b, the latter term may be neglected with respect to $V'_{b'e}$.

It follows that, using equation (6) and indicating with V'_{be1}, I'_{c1} the measured rates of change with constant I_b and constant V_{ce}, the parameters I'_{bT} and $I'_{cT}{}^*$ are given by the following relations

$$\left.\begin{aligned}I'_{bT} &\simeq -g_{bb}V'_{be1}\\ I_{cT}{}'^* &= I'_{c1}\end{aligned}\right\}\ \ldots\ldots\ldots\ldots \quad (29)$$

The value of I'_{cT} follows from the equation $I'_{cT} = I'_{cT}{}^* + \alpha_{cb}I'_{bT}$.

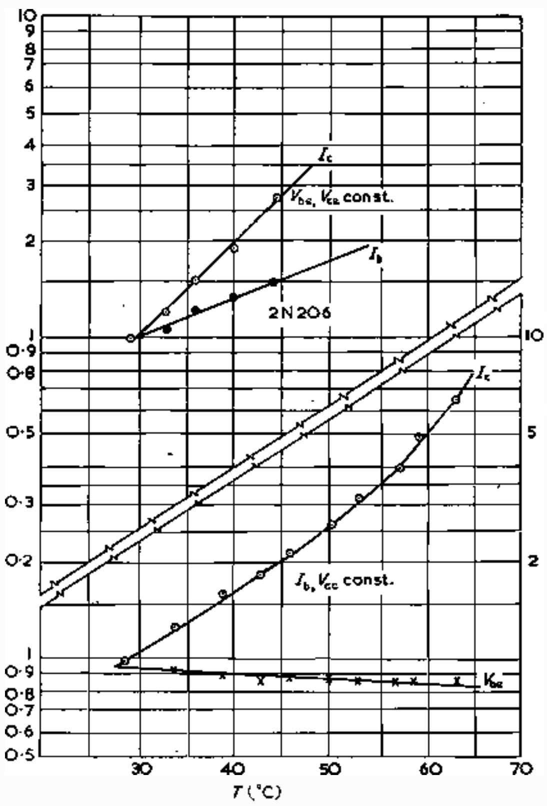

Fig. 5. Thermal behaviour of 2N206 at constant I_b, V_{ce} and at constant V_{be}, V_{ce}

If the values of I_b and of I_c are measured as a function of temperature, maintaining constant V_{be} and V_{ce}, the following results are obtained: indicating with I'_{b2}, I'_{c2} the corresponding measured values, one has from application of equations (5) and (1)

$$\begin{aligned}V'_{b'e} &= -(r_{b'}I'_{b2} + r'_{b'}I_b) \simeq -r_{b'}I'_{b2}\\ I'_{bT} &\simeq I'_{b2}\,(1 + g_{bb}r_{b'}) \qquad\qquad \ldots \quad (30)\\ I'_{cT} &\simeq I'_{c2} + g_{cb}r_{b'}I'_{b2}\end{aligned}$$

Furthermore one has

$$I'_{cT}{}^* = I'_{cT} - \alpha_{cb}I'_{bT} = I'_{c2} - \alpha_{cb}I'_{b2} \ \ldots\ldots\ldots \quad (31)$$

Therefore the values of I'_{c2} and of I'_{b2} provide directly an evaluation of $I'_{cT}{}^*$, independently of the rate of change of $r_{b'}$ with temperature. On the other hand the computation of I'_{bT} and of I'_{cT} requires that the value of $r_{b'}$ be known.

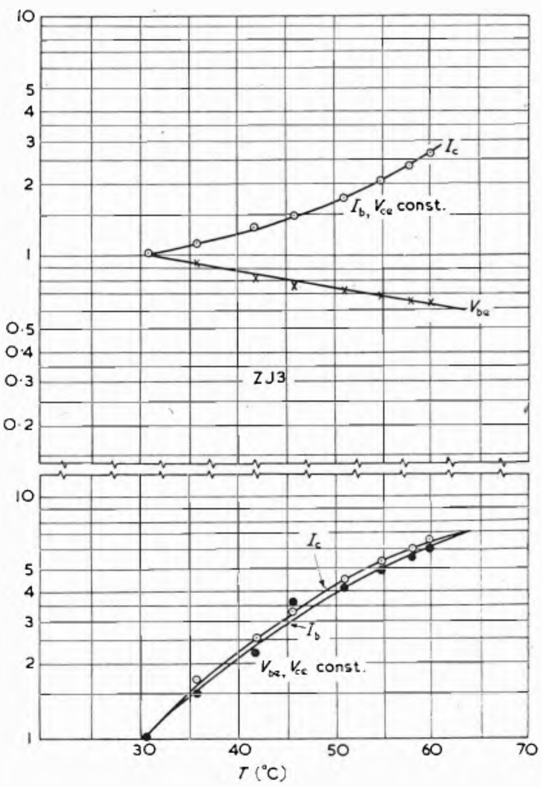

Fig. 6. Thermal behaviour of ZJ3 at constant I_b, V_{ce} and at constant V_{be}, V_{ce}

Finally it may be noted that, if resistance $r_{b'}$ is artificially increased with an external resistance $R_b \gg r_{b'}$ and the measurement procedure described by equations (30) repeated in these conditions, the values of I'_{bT}, I'_{cT} are obtained from the equations

$$\left.\begin{aligned}I'_{bT} &\simeq I'_{b2}(1 + g_{bb}R_b)\\ I'_{cT} &\simeq I'_{c2} + g_{cb}R_bI'_{b2}\end{aligned}\right\}\ \ldots\ldots\ldots\ldots \quad (32)$$

There may be an advantage in using the latter procedure of measurement instead of the procedure with constant I_b, constant V_{ce}, since in the latter case the measurement of V_{be} must be obtained with a voltmeter of uncommonly high values of internal resistance and sensitivity.

Finally, another procedure of measurement is based on using the network of Fig. 4 (where $g_{11} = G_F$, $g_{12} = g_{21} = -G_F$, $g_{22} = G_3 + G_F$), in conjunction with an experimental determination of I'_b and I'_c and use of equations (10).

Measurements of thermal incremental parameters have been performed on a number of transistor samples. In

Table 1 the results obtained for a 2N206 (Fig. 5), 2N28, ZJ3 (Fig. 6) are shown. In these experiments the values of g_{bb} and of g_{cb} were obtained from the relations

$$g_{bb} = -\Lambda(I_b - I_{bs}),\ g_{cb} = -\Lambda(I_c - I_{cs})$$

and the thermal rates of change were computed by means of equations (11).

It is of interest to note that some of the values of $V'_{b'e}$ have been found to be lower than the theoretical value 2mV/°C.

For the transistors investigated the values of I'_{bT} have been found to be of the order of 0·06 to 0·08 times the value of $I_b(T_o)$; on the other hand the parameter $I'_{cT}{}^*$ is somewhat higher, but of the same order of magnitude of I'_{ceo}. These observations may be useful either for a check or for a rough estimate of the values of the thermal incremental parameters.

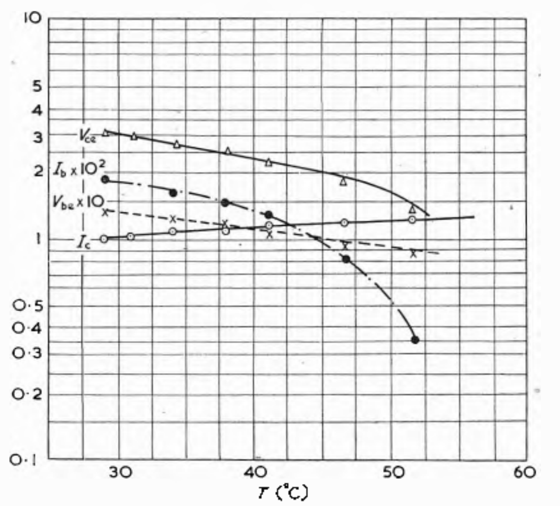

Fig. 7. Thermal behaviour of bias configuration of Fig. 3(a) for a 2N206, with $E_{bb} = -1$V

TABLE 1

Measurement of Transistor Thermal Incremental Parameters

	2N206	2N28	ZJ3
V_{be} ..	−0·13V	+0·145V	−0·105V
I_b ..	−18μA	+67μA	−26μA
V_{ce} ..	−3V	+3·0V	−3·0V
I_c ..	−1mA	+1·0mA	−1·0mA
$V'_{b'e}$ ..	+1·2mV/°C	−1·7mV/°C	+2mV/°C
I'_{c1} ..	−47μA/°C	+20μA/°C	−22μA/°C
g_{bb} ..	820μmhos	2·64mhos	1·05mhos
g_{cb} ..	0·038mhos	0·038mhos	0·038mhos
α_{cb} ..	47	14·5	41
I'_{bT} ..	−1μA/°C	+4·5μA/°C	−2·1μA/°C
$I'_{cT}{}^*$ ..	−47μA/°C	+30μA/°C	−22μA/°C
$I'_{cT}{}^*$ ..	−94μA/°C	+95μA/°C	−108μA/°C
I'_{b2} ..	−0·4μA/°C	+3·0μA/°C	−1·9μA/°C
I'_{c2} ..	−53μA/°C	+78μA/°C	−105μA/°C
$I'_{cT}{}^*$ ..	−34μA/°C	+34·5μA/°C	−27μA/°C

Practical Design of Bias Circuits

The design of the bias circuits shown in Fig. 3 has been investigated using the transistor 2N206, whose incremental properties are listed in Table 1.

For the network of Fig. 3(a) the following data were assumed

$E_{cc} = -10$V, $I_c = -1$mA, $V_{ce} = -3$V, $V_{be} = -0·13$V, $I_b = -18\mu$A at room temperature. A thermal incremental rate of change $I'_c = -10\mu$A/°C was prescribed.

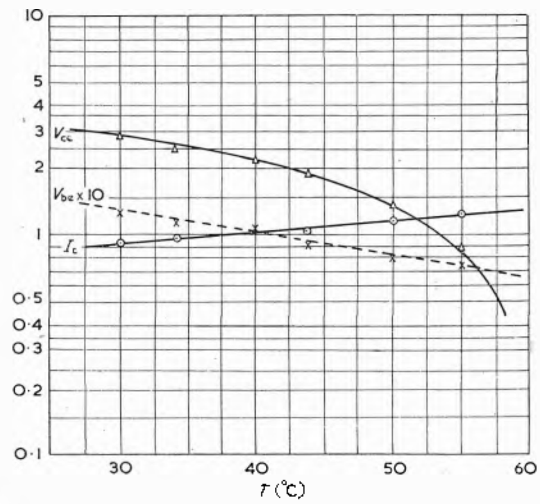

Fig. 8. Thermal behaviour of the bias network of Fig. 4(a) for a 2N206

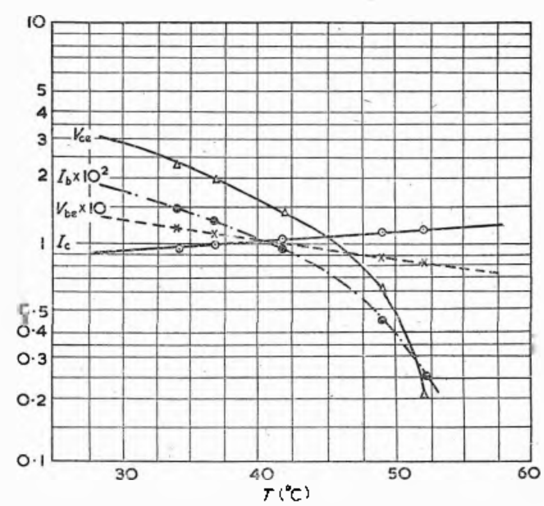

Fig. 9. Thermal behaviour of the bias network of Fig. 3(b) with a 2N206, in this case $R_e = 0$

TABLE 2

Bias Design for the Network of Fig. 3(a)

E_{bb}	−0·236V	−0·5V	−1V	−5V	−8·73V
R_b	0	1 230	5 910	44 200	86 500
R_e	106	347	761	4 050	7 000
R_c	6 894	6 653	6 239	2 950	0
R''_b	—	24 600	59 100	88 400	99 000
R'_b	—	1 295	8 100	88 400	685 000
I'_c	—	−8·3μA/°C	−8·5μA/°C	−8·5μA/°C	—

From equations (17) and (18) the limiting values of E_{bb} are found to be

$$E_{bb1} = -0{\cdot}236\text{V},\ E_{bb2} = -8{\cdot}73\text{V}$$

Theoretically the design may be performed with any choice of E_{bb} within the above limits. In Table 2 some designs are shown for a number of values of E_{bb}. These designs have been tried in the range of 30 to 60°C and have produced the values of I'_c shown in the last row of Table 2. One of these experimental investigations is reported in detail in Fig. 7. These values of I'_c are slightly less than the prescribed 10μA/°C (a divergence probably due to inaccuracy in the evaluation of the transistor conductance parameters), but they are very consistent.

The network of Fig. 3(b) was designed first for the case shown in Fig. 4. Using the same bias and I'_c data it was found that

$G_F = 1/155\,000$mhos, $G_3 = 1/13\,620$mhos, $E_{cc} = -16{\cdot}6$V

Assuming that the available supply voltage is larger than 16·6V in absolute value, the configuration of Fig. 4(b) could be used. The experimental investigation on this network is shown in Fig. 8 and produced a rate of change I'_c of about 7·5μA/°C.

Finally the design of the network of Fig. 3(b) with $R'_b \neq 0$, but with $R_e = 0$, was obtained. Again using the same bias values and assuming that $E_{cc} = -10$V, it was found that

$G_F = 1/51\,000$mhos, $G_c = 1/6\,650$mhos, $G_b = 1/3\,400$mhos

The experimental investigation of this network (Fig. 9) gave a resulting rate of change of I'_c of about -8μA/°C.

Conclusion

The experimental investigations reported in this article were performed primarily with the aim of checking the validity of the theory. The results obtained are considered satisfactory, since they have shown that a reliable thermal design is feasible. The thermal incremental properties of transistors can be measured simply; their orders of magnitude can be predicted.

The study could be further extended to investigate the sensitivity of the design to the spread of the transistor parameters.

Acknowledgments

This work was performed under a contract sponsored by the O.N.R., (U.S.A.). The author wishes to thank Mr. R. F. Shea for his advice.

APPENDIX—Recapitulation of Design Procedures.

Check that $|I'_c| > |I'_{cT}|/(1 + \alpha_{cb})$.

For the circuit of Fig. 3(a) compute E_{bb1}, E_{bb2} from equations (17) and (18). If R_c is given, compute R_b from equation (16) and R_e from equation (15). Then evaluate E_{bb} and R'_b, R''_b.

For the circuit of Fig. 3(b), if R'_b is infinite, compute G_F from the first of equations (20), $G_3 = 1/(R_c + R_e)$ from equation (21) and E_{cc} from the second of equations (20). Use any convenient part of $1/G_3$ as R_c.

If R'_b is finite, $R_e = 0$, and if G_c is prescribed, find G_F from equation (24), where

$$a = (I'_{cT} - I'_c)(V_{ce} - V_{be})/V_{be}$$

$$b = \left(G_c\left(\frac{V_{ce} - 2V_{be}}{V_{be}}\right) - \frac{I_b}{V_{be}}\right)(I'_{cT} - I'_c) + g_{cb}I'_c - g_{bb}(I'_{cT}{}^* - I'_c)$$

$$c = -G_c\left(g_{bb}(I'_{cT}{}^* - I'_c) + (I_b/V_{be})(I'_{cT} - I'_c)\right)$$

Check that $G_F > I_b/(V_{ce} - V_{be})$; compute G_b and E_{cc} from equations (23).

If R'_b is infinite, $R_e \neq 0$, use equations (27) to obtain the design from the network of Fig. 3(a).

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A Mobile Radar Demonstration Unit

A new mobile marine demonstration unit has recently been constructed by Decca Radar Ltd and the Decca Navigator Company Ltd.

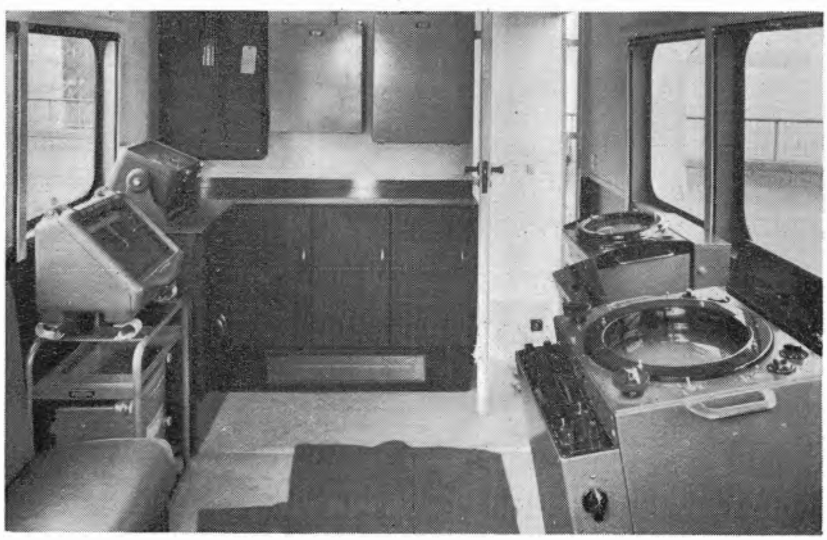
Built on a Commer 5-ton chassis, the equipment comprises two of the new D7 marine radar units, the TM909 and the D404; the Mark V Navigator unit and the Marine Track Plotter.

A hydraulic aerial mast has been fitted to enable the scanner to be lowered and stowed at roof level when travelling. The same scanner is used for both radar sets, a waveguide changeover switch being employed to select the 75kW transmitter of the TM909 or the 20kW transmitter of the D404 set.

Power for the equipment is obtained from a 5kW generator mounted beneath the chassis and driven by the main engine.

The vehicle, named "Polaris" will be shipped to Europe early next month to begin a six months' demonstration tour, mainly in countries inaccessible to the Company's demonstration yacht "Navigator".

The interior of the demonstration vehicle with the Mark V Navigator on the left and on the right the D404 and TM909 units.



Approximate Models for Transmission-Lines and their Errors

By R. R. Vierhout*

It is often convenient in experimental work and in calculations to replace transmission lines by simple models such as the T- and π-networks. In this article simple formulæ are derived which give immediate indication of the magnitude of error which is introduced by using these models.

(Voir page 121 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 128)

TRANSMISSION-LINES are often replaced in calculations or experiments by simple models, such as T- and π-networks of capacitors, inductors and resistors. There is a need for those who want to replace transmission-lines by simple circuits for calculation or experimental purposes, to have a simple formula, which at once provides information about the errors one is making. The following can be applied to electrical, hydrodynamical, mechanical, accoustical, optical and thermal transmission-lines.

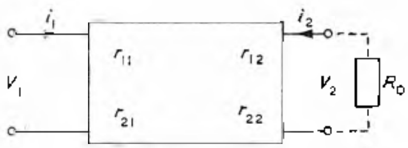


Fig. 1. General four-pole network

General Theory of Four-Poles

For a good understanding the reader may be reminded of the theory of four-poles, by which a four-pole (Fig. 1) is characterized by the matrix:

$$\begin{bmatrix} r_{11} & r_{12} \\ r_{21} & r_{22} \end{bmatrix} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

following from the equations:

$$\left.\begin{aligned} V_1 &= r_{11}\,.\,i_1 + r_{12}\,.\,i_2 \\ V_2 &= r_{21}\,.\,i_1 + r_{22}\,.\,i_2 \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots \quad (2)$$

The coefficients $r_{11}, r_{12}, r_{21}, r_{22}$ have the dimensions of an impedance and generally are of a complex nature. The variables V and i are represented in the ω-domain (ω is the angular frequency). As the relation between two variables out of four (V_1, V_2, i_1, i_2) is of importance, three equations are necessary. Hence, the two equations (2) must be completed by a third one, generally the relation between the output voltage and the output current, V_2 and i_2:

$$-(V_2/i_2) = R_0 \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

where R_0 is the load of the four-pole.

From the equations (2) and (3) follows:

$$\frac{V_1}{\Delta + r_{11}\,.\,R_0} = \frac{V_2}{r_{21}\,.\,R_0} = \frac{i_1}{r_{22} + R_0} = (i_2/ - r_{21}) = f(\omega) \quad \ldots\ldots\ldots\ldots \quad (4)$$

where Δ is the value of the determinant of equation (1) or $\Delta = r_{11}\,.\,r_{22} - r_{12}\,.\,r_{21}$.

The formulae for the input- and transfer-impedances which are of interest for calculations and experiments are:

$$V_1/i_1 = \frac{\Delta + r_{11}\,.\,R_0}{r_{22} + R_0} \quad \ldots\ldots\ldots\ldots \quad (5)$$

$$V_2/i_1 = \frac{r_{21}\,.\,R_0}{r_{22} + R_0} \quad \ldots\ldots\ldots\ldots\ldots \quad (5a)$$

and for the transfer-functions:

$$V_2/V_1 = \frac{r_{21}\,.\,R_0}{\Delta + r_{11}\,.\,R_0} \quad \ldots\ldots\ldots\ldots \quad (6)$$

$$i_2/i_1 = \frac{-r_{21}}{r_{22} + R_0} \quad \ldots\ldots\ldots\ldots\ldots \quad (6a)$$

Now, if in the case of passive four-poles there will be an error in the matrix-elements or Δ of, at most, p per cent, than the error in V_2/i_1, i_2/i_1, V_1/i_1 etc. will also be at most p per cent.

The Transmission-Line as a Four-Pole

The transmission-line is a four-pole with the matrix:

$$\begin{vmatrix} Z\,.\coth\gamma & Z/\sinh\gamma \\ Z/\sinh\gamma & Z\,.\coth\gamma \end{vmatrix} \quad \ldots\ldots\ldots\ldots \quad (7)$$

where $\gamma = \sqrt{[(R + j\omega L)(G + j\omega C)]}$ and $Z = \sqrt{\left[\dfrac{R + j\omega L}{G + j\omega C}\right]}$

R, L, C and G are the total resistance, inductance, capacitance and leak-admittance of the transmission-line. The matrix is symmetrical because the transmission-line is symmetrical. The determinant $\Delta = Z^2$.

The Transmission-Line as a Four-Pole

The T-network of Fig. 2(a) is the same as the T-network of the Fig. 2(b), because $R + j\omega L = Z\,.\,\gamma$ and $1/(G + j\omega C) = Z/\gamma$ using the same notation as above.

From Fig. 2 one obtains:

$$\begin{aligned} V_1 &= (Z\gamma/2)\,.\,i_1 + (Z/\gamma)(i_1 + i_2) = Z(\gamma/2 + 1/\gamma)\,i_1 \\ &\qquad + (Z/\gamma)\,.\,i_2 = t_{11}\,.\,i_1 + t_{12}\,.\,i_2 \\ V_2 &= (Z\gamma/2)\,.\,i_2 + (Z/\gamma)(i_1 + i_2) = (Z/\gamma)\,.\,i_1 + \\ &\qquad Z(\gamma/2 + 1/\gamma)\,i_2 = t_{21}\,.\,i_1 + t_{22}\,.\,i_2 \end{aligned} \quad \ldots\ldots\ldots\ldots \quad (8)$$

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Hence, the matrix of this four-pole is:

$$\begin{vmatrix} t_{11} & t_{12} \\ t_{21} & t_{22} \end{vmatrix} \equiv \begin{vmatrix} Z(\gamma/2 + 1/\gamma) & Z/\gamma \\ Z/\gamma & Z(\gamma/2 + 1/\gamma) \end{vmatrix} \quad \ldots\ldots (9)$$

The determinant $\Delta = (1 + (\gamma^2/4)) \,.\, Z^2$

$$\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (9a)$$

If a simplification is wanted of a transmission-line, equation (7), by substituting a symmetrical T-network, equation (9), the error in the matrix-elements can be calculated by developing $\coth\gamma$ and $\sinh\gamma$ in sequences.

$$t_{11} = t_{22} = Z \,.\, \coth\gamma \; \frac{(\gamma/2 + 1/\gamma)\,(\gamma + (\gamma^3/6) + (\gamma^5/120) + \ldots)}{(1 + (\gamma^2/2) + (\gamma^4/24) + \gamma^6/720) + \ldots)} =$$

$$Z \,.\, \coth\gamma \,.\, (1 + (\gamma^2/6) - (\gamma^4/30) \ldots) \quad \ldots\ldots\ldots\ldots (10)$$

$$t_{12} = t_{21} = (Z/\sinh\gamma) \,.\, \frac{\gamma + (\gamma^3/6) + (\gamma^5/120) + \ldots}{\gamma} =$$

$$Z/\sinh\gamma \,.\, (1 + (\gamma^2/6) + (\gamma^4/120) + \ldots) \quad \ldots\ldots\ldots (11)$$

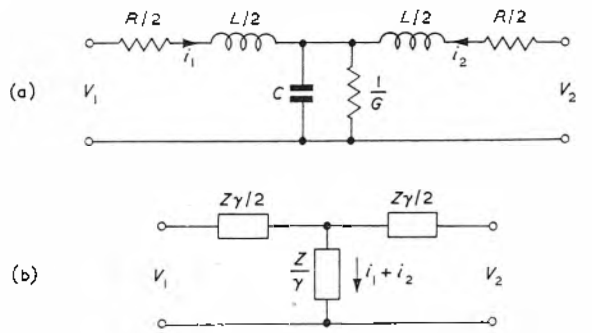

Fig. 2. The T-network as a simplification of a transmission-line

The matrix-elements of the T-network deviate for small values of γ about $100 \,.\, (\gamma^2/6)$ per cent from the matrix-elements of the transmission-line. That means, according to the rule in the second paragraph that the errors in the values of equations (5a) or (6a) are at most:

$$100/6 \,.\, \gamma^2 \text{ per cent} \simeq 17 \,.\, |R + j\omega L|\, |G + j\omega C| \text{ per cent} \quad \ldots (12)$$

According to equation (9a), the error in Δ will be $\gamma^2/4$. Hence for the calculation of (16), the error will be at most:

$$\gamma^2/4 - (\gamma^2/6) = (\gamma^2/12)$$

or:

$$100/12\ \gamma^2 \text{ per cent} \simeq 8\, |R + j\omega L|\, |G + j\omega C| \text{ per cent} \quad \ldots (12a)$$

For equation (5), the error will be:

$$100\gamma^2/4 = 25\, |R + j\omega L|\, |G + j\omega C| \text{ per cent} \quad \ldots\ldots (12b)$$

The error here is given as the radius of the circle of an error-domain, around the end-point of the true vector, divided by the absolute value of this true vector. The formulae are valid for small values of γ.

Because the matrix-elements of the T-network all deviate in the same direction with the error $+(\gamma^2/6)$, it is possible to make the error in the model-calculations slightly smaller. One has, then, to divide the values of R and L by and to multiply the values of C and G by $(1 + (\gamma^2/6))$. In the present case Z is divided by $(1 + (\gamma^2/6))$, while γ remains the same. The formulae (5a) and (6a) give in this case the model-error of $(\gamma^4/30) \,.\, 100$ per cent, or:

$$3{\cdot}3 \,.\, |R + j\omega L|^2\, |G + j\omega C|^2 \text{ per cent} \quad \ldots\ldots (13)$$

However, in the examples (5) and (6), the determinant Δ appears and the correction on this is:

$$(1 + (\gamma^2/4))\,\frac{Z^2}{(1 + (\gamma^2/6))^2} = Z^2(1 - (\gamma^2/12) \ldots) \quad \ldots\ldots (14)$$

Hence, the model-error of equation (5) and (6) will be at most: $\gamma^2/12$ or:

$$8 \,.\, |R + j\omega L|\, |G + j\omega C| \text{ per cent} \quad \ldots\ldots\ldots (14a)$$

So, the advantage of this procedure is small.

If a π-network is chosen as a model, then the circuit of Fig. 3(a) can be replaced by the circuit of Fig. 3(b).

The matrix of this four-pole is again symmetrical:

$$\begin{vmatrix} \pi_{11} & \pi_{12} \\ \pi_{21} & \pi_{22} \end{vmatrix} \equiv \begin{vmatrix} \dfrac{Z(1 + (\gamma^2/2))}{\gamma(1 + (\gamma^2/4))} & \dfrac{Z}{\gamma(1 + (\gamma^2/4))} \\ \dfrac{Z}{\gamma(1 + (\gamma^2/4))} & \dfrac{Z(1 + (\gamma^2/2))}{\gamma(1 + (\gamma^2/4))} \end{vmatrix} \quad \ldots\ldots (15)$$

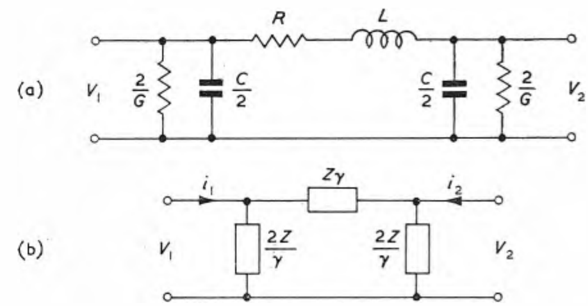

Fig. 3. The π-network as a simplification of a transmission-line

Developing again $\coth\gamma$ and $\sinh\gamma$ in sequences, one obtains:

$$\pi_{11} = \pi_{22} = Z \,.\, \coth\gamma \,.\, (1 - (\gamma^2/12) - (\gamma^4/80) \ldots)$$

$$\pi_{12} = \pi_{21} = \frac{Z}{\sinh\gamma} \,.\, (1 - (\gamma^2/12) + (7\gamma^4/240) \ldots) \quad \ldots (16)$$

and:

$$\Delta = \frac{Z^2}{1 + (\gamma^2/4)} = Z^2(1 - (\gamma^2/4) - (\gamma^4/16) \ldots) \quad \ldots (16a)$$

If one divides L and R and multiplies G and C in the π-circuit by $(1 - (\gamma^2/12))$, then the error in equations (5a) and (6a) will be at most:

$$100 \,.\, (7\gamma^4/240) \text{ per cent} = 2{\cdot}9\, |R + j\omega L|^2\, |G + j\omega C|^2 \text{ per cent}$$

$$\ldots\ldots\ldots\ldots\ldots (17)$$

In the formulae (5) and (6) Δ appears, which now has the error:

$$\frac{Z^2}{(1 + (\gamma^2/4))\,(1 - (\gamma^2/12))^2} = Z^2/(1 - (\gamma^2/12) \ldots) \quad \ldots (16b)$$

Hence, the error in equations (5) and (6) will be again at most

$$100 \,.\, (\gamma^2/12) \text{ per cent} = 8\, |R + j\omega L|\, |G + j\omega C| \text{ per cent}$$

$$\ldots\ldots\ldots\ldots\ldots (17a)$$

For the examples (5) and (6), the error introduced by the π-network is equal to the error, introduced by the T-network, if both circuits are corrected. A correction for model-inductances, capacitances, resistances and admittances is recommended for those formulae, in which Δ does not appear.

A 1Mc/s Transistor Decade Counter

By C. G. Bradshaw*, B.Eng., Ph.D.

A transistor decade counter is described which accepts input pulses at a maximum repetition frequency of 1·3Mc/s. It employs four binary stages with a feedback circuit which reduces the sixteen possible states to ten. This method is well known in pulse-counter techniques.

A typical input shaping circuit which provides the required pulses for the counter is shown.

The counter, designed for a nominal 12-volt supply, functions correctly over the range 9 to 15V at temperatures up to 55°C.

Satisfactory laboratory operation for some 150 hours has so far been achieved.

(Voir page 121 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 128)

THE use of valves and transistors in frequency dividing and pulse counting circuits has already been well exploited[1,2,3,4,5], but until comparatively recently, however, transistors have not been readily available in this country to enable circuits to operate over frequency ranges covered by their valve counterparts. Fortunately this situation is now greatly improved and the present manufacture of reliable and consistent high-frequency transistors has made possible the design of fast switching circuits vital in the digital computer and pulse counter fields.

A decade counter is described which will accept input pulses at a maximum repetition frequency of 1·3Mc/s and will provide a direct output to a following decade circuit without the need for impedance matching between the decades.

Details of the Circuit

The complete decade circuit is shown in Fig. 1 The methods of modifying a binary scaling system to produce a decade system are well known, the principle being that the sixteen stable states obtained by using four series binary stages are reduced to ten by means of feedback loops inserted at appropriate points.

Each binary stage consists of a conventional Eccles-Jordan bistable circuit with an additional input steering circuit to give the required binary action. The diodes connected between each of the two bases and the common emitters limit the positive excursions of the base and promote quicker switching speeds. Essential repeatability of the complete circuit prompted the choice of OC44 transistors and, in fact, three such decade circuits have given the same frequency range for given input conditions.

A brief description of the circuit's operation can be given with reference to Figs. 1 and 2. The convention is adopted that all the binaries are in their 'zero' state when each even-number transistor is conducting. When these transistors are delivering their respective number of units of current into the indicating circuit they are therefore in their 'one' state. The ideal waveforms appearing at the collectors of X_2, X_4, X_6, X_7 and X_8 are shown in Fig. 2 and the units of current associated with each stage are indicated. The operation of the circuit up to the eighth input pulse follows the normal binary pattern and the addition of the units of current at any time gives the appropriate number of pulses counted. On the eighth pulse transistor X_7 changes to its 'zero' state thereby producing in the feedback circuit a positive pulse which is passed simultaneously to stages 2 and 3, or to the bases of X_4 and X_6. The effect of this pulse is to turn X_4 and X_6 to their 'one' states immediately after their normal sequential change. It is then seen that all the binaries are returned to their 'zero' states on the tenth input pulse and the cycle is ready to be repeated.

Further comment is perhaps needed regarding the action

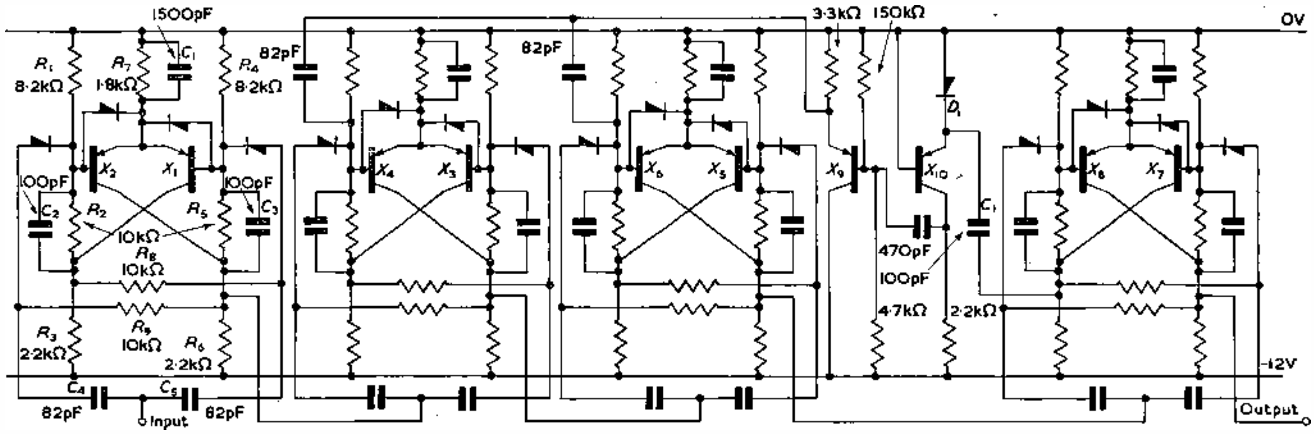

Fig. 1. Complete decade counter circuit

Values common to all stages except C_2 and C_3 = 150pF in last 3 stages

* *S.R.D.E., Ministry of Supply*

of the 'feedback' transistors X_9 and X_{10}. When X_7 is in its 'one' state the emitter-base diode of X_{10} is cut-off but capacitor C_1 charges through diode D_1. Coinciding with the change of X_7 to its 'zero' state the capacitor then discharges into the low emitter-base resistance with a resulting low time-constant and a brief pulse of collector current is obtained. Transistor X_9, connected in the common-collector configuration, merely reproduces this pulse on its emitter, use being made of this transistor's buffer action to minimize the loading effect of the bases of X_4 and X_6 on the feedback pulse.

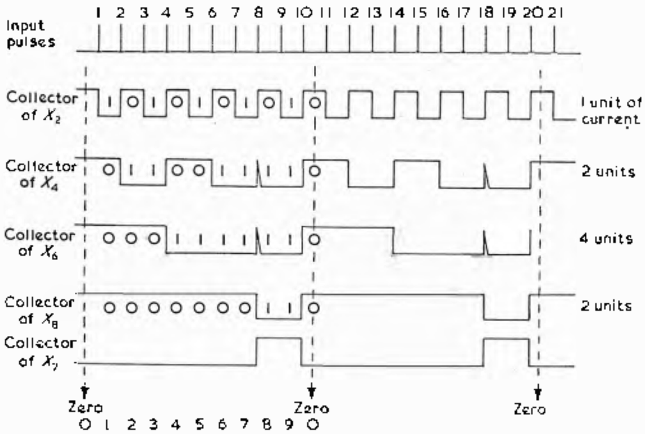

Fig. 2. Waveform operation of the circuit showing the arrangement of the decade count

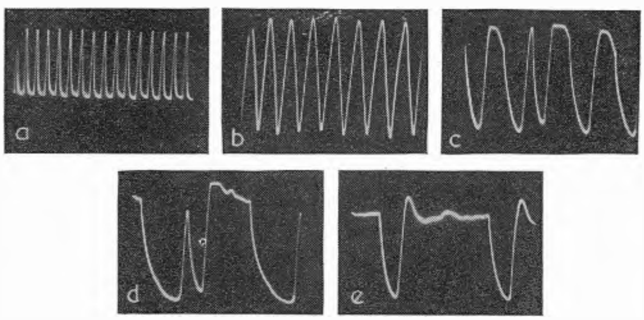

Fig. 3. Decade counter waveforms for an input frequency of 1Mc/s

(a) Input pulses (repetition frequency 1Mc/s)
(b) Waveform at collector of X_2
(c) Waveform at collector of X_4
(d) Waveform at collector of X_6
(e) Waveform at collector of X_8

Fig. 3 (a) to (e) shows the input pulses and the waveforms actually appearing at the collectors of X_2, X_4, X_6 and X_8 for an input frequency of 1Mc/s. The action of the positive feedback pulse on transistors X_4 and X_6 is clearly shown in (c) and (d).

The decade count is indicated on a sensitive current meter, say, one of 100μA full scale. The very low current drain from each collector for indication purposes ensures that the operation of the circuit is not impaired. One unit of current referred to above is 10μA and the meter then steps 0 to 9 and resets to 0 on every tenth input pulse.

A simple resetting arrangement can easily be added to the counter. This involves connecting the R_1 components associated with X_2, X_4, X_6 and X_8 in Fig. 1 not to the positive supply line but to a common point and thence through a resistance, say, 33kΩ to the positive supply line. The voltage drop across this resistance caused by the base currents flowing through it pulls the bases of the appropriate transistors sufficiently negative to allow them to conduct. The counter is then in its 'zero' state but for normal operation the reset resistor is short-circuited giving the circuit connexions shown in the diagram.

Input Shaping Circuit

The decade circuit requires an input pulse of amplitude approximately 3V and rise-time 0·2μsec. An input pulse shaping circuit is essential and is shown in Fig. 4. The operation of this circuit is similar to that of X_{10} in the decade circuit and need not be further described. The shaping circuit is operated by a square-wave of minimum amplitude 2V and produces the required input conditions for the decade circuit. Consequently the use of this decade in a counter chronometer, in which a 1Mc/s crystal oscillator is the frequency source, requires some form of squaring circuit, say, of the Schmitt type.

Performance of Decade Circuit

The circuit was designed to operate on a 12V supply but will function correctly over the range 9 to 15V.

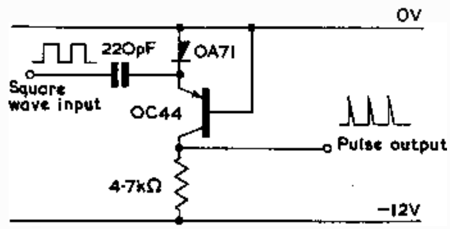

Fig. 4. Input shaping circuit

It will cover the frequency range 10c/s to 1·3Mc/s assuming that any ancillary shaping circuit always provides the input pulse already specified.

One decade unit has been subjected to an elevated temperature test, correct operation being maintained up to 55°C.

Conclusions

The decade counter circuit described has operated satisfactorily and reliably in the laboratory for some 150 hours and it is now intended to include the unit in pulse counting equipment.

Furthermore it should be mentioned that the circuit has been adopted as a basis for development by several commercial concerns for subsequent inclusion in counter chronometer instruments.

It would seem to be a distinct possibility that the availability of transistors with an even greater cut-off frequency than those used in the present circuit will enable the operation of this type of decade counter to approach 10Mc/s.

Acknowledgments

The author wishes to thank Mr. H. Walker and Mr. S. A. Goold, who, during recent summer vacations from the University of Nottingham, contributed to the design and testing of the circuits.

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An Introduction to Matrices
and their use in Transistor Circuit Analysis

By J. S. Bell* and K. Brewster†, B.Sc.

This article is written with the object of introducing matrix methods into the analysis of transistor circuits to those who are meeting the subject for the first time. The mathematical approach has been made as practical as possible as it was felt that too rigorous a treatment tends to obscure the ultimate issue—the analysis of practical circuits.

(Voir page 121 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 128)

SEVERAL methods of calculating the main parameters of a simple transistor amplifying stage have been developed. In some of these a general formula is derived by consideration of the equivalent circuit, but the expressions derived are usually unwieldy though more simple formulae[1] have been obtained by regarding the transistor as a 'black box'.

Although these various formulae, especially those derived using the 'black box methods', are extremely useful in particular applications, they are not applicable (without tedious modification) to, for instance, the analysis of a practical amplifying stage where there may be stabilizing and/or feedback networks incorporated. A further difficulty in using conventional analytical methods is that, unlike valve circuits, transistor stages cannot be separated from each other. To be more explicit, there is no valve buffer action between stages, that is succeeding transistor stages are not independent of each other. (A moment's thought will illustrate that filter networks are in the same category). The use of matrices reduces these difficulties.

In order to use matrix methods it is necessary to have some knowledge of the rules of matrix algebra. Then, by consideration of simple four terminal networks it is possible to derive the matrix for a particular circuit. Circuit analysis thus becomes analagous to the evaluation of an arithmetical expression using logarithms. There is little advantage in being able to use logarithms unless a table of logarithms is available. The same applies to matrix analysis; for most work it is essential to have a table of matrices available.

As many excellent works of reference exist upon matrix algebra there is no necessity to delve into the fundamental theory. It is therefore sufficient to state and to illustrate the basic rules governing their manipulation and then to derive the matrices of particular four terminal networks.

A matrix may be considered as a symbolic method of writing simultaneous equations. It is in fact a display of information and it is important to note that, unlike a determinant, a matrix cannot be worked out. Consider the general simultaneous equation below:

$$\begin{aligned} a_{11}x_1 + a_{12}x_2 \ldots\ldots a_{1n}x_n &= y_1 \\ a_{21}x_1 + a_{22}x_2 \ldots\ldots a_{2n}x_n &= y_2 \\ \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots & \\ a_{m1}x_1 + a_{m2}x_2 \ldots a_{mn}x_n &= y_m \end{aligned}$$

This may be expressed in symbolic form as:

$$\begin{bmatrix} a_{11} & a_{12} & \ldots\ldots\ldots & a_{1n} \\ a_{21} & a_{22} & \ldots\ldots\ldots & a_{2n} \\ \ldots & \ldots & \ldots\ldots\ldots & \ldots \\ a_{m1} & a_{m2} & \ldots\ldots & a_{mn} \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ \ldots \\ x_n \end{bmatrix} = \begin{bmatrix} y_1 \\ y_2 \\ \ldots \\ y_m \end{bmatrix}$$

or more concisely as:

$$[a]\ [x] = [y]$$

The terms in the square brackets are known as matrices and the few basic rules used to manipulate them are given below.

Basic Rules of Matrix Algebra

EQUALITY

For two matrices to be equal, all their corresponding elements must be equal.

$$[a] = [b] \text{ if, and only if, } a_{ij} = b_{ij}$$

ADDITION

The two matrices to be added must have the same number of columns and the same number of rows.

$$[a] + [b] = [c] \text{ where } c_{ij} = a_{ij} + b_{ij}$$

e.g. $$\begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix} + \begin{bmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{bmatrix} = \begin{bmatrix} a_{11}+b_{11} & a_{12}+b_{12} \\ a_{21}+b_{21} & a_{22}+b_{22} \end{bmatrix}$$

MULTIPLICATION

A necessary condition is that the second matrix must have as many rows as the first has columns. There is no limit to the number of rows the first has, or to the number of columns in the second.

$$[a]\ [b] = [c]$$

Consider the two matrices $[a]$ and $[b]$ where the product matrix $[c]$ is required. In order to find the value of the element in the i^{th} row and the j^{th} column in $[c]$ the elements of the i^{th} row of $[a]$ are multiplied into the j^{th} column of $[b]$ and the results are summed.

Symbolically $$c_{ij} = \sum_{k=1}^{n} a_{ik}\ b_{kj}$$

e.g.

$$\begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix} \begin{bmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{bmatrix} = \begin{bmatrix} a_{11}b_{11}+a_{12}b_{21} & a_{11}b_{12}+a_{12}b_{22} \\ a_{21}b_{11}+a_{22}b_{21} & a_{21}b_{12}+a_{22}b_{22} \end{bmatrix}$$

$$\begin{bmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{bmatrix} \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix} = \begin{bmatrix} b_{11}a_{11}+b_{12}a_{21} & b_{11}a_{12}+b_{12}a_{22} \\ b_{21}a_{11}+b_{22}a_{21} & b_{21}a_{12}+b_{22}a_{22} \end{bmatrix}$$

Note that $[a]\ [b] \neq [b]\ [a]$

The above examples illustrate that matrix multiplication is non-commutative. Only in a few isolated instances is this not true and so great care must be exercised in determining whether pre- or post-multiplication is required.

The Determination of the Matrix Parameters of a Linear Passive Network

In order to give the preceding theory some electrical significance consider the application to the fourpole illustrated in Fig. 1. The output voltage V_2 is a function of the input voltage V_1 and the current flowing through Z_1.

* *Formerly Blackburn and General Aircraft Ltd.*
† *Blackburn and General Aircraft Ltd.*

Similarly, the output current I_2 is a function of the input current I_1, in this case they are equal. Hence the following two equations define the voltage and current conditions respectively.

$$V_1 = 1 \, . \, V_2 + Z_1 \, . \, I_2 \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

$$I_1 = 0 \, . \, V_2 + 1 \, . \, I_2 \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

or in matrix notation

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} 1 \, . \, V_2 + Z_1 \, . \, I_2 \\ 0 \, . \, V_2 + 1 \, . \, I_2 \end{bmatrix} \ldots\ldots\ldots\ldots \quad (3)$$

Hence,

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} 1 & Z_1 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix} \ldots\ldots\ldots\ldots \quad (4)$$

This result may be checked by multiplication.

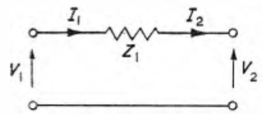


Fig. 1. Fourpole representing series impedance

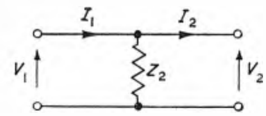


Fig. 2. Fourpole representing shunt impedance

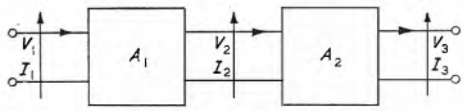


Fig. 3. Cascaded fourpoles

Similarly, by consideration of Fig. 2, the equations governing the conditions appertaining to voltage and current may be expressed in the form

$$V_1 = 1 \, . \, V_2 + 0 \, . \, I_2 \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

$$I_1 = V_2/Z_2 + 1 \, . \, I_2 \quad \ldots\ldots\ldots\ldots\ldots \quad (6)$$

Hence,

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 1/Z_2 & 1 \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix} \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

(Note that the determinant of the four element matrices of (4) and (7) is unity).

Equation (4) and (7) gives the A matrices of the networks concerned. Extending the arguments to cascaded networks (Fig. 3) it may be shown that

$$[A'] = [A_1] \times [A_2]$$

where $[A']$ is the matrix of the equivalent network.

There are, however, several other ways of defining the behaviour of a four pole. For instance, a matrix may have been derived in the following form

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} \ldots\ldots\ldots\ldots \quad (8)$$

i.e.

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = [Y] \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

Referring to Fig. 4, it may be shown that

$$[Y'] = [Y_1] + [Y_2]$$

where $[Y']$ is the matrix of the equivalent network.

By further consideration of the data obtained from Figs. 1 and 2 it is possible to derive the A and Y matrices of standard networks. These are illustrated in Table 1. Other types of matrix exist and may be referred to in other work[2], but to avoid undue complications these are not considered in this article.

The chief use of A matrices is in determining the characteristics of cascaded networks, while Y matrices are required when it is necessary to parallel fourpoles.

Example Using A Matrices

Before developing the matrix representation of transistor parameters it would be as well to determine the insertion loss ratio of the network indicated in Fig. 5 using A

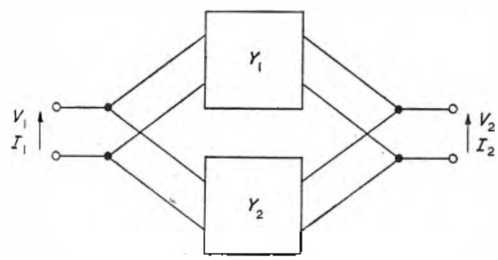
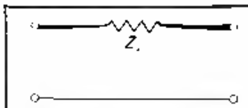
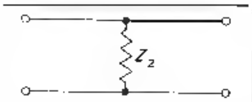
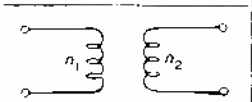
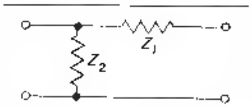
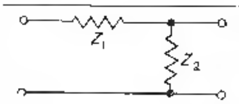
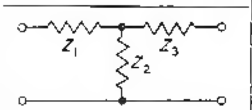
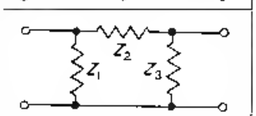
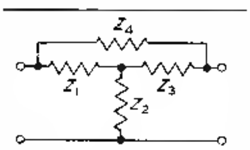


Fig. 4. Parallel fourpoles

TABLE 1.

Network	$[A]$	$[Y]$
	$\begin{matrix} 1 & Z_1 \\ 0 & 1 \end{matrix}$	$\begin{matrix} 1/Z_1 & -1/Z_1 \\ 1/Z_1 & -1/Z_1 \end{matrix}$
	$\begin{matrix} 1 & 0 \\ 1/Z_2 & 1 \end{matrix}$	$\begin{matrix} \infty & \infty \\ \infty & \infty \end{matrix}$
	$\begin{matrix} n_1/n_2 & 0 \\ 0 & n_2/n_1 \end{matrix}$	$\begin{matrix} \infty & -\infty \\ \infty & -\infty \end{matrix}$
	$\begin{matrix} 1 & Z_1 \\ 1/Z_2 & 1-Z_1/Z_2 \end{matrix}$	$\begin{matrix} \frac{Z_1-Z_2}{Z_1Z_2} & -1/Z_1 \\ 1/Z_1 & -1/Z_1 \end{matrix}$
	$\begin{matrix} 1+Z_1/Z_2 & Z_1 \\ 1/Z_2 & 1 \end{matrix}$	$\begin{matrix} 1/Z_1 & -1/Z_1 \\ 1/Z_1 & -\left(\frac{Z_1+Z_2}{Z_1Z_2}\right) \end{matrix}$
	$\begin{matrix} 1+Z_1/Z_2 & K/Z_2 \\ 1/Z_2 & 1-Z_3/Z_2 \end{matrix}$	$\begin{matrix} \frac{Z_2+Z_3}{K} & -Z_2/K \\ Z_2/K & -\left(\frac{Z_1+Z_2}{K}\right) \end{matrix}$
	$\begin{matrix} 1+Z_2/Z_3 & Z_2 \\ \frac{Z_1+Z_2+Z_3}{Z_1Z_3} & 1-Z_2/Z_1 \end{matrix}$	$\begin{matrix} \frac{Z_1+Z_2}{Z_1Z_2} & -1/Z_2 \\ 1/Z_2 & -\left(\frac{Z_2+Z_3}{Z_2Z_3}\right) \end{matrix}$
	$\dfrac{\begin{matrix} K-(Z_1+Z_2)Z_4 & KZ_4 \\ Z_1+Z_2+Z_4 & K-(Z_2-Z_3)Z_4 \end{matrix}}{K+Z_2Z_4}$	$\dfrac{\begin{matrix} K+(Z_2+Z_3)Z_4 & -(K+Z_3+Z_4) \\ K+Z_2Z_4 & -[K+(Z_1+Z_2)Z_4] \end{matrix}}{KZ_4}$

NOTE $K = Z_1Z_2 + Z_2Z_3 + Z_1Z_3$

matrices. This will serve to illustrate the principles outlined so far and will assist in the analysis of further circuits incorporating transistors.

Representing the circuit in matrix form

$$\begin{bmatrix}V_1\\ I_1\end{bmatrix} = \begin{bmatrix}1+(R_1+j\omega L)j\omega C & R_1+j\omega L\\ j\omega C & 1\end{bmatrix}\begin{bmatrix}1 & 0\\ 1/R_2 & 1\end{bmatrix}\begin{bmatrix}V_2\\ I_2\end{bmatrix}$$

$$\begin{bmatrix}V_1\\ I_1\end{bmatrix} = \begin{bmatrix}1+(R_1+j\omega L)j\omega C+(R_1+j\omega L)/R_2 & R_1+j\omega L\\ (1/R_2)+j\omega C & 1\end{bmatrix}\begin{bmatrix}V_2\\ I_2\end{bmatrix}$$

Only the voltage relationships are required, thus

$$V_1 = \left(1+j\omega CR_1-\omega^2LC+\frac{R_1+j\omega L}{R_2}\right)V_2+(R_1+j\omega L)\ I_2$$

Now I_2 is zero (see Fig. 5).

Therefore

$$V_1/V_2 = 1 - \omega^2LC + R_1/R_2 + j\omega(CR_1 + L/R_2)$$

i.e., insertion loss ratio =

$$1 - \omega^2LC\frac{R_2}{R_1+R_2} + j\omega\left(\frac{R_1R_2C}{R_1+R_2} + \frac{L}{R_1+R_2}\right)$$

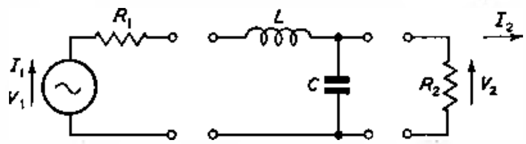


Fig. 5. Simple filter network

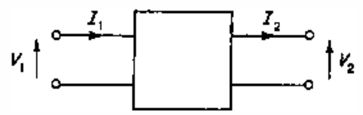


Fig. 6. 'Black box' representation of a transistor

Relation Between A and Y Matrices and h-Parameters

A transistor may be considered as a 'black box' (Fig. 6) and for this box one may write a number of sets of equations. The h-parameters can be defined according to the equations.

$$\left.\begin{aligned} V_1 &= h_{12}V_2 + h_{11}I_1\\ -I_2 &= h_{22}V_2 + h_{21}I_1\end{aligned}\right\} \quad \ldots\ldots\ldots (9)$$

where V_1, V_2, I_1 and I_2 are small a.f. signals. In manufacturer's data the parameters of a transistor are frequently expressed in terms of its h-parameters. However, they do not lend themselves well to manipulation (except in the simplest cases) and it is desirable to derive more convenient parameters such as the a parameters mentioned previously.

i.e.

$$\left.\begin{aligned} V_1 &= a_{11}V_2 + a_{12}I_2\\ I_1 &= a_{21}V_2 + a_{22}I_2\end{aligned}\right\} \ldots\ldots\ldots\ldots (10)$$

From equation (10) it can be seen that if $I_2 = 0$.

$$a_{11} = V_1/V_2$$

Or in symbolic notation $a_{11} = (V_1/V_2)_{I_2=0}$

Substituting the condition $I_2 = 0$ in equation (9),

$$V_1 = h_{12}V_2 + h_{11}I_1$$
$$0 = h_{22}V_2 + h_{21}I_1$$

or, eliminating I_1,

$$V_1 = h_{12}V_2 - h_{11}.(h_{22}/h_{21}).V_2$$

$$\therefore a_{11} = (V_1/V_2)_{I_2=0} = \frac{-h_{11}h_{22} - h_{12}h_{21}}{h_{21}}$$

$$= -|h|/h_{21}$$

where $|h| = h_{11}h_{22} - h_{12}h_{21}$

Similarly $a_{12} = (V_1/I_2)_{V_2=0}$

$$= -(h_{11}/h_{21})$$

Also

$$a_{21} = -(h_{22}/h_{21})$$
$$a_{22} = -(1/h_{21})$$

A set of y parameters can also be defined

$$I_1 = y_{11}V_1 + y_{12}V_2$$
$$I_2 = y_{21}V_1 + y_{22}V_2$$

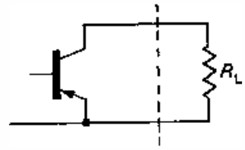


Fig. 7. A.C. circuit of transistor with load

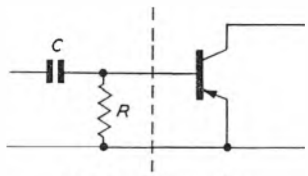


Fig. 8. *RC* coupling

or in matrix notation

$$\begin{bmatrix}I_1\\ I_2\end{bmatrix} = [Y]\begin{bmatrix}V_1\\ V_2\end{bmatrix} \text{ where } Y = \begin{bmatrix}y_{11} & y_{12}\\ y_{21} & y_{22}\end{bmatrix}$$

The y parameters can then be obtained in terms of the h-parameters as was done for the a case.

$$[Y] = \begin{bmatrix}y_{11} & y_{12}\\ y_{21} & y_{22}\end{bmatrix} = \begin{bmatrix}1/h_{11} & -h_{12}/h_{11}\\ -h_{21}/h_{11} & -|h|/h_{11}\end{bmatrix} \quad \ldots (11)$$

$$[A] = \begin{bmatrix}a_{11} & a_{12}\\ a_{21} & a_{22}\end{bmatrix} = \begin{bmatrix}-|h|/h_{21} & -h_{11}/h_{21}\\ -h_{22}/h_{21} & -1/h_{21}\end{bmatrix} \quad \ldots (12)$$

Similarly

$$[Y] = \begin{bmatrix}a_{22}/a_{12} & -|a|/a_{12}\\ 1/a_{12} & -a_{11}/a_{12}\end{bmatrix} \quad \ldots\ldots\ldots\ldots (13)$$

$$[A] = \begin{bmatrix}-y_{22}/y_{21} & 1/y_{21}\\ -|y|/y_{21} & y_{11}/y_{21}\end{bmatrix} \quad \ldots\ldots\ldots\ldots (14)$$

It is usual to denote the h-parameters for the common base connexion by h as in the previous treatment. If the common emitter connexion is used h is replaced by h' and by h'' for the grounded collector connexion.

Modification of a and y Parameters due to Practical Considerations

The modification to the transistor matrix caused by adding a load to the grounded emitter stage will now be considered. The a.c. circuit is illustrated in Fig. 7.

The resultant A matrix $[A_R] = [A_T]\,[A_{RL}]$

where $[A_T]$ is the A matrix of the transistor

$[A_{RL}]$ is the A matrix of the load R_L

$$A_R = \begin{bmatrix} -|h'|/h_{21}' & -h_{11}'/h_{21}' \\ -h_{22}'/h_{21}' & -1/h_{21}' \end{bmatrix} \begin{bmatrix} 1 & 0 \\ 1/R_L & 1 \end{bmatrix} \quad \ldots\ldots\ (15)$$

$$= \begin{bmatrix} -|h'|/h_{21}' - h_{11}'/R_L h_{21}' & -h_{11}'/h_{21}' \\ -h_{22}'/h_{21}' - 1/h_{21}'R_L & -1/h_{21}' \end{bmatrix} \quad \ldots\ldots\ (16)$$

Note that $[A_R]$ may be used to derive the well-known formulae for stage gain etc. that is often quoted for simple a.f. stages. This is considered for the more general case later in this article.

The effect of RC coupling is also obtained directly from A matrices. This network is illustrated in Fig. 8.
Resultant $[A_R] = [A_{CR}]\ [A_T]$
where $[A_{CR}]$ is the A matrix of the CR network.

$$[A_R] = \begin{bmatrix} 1+(1/j\omega CR) & 1/j\omega C \\ 1/R & 1 \end{bmatrix} \begin{bmatrix} -|h'|/h_{21}' & -h_{11}'/h_{21}' \\ h_{22}'/h_{21}' & 1/h_{21}' \end{bmatrix} \quad \ldots\ldots\ (17)$$

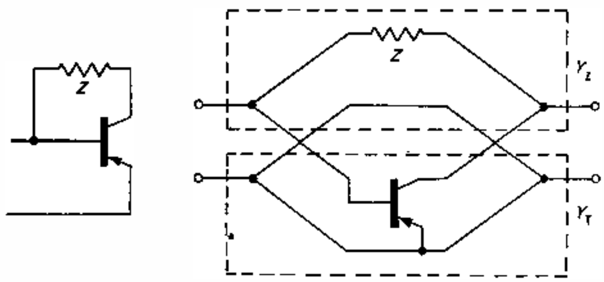


Fig. 9. Voltage feedback and equivalent fourpoles

$$= \begin{bmatrix} \left(1+\frac{1}{j\omega CR}\right)|h'|/h_{21}' - \frac{h_{22}'}{j\omega C\, h_{21}'} & -\left(1+\frac{1}{j\omega CR}\right)h_{11}'/h_{21}' - \frac{h_{11}'}{j\omega C\, h_{21}'} \\ -1/R\,.\,|h'|/h_{21}' - h_{22}'/h_{21}' & -h_{11}'/Rh_{21}' - 1/h_{21}' \end{bmatrix} \quad \ldots\ldots\ (18)$$

For voltage feedback it is necessary to use the y parameters (it is assured that the working conditions are not altered). (Fig. 9.)

$$[Y_R] = [Y_Z] + [Y_T]$$

where $[Y_R]$ is the resultant Y matrix

$[Y_Z]$ is the Y matrix of the feedback network

$[Y_T]$ is the Y matrix of the transistor

$$[Y_R] = \begin{bmatrix} 1/h_{11}' & -h_{12}'/h_{11}' \\ -h_{21}'/h_{11}' & -|h'|/h_{11}' \end{bmatrix} + \begin{bmatrix} 1/Z & -1/Z \\ 1/Z & -1/Z \end{bmatrix} \quad \ldots\ldots\ (19)$$

$$= \begin{bmatrix} 1/h_{11}' + 1/Z & -h_{12}'/h_{11}' \quad 1/Z \\ -h_{21}'/h_{11}' + 1/Z & -|h'|/h_{11}' - 1/Z \end{bmatrix} \quad \ldots\ (20)$$

Thus the characteristics of the transistor and its associated circuit elements can be collected together into a single matrix.

The effect of an impedance Z in the emitter lead (Fig. 10) can also be calculated as above but as this requires other parameters, not previously used, only the results are stated.

The resulting A matrix of the transistor A_Z is given by

$$[A_Z] = \begin{bmatrix} \frac{|h'|+Zh_{22}'}{Zh_{22}'-h_{21}'} & \frac{h_{11}'(1+h_{22}'Z)+Z(1+h_{21}')}{Zh_{22}'-h_{21}'} \\ -\frac{h_{22}'}{h_{21}'-h_{22}'Z} & \frac{1+h_{22}'\,.\,Z}{Zh_{22}'-h_{21}'} \end{bmatrix} \quad \ldots\ldots\ (21)$$

In practice, $[A_Z]$ may frequently be simplified by using the approximations

$$h_{22}'Z \ll 1$$
$$h_{22}'Z \ll h_{21}$$

Thus

$$A_Z = \begin{bmatrix} -\left(\frac{|h'|+Zh_{22}'}{h_{21}'}\right) & -\left(\frac{h_{11}'+Z(1+h_{22}')}{h_{21}'}\right) \\ -h_{22}'/h_{21}' & -1/h_{21}' \end{bmatrix} \quad \ldots\ldots\ (22)$$

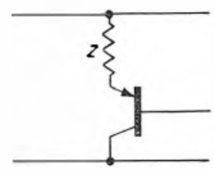


Fig. 10. Impedance in emitter lead

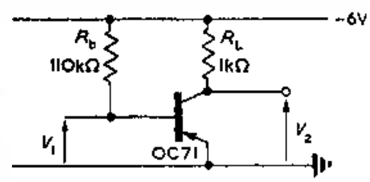


Fig. 11. Simple a.f. amplifier

The Calculation of Dynamic Performance from the *a* Parameters

The configuration illustrated in Fig. 6 may, to be more general, be regarded as a transistor and its associated circuits. If all are reduced to a single A matrix, then

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = [A] \begin{bmatrix} V_2 \\ I_2 \end{bmatrix} \quad \ldots\ldots\ (23)$$

where V_1 and I_1 are the input voltage and current and V_2 is the output voltage. As the load is included inside the A matrix I_2 is zero.

Thus

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = [A] \begin{bmatrix} V_2 \\ 0 \end{bmatrix} \quad \ldots\ldots\ (24)$$

or

$$V_1 = a_{11} V_2 \quad \ldots\ldots\ (25)$$
$$I_1 = a_{21} V_2 \quad \ldots\ldots\ (26)$$

From equation (25),

$$\text{voltage gain } A_V = V_2/V_1 = 1/a_{11} \quad \ldots\ldots\ (27)$$

If Z_L is the load, then where I_L is the load current

$$V_2 = I_L\,.\,Z_L$$

From equation (26) $I_1 = a_{21} \cdot I_L \cdot Z_L$

$$\therefore \text{ Current gain } A_1 = I_L/I_1 = \frac{1}{a_{21} \cdot Z_L} \quad \ldots\ldots\ldots\ldots\ldots \quad (28)$$

Dividing equation (25) by equation (26)

$$\text{Input impedance } Z_{in} = V_1/I_1 = a_{11}/a_{21} \quad \ldots\ldots\ldots\ldots \quad (29)$$

The output impedance Z_{out} may be determined by the open-circuit voltage, short-circuit current technique. The result only is quoted.

$$Z_{out} = \frac{a_{12} + a_{22} R_s}{a_{11} + a_{21} R_s} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (30)$$

where R_s is the source resistance.

$$\text{Power gain} = \text{voltage gain} \times \text{current gain}$$
$$= \frac{1}{a_{11} \cdot a_{21} \cdot Z_L}$$

Practical Examples

Consider Fig. 11 and suppose it is required to determine the voltage gain and input impedance of the simple circuit shown.

Now the h-parameters of a transistor are functions of collector current, or alternatively of collector voltage, and it is necessary to determine one or the other as the first step. Using the output characteristics of the OC71 it will be seen that the connexions correspond to a collector current of 3mA. Upon consulting the manufacturer's data it is found that

$$h_{11} = 800\Omega \qquad h_{12}' = 5{\cdot}4 \times 10^{-4}$$
$$h_{21} = 47 \qquad h_{22}' = 80 \text{ micromhos}$$
$$|h'| = h_{11}' h_{12}' - h_{12}' h_{21}'$$
$$= 39 \times 10^{-3}$$

Using equation (16) in conjunction with equation (27)

$$A_V = 1/a_{11} = \left\{ -(|h'|/h_{21}' + h_{11}'/R_L h_{21}') \right\}^{1}$$
$$= \left\{ -1/47\ (39{\cdot}10^{-3} + 800/1\,000) \right\}^{-1}$$
$$= -56$$

$$Z_{in} = a_{11}/a_{21} = \frac{|h'|\, R_L + h_{11}'}{h_{22}'\, R_L + 1}$$
$$= 839/1{\cdot}08$$
$$= 778\Omega.$$

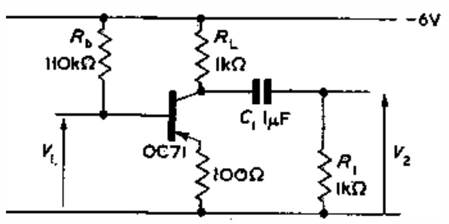


Fig. 12. A.F. amplifier with current feedback and coupling network

Fig. 12 illustrates a more complicated a.f. amplifier. For convenience the working conditions have been kept the same as in the previous example. As before, let it be supposed that the voltage gain and input impedance are required. The resultant circuit matrix is given by

$$[A_R] = [A_{Rb}]\,[A_{TZ}]\,[A_{RL}]\,[A_{C1R1}]$$

$$= \begin{bmatrix} 1 & 0 \\ 1/R_b & 1 \end{bmatrix} \begin{bmatrix} -\left(\dfrac{|h'| + Zh_{22}'}{h_{21}'}\right) & -\left(\dfrac{h_{11}' + Z(1 + h_{21}')}{h_{21}'}\right) \\ -h_{22}'/h_{21}' & -1/h_{21}' \end{bmatrix}$$

$$\begin{bmatrix} 1 & 0 \\ 1/R_L & 1 \end{bmatrix} \begin{bmatrix} 1 + \dfrac{1}{j\omega C_1 R_1} & \dfrac{1}{j\omega C_1} \\ 1/R_1 & 1 \end{bmatrix}$$

$$= \begin{bmatrix} 1 & 0 \\ 1/(110 \,.\, 10^3) & 1 \end{bmatrix} \begin{bmatrix} -10^{-3} & -119 \\ -1{\cdot}7 \,.\, 10^{-6} & -0{\cdot}0213 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ 10^{-3} & 1 \end{bmatrix}$$

$$\begin{bmatrix} 1 + \dfrac{10^3}{j} & 10^6 \\ 10^{-3} & 1 \end{bmatrix}$$

$$= \begin{bmatrix} -(239 \,.\, 10^{-3} + (120/j\omega)) & -\left(119 + \dfrac{120 \,.\, 10^3}{j\omega}\right) \\ -\left(46{\cdot}3 \,.\, 10^{-6} + \dfrac{24 \times 10^{-3}}{j\omega}\right) & -(22{\cdot}3 \,.\, 10^{-3} + (24/j\omega)) \end{bmatrix}$$

Assuming an input frequency of 100c/s.

$$\text{Voltage gain} = 1/a_{11} = -\frac{1}{0{\cdot}239 - j\,0{\cdot}191}$$
$$= -3{\cdot}3 \angle 39°$$
$$= 3{\cdot}3 \angle 219°$$

$$Z_{in} = a_{11}/a_{21} = \frac{1}{3{\cdot}3 \angle 39°} \cdot \frac{1}{0{\cdot}06 \,.\, 10^{-3} \angle -40°}$$
$$= \frac{1}{0{\cdot}198 \angle -1°}$$
$$\simeq 505\Omega.$$

Conclusion

Brief, but sufficient details of a system for analysing both simple and complex circuits containing transistors have been given.

Transistor parameters vary widely between specimens and the analysis described must only be regarded as approximate unless the individual parameters are known for the particular specimen which is to be used. Linear operation is assumed and care must be taken that the working conditions of a transistor are not unduly upset, by for instance, voltage feedback connexions. Such precautions lie within the realms of common circuit practice but are apt to be overlooked if a too rigorous mathematical approach is adopted.

Matrices have been found suitable in the analysis of transmission lines[3] and thermionic valve circuits[4] and thus they embrace a comprehensive field.

In the transistor cases considered in this article it is emphasized that the results are perfectly general. That is, in order to use the formulae derived it is only necessary to insert the h-parameters appertaining to the selected connexion, e.g., grounded collector, grounded base or grounded emitter. These are all available from manufacturers literature.

Acknowledgments

The authors are indebted to Dr. H. Fuchs, Chief Electronic Engineer, Blackburn and General Aircraft Ltd for his kind permission to publish this article.

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LETTERS TO THE EDITOR

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

Tuned Transistor Audio Amplifier

DEAR SIR,—The following circuit was designed to serve as part of a portable phase-switched radio-astronomy receiver. The principal requirements were high gain and appreciable selectivity combined with small size, light weight and low current consumption. The complete receiver uses two such stages, the output of the final stage being fed via a phase-sensitive detector to a pen recorder.

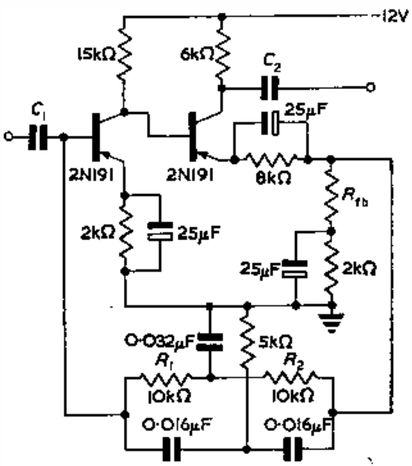
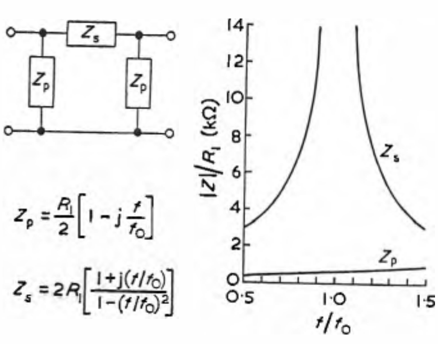


Fig. 1. Circuit of amplifier

The complete circuit diagram, shown in Fig. 1, consists of a direct coupled amplifier, with frequency selective feedback from the emitter of the second stage to the base of the first. The feedback network is a twin-T[1], tuned to reject 1kc/s. The use of this network with junction transistors has been proposed previously[2]. In the present circuit the twin-T network also provides effective bias stabilization through the series resistors R_1 and R_2.

The design procedure for the amplifier is conventional, the collector current of each transistor (2N191 or 2N175) being 0·5mA in the interest of low overall current consumption. This restricts the maximum undistorted output voltage to approximately 2V r.m.s., which is adequate in the present application. The method of biasing is such that interchange of transistors is possible without any appreciable change in response or gain.

Fig. 2. Response of Twin-T Network



The size of the elements used in the twin-T network must be chosen so that there is little loading on the input and output sections of the amplifier. At the same time maximum off-resonance feedback is desired. Fig. 2 shows the π equivalent circuit of a symmetrical twin-T network with curves showing how the shunt and series impedances vary with frequency. With the component values

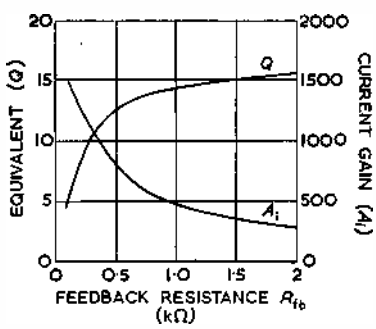


Fig. 3. Effect of feedback resistance on gain and selectivity

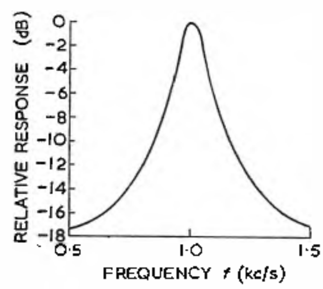


Fig. 4. Response of tuned amplifier

used, the variation of the amount of feedback with frequency is virtually a function of the series impedance alone.

Fig. 3 shows the effect of the feedback resistance R_{fb} on the gain and selectivity of the amplifier. As R_{fb} is increased, the amount of feedback increases rapidly resulting in an increase in selectivity (measured by $Q = f_0/\Delta_f$) and a reduction in gain. This continues until the input impedance of the second stage is approximately equal to the output impedance of the first stage. Thereafter the second stage acts as an emitter-follower and the changes in gain and response are small. It can be seen that R_{fb} provides a simple means of gain control, provided it is not necessary to reduce the gain to zero.

Fig. 4 shows the response curve of the amplifier for $R_{fb} = 300\Omega$. The power gain at resonance is about 60dB. It should be noted that a high impedance source is required for maximum selectivity, resulting in a slight loss due to mismatching. However, the effect of selectivity does not become appreciable until the source resistance is less than 2kΩ, hence two such amplifiers may be connected in cascade with little loss in overall power gain or selectivity.

Yours faithfully,

W. D. RYAN,

Department of Electrical Engineering,
Royal Military College of Canada,
Kingston,
Ontario.

REFERENCES

1. SMITH, D. H. The Characteristics of Parallel-T.R.C. Networks *Electronic Engng. 29*, 71 (1957).
2. SOKRABH, N. RC Filters and Oscillators Using Junction Transistors *Electronic Engng. 29*, 606 (1957).

A Four Terminal Network

DEAR SIR,—It has occurred to the writer that one statement in his recent letter on measuring the admittances of a four terminal network[1] might be rather misleading. The letter showed how one could determine the four-terminal parameters of the network by measuring two-terminal admittances, one of the latter being the admittance to the two ends of the network fastened together. The discussion might have been taken to imply that the only reason for this is one of practical convenience; and that the properties can always be determined by measuring the admittance at each end of the network, the other end being first open-circuited and then shorted.

Actually, of course, this is not quite so. Of the four such measurements which might be made, three would suffice to determine the parameters of a reciprocal network. For a non-reciprocal network, however, the fourth would give another expression for the product $y_{12}y_{21}$, the product of the transfer admittances, but no way of separating these two factors. For a reciprocal network, of course, since $y_{12} = y_{21}$, this difficulty does not arise. On the other hand, the set of measurements which includes the admittance to the input and output fastened together will enable one to find all the parameters in every case.

Yours faithfully,

H. L. ARMSTRONG,

Queen's University,
Kingston,
Ontario.

REFERENCE

1. ARMSTRONG, H. L. *Electronic Engng. 30*, 669, (1958).

Short News Items

The Royal Society has announced the awards of grants by the Paul Instrument Fund Committee as follows.

£3 835 to Professor A. L. Cullen, Professor of Electrical Engineering in the University of Sheffield, for the construction of a detector in which radiation pressure is used to convert a microwave signal to an audio or intermediate frequency.

£545 to Dr. R. Feinberg, of the Electrical Engineering Department, Manchester College of Science and Technology, for the construction of an instrument for measuring the intensity of soft X-radiation generated in experimental high voltage valves.

£1 000 to Mr. R. M. Sillitto, lecturer in natural philosophy, University of Edinburgh, for the development of a 30-channel pulse-height analyser using transistors in place of thermionic valves.

The Paul Instrument Fund Committee, composed of representatives of the Royal Society, the Physical Society, the Institute of Physics and the Institution of Electrical Engineers, was set up in 1945 'to receive applications from British subjects who are research workers in Great Britain for grants for the design, construction and maintenance of novel, unusual or much improved types of physical instruments and apparatus for investigations in pure or applied physical science'.

The CIBA Fellowship Trust has been founded for the purpose of improving and increasing the interchange of ideas between scientists in the United Kingdom and on the Continent. Several Fellowships will be awarded for tenure during the academic year 1959-1960 at Continental universities or institutions for research in chemistry, physics or some other allied scientific subject. They will be awarded to graduates of universities situated in the United Kingdom of Great Britain and Northern Ireland or in the Republic of Ireland, or to members of those universities who graduated in 1958. It is anticipated that some Fellowships will be awarded to recent graduates and other Fellowships to candidates who have already taken their Ph.D., or who have already spent some time in industry. The basic award for Fellows who have already undergone training in research and wish to continue Post Doctorate studies will be £800 per annum plus allowances including cost of living and children's allowance. For candidates who have obtained a First Degree in Science and wish to undergo training in research, the basic award will be £500 per annum plus allowances including cost of living, etc. Further details are available from the Secretary of the CIBA Fellowship Trust, CIBA (A.R.L.) Ltd, Duxford, Cambridge.

The Institute of Physics is arranging a one-day symposium entitled 'Current Developments in the Production of High Vacua' to take place in London on 17 April. There will be three sessions: (a) Chemical and ionic pumping in kinetic vacuum systems; (b) Problems in the production of high vacua in large equipment; (c) Analysis of residual gases in kinetic vacuum systems. All communications regarding the symposium should be sent to the Secretary, the Institute of Physics, 47 Belgrave Square, London, S.W.1.

An International Congress on Microwave Tubes will be held in Munich from 7-11 June, 1960. It is to be organized by Verband Deutscher Elektrotechniker and further information may be obtained from the Reception and Information Office, Brienner Strasse 40, Munich 37.

An Information Centre has been opened at the London Office of the United Kingdom Atomic Energy Authority. Its purpose is to provide a convenient centre in London where members of the public, commercial firms and other organizations may consult published unclassified material (i.e. with no security restrictions) and seek advice on sources of information on United Kingdom atomic energy matters. A collection of U.K.A.E.A. Unclassified Reports is available for reference. Indices and abstracts of atomic energy literature are filed and publications about the work of the United Kingdom Atomic Energy Authority are available. The Centre forms part of the Public Relations Branch of the Authority. The facilities which it offers are complementary to the already well established information services provided by the Authority at Harwell, Risley and, to a limited extent, Aldermaston, for dealing with scientific and technical inquiries. The Information Centre and the Photographic Library are open from Mondays to Fridays inclusive between 9.30 a.m. and 5 p.m. at the United Kingdom Atomic Energy Authority, 11 Charles II Street, London, S.W.1.

The Department of Scientific and Industrial Research announces that expenditure on research will be nearly doubled in the next five years. Under its second five-year plan, for the period 1959-64, approximately £61M. will be made available to the Department compared with £36M. for the first quinquennium which ends on 31 March. The largest expansion is planned to take place in the field of scientific grants to the universities. Post-graduate awards to students will be increased by about 10 per cent each year until in 1963-64 it is hoped some 3 800 students will be receiving D.S.I.R. grants for research training. In the same year it is expected that D.S.I.R. support for special research in the research departments of universities will be operating on a scale of about £1½M. per annum. In support of additional research carried out in the Department's own laboratories, expansion of staff at the rate of about 6 per cent per annum—or approximately 30 per cent over five years—is included in the plan.

The Plessey Co. Ltd mentions that an agreement for the interchange of technical information concerning the manufacture of cast permanent magnets of high energy, for use in the electronic and electrical industries, has been concluded with the Arnold Engineering Co of Illinois. Plessey has for many years been concerned with the development and production of magnetic materials and the new agreement will ensure that adequate technical progress is maintained in the future. These magnets, which will be marketed under the registered trade mark of 'Magloy', have a variety of applications especially in radar equipment, loudspeakers, electrical meters, magnetos, and electric motors and generators. The magnets will be manufactured by Preformations Ltd, a company formed by Plessey for the development and production of permanent magnets in this country. The company is located with other Plessey Group plants on the new Cheney Manor factory estate at Swindon, Wiltshire, where the build-up of the highly specialized plant needed for magnet production has now reached an advanced stage.

Notes from ———

NORTH AMERICA

Conventions and Symposiums

I.R.E. National Convention

The Institute of Radio Engineers' National Convention will be held from 23 to 26 March at the Waldorf-Astoria Hotel, and the New York Coliseum in New York City. More than 55 000 engineers and scientists from 40 countries are expected to attend.

A comprehensive programme of 275 papers, covering the most recent developments in the fields of all 28 I.R.E. Professional Groups, will be presented in 54 sessions at the Waldorf-Astoria and the Coliseum. A feature of the programme will be a special symposium on 'Future Developments in Space' to be held Tuesday evening, 24 March. The symposium will be conducted by Eric A. Walker, president of Pennsylvania State University, and a panel of leading experts. The complete programme will be announced later.

The Radio Engineering Show, occupying all four floors of the Coliseum, will accommodate approximately 850 exhibitors.

Aeronautical Electronics Conference

The Eleventh National Aeronautical Electronics Conference (NAECON) will be held from 5 to 6 May at the Biltmore and Miami Hotels, Dayton, Ohio. This conference is sponsored by both local and national technical groups of the Institute of Radio Engineers.

The theme of the conference is, 'Electronics Systems in the Space Age' and, in addition to comprehensive technical presentations, a forum to consist of some of the best-informed figures in the aeronautical electronics field is again planned to exchange ideas on better electronic systems management.

Exhibit space has been increased by one-third of the 1958 NAECON and leading aeronautical equipment concerns are again scheduled to present their latest equipment designs.

Synthetic Quartz Production

Details of pilot plant production of synthetic quartz have recently been announced jointly by the Western Electric Company and Bell Telephone Laboratories.

Synthetic quartz provides a number of advantages over natural quartz in addition to its availability in any quantity and size. Seeds can be cut to provide grown crystals which can be sawn in the most efficient manner. Also, synthetic quartz has natural crystal faces which allow easier orientation of the stock for cutting into crystal units. It has none of the foreign inclusions which usually occur in natural quartz, and it can be produced without either optical or electrical twinning. It is estimated that the greater yield resulting from these improvements is at present at least 2½ times that of natural quartz.

The hydrothermal process has been widely investigated on a laboratory scale for several years, and quartz crystals about ¾in square by 1-½in long and larger have been grown at Bell Laboratories. The processing equipment has been scaled up for pilot plant investigation and quartz crystals up to two to three inches in each cross-section dimension and five to six inches long have been grown in this equipment. Operations have not been directed toward growing the biggest possible crystals, but toward determining the optimum processing parameters.

In the hydrothermal process, a long narrow autoclave mounted vertically is filled with an alkaline solution, usually sodium hydroxide. Small pieces of readily available natural quartz are then placed in the bottom of the vessel to provide the nutrient. Future production will utilize even less costly and more readily available material such as high quality sand as the nutrient. Seed plates cut from either natural quartz, or previously grown synthetic crystals, are hung from a rack in the upper section of the vessel. After sealing, the autoclave is heated to the required temperature, and maintained under a constant temperature differential from bottom to top for the requisite processing time. This period has varied from a week to several weeks, depending on the experiment being performed.

Economical growth rates considerably in excess of 0·06in/day have been achieved. Since growth occurs primarily in one dimension, flat seed plates are used which effectively become thicker but change little in other dimensions. The rate of growth varies with each of a number of conditions, including the temperature differential between nutrient and seed plate area, the alkalinity of the solution, the temperature, and per cent fill (or pressure).

Operation of the hydrothermal process depends on the maintenance of a temperature differential between the nutrient area and the seed plate area. The nutrient dissolves on the hotter lower region and is carried by convection currents to the cooler upper region. The lower temperature here leads to a supersaturated condition in the nutrient solution, which causes the dissolved quartz to redeposit on to the seed plates in single crystal form.

One of the main problems which appeared in scaling up the pilot plant equipment was that of finding a suitable closure for the autoclaves. The laboratory vessels were simply welded tubes supported by capped, high pressure piping. A repetitive closure was required for the larger autoclave and a satisfactory closure was developed which gives a tighter seal as the pressure inside builds up.

The combination of high temperature and pressure conditions present in this process is believed to be one of the most severe encountered in any process industry at present. This gave rise to a number of material selection and equipment design problems.

Submarine Cable Testing

A dry-land 'ocean', with the environment found two nautical miles deep in the sea, is under construction by Bell Telephone Laboratories for long term testing of underwater cables.

The man-made ocean, believed to be the only one of its kind in the world, will help examine cable 'ageing', the minute changes in electrical characteristics which may occur as the telephone cable rests on the floor of an ocean.

It was estimated that a simulated ocean would be about 75 per cent cheaper than the cost of one-time operation in the deep sea, and would permit much more accurate control and measurement of cable samples.

The 'ocean' at Chester is 315ft long, and buried 7ft under the ground. The 'bed' of the simulated sea is a concrete trough 3ft in depth and about 8ft in width, with pre-cast concrete slabs for a cover. On top of the trough is 4ft of soil. Going to the 7ft depth assures reasonably constant earth temperatures the year round.

The trough will be filled with water maintained at an ocean bottom temperature of 37°F by means of a network of cooling pipes cast within the trough walls. To obtain meaningful measurements, this temperature must be held to within one-tenth of a degree during measurements.

In the trough will be 10 lengths of steel tubing arranged in pairs, each tube large enough in diameter to hold a typical undersea cable plus salt water with the approximate salinity found two miles below the surface of the ocean.

An hydraulic system will maintain a pressure of 5 000lb/in^2 inside each tube to duplicate conditions at a depth of 12 000 ft. Each cable sample, 630ft in length, will be looped in a pair of pipes so that both ends will be available at one point for measurement.

Ground anchors at one end of the installation will provide for cable tensioning to simulate cable-laying conditions. Tension will be applied for a brief period after the cable is placed in its tube and will be reduced slowly to provide conditions duplicating a trip from ship to ocean bottom.

The cable which finally is left in the simulated ocean will be subjected to tests lasting from five to ten years.

BOOK REVIEWS

Television Engineering—Principles and Practice. Vol. 4 General Circuit Techniques

By S. W. Amos and D. C. Birkinshaw. 268 pp. 149 figs. Demy 8vo. Iliffe & Sons Ltd. 1958. Price 35s.

THIS book is the fourth in a series of BBC engineering training manuals. Although it is stated to be the last, it is hoped that this is not so.

While primarily written for BBC engineers and therefore dealing mainly with circuits and techniques used at the transmitting end of television systems, there is much to interest circuit engineers.

After dealing with counter circuits and frequency dividers d.c. clamping and restoration is dealt with. A chapter on gamma control follows and a number of circuits are described. Photographs illustrate correct and incorrect gamma. They are not particularly good examples and step wedges might have been better. A very interesting and complete chapter on the theory, design and application of delay lines leads nicely to a chapter on equalizers. The development of equalizers is taken step by step up to variable, Bode type ones and a complete section with all values is shown.

Chapters X, XI and XII deal with scanning problems in field and line output stages. As in the rest of the book theory is developed in easy stages and the mathematics are straightforward, easily followed and practical examples are fully worked out.

The final chapters deal with the eternal problem of amplifier response and holding off the enemy, capacitance. Shunt regulated amplifiers are dealt with starting with simple basic circuits, followed by ones using compensating networks and the improvements possible by such methods are shown. Some of the circuits seem to be rather over-praised. For instance, when discussing the several advantages of series instead of parallel valves, it is pointed out that the mean current is reduced by one-half and a saving of 2kW is quoted without mentioning that the h.t. supply would need to be higher. Loads are assumed to be simple capacitors which might not always be the case. However, as elsewhere, these chapters are full of practical information.

It is difficult to find fault with such a well written, illustrated and produced book, small complaints are that apart from in "brute force" output stages, no line linearity methods are even mentioned, nor is the linearity problem of flat faced tubes, particularly wide angle ones.

D.A.P. occurs here and there, surely a relic now, p-p being more widely used.

Perhaps the description of the operation of an efficiency diode would have been clearer if electron flow only had been described. As it is current is said to "flow through C and the diode and upwards through the primary winding (this is the direction of conventional current flow; electron flow is in the opposite direction)".

This reviewer enjoyed the book, a better understanding of a number of circuits resulted, and parts of it were duly copied into the notebook.

C. H. BANTHORPE

Automatic Control: Principles and Practice

By Werner G. Holzbock. 258 pp. 110 figs. Demy 8vo. Reinhold Publishing Corporation, New York. Chapman & Hall Ltd. Ltd, London, 1958. Price 60s.

AS an essentially practical introduction to the techniques used in current automatic control practice, this book is very good. The mathematics used have been reduced to an absolute minimum. Since both the practical application and derivation of the automatic control techniques are covered this volume should appeal to the student and the qualified engineer.

One ever present difficulty in a text of this nature is the difference between American and British terminologies in control engineering. The author uses the first chapter to introduce the American symbols and terms. Even at this early stage the importance of stability in a control system is discussed.

Static characteristics are well treated and include a brief mention of several of the common non-linearities including dead time and hysteresis effects.

When dynamic process characteristics are considered both step function response and frequency response methods are introduced and their effectiveness in process analysis critically compared. This section also includes a method of adjusting the settings on industrial controllers to obtain optimum control. Phase and gain margins as a measure of stability and also Nyquist's stability criterion are discussed without the use of mathematics.

Over half of the volume is devoted to describing in considerable detail the physical equipment which is utilized in the measurement and control of many process variables. Details of operation are given for most existing equipments. Control of temperature is discussed at great length, pointing out the difficulties of measurement in particular.

Rather more complex control systems are introduced in the penultimate chapter. The component parts used in composite control schemes are well described including cascade, ratio and programme (Time Schedule) control.

Several industrial applications of control are briefly described in the final chapter.

It is unfortunate that very few references to existing works are made during the text. Also no bibliography or recommended sources of more specialized information is given.

The illustrations used are very good and the book is generally well produced.

E. F. SNARE

Transistor Technology—Vol. II

Bell Laboratories Series. Edited by F. J. Biondi. 701 pp. 300 figs. Demy 8vo. D. Van Nostrand Company Ltd, New York, Toronto, London. 1958. Price 131s. 6d.

THIS book is the second of a series of three volumes which deal with semiconductor technology and devices. It consists of a collection of papers which appeared in the technical press, mostly during the years 1954, 1955 and 1956; these were the years when the rate of progress in the understanding of the transistor art and in the evolution of new transistor structures was greatest. The papers are mainly by Bell Telephone Laboratories engineers but a few by engineers from other organizations are included. Where necessary, the subject matter has been brought up to date by the addition of further material.

The book is in two parts. Part I—Technology of Materials—has nine papers collected into three chapters which deal with effects of impurities and imperfections in germanium and silicon with special emphasis on the latter.

Part II—Principles of Transistor Design—has twenty-nine papers collected into eight chapters. Chapter IV is concerned with forward characteristic, reverse breakdown mechanisms and speed of response of junction diodes. Chapter V deals with junction transistors wtih emphasis on the large signal behaviour and high frequency design parameters. Chapter VI discusses avalanche transistors and p-n-p-n devices, while Chapter VII discusses high frequency junction tetrodes. The remaining three chapters are concerned with radiation sensitive devices, field effect devices, noise in junction transistors and the effect of surface phenomena on the characteristics of p-n junctions. A fairly extensive list of references is given at the end of each paper.

Appendix I gives the list and definition of semiconductor terms which was published in the Proceedings of the I.R.E. in 1954, while Appendix II gives the I.R.E. Standard Letter Symbols published in 1956.

The Bell Telephone Laboratories have performed a great service in collecting these important papers together in book form. Any engineer or physicist who is concerned with semiconductor research, with the development or manufacture of semiconductor devices, with the development of equipment incorporating semiconductor devices or with the study of the subject at a university will find the book invaluable for reference purposes.

E. G. JAMES

Radioisotopes in Scientific Research

Volume 1 Research with Radioisotopes in Physics and Industry

Edited by R. C. Extermann. 761 pp. 300 figs. Royal 8vo. Pergamon Press, New York, Los Angeles, Paris, London. 1958. Price £7 a volume. (Complete set of four volumes £28.)

THIS series includes the proceedings of the first UNESCO International Conference on Isotopes held in Paris in September 1957.

This was the largest Conference on isotopes ever held. More than 200 papers were read and discussed. Some 1 200 scientists from 61 countries attended.

In the physical sciences sessions more than 100 papers were read, recording the great progress which has been made since the last Geneva Conference almost three years ago.

In the biological sciences sessions all the major fields of biological study in which isotopes are used, excepting only clinical diagnosis and the biological effects of radiation, were embraced in more than 100 papers read at 17 sessions.

English-Russian Russian-English Electronics Dictionary

943 pp. Royal 8vo. McGraw Hill Publishing Co. Ltd., New York, London. 1958. Price 62s.

THIS dictionary translates about 22 000 Russian terms and abbreviations into English and 25 000 terms from English into Russian. It was compiled from all modern sources available in the U.S.A., U.S.S.R. and U.K.—including also many terms obtained from Soviet factories, research institutions, and individual scientists. The dictionary was prepared by the Department of the Army, Washington, D.C.

An Introduction to Electronic Theories of Organic Chemistry

By G. I. Brown. 209 pp. 40 figs. Demy 8vo. Longmans, Green & Co. Ltd., New York, Toronto, London. 1958. Price 15s.

THE author sets forward in a simple, concise, and mainly qualitative form a detailed account of the electronic theories on which organic chemistry is based. His aim has been to introduce the theories to the student by explaining the theoretical ideas involved and by showing how they can be applied to give a clearer understanding of organic chemistry. Mathematical considerations and experimental details have been reduced to a minimum. The work is intended for use by advanced sixth-form pupils, by university students, and by chemists who may be interested in modern developments.

A First Course in Television

By 'Decibel'. 149 pp. 50 figs. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1958. Price 15s.

THIS book is by the author of *A First Course in Wireless*. It has been written on similar lines for the reader who possesses a general knowledge of the fundamental principles of wireless and wishes to acquire a reasonable understanding of the principles of television. Mathematics beyond simple arithmetic has been avoided.

Principles and Applications of Random Noise Theory

By J. S. Bendat. 431 pp. 106 figs. Demy 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1958. Price 88s.

BOOKS on the subject of noise range from the one extreme of dealing mainly with the electrical fluctuations produced in circuits by the active and passive elements such as valves, resistors, etc. to the other of concern with the mathematical basis of random processes. Dr. Bendat's book, as might be expected from one closely associated with systems analysis and engineering, tends towards the latter class. The circuit engineer who takes up the book will find a most clear exposition of the basic ideas of correlation functions and power spectra and their advanced uses in many fields of communication. The valve engineer, however, will seek in vain for discussion or even mention of the noise factor of amplifiers, or of noise in electron beams.

Starting with a first chapter on random processes and correlation functions, the book passes by way of a consideration of Fourier series and integrals, to the subject of power spectra and their relation to correlation functions. In Chapter III probability distributions are introduced, and a preliminary discussion is given of the zero crossing problem, which is extended to much greater detail in the last part of the book. Linear prediction and optimum filtering, as treated by the author in Chapter IV, includes Wiener's work as special cases. A time domain procedure is introduced, which again is treated in much greater detail later in the book. Following the demonstration in Chapter V that most of the naturally occurring random processes have autocorrelation functions of exponential or exponential-cosine form, the remainder of the book is concerned with manipulation of these functions. The use of analogue computer techniques for handling the functions is dealt with in some detail, and chapters are devoted to the statistical errors which can occur in autocorrelation measurements of video noise and envelope detected noise.

It is clear that in writing this book the reader has been kept well in mind. Almost every chapter has not only an introduction to excite interest in the physical processes covered by the following mathematical procedure, but also a summary to indicate the field covered and to give its relation to subsequent material. Indeed, the author is so anxious to ensure that the book should be used, that a separate five page summary of the whole contents is included.

It is a general criticism of books on noise that there seems to be little or no correlation between them on the symbols employed for various functions. In the present volume the author's correct insistence on the difference between time average and ensemble average leads to a more complicated symbolism than usual. This is more than compensated by the general attention to the physical explanation of the various mathematical steps.

The book is well produced and free from errors, and can be thoroughly recommended to communication and systems engineers.

W. H. ALDOUS

Automation in Production Engineering

By J. A. Oates. 326 pp. 300 figs. Demy 8vo. Geo. Newnes Ltd. 1958. Price 45s.

IN this book, automation in the machine shop has been divided into two main sections; first, the use of machines for carrying out automatically a series of operations on a component; and, secondly, single machines capable of automatically machining a component from data supplied by punched cards and tape, and magnetic tape.

The author deals progressively with pneumatic, hydraulic and electrical equipment for loading, clamping and control. Then follow chapters on automatic loading and unloading installations; copying lathes; gear hobbing; and the automatic gauging and correction of machines. The next sections deal with unit type machines and the various types of transfer machines, including examples of typical installations. Finally, there is an extensive section dealing with machines controlled by punched cards or tapes, or magnetic tapes.

THE PHYSICAL SOCIETY EXHIBITION

A description, compiled from information supplied by the manufacturers, of a small selection of the electronic equipment, exhibited at the 43rd Physical Society Exhibition, held in London from 19-22 January.

(Voir page 115 pour la traduction française: Deutsche Ubersetyung auf Seite 122)

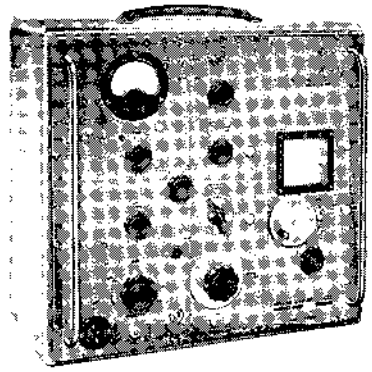
ADVANCE COMPONENTS LTD

Roebuck Road, Hainault, Ilford, Essex.

F.M.-A.M. SIGNAL GENERATOR

(Illustrated below)

The signal generator type 60 provides an accurately controlled signal in the range of 4 to 230Mc/s, frequency and/or amplitude modulated. The carrier frequency calibration may be checked against an internal 5Mc/s crystal calibrator. The attenuator consists of a continuously variable inductive potentiometer followed by a four-step decade attenuator, which together provide a range of 100dB. The carrier output level is monitored by means of a crystal voltmeter.



This instrument employs a Colpitts oscillator and the frequency modulation is effected by a variable reactance connected across the tuned circuit. The frequency deviation is a direct function of the modulating voltage controlling the variable reactance. As different amplitudes are necessary at different frequencies to produce the same frequency deviation, a deviation equalizer is included in the circuit, bringing the modulating voltage to a value dependent on the position of the main tuning capacitor and frequency range used. The output of the carrier oscillator is fed to an r.f. amplifier and its tuned circuit is ganged with that of the amplifier. Amplitude modulation is obtained by applying the modulating voltage to the input of the amplifier. The carrier and modulating voltage are brought to different electrodes of the amplifier valve to minimize frequency variations.

The output level setting is controlled by an inductive potentiometer loosely coupled to the amplified output so that the setting has little effect on the signal frequency. By these means the signal frequency is made virtually independent of both the output setting and amplitude modulation depth. The incidental frequency modulation caused by amplitude modulation is very small. An oscillator working at 1kc/s provides a source of internal modulation and is switched to effect amplitude or frequency modulation according to requirements. If simultaneously both types of modulation are to be applied to the carrier wave an external source of modulation voltage is necessary.

EE 6751 *for further details*

AIRMEC LTD

High Wycombe, Buckinghamshire.

PROXIMITY DETECTOR

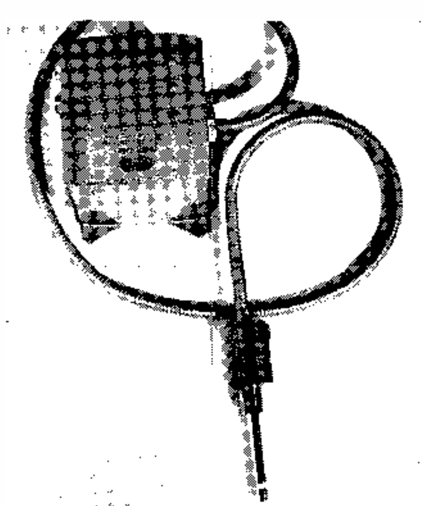
(Illustrated below)

The proximity detector type N266 is a capacitive switch which operates a relay system when any substance is brought within a pre-determined distance from the sensing probe. It has a wide range of applications and is suitable for insertion in liquids, powders etc. Printed circuits using transistors are employed in the amplifier, which may be operated from mains or battery power supplies. The power supply and amplifier units are bolted together and connected by a printed circuit strip and socket to allow for easy interchange of power and battery units as required. The useful life on battery operation is in excess of 200 hours.

The sensing probe may be mounted either on the case of the instrument or on the end of an armoured extension cable which may be up to 25ft (7·5m) long.

Probes of various sizes are available in waterproof covers suitable for use in liquids at temperatures up to 100°C.

The circuit remains quiescent when the capacitance to the probe is constant at a preset value, and operates when the capacitance becomes greater or smaller than the preset value. The sensitivity of the instrument is equivalent to a change in capacitance of 0·02pF, but this may be varied to suit any particular application. The capacitance of the probe at balance is 8pF maximum and may be increased by adding capacitance inside the instrument as required.



EE 6752 *for further details*

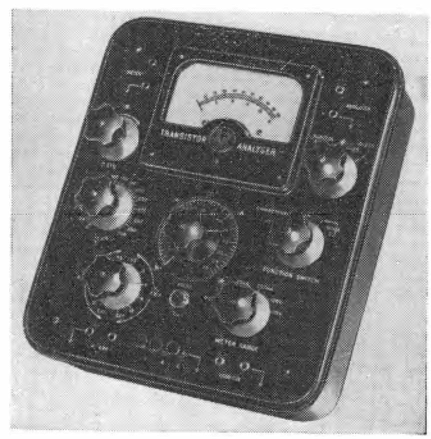
AVO LTD

Avocet House, 92-96 Vauxhall Bridge Road, London, S.W.1.

TRANSISTOR ANALYSER

(Illustrated below)

This is a portable compact, battery operated transistor tester suitable for testing small and medium npn or pnp transistors either in situ or removed from associated apparatus. The instrument



meets the requirements of the U.K. Climatic and Durability Specification K.114.

The instrument is suitable for taking I_b/I_c characteristics with a choice of V_c with the transistor in the grounded emitter configuration. α' can be measured at any predetermined point on the I_b/I_c characteristic and I_{co}' is directly indicated in microamperes.

Provision has been made for the measurement of transistor noise by comparing peak noise over the approximate range 100c/s to 10kc/s, with an equivalent input signal calibrated from the internal 1kc/s oscillator. The instrument can also be used as a multi-range test meter for servicing transistorized apparatus.

EE 6753 *for further details*

BRITISH PHYSICAL LABORATORIES

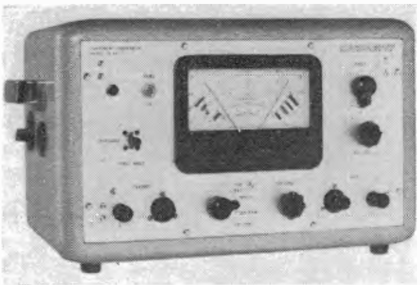
Houseboat Works, Radlett, Hertfordshire.

COMPONENT COMPARATOR

(Illustrated above right)

This inductance, resistance and capacitance component comparator type CZ. 457/1 is operated at 100kc/s and is a development of the model 457 which operates at 1kc/s.

This instrument has been specifically



designed for the testing of capacitors, resistors and inductors under mass production conditions. Without additional mechanical aids test speeds of 1 500 to 2 000 components per hour are readily achieved.

The equipment compares the component under test with a standard unit and displays the percentage deviation directly on a 5in (127mm) scale meter. Plus or minus tolerances from 0·5 per cent to 25 per cent are clearly marked. In addition, capacitors and inductors can be checked for phase-angle, the deviation as compared with the standard component being directly indicated on the scale of the meter.

A special feature of the instrument consists of the automatic overload protection of the meter. Both under open-circuit and short-circuit conditions of the component under test, the meter deflexion does not exceed by more than 20 per cent the maximum value of the meter range.

The equipment is completely self-contained and controls are incorporated for checking the accuracy of the calibration, no extra scales or correction data being required.

It will compare capacitors within the range 20pF to 0·1μF and inductors from 20μH to 100μH, the accuracy being to 3 per cent of the indicated value.

EE 6754 *for further details*

THE BRITISH THOMSON-HOUSTON CO. LTD

Rugby, Warwickshire.

SUPERHETERODYNE CENTIMETRE SPECTROMETER

This spectrometer is intended primarily for the analysis of the spectra of pulsed oscillators by displaying on a cathode-ray tube the frequency distribution of the r.f. power. The range of pulse width is 0·1μsec to 5μsec. Also suitable for the measurement of frequency pulling and the measurement of standing wave ratios of signals which are very low in amplitude, the spectrometer consists of a narrow-band superheterodyne receiver whose local oscillator is varied in frequency by means of a sawtooth waveform applied to the reflector. This waveform is also applied to the X plates of the c.r.t.

The input signal is applied to the mixer through an r.f. attenuator, and the local oscillator is modulated by means of the sawtooth waveform and is also fed to the mixer. The resulting intermediate frequency is amplified, detected and the output applied to the Y plates of the c.r.t. through a video amplifier. The i.f. amplifier operates at a frequency of 48Mc/s and has a bandwidth of 50c/s. A crystal controlled oscillator is included to provide a frequency calibration of the c.r.t. trace at 1Mc/s intervals. The centimetre components are mounted on a detachable portion of the front panel. Interchangeable panels make the spectrometer suitable for the sizes of centimetre wavebands which are most used.

EE 6755 *for further details*

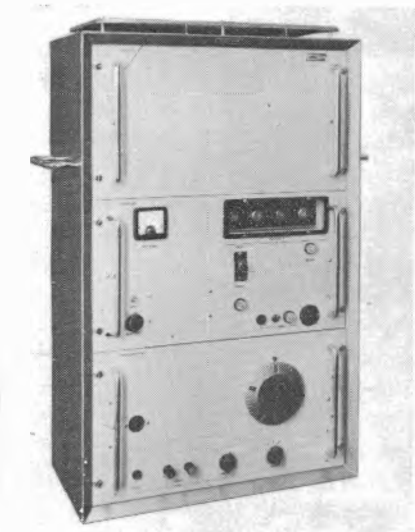
CAWKELL RESEARCH & ELECTRONICS LTD

Scotts Road, Southall, Middlesex.

DAMPING CONSTANT METER

(Illustrated below)

This instrument provides a sensitive, non-destructive means of obtaining information about the internal state of any engineering material. The basic principle is to accurately measure the decay time, damping constant, logarithmic decrement or 'Q' of mechanical vibration in the material under test in order to obtain information about the internal state, flaws, non-homogeneity, chemical composition etc. The specimen is exited into mechanical vibration at its natural frequency or at some harmonically related mode. When steady oscillations are established the exitation is cut off and the decay in oscillation is 'observed' by a suitable transducer and counted during a predetermined time interval. The count is presented as a four digit number on Dekatrons. In the latest version of this instrument measurements can be made between 50c/s and 50kc/s.



EE 6756 *for further details*

DAWE INSTRUMENTS LTD

99-101 Uxbridge Road, Ealing, London, W.5.

PORTABLE AUDIO OSCILLATOR

(Illustrated above right)

The type 421 audio oscillator may be battery operated and uses transistors throughout. The frequency is continuously variable over the range 20c/s to 20kc/s which is covered in three decade steps.

These features make it particularly suitable for field work; for example, the calibration of sound level meters. Where continuous bench use is required, a mains power unit is incorporated.

A novel push-pull Wien bridge oscillator is employed to overcome the difficulties usually experienced with the low impedances encountered in transistor circuits. Each push-pull section of the oscillator employs two grounded collector stages and one stabilized grounded emitter stage. Constancy of amplitude is maintained by the use of a newly developed very low power thermistor. The signal from the oscillator is fed to two push-pull transistor output pairs, each of the 'emitter-squared follower' type giving an output of up to 2·5V into 600Ω. An output voltmeter is incorporated.

The accuracy is ±2 per cent and the distortion less than 1 per cent. The output is variable up to 2·5V into 600Ω.

EE 6757 *for further details*

THE DISTILLERS CO. LTD

172/3 Tottenham Court Road, London. W.1.

PANORAMIC TEMPERATURE DISPLAY UNIT

This instrument displays the outputs of up to 24 thermocouples in the form of a profile on a cathode-ray tube. The 24 thermocouples are scanned successively, vertical displacement of the spot representing the temperature each time a thermocouple is switched in. The display thus consists of a number of vertical lines equally spaced apart in a horizontal direction. The envelope of this pattern represents the temperature distribution at the points at which the thermocouples are located. Scanning of 24 thermocouples takes 1·5sec.

There is provision for backing-off all inputs by the same amount so that a small variation about a high mean temperature may be displayed with a high sensitivity. This feature also permits individual temperatures to be measured. In addition each input may be backed-off individually so that when a desired temperature distribution is achieved all inputs are effectively zero and a straight line display results. The sensitivity may then be increased so that any deviation from the desired condition may be easily detected. If required an alarm system may be operated when the deviation exceeds preset limits, and a set of indicator lamps shows which points have given rise to alarm conditions.

Although this instrument is designed primarily for use with thermocouples, any type of transducer which provides an electrical output signal may be used.

EE 6758 *for further details*

EDWARDS HIGH VACUUM LTD
Manor Royal, Crawley, Sussex.

C.R.T. Aluminizing Plant

The attainment of a uniform aluminium film on the back of the screens of 110° cathode-ray tubes presents a difficult problem due to the short length of the envelope and small neck diameter.

A special plant has been designed to improve the uniformity of an aluminium film deposited from a single tungsten filament. The filament is mounted on an arm which is hinged for easy insertion in the neck of the tube. The arm is held in position by a spring, and is rotated inside the tube to give a very uniform deposit on the inner face. The radial distance and position of the source are adjustable to accommodate several tube sizes.

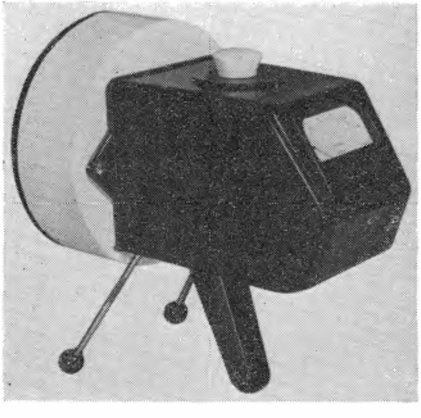
EE 6759 *for further details*

EKCO ELECTRONICS LTD
Southend-on-Sea, Essex.

Radiation Monitors

(Illustrated below)

The Ekco type N596 beta-gamma survey meter is a new portable radiation monitor designed to meet the requirements of the impending Factories Act for beta and gamma radiation. Ranges of 0 to 3, 0 to 30 and 0 to 300mRad/h are provided together with an integration facility up to 30mRad. The air-equivalent ionization chamber can be changed readily for an alternative chamber of one tenth the size for very localized dose rate measurement. The instrument is also suitable for measurement of soft X-rays and its virtual immunity to r.f. interference makes it particularly suitable in connexion with high-voltage valves, cathode-ray tubes etc.



The type N645 universal low activity monitor is a lightweight, portable equipment for tracing very small quantities of any radioactive source. It comprises a transistorized, battery-operated ratemeter which is carried on a shoulder strap and connected to a scintillation counter probe which is held in the hand. Interchangeable phosphors cater for alpha and beta particles, gamma rays, slow and fast neutrons. The counting ranges provided on the ratemeter are 0 to 10, 0 to 100 and 0 to 1000 counts/sec.

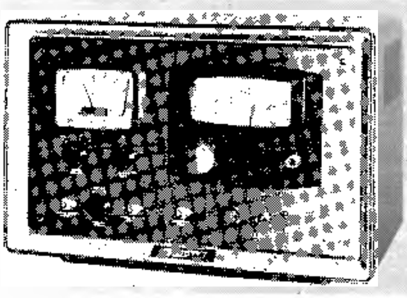
EE 6760 *for further details*

ELLIOTT BROS (LONDON) LTD
Century Works, Lewisham, London, S.E.13.

U.H.F. Test Oscillator

(Illustrated below)

The u.h.f. test oscillator type B.689 comprises a coaxial oscillator cavity designed for operation from 7 to 12kMc/s using a CV2346 klystron oscillator, a stabilized power supply and a



square-wave modulator for 400c/s to 4Mc/s. The r.f. energy is extracted through a coaxial plug and adequate power output is obtained over a wide band for all normal microwave bench measurements. A direct reading frequency dial is incorporated which gives unambiguous indications of the operating frequency. Provision is also made for external modulations.

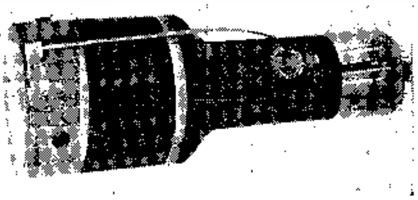
EE 6761 *for further details*

ENGLISH ELECTRIC VALVE CO. LTD
Chelmsford, Essex.

Daylight Viewing Storage Tubes

(Illustrated above right)

The E702 has electrostatic deflexion and is intended primarily for use in oscilloscopes where storage of the image is required. Because the image persists without deterioration for up to two minutes, the tube is applicable equally to the display of slowly varying waveforms or fast transients. The maximum writing speed is approximately 8mm/μsec. The writing gun is operated at 2·2kV, which makes the deflecting power small. The screen is operated at 10kV, giving the very high brightness which is a characteristic of these tubes and makes the use of hoods unnecessary, even in high ambient light.



The E703 is also intended for oscilloscopes, and has two writing guns, which allows simultaneous recording of two independent waveforms.

The E701 uses magnetic deflexion, and is designed for radar installations, where its high brightness can be used to advantage in displays mounted in a cockpit or on the bridge. All these tubes have useful screen diameter of at least 4in (102mm).

EE 6762 *for further details*

E.M.I. ELECTRONICS LTD
Hayes, Middlesex.

Digital Drum Stores

(Illustrated below)

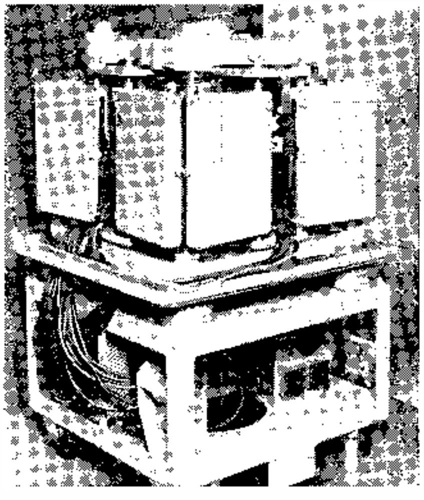
These stores, which have been produced to meet the needs of the digital computer engineer, incorporate a unique design of highly sensitive magnetic head block. The printed circuit head switching units, which form an integral part of the drum assembly, are easily accessible.

The storage capacity is 8 000 or 16 000 words of 36 digit length, on 128 + 8 spare tracks or 256 + 16 spare tracks. There are five clock tracks and spares. Interleaving blocks are of 16 read-write heads and one spare.

All heads are suitable for reading or writing. The five clock heads are suitable for simultaneous reading and writing. There is high efficiency screening from external interference on all heads.

The drums are 8in (203mm) in diameter. There are approximately 7in (178mm) length of drum surface on the 8 000 word drum 12·75in (324mm) on the 16 000 word drum. It is claimed that at least 20 000 hours' continuous running at 3 000 rev/min under normal conditions can be expected.

EE 6763 *for further details*



FERRANTI LTD
Hollinwood, Lancashire.

'WATTLESS' TRANSISTOR REGULATORS

These regulators employ one or more transistors in parallel acting as a switch in a circuit with a diode and choke-capacitor filter. In this circuit the voltage applied to the input of the choke capacitor filter is alternatively the supply voltage or zero, as the transistor is switched on and off. The d.c. output from the filter is equal to the mean value of the input waveform. A control loop compares the output voltage with a reference potential such as a Zener diode and automatically adjusts the mark-to-space ratio of the signal to the switch so as to stabilize the output for variations in input voltage and load. Efficiencies of the order of 90 per cent are obtained, and in a typical unit employing a simple circuit input voltage variation is reduced one hundred times.

EE 6764 *for further details*

THE GENERAL ELECTRIC CO. LTD
Magnet House, Kingsway, London, W.C.2.

INDUCTION DIGITIZER

This is an entirely new type of transducer for use with digital data handling and computer systems. It gives a numerical representation of the position of a shaft or slide in the form of coded electrical signals without using slip-rings, commutators or photocells.

It is basically a transformer with a single exciting winding and a number of secondary windings corresponding to the number of digits. The position of a moving yoke decides the polarity of coupling between the exciting winding and the various digit windings, which are arranged to conform with the requirements of the code. Pulse or sine wave excitation can be used.

Shaft digitizers can be operated in coarse-fine combinations; an additional winding preventing ambiguity between the digitizers. Instruments for 5-digit read-out in a size 11 standard synchro housing and for 7-digit read-out in a size 23 housing have been made.

EE 6765 *for further details*

GRIFFIN & GEORGE LTD
Ealing Road, Alperton, Wembley, Middlesex.

RESISTANCE-REACTANCE BRIDGE

The Griffin Panmetron bridge is a mains-operated instrument which serves not only as an a.c. bridge for comparing resistance and reactance at 50c/s using external standards, but also as a high gain detector amplifier for use with external bridge circuits.

For energizing the latter a 2V and 4V source of a.c. is incorporated. The out-of-balance current is applied to the input side of a high gain amplifier using an EM84 as balance indicator. With the control switch in the highest position a good indication of balance is provided with an input of only 0·1μV.

EE 6766 *for further details*

ISOTOPE DEVELOPMENTS LTD
Beenham Grange, Aldermaston Wharf, Nr. Reading, Berkshire.

SCALER

(Illustrated below)

The principal new addition to the I.D.L. range is the scaler type 1700. This incorporates a complete and automatic counting apparatus in a single table-top unit. It requires only the addition of a radiation detector to provide for any type of radioactivity measurement.

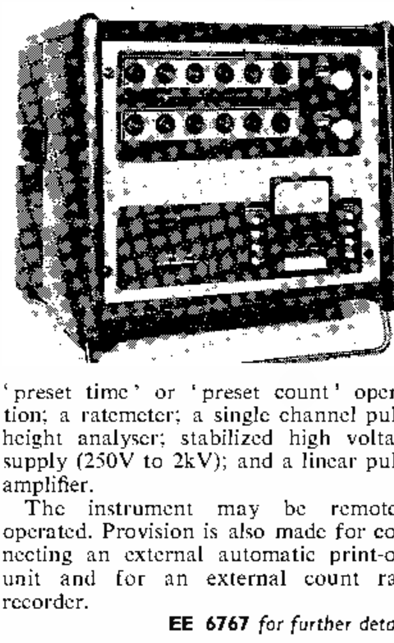
In the design of the scaler type 1700, a new approach has been made to the problem of servicing. With the recommended spares pack, out of service time can be greatly reduced and servicing, even by unskilled personnel, can be quickly carried out in any part of the world.

Apart from the power pack, the scaler type 1700 incorporates a simple printed circuit system with individual plug-in units. Thus any one part of the circuit can be replaced instantly by a spare should a fault occur.

Complete access to the interior is also possible in a few seconds by means of quick release dust covers which are provided at the top and bottom of the instrument.

The power pack is mounted vertically and is separate from all main counting circuits, while its chassis is hinged to give quick and easy access.

Incorporated in the scaler are a fast scaler; an electronic timer; facilities for



'preset time' or 'preset count' operation; a ratemeter; a single channel pulse height analyser; stabilized high voltage supply (250V to 2kV); and a linear pulse amplifier.

The instrument may be remotely operated. Provision is also made for connecting an external automatic print-out unit and for an external count rate recorder.

EE 6767 *for further details*

KELVIN & HUGHES LTD
New North Road, Barkingside, Essex.

LOW FRICTION SYNCHRO SYSTEM

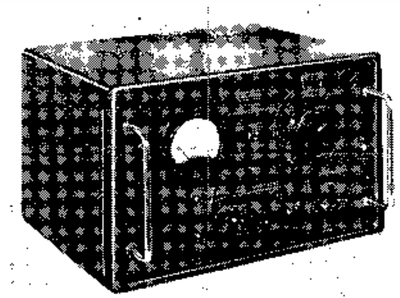
The Syncrotel is an angular position transmitting element of the synchro family, but its extremely low friction and inertia enable it to be employed in cases where the available driving power is very small, e.g. in capsule-operated mechanisms.

The stator is a conventional three-phase synchro type, but instead of the usual rotor a fixed single-phase winding on a bobbin is used, with its axis coincident with the spindle. The rotating portion consists of an oblique section of a hollow cylinder of aluminium, and its shaft is supported on jewel bearings.

Syncrotels may be used directly in back-to-back pairs in certain applications, but are usually employed in servo systems with an ordinary synchro receiving element.

The rotor weight is 0·76g and the moment of inertia 0·11g-cm^2.

EE 6768 *for further details*



MUIRHEAD & CO. LTD
Beckenham, Kent.

H.F. ANALYSER-OSCILLATOR

(Illustrated above)

Having a frequency range of 200c/s to 650kc/s, this new, dual purpose instrument extends the frequency coverage of Muirhead oscillators and analysers into the carrier telephone communication band.

By operating one switch, the instrument can be used either as a sensitive, constant gain frequency analyser (with a Q-factor of not less than 70, an octave discrimination of 40dB and a minimum indication of 50μV) or an oscillator (with an output level of 2·5V into 600Ω). In either of its roles, the frequency accuracy is 2 per cent.

EE 6769 *for further details*

MULLARD EQUIPMENT LTD
Mullard House, Torrington Place, London, W.C.1

HIGH SPEED ANALOGUE-DIGITAL CONVERTOR

This equipment is designed for converting at high speed an analogue input voltage into a binary output of 10 'bits', available in parallel or sequentially and in true binary or binary-coded decimal form. The sampling period is less than 1msec, samples being taken on receipt of input pulses, which may be at any frequency up to 1 000 a second. The analogue input voltage range is 0 to 10V (positive). For sequential output the maximum output pulse rate is approximately 12kc/s, or it may be locked to any external source of pulses provided the p.r.f. does not exceed 12kc/s. For parallel output, the pulse rate will be coincident with the succeeding input sampling pulse. The convertor is completely transistorized, uses printed wiring boards and has a very high degree of reliability. It is particularly suitable for data handling processes where large amounts of data have to be dealt with in a very short time, as, for example, in making dynamic measurements on rocket motors or in testing aircraft structures in supersonic wind tunnels.

EE 6770 *for further details*

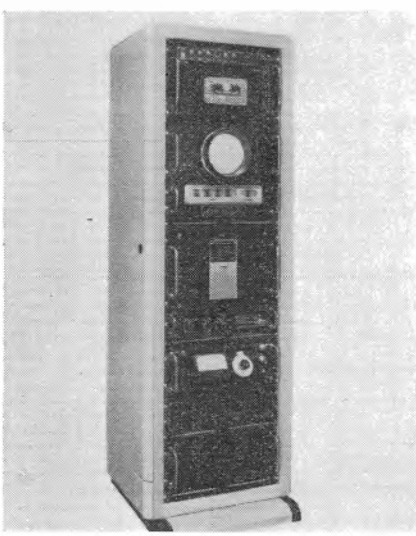
MULTI-CHANNEL PULSE SPECTROMETER

(Illustrated below)

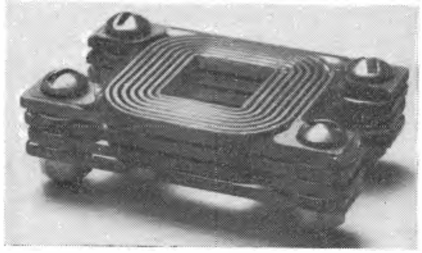
The L.332 is a 100-channel pulse spectrometer providing fast, accurate analyses of the energy spectra of nuclear radiations. Pulses from a proportional detector are counted and sorted, according to their relative amplitudes, into any of the 100 channels, where they are stored in a magnetic matrix memory system. Since the amplitudes of the pulses produced by the proportional detector are directly related to the energy of the incident radiation, the information stored in the memory provides a histogram showing the relative abundance of radiation at each of 100 energy levels within the spectrum being investigated. This information is displayed on a 6in (152mm) c.r.t. as a continuous, linear, analogue display of the spectral curve, the display being available both during and after the counting period. A direct read-out of the counts registered in a particular channel is also available, from 'in-line' neon numerical indicators. The read-out is non-destructive. The matrix memory contains 100 × 16 ferroxcube cores, giving each channel a storage capacity of 65 536 counts. Maximum counting rate is limited only by the percentage of lost counts due to the dead-time, which is constant at 250μsec for accepted pulses.

The basic input range of the spectrometer is 0·5 to 10V, each channel having a width of approximately 0·1V, and up to 30V back-bias may be applied giving a total coverage of equivalent to 400 channels. Channel width uniformity is better than 1 per cent and the stability of channel positions is better than 10mV over the basic input range. Random jitter is less than 3mV.

The L.332 has high reliability and makes extensive use of transistors. It is based on a design of the A.E.R.E., Harwell.



EE 6771 *for further details*



G. V. PLANER LTD

Windmill Road, Sunbury-on-Thames, Middlesex.

COPPER-ON-GLASS PRINTED CIRCUITS

(Illustrated above)

In this type of printed electronic circuit copper patterns are firmly bonded to a glass base. The advantages of this system are suitability for high temperature operation and inertness due to the absence of organic substrate or bonding constituents. Examples shown included inductors made up from ultrasonically cut glass substrates, each carrying a copper pattern on either side. The plates may be stacked to form multiple layer inductors, interconnexions being made across the edges of the glass. An inductor made by this method is shown in the accompanying illustration at approximately ×2 magnification.

EE 6772 *for further details*

W. G. PYE & CO. LTD

Granta Works, Cambridge.

PRECISION DECADE POTENTIOMETER

This refined and highly accurate potentiometer has been designed to offer high performance without maintenance and to present information in the most convenient form. Decade switching is used throughout and slidewires are eliminated. The first decade has been combined with the range-change switch thus making the instrument direct reading on all ranges. A total voltage range extending from 1·99999V down to $-0{\cdot}1\mu$V is available. An accurate fine decade current control is included giving variations in current in steps of 1 part in 10 down to 1 part in 10^5. Switching facilities are provided to enable any one of five external circuits to be selected and to transfer the standard cell from the correction circuits to the potential divider for calibration purposes. A system reversing switch is fitted to allow account to be taken of any spurious e.m.f.'s (less than 0·1μV) which may be present in the galvanometer or output circuit.

The three ranges are as follows:

-10μV to 1·99999V in steps of 10μV

-1μV to 0·199999V in steps of 1μV

$-0{\cdot}1\mu$V to 0·0199999V in steps of 0·1μV

The accuracy of potential division is to ±0·0001 per cent and this accuracy will be obtained in comparing two e.m.f.'s. The absolute accuracy of measurement is to ±0·002 per cent.

The standard cell correction covers a temperature range of from 10°C to 37°C. The smallest step is 10μV equivalent to 0·25°C.

EE 6773 *for further details*

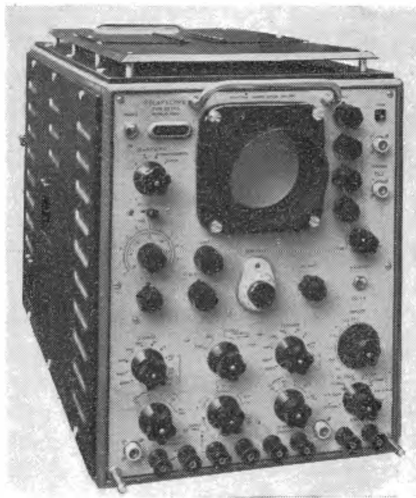
THE SOLARTRON ELECTRONIC GROUP LTD

Solartron Works, Thames Ditton, Surrey.

DOUBLE BEAM OSCILLOSCOPE

(Illustrated below)

The CD711S. 2 is a further development of the CD711 and the CD711S. Both ruggedness and reliability have been improved, in order to withstand Service conditions. The instrument has received Service approval and been designated CT414. In addition, the internal layout of the oscilloscope has been revised, in order to give improved accessibility.



The instrument employs a beam switching technique, with 150kc/s chopping facilities provided for low time-base sweep speeds. Both vertical deflexion channels have identical characteristics, with a bandwidth of 7Mc/s and a maximum sensitivity of 3mV/cm.

A calibrated 'Y' shift control which may be switched to either channel is used to give amplitude calibration to an accuracy of ± 5 per cent.

The time-base has a range of sweep speeds from 3cm/μsec to 0·3cm/sec.

The time-base calibration can be standardized against an internal crystal oscillator and frequency divider system.

EE 6774 *for further details*

PROCESS RESPONSE ANALYSER

(Illustrated above right)

The process response analyser type JY743 provides a number of special facilities for very low frequency dynamic analysis.

The frequency range is from 0·0001c/s to 100c/s with a frequency accuracy of ±1·5 per cent and an amplitude stability better than 1 per cent over the entire frequency range. Square wave and triangular wave outputs are provided in addition to four-phase and three-phase sine wave outputs. The d.c. content of the sine wave outputs is less than 0·1 per cent of the peak-to-peak a.c. output and the harmonic distortion is less than 0·5 per cent.

Relays are provided for priming, starting, holding and releasing the oscillator action at frequencies below 10c/s. These relays can be operated manually, or externally—from e.g. the 'compute', 'hold' and 'reset' switches of a computer or simulator. Provision is also made for digital monitoring of the instantaneous oscillator voltage.

In the display unit two 5in (127mm) meters (reference and quadrature) are provided which indicate the amplitude and sign of the real and imaginary components. Digital read-out facilities are also provided for automatic plotting of frequency and amplitude. The maximum sensitivity is 3mV for full scale deflexion and the harmonic rejection is greater than 40dB.

The special technique employed in the display unit will give steady indications, irrespective of frequency of fundamental signals. The use of integration for any given measurement is, therefore, determined by the noise content. Smoothing facilities are provided at time-constants of 0·1, 0·3, 1, 3 and 10sec and single cycle integration is provided for frequencies between 1c/s and 1 cycle per hour. There is also an arrangement for manual control of integration period.

EE 6775 *for further details*

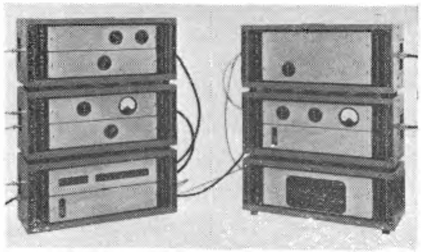
STANDARD TELEPHONES AND CABLES LTD

Connaught House, Aldwych, London, W.C.2.

'White Noise' Testing Equipment

(Illustrated above right)

This equipment is designed to measure basic noise and intermodulation noise on multi-channel telephone systems having a bandwidth up to 4Mc/s (16 supergroups), it generates a band of random noise having an approximately uniform energy distribution over the desired frequency band which can correspond to 2, 4, 10 or 16 supergroups as desired. The noise is passed through a set of filters to determine the width of the noise band and to provide a narrow 'quiet' band within it, simulating a loaded system with one quiet channel.



This signal can be applied to a multi-channel telephone system (in out-of-traffic conditions) at a power level simulating actual traffic loading, permitting measurements of crosstalk and noise to be made in the narrow stop-band at the far end of the system. Noise-power ratios up to 70dB can be measured, and back-to-back operation of the generator and receiver will produce noise-power ratio figures of 75dB or more for each band of frequencies received. The minimum absolute white noise density which can be measured is −118dBm/kc/s, corresponding to −114·5dBm/channel when psophometrically weighted with a C.C.I.F. (1951) telephone weighting network. The equipment comprises a generator consisting of three portable units each measuring 22in × 9⅞in × 11¾in (559mm × 251mm × 298mm), and receiver consisting of two units of this size plus one unit measuring 22in × 8⅛in × 11¾in (559mm × 206mm × 298mm). The generator and the receiver each incorporates its own power units operating from a.c. mains.

EE 6776 *for further details*

SUNVIC CONTROLS LTD

Crown House, Aldwych, London, W.C.2.

Pulse Height Analyser

(Illustrated below)

The new Sunvic pulse height analyser is available in either console or rack mounted form.

Largely transistorized, it uses a ferrite core storage providing 100 channels each with a maximum capacity of 65 535. Counts can be displayed as a c.r.t. analogue trace in which horizontal deflexion represents channel number and vertical deflexion is proportional to the count. Dekatrons are used to indicate, in decimal form, the total of pulses in any channel selected either manually or automatically in sequence. Totals can also be printed automatically.

EE 6777 *for further details*

VENNER ELECTRONICS LTD

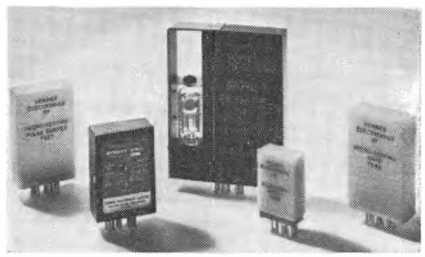
Kingston By-Pass, New Malden, Surrey.

Packaged Transistor Stages

(Illustrated below)

The latest additions to the Venner range of plug-in stages include the crystal oscillator type TS25, which will function at frequencies between 100kc/s and 1Mc/s (depending on the crystal employed). This unit can be used to give either a sine wave or pulse waveform output. It is the same size as the 10kc/s crystal oscillator type TS5. A further addition to the range of crystal oscillators is the TS30 which is a low frequency oscillator for use with XY flexure crystals. This block will cover the range from 5kc/s to 100kc/s.

Another new item is the 1Mc/s binary unit which is available in two forms, the TS2A/HF whch is suitable for divide-down chains where a division ratio other than in multiples of two is required, and the TS2B/HF which is particularly designed for use with the microsecond gate type TS36. This gate can be opened and closed by varying



the d.c. level of the control pin from −9V to −4V with a 10V supply, these levels being conveniently provided by the binary unit type TS2B/HF. Two such binary units connected to the microsecond gate can be arranged to provide an interlock action so that the first control pulse recived opens the gate, the second closes it and ensuing pulses have no further action.

The microsecond pulse shaper type TS27 is designed to provide a pulse suitable for triggering binary and decade units from a wide range of waveforms at frequencies up to 1Mc/s. The output consists of positive-going pulses of approximately 8V amplitude, 0·4μsec width, and having a rise-time less than 0·2μsec at 1Mc/s. It may also, of course, be used to drive the microsecond gate.

Also available is the 1Mc/s decade unit type TS10/HF which is similar in operation to the standard decade unit type TS10/5 in that it drives a meter calibrated 0 to 9 to give indication of the count which has been achieved. It will operate at any frequency up to 1Mc/s when driven from, say, the microsecond pulse shaper type TS27.

The electronic reset unit type TS32 is intended for use with any binary or decade stage and is capable of resetting these within 0·4μsec when triggered by a negative-going pulse. No contacts are employed, the operation being entirely electronic.

EE 6778 *for further details*

Meetings this Month

BRITISH INSTITUTION OF RADIO ENGINEERS

All London meetings, unless otherwise stated, will be held at The Institution, 9 Bedford Square, London, W.C.1, commencing at 6.30 p.m.
Date: 13 February
Lecture: Some Instrumentation Problems in Medical Electronics with particular reference to Electromyography.
By: Peter Styles.
Date: 25 February.
Lecture: Patents and the Radio Engineer.
By: E. D. Swann.

South Midlands Section

Date: 27 February. Time: 7 p.m.
Held at: North Gloucestershire Technical College, Cheltenham.
Lecture: Micro-miniaturization.
By: G. W. A. Dummer.
Date: 28 February. Time: 7 p.m.
Held at: The Winter Gardens, Malvern.
Lecture: Industrial and Underwater Television.
By: B. V. Somes-Charlton.

West Midlands Section

Date: 11 February. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire College of Technology, Wulfruna Street, Wolverhampton.
Lecture: Some Aspects of the Control of Nuclear Reactors.
By: L. W. J. Newman.

Merseyside Section

Date: 19 February. Time: 7 p.m.
Held at: The University Club, Liverpool.
Lecture: Electronic Welding Controls.
By: C. R. Bates.

North Western Section

Date: 4 February. Time: 6.30 p.m.
Held at: The Reynolds Hall, College of Technology, Sackville Street, Manchester, 1
Lecture: Recent Astronomical Research using Radio Waves.
By: H. P. Palmer.

North Eastern Section

Date: 11 February. Time: 6 p.m.
Held at: The Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle-on-Tyne.
Lecture: Stereophonic Sound from Records.
By: P. B. Cooper.

South Wales Section

Date: 11 February. Time: 6.30 p.m.
Held at: Glamorgan College of Technology, Treforest, Glamorgan.
Lecture: Industrial Television.
By: E. A. Naef.

Scottish Section

Date: 19 February. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: True Motion Radar.
By: J. H. Beattie.
Date: 20 February. Time: 7 p.m.
Held at: The Department of Natural Philosophy, The University, Drummond Street, Edinburgh.
Repeat of above lecture.

BRITISH SOUND RECORDING ASSOCIATION

Date: 20 February. Time: 7.15 p.m.
Held at: The Royal Society of Arts, John Adam Street, London, W.C.2.
Lecture: The Design of an Electronic Organ for the Home.
By: G. W. Barnes.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Medical Electronics Discussion Group

Date: 6 February. Time: 6 p.m.
Discussion: Problems of Storing Transient Phenomena for Subsequent Analysis.
Opened by: P. Bauwens.

Radio and Telecommunication Section

Date: 18 February.
Lecture: Ultrasonic Iconoscopes.
By: C. N. Smyth and J. Sayers.

Radio and Telecommunication Section

Date: 23 February.
Lectures: A Simple Investigation of the Cross-Modulation Distortion Arising from the Pulling Effect in a Frequency-Modulated Klystron.
By: D. Gjessing.
Theory and Behaviour of Helix Structures for a High-Power Pulsed Travelling-Wave Tube.
By: G. W. Buckley and J. Gunson.
A Multi-Cavity Klystron with Double-Tuned Output Circuit.
By: H. J. Curnow and L. E. S. Mathias.
A Method for the Measurement of Very High Q-Factors of Electromagnetic Resonators.
By: F. H. James.

Cambridge Radio and Telecommunication Group

Date: 17 February. Time: 8 p.m.
Held at: The Cavendish Laboratory, Free School Lane, Cambridge.
Informal evening: Some Recent Developments in X-ray and Electron Microscopy.
Talk by: P. Duncumb.

Mersey and North Wales Centre

Date: 26 February. Time: 6.45 p.m.
Held at: Liverpool Philharmonic Hall.
Faraday Lecture: Automation.
By: H. A. Thomas.

North-Eastern Radio and Measurement Group

Date: 2 February. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

North-Western Centre

Date: 24 February. Time: 7.30 p.m.
Held at: The Free Trade Hall, Manchester.
Faraday Lecture: Automation.
By: H. A. Thomas.

North-Western Radio and Telecommunication Group

Date: 11 February. Time: 6.15 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: Silicon Transistors, their Manufacture and Field of Operation.
By: J. T. Kendall.

South Midland Radio and Measurement Group

Date: 10 February. Time: 6.30 p.m.
Held at: The Technical College, Coventry.
Lecture: Eurovision.
By: J. Treeby Dickinson.
Date: 23 February. (Time and place as above).
Lecture: Computing Principles and Applications.
By: A. St. Johnston.

Rugby Sub-Centre

Date: 25 February. Time: 6.30 p.m.
Held at: Rugby College of Technology and Arts.
Lecture: Storage and Manipulation of Information in the Brain.
By: R. L. Beurle.

Southern Centre

Date: 5 February. Time: 6.30 p.m.
Held at: Brighton Technical College. (Joint meeting with the Southern Association of the Institution of Civil Engineers.)
Lecture: Jodrell Bank Radio Telescope.
By: H. C. Husband.
Date: 9 February. Time: 6.30 p.m.
Held at: The Guildhall, Southampton.
Faraday Lecture: Automation.
By: H. A. Thomas.
Date: 20 February. Time: 6.30 p.m.
Held at: The S.E.B. Showrooms, Newport, Isle of Wight.
Lecture: Installation of a BBC Television Transmission Station.
By: E. P. Metcalf.
Date: 27 February. Time: 6.30 p.m.
Held at: South Dorset Technical College, Weymouth.
Lecture: Survey of Recent Developments in Marine Radar.
By: A. L. P. Milwright.

West Wales (Swansea) Sub-Centre

Date: 12 February. Time: 6 p.m.
Held at: The Conference Room, S.W.E.B. Showrooms, The Kingsway, Swansea.
Lecture: Semiconductors and their Application.
By: W. Fishwick.

THE INSTITUTION OF ELECTRONICS

Date: 27 February. Time: 7 p.m.
Held at: Reynolds Hall, Manchester College of Science and Technology.
Lecture: Electronics in Industry, including Computer Applications.
By: R. S. Evans.

THE PHYSICAL SOCIETY

Optical Group

Date: 5 February.
One-day meeting at R.A.E. Farnborough: Visual Simulation of Flight.
(Further details from The Physical Society, 1 Lowther Gardens, Prince Consort Road, London, S.W.7.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

Date: 11 February. Time: 6 p.m.
Held at: Manson House, Portland Place, London, W.1.
Lecture: A Digital Instrumentation System for use in the Testing of Jet Engines.
By: L. Airey.
Date: 24 February. (Time and place as above).
Symposium on Automatic Weight Control in Industry.

THE TELEVISION SOCIETY

Date: 6 February. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: Master Control Room Techniques.
By: B. Marsden.
Date: 26 February. (Time and place as above).
Lecture: A Colour Signal Encoder for Laboratory Use.
By: S. H. Cohen and P. C. Kidd.

PUBLICATIONS RECEIVED

INDUSTRIAL TELEVISION is the subject of a new folder produced by E.M.I. Electronics Ltd to introduce to business executives the company's closed-circuit television equipment which has proved so successful in various fields of industry. Various applications are described, ranging from visual control of several widely-separated operations, observation of dangerous processes, training and surveillance to such uses as sales conferences, demonstrations in department stores and the transmission of blue-prints, signatures and other data. Emphasis is placed on E.M.I. television's plug-in construction and printed circuit techniques, and the last page is devoted to a specification and details of various accessories. E.M.I. Electronics Ltd, Hayes, Middlesex.

SYMBOLS OF SOME IMPORTANCE introduces a regular series of bulletins offering brief information on the new components and instruments produced by those firms for whom Aveley Electric Ltd are British agents. They have a programme of several hundred information sheets to be produced in the next two years. Aveley Electric Ltd, Ayron Road, Aveley Industrial Estate, South Ockendon, Essex.

THE BRITISH COUNCIL ANNUAL REPORT 1957-58 mentions that since the war the Council has arranged programmes for more than 50 000 professional visitors from overseas studying British developments and achievements. The yearly number, the report states, has more than doubled to nearly 5 000 in 1957-58. Only a fifth were financed by the Council. About another fifth, 1 012 (mainly United Nations Fellows and Colombo Plan Trainees) were financed under intergovernmental schemes, compared with 191 in 1950. About £380 000 a year is administered by the Council for international and overseas organizations. An article by Sir Charles Snow, a member of the Council's Executive Committee, points out that about a million and a half books are borrowed yearly from the Council's libraries for professional students overseas, and most of the Council's expenditure on the printed word goes on supplies of books and periodicals. He stresses the need for far more British books overseas. The Council has expanded in various countries, including Nigeria, Iran, India, Pakistan and Poland, and next year will take over British educational and cultural work in Germany. The British Council, 65 Davies Street, London, W.1.

I.C.I. SILICONE RUBBERS is a booklet which contains a summary of the grades, properties and applications of I.C.I. silicone rubbers. It is intended to show how silicone rubbers are used in various industries, and it is hoped that the information given may suggest new applications or improvements to existing plant, machinery, instruments, etc. Imperial Chemical Industries Ltd, Nobel Division, Silicones Department, Ardeer Factory, Stevenston, Ayrshire.

LIBRARY SERVICES AND TECHNICAL INFORMATION FOR THE RADIO AND ELECTRONICS ENGINEER is the title of a handbook produced by the British Institution of Radio Engineers. It includes a catalogue of the contents of the Institution Library, and also embodies Institution reports on radio and electronic engineering literature and good engineering practice which have been brought completely up to date. Information on international bodies, translating services, post graduate fellowships and awards, and technical films on radio and electronics, provide a most useful section. The British Institution of Radio Engineers, 9 Bedford Square, London, W.C.1. Price 2s. 6d.

(Pages 115 to 128 French and German section)

Le Salon de la PHYSICAL SOCIETY de Londres

Une description, basée sur des renseignements fournis par les fabricants, d'un choix de matériels électroniques exposés au 43ème Salon de la Physical Society, tenu à Londres du 19 au 22 janvier 1959. *(Traduction des pages 108 à 113)*

ADVANCE COMPONENTS LTD

Roebuck Road, Hainault, Ilford, Essex.

GENERATEUR DE SIGNAUX M.F.-M.A.

(Illustration à la page 108)

Le générateur de signaux type 60 fournit un signal commandé avec précision dans la bande de 4 à 230 MHz, avec modulation de fréquence ou d'amplitude. L'étalonnage de la fréquence porteuse peut être vérifié en fonction d'un calibre piézoélectrique interne de 50 MHz. L'atténuateur consiste en un potentiomètre inductif à variation continue, suivi d'un atténuateur de décades à quatre plots qui assurent conjointement une gamme de 100 dB. Le niveau de sortie de la fréquence porteuse est contrôlé au moyen d'un voltmètre à cristal.

Cet instrument utilise un oscillateur Colpitts et la modulation de fréquence s'effectue par une réactance variable reliée à travers le circuit accordé. La déviation de fréquence dépend directement de la tension modulatrice régissant la réactance variable. Vu que l'amplitude doit différer en fonction du changement de fréquence afin de produire le même écart de fréquence, le circuit comprend également un compensateur de déviation, portant la tension modulatrice à une valeur dépendant de la position du condensateur d'accord principal et de la gamme de fréquences utilisée. La sortie de l'oscillateur porteur alimente un amplificateur haute fréquence et son circuit accordé est accouplé à celui de l'amplificateur. La modulation d'amplitude s'obtient en appliquant la tension modulatrice à l'entrée de l'amplificateur. L'onde porteuse et la tension modulatrice sont dirigées vers des électrodes différentes de la lampe de l'amplificateur afin de réduire au minimum les variations de fréquence.

Le réglage du niveau de sortie est contrôlé par un potentiomètre inductif relié de façon lâche à la sortie amplifiée, de manière à ce que le réglage ait peu d'effet sur la fréquence de signal. Par ce moyen, la fréquence de signal est rendue virtuellement indépendante tant du réglage de sortie que de la profondeur de la modulation d'amplitude. La modulation de fréquence fortuite pouvant être causée par la modulation d'amplitude est minime. Un oscillateur fonctionnant à 1 KHz fournit une source de modulation intérieure et peut être branché soit pour la modulation de fréquence soit pour celle d'amplitude, selon les besoins. Au cas où l'on voudrait appliquer simultanément les deux types de modulation à l'onde porteuse, il y aura lieu d'avoir une source extérieure de tension de modulation.

EE 6751 *pour plus amples renseignements*

AIRMEC LTD

High Wycombe, Buckinghamshire.

DETECTEUR DE PROXIMITE

(Illustration à la page 108)

Le Détecteur de proximité type N.266 est un interrupteur capacitif actionnant un système de relais lorsqu'une matière quelconque est placée à une distance prédéterminée de la sonde. Il s'applique à une variété étendue d'usages et se prête, en particulier, à l'introduction dans les liquides, poudres, etc. L'amplificateur, qui fonctionne soit sur secteur soit sur pile, comporte des circuits imprimés à transistors. Les blocs d'alimentation et d'amplification sont boulonnés l'un à l'autre et reliés par une baguette et une prise de circuit imprimé afin de permettre d'échanger facilement les blocs d'alimentation secteur ou pile, selon les besoins. La durée utile du fonctionnement sur pile dépasse 200 heures.

La sonde peut être fixée soit au coffret soit à l'extrémité d'un câble prolongateur armé pouvant atteindre jusqu'à 7,5 m. de long.

Des sondes de différentes dimensions peuvent être fournies sur demande. Etant sous enveloppe imperméable, elles peuvent être utilisées dans des liquides jusqu'à des températures de 100°C.

Le circuit demeure silencieux lorsque la capacité vers la sonde reste constante à sa valeur préréglée. Il s'anime, par contre, lorsque la capacité est plus grande ou plus petite que la valeur préréglée. La sensibilité de l'instrument équivaut à un changement de capacité de 0,02 pF, mais cette valeur peut être modifiée pour toute application spéciale. La capacité de la sonde équilibrée est de 8 pF au maximum. Cette valeur peut être accrue lorsqu'on augmente la capacité à l'intérieur de l'instrument, selon les besoins.

EE 6752 *pour plus amples renseignements*

AVO LTD

Avocet House, 92-96 Vauxhall Bridge Road, London, S.W.1.

ANALYSEUR DE TRANSISTORS

(Illustration à la page 108)

Cet appareil compact et portatif, fonctionnant sur pile, a été conçu pour le contrôle des petits et moyens transistors npn ou pnp, soit en place soit hors de l'installation dont il fait partie. Il répond rigoureusement aux conditions de la "United Kingdom Climatic and Durability Specification K.114".

L'Analyseur de transistors permet de relever les caractéristiques I_b/I_c avec choix de V_c, le transistor se trouvant dans un émetteur mis à la terre. On peut mesurer a' en n'importe quel point prédéterminé de la caractéristique I_b/I_c tandis que la valeur Ico' est indiquée directement en microampères.

Il permet aussi de mesurer les parasites de transistors en comparant les parasites de pointe, sur une gamme de 100 Hz à 10 KHz, avec un signal d'entrée équivalent, étalonné par l'oscillateur intérieur de 1 KHz. Enfin, l'analyseur peut être utilisé comme appareil de contrôle à plusieurs sensibilités pour réparer les matériels transistorisés.

EE 6753 *pour plus amples renseignements*

BRITISH PHYSICAL LABORATORIES

Houseboat Works, Radlett, Hertfordshire.

COMPARATEUR DE PIECES DETACHEES

(Illustration à la page 109)

Ce comparateur d'inductance, résistance et capacitance, type CZ 457/1, fonctionne sur 100 KHz et constitue un dévelopement du modèle 457 fonctionnant sur 1 kHz.

Il a été spécialement conçu pour le contrôle de condensateurs, résistances et inducteurs en cours de production en masse. Sans l'aide d'aucun accessoire mécanique, il permet d'atteindre aisément des vitesses de contrôle de 1 500 à 2 000 pièces par heure.

L'instrument compare la pièce détachée sous examen avec une pièce-étalon et indique directement le pourcentage de déviation sur un enregistreur à échelle de 125 mm. Les tolérances, en plus ou en moins, de 0,5% à 25%, sont clairement indiquées. En outre, il permet de vérifier l'angle de phase des condensateurs et des inducteurs, dont la déviation, telle que comparée avec la pièce-étalon, est indiquée directement sur l'échelle de l'enregistreur.

Une des particularités les plus remarquables de l'instrument consiste en un disjoncteur à maxima automatique dont est muni l'enregistreur. Que la pièce sous examen se trouve en circuit ouvert ou en court-circuit, la déviation de l'enregistreur ne dépassera pas de plus de 20% la valeur maximum de la gamme de l'enregistreur.

L'instrument est entièrement auto-contrôlé et comporte des commandes pour vérifier la précision de l'étalonnage, sans qu'il soit besoin d'échelles supplémentaires ou de références de correction.

Il effectue les comparaisons de condensateurs dans la gamme de 20 pF à 0,1 μF et celles d'inducteurs de 20 μH à 100 μH, le degré de précision étant à 3% près de la valeur indiquée.

EE 6754 *pour plus amples renseignements*

THE BRITISH THOMSON-HOUSTON CO. LTD

Rugby, Warwickshire.

SPECTROMETRE CENTIMETRIQUE A SUPER-HETERODYNE

Ce spectromètre a été étudié principalement en vue de l'analyse des spectres d'oscillateurs à impulsions, en indiquant sur un tube cathodique la distribution des fréquences de la puissance haute-fréquence. La gamme des largeurs d'impulsions va de 0,1 microseconde à 5 microsecondes.

Le spectromètre se prête également à la mesure des variations de fréquence et à la mesure des rapports d'amplitude de signaux à très faible amplitude. Il consiste, en effet, d'une superhétérodyne à bande étroite dont on peut varier la fréquence de l'oscillateur local au moyen d'une forme d'onde en dents de scie. Cette forme d'onde est aussi appliquée aux plaques X du tube à rayons cathodiques.

Le signal d'entrée est appliqué au mélangeur par un atténuateur haute fréquence, cependant que l'oscillateur local est modulé par la forme d'onde en dents de scie et qu'il alimente également le mélangeur. La fréquence intermédiaire qui en résulte est détectée et amplifiée. La fréquence de sortie est ensuite appliquée aux plaques Y du tube à rayons cathodiques au moyen d'un amplificateur vidéo. L'amplificateur de fréquence intermédiaire fonctionne à une fréquence de 48 MHz et sa largeur de bande est de 50 Hz. L'appareil comprend un oscillateur piloté par quartz afin d'assurer un étalonnage de fréquence de la trace du tube cathodique à intervalles de 1 MHz. Les pièces centimétriques sont montées sur la partie démontable du panneau avant. Les panneaux interchangeables rendent le spectromètre approprié aux longueurs d'ondes centimétriques les plus utilisées.

EE 6755 *pour plus amples renseignements*

CAWKELL RESEARCH & ELECTRONICS LTD

Scotts Road, Southall, Middlesex.

APPAREIL POUR MESURER LA CONSTANTE D'AMORTISSEMENT

(Illustration à la page 109)

Cet appareil offre un moyen sensible et non-destructif de s'assurer de l'état intérieur de n'importe quel matériau de construction. Son principe de base est la mesure précise de la période d'extinction, de la constante d'amortissement, du décrément logarithmique ou "Q" de vibration mécanique dans le matériau sous examen, afin d'obtenir des données sur son état intérieur, ses défauts, son hétérogénéité, sa composition chimique, etc. Le spécimen est excité en une vibration mécanique à sa fréquence naturelle ou à une fréquence quelconque harmoniquement analogue. Lorsque des oscillations régulières sont produites, l'excitation est interrompu eet la diminution d'oscillations est "observée" par un transducteur approprié et enregistrée pendant un intervalle de temps réglé à l'avance. L'enregistrement est présenté sous forme de chiffre à quatre signes sur des dékatrons. Avec la toute dernière version de cet appareil, on peut effectuer des mesures entre 50 Hz et 50 kHz.

EE 6756 *pour plus amples renseignements*

DAWE INSTRUMENTS LTD

99-101 Uxbridge Road, Ealing, London, W.5.

GENERATEUR BASSE FREQUENCE PORTATIF

(Illustration à la page 109)

Ce générateur de basse fréquence, type 421, fonctionne sur batterie, s'il y a lieu, et il est entièrement transistorisé. La fréquence est à variation continue dans la gamme de 20 Hz à 20 kHz, couverte en trois étages de décades.

Ces caractéristiques le rendent particulièrement indiqué pour les travaux extérieurs comme, par exemple, l'étalonnage d'enregistreurs de niveau de son. Pour les travaux sur banc d'essai continu, on emploie le bloc d'alimentation secteur incorporé à l'appareil.

Un nouvel oscillateur symétrique à pont de Wien peut être employé pour surmonter les difficultés qu'on éprouve habituellement avec les basses impédances des circuits transistorisés. Chaque section symétrique de l'oscillateur emploie deux plaques collectrices à la terre et une plaque transmetteuse stabilisée à la terre. La constance d'amplitude est maintenue par l'emploi d'un thermistor à très faible puissance, nouvellement conçu. Le signal de l'oscillateur alimente deux paires de sorties symétriques à transistors, du type des circuits de sortie à charge cathodique par transistor, donnant une tension de sortie de 2,5 volts dans 600 Ohms. Un voltmètre de sortie est également incorporé.

La précision est de $\pm 2\%$ et la distorsion inférieure à 1%. La sortie varie jusqu'à 2,5 volts dans 600 Ohms.

EE 6757 *pour plus amples renseignements*

THE DISTILLERS CO. LTD

172/3 Tottenham Court Road, London, W.1.

BLOC A INDICATEUR PANORAMIQUE DE TEMPERATURE

Cet instrument indique, sous forme de profil, sur un tube à rayons cathodiques, les sorties jusqu'à 24 couples thermoélectriques. Les 24 couples thermoélectriques sont explorés à la file, le déplacement vertical du spot représentant la température chaque fois qu'un couple thermoélectrique est mis en circuit. L'image consiste donc d'un nombre de lignes verticales, également séparées l'une de l'autre dans une direction horizontale. Le pourtour de cette image représente la répartition de la température aux points où ces couples thermoélectriques sont situés. L'exploration de 24 couples thermoélectriques dure 1,5 seconde.

Toutes les entrées peuvent être compensées dans la même mesure, de sorte qu'une légère variation de haute température moyenne peut être indiquée avec une grande sensibilité. Cette particularité permet aussi de mesurer les températures individuelles. En outre, chaque entrée peut être refoulée individuellement, de sorte que lorsqu'une répartition voulue de température est réalisée, toutes les entrées sont effectivement à zéro et il en résulte une image à lignes droites. La sensibilité peut alors être augmentée de manière que toute déviation de la position requise peut être facilement décelée. On peut, si l'on veut, actionner un système avertisseur lorsque la déviation dépasse les limites préréglées, une série de lampes de signalisation indiquant, à ce moment-là, les divers points ayant donné lieu à l'état d'alarme.

Quoique cet instrument ait été étudié principalement pour être employé avec les couples thermoélectriques, on peut utiliser tout genre de transducteur produisant un signal de sortie électrique.

EE 6758 *pour plus amples renseignements*

EDWARDS HIGH VACUUM LTD

Manor Royal, Crawley, Sussex.

INSTALLATION POUR L'ALUMINIAGE DE TUBES A RAYONS CATHODIQUES

L'application d'un film d'aluminium uniforme au dos de l'écran d'un tube à rayons cathodiques de 110° constitue un problème fort ardu en raison de la longueur réduite de l'enveloppe et du diamètre également réduit du cou.

Une installation spéciale a donc été conçue en vue d'améliorer l'uniformité du film d'aluminium déposé par un filament de tungstène. Ce dernier est fixé sur un bras à charnière afin de l'introduire plus facilement dans le cou du tube. Le bras est maintenu en place par un ressort et on le fait tourner à l'intérieur du tube afin d'appliquer une couche parfaitement uniforme à la face interne. La distance radiale et la position de la source sont réglables de façon à pouvoir y adapter des tubes de différentes dimensions.

EE 6759 *pour plus amples renseignements*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex.

DETECTEUR DE RADIATIONS

(Illustration à la page 110)

Le Détecteur beta-gamma Ekco type N596 est un nouveau Détecteur de radiations portatif, étudié pour répondre aux conditions exigées par les règlements devant entrer en vigueur prochainement au sujet des radiations beta et gamma dans les usines. Des gammes de lecture de 0 à 3, 0 à 30 et 0 à 300 mRad/h sont prévues par l'appareil, ainsi qu'un dispositif d'intégration jusqu'à 30 mRad. La chambre d'ionisation équivalente à l'air peut facilement être remplacée par une autre chambre d'un dixième de la dimension de la première, pour la mesure très localisée du taux de dosage. L'appareil se prête également à la mesure des rayons X faibles et son insensibilité virtuelle aux parasites de haute fréquence le rend particulièrement utile en ce qui à trait aux lampes haute tension, tubes à rayons cathodiques, etc.

Le Détecteur Universel N645 de faible activité est un appareil portatif poids léger, étudié pour la détection de minuscules quantités de particules radioactives émanant d'une source quelconque. Il comprend un compteur transistorisé pouvant être porté en bandoulière et fonctionnant sur batterie. Ce compteur est relié à une jauge à scintillations tenue en main. Les tungstates phosphorescents interchangeables détectent les particules alpha et beta, les rayonnements gamma ainsi que les neutrons lents et rapides. Les gammes de comptage prévues sur le compteur vont de 0 à 10, 0 à 100 et 0 à 1 000 enregistrements par seconde.

EE 6760 *pour plus amples renseignements*

ELLIOTT BROS (LONDON) LTD

Century Works, Lewisham, London, S.E.13.

HETERODYNE DE SERVICE A HYPERFREQUENCES

(Illustration à la page 110)

L'Hétérodyne de service à hyperfréquences type B689 comprend une cavité d'oscillateur coaxial étudiée pour fonctionner entre 7 et 12 kMHz, en employant un oscillateur à klystron CV 2346, une alimentation stabilisée et un modulateur à ondes carrées, de 400 Hz à 4 MHz. L'énergie haute fréquence est extraite au moyen d'une prise coaxiale et l'on obtient ainsi une puissance de sortie suffisante, sur toute la largeur de la bande, pour toutes les mesures normales d'hyperfréquences sur banc d'essai. Un cadran de lecture directe de fréquences, donnant des indications précises sur la fréquence de fonctionnement, est incorporé à l'appareil, qui est également prévu pour les modulations extérieures.

EE 6761 *pour plus amples renseignements*

ENGLISH ELECTRIC VALVE CO. LTD

Chelmsford, Essex.

TUBES D'ENREGISTREMENT POUR VISION A LA LUMIERE DU JOUR

(Illustration à la page 110)

Le tube cathodique E702 à déviation électrostatique a été étudié essentiellement pour être utilisé dans des oscilloscopes où l'on veut "emmagasiner" l'image. Etant donné que la persistance de l'image peut durer près de 2 minutes, ce tube peut être utilisé tout aussi bien pour la projection des formes d'ondes à variation lente que pour celle d'images transitoires rapides. La vitesse d'enregistrement maximum est d'environ 8mm/microseconde. Le canon de parcours électronique est actionné à 2,2 kV, ce qui a pour effet de réduire la puissance de déviation. L'écran lui-même, est actionné à 10 kV, lui donnant cette grande brillance qui caractérise ces tubes et rend l'usage de masques inutiles, même sous une forte lumière ambiante.

Le tube E703 a également été étudié à l'intention des oscilloscopes et il est muni de deux canons de passage électronique permettant l'enregistrement simultané de deux formes d'ondes indépendantes.

Le Tube E701 à déviation magnétique est destiné aux installations de radar où sa haute brilliance est utilisée avec avantage dans des indicateurs installés dans un poste de pilotage ou sur une passerelle.

Tous ces tubes cathodiques ont un diamètre utile d'écran d'au moins 102 mm.

EE 6762 *pour plus amples renseignements*

E.M.I. ELECTRONICS LTD

Hayes, Middlesex.

CYLINDRES ENREGISTREURS POUR CALCULATEURS ELECTRONIQUES

(Illustration à la page 110)

Ces cylindres qui ont été produits pour répondre aux besoins de l'ingénieur d'ordinateurs électroniques, comprennent des blocs de tête magnétiques à haute sensibilité d'un modèle nouveau. Les éléments de commutation à circuits imprimés, faisant corps avec l'ensemble cylindrique, sont d'un accès aisé.

La capacité d'enregistrement—ou "d'emmagasinage"—est de 8 000 ou 16 000 mots d'une longueur de 36 signes, sur 128 + 8 lignes de réserve ou 256 + 16 lignes de réserve. Il y a cinq lignes-montre et réserves. Les blocs entrelacés comprennent 16 têtes lecture-écriture et 1 tête de réserve.

Toutes les têtes se prêtent à la lecture et à l'écriture. Les cinq têtes de montre permettent la lecture et l'écriture simultanées. Un écran de grande efficacité protège chaque tête contre toute perturbation de l'extérieur.

Les cylindres ont un diamètre de 203 mm. Le cylindre de 8 000 mots a une surface en tambour d'environ 178 mm, tandis que celui de 16 000 mots en a une de 324 mm. Un fonctionnement continu d'au moins 20 000 heures, à 3 000 t.p.m. dans des conditions normales, pourrait être assuré au dire des constructeurs.

EE 6763 *pour plus amples renseignements*

FERRANTI LTD

Hollinwood, Lancashire.

REGULATEURS DEWATTES A TRANSISTORS

Ces régulateurs emploient un ou plusieurs transistors en parallèle, faisant fonction d'interrupteurs dans un circuit à diode et bobine de filtrage condensateur. Dans ce circuit, la tension apliquée à l'entrée de la bobine de filtrage-condensateur constitue soit la tension d'alimentation soit zéro, selon que le transistor est mis en circuit ou hors circuit. La sortie de courant continu du filtre est égale à la valeur moyenne de la forme d'onde d'entrée. Un circuit bouclé de contrôle vérifie la tension de sortie sur un potentiel de référence, comme la diode Zener, et règle automatiquement le rapport marque/espace du signal à l'interrupteur, de façon à stabiliser la sortie en cas de fluctuations de tension d'entrée ou de charge. On peut obtenir des rendements de l'ordre de 90%, et dans le cas d'un régulateur type employant un simple circuit, les variations de tension d'entrée sont réduites au centuple.

EE 6764 *pour plus amples renseignements*

THE GENERAL ELECTRIC CO. LTD

Magnet House, Kingsway, London, W.C.2.

INDICATEUR A INDUCTION

Cet appareil est un transducteur d'un modèle entièrement nouveau, étudié pour l'emploi avec des installations de calculateurs électroniques. Il donne une indication numérique de la position d'un axe ou d'un curseur, sous forme de signaux électriques chiffrés, sans qu'il soit besoin de faire usage de collecteurs, de commutateurs ou de cellules photoélectriques.

Fondamentalement, il s'agit d'un transformateur comprenant un enroulement d'excitation unique et un nombre d'enroulements secondaires correspondant aux nombre de signes. La position d une culasse mobile détermine la polarité de l'accouplage entre l'enroulement d'excitation et les différents enroulements de signes, qui sont disposés de façon à se conformer aux conditions du code. L'excitation s'effectue soit par impulsion soit par onde sinusoïdale.

Les indicateurs à induction peuvent être employés pour des combinaisons de réglages approximatifs ou précis, un enroulement supplémentaire empêchant toute confusion entre les indicateurs. Ces appareils ont été construits soit pour les indications à cinq signes dans un carter synchrone, dimension II, soit pour les indications à sept signes dans un carter de la dimension 23.

EE 6765 *pour plus amples renseignements*

GRIFFIN & GEORGE LTD

Ealing Road, Alperton, Wembley, Middlesex.

PONT DE RESISTANCE-REACTANCE

Le Pont Griffin Panmetron, fonctionnant sur secteur, sert non seulement de pont de courant alternatif pour comparer la résistance et la réactance à 50 Hz, en employant l'étalon externe, mais aussi d'amplificateur-détecteur de grande amplitude pour l'emploi avec des circuits de pont externes. Pour mettre des derniers sous tension, une source de courant alternatif de 2 et 4 volts est incorporée à l'instrument. Le courant déréglé est appliqué sur le côté d'entrée d'un amplificateur de grand gain, en employant un EM84 comme indicateur d'équilibre. Lorsque l'interrupteur de contrôle est à sa plus haute position, on obtient une bonne indication d'équilibre avec une tension d'entrée de 0,1 μV seulement.

EE 6766 *pour plus amples renseignements*

ISOTOPE DEVELOPMENTS LTD

Beenham Grange, Aldermaston Wharf, near Reading, Berkshire.

ECHELLE AUTOMATIQUE

(Illustration à la page 111)

L'Echelle type 1700 constitue un nouvel instrument des plus intéressants de la série des productions de la Société Isotope Developments Ltd. Elle se compose d'un mécanisme complet de comptage automatique ne formant qu'un seul bloc pouvant se poser sur une table. Il suffit de lui adjoindre un détecteur de radiations pour pouvoir effectuer n'importe quelle mesure de radioactivité.

L'Echelle Type 1700 a résolu, par sa construction particulière, le problème des réparations. Grâce, en effet, à la trousse de pièces de rechange fournie avec l'appareil, on peut réduire considérablement les pertes de temps dues à l'inutilisabilité de ce dernier, et toute réparation nécessaire pourra être exécutée facilement en n'importe quel lieu du monde, même par une personne inexpérimentée.

En dehors du bloc d'alimentation, l'Echelle Type 1700 comprend un simple système à circuit imprimé avec des éléments à prises individuels. Ainsi, n'importe quelle partie du circuit peut être remplacée instantanément par sa pièce de rechange au cas où une panne se produirait.

On peut également vérifier l'intérieur en quelques secondes au moyen de couvercles de protection contre la poussière, au dessus et au dessous de l'appareil, se détachant aisément.

Le bloc d'alimentation est monté verticalement et il est isolé de tous les principaux circuits de comptage. Quant au châssis, il est fixé sur charnières afin qu'il puisse être ouvert facilement et rapidement.

Une échelle "rapide" est incorporée à l'Echelle type 1700, ainsi qu'une minuterie électronique, des dispositifs de temps préréglé ou de comptage préréglé, un compteur, un analyseur de hauteur d'impulsions à canal unique, une alimentation de haute tension stabilisée (250 V à 2 KV) et un amplificateur d'impulsions linéaires.

L'appareil peut être télécommandé. On peut aussi le relier à un bloc d'impression automatique extérieur et à un enregistreur extérieur du taux de comptage.

EE 6767 *pour plus amples renseignements*

KELVIN & HUGHES LTD

New North Road, Barkingside, Essex.

SYSTEME DE TELECOMMANDE A AUTO-SYNCHRONISATION

Le Syncrotel est un élément de transmission à position angulaire de la série des appareils de synchronisation, mais dont la friction et l'inertie extrêmement basses permettent son emploi dans les cas où la force électromotrice existante est fort réduite, comme, par exemple, dans les mécanismes fonctionnant par capsule.

Le stator de synchronisation est du type courant à trois phases mais le rotor a été remplacé par un enroulement monophasé sur bobine dont l'axe coincide avec la broche. La partie mobile consiste d'une section oblique de cylindre creux d'aluminium, et son arbre est soutenu par des roulements à rubis.

Les Syncrotels peuvent être utilisés directement en paires dos à dos dans certaines applications, mais ils sont habituellement utilisés dans les servo-systèmes avec un élément récepteur ordinaire de synchronisation.

Le poids du rotor est de 0,76 grammes et le moment d'inertie de 0,11 g-cm².

EE 6768 *pour plus amples renseignements*

MUIRHEAD & CO. LTD

Beckenham, Kent.

ANALYSEUR OSCILLATEUR DE HAUTE FREQUENCE

(Illustration à la page 111)

Ce nouvel instrument, à double usage, a une gamme de fréquences de 200 Hz à 650 kHz. Il étend ainsi les fréquences couvertes par les analyseurs et les oscillateurs Muirhead à la bande des transmissions téléphoniques par courants porteurs.

En actionnant un seul commutateur, on peut employer l'instrument soit comme analyseur sensible et constant de l'amplitude de fréquence (avec un coefficient de surtension de pas moins de 70, une discrimination d'octave de 40 dB et une indication minimum de 50 microvolts), soit comme oscillateur (avec un niveau de sortie de 2,5 Volts dans 6 000 Ohms). Dans l'un et l'autre de ces rôles, la précision de fréquence est de 2 %.

EE 6769 *pour plus amples renseignements*

MULLARD EQUIPMENT LTD

Mullard House, Torrington Place, London, W.C.1.

CONVERTISSEUR ANALOGIQUE-NUMERAL DE GRANDE VITESSE

Cet appareil a été étudié pour la conversion à grande vitesse d'une tension d'entrée analogique en une tension binaire de 10 parties, produite en parallèle ou en série et sous forme binaire réelle ou décimale binaire-chiffrée. La durée de la prise d'échantillons est de moins d'une microseconde, les prises étant effectuées à la réception d'impulsions d'entrée, à n'importe quelle cadence jusqu'à 1 000 par seconde. La gamme de tensions d'entrée analogiques va de 0 à 10 volts (positifs). Pour la sortie en série, le taux d'impulsions de sortie maximum est d'environ 12 kHz, mais on peut aussi bien le bloquer à n'importe quelle source extérieure d'impulsions, à condition que la fréquence d'impulsion ne dépasse pas 12 kHz. Pour la sortie en parallèle, le taux d'impulsions coincide avec l'impulsion de prise d'échantillons d'entrée suivante. Le convertisseur est entièrement transistorisé et muni de planches de circuits imprimés et il est donc d'un fonctionnement très sûr. Il est particulièrement indiqué pour les procédés dits de maniement de données où l'on doit manier de très grosses quantités de données en un très court laps de temps comme, par exemple, pour les mesures dynamiques de moteurs de fusées ou pour les essais de structures d'avions dans des souffleries aérodynamiques pour avions supersoniques.

EE 6770 *pour plus amples renseignements*

SPECTROMETRE D'IMPULSIONS MULTI-CANAUX

(Illustration à la page 112)

Le Spectromètre d'impulsions L.332, à cent canaux, effectue des analyses rapides et précises des spectres d'énergie de radiations nucléaires. Les impulsions d'un détecteur proportionnel sont comptées et triées, selon leur amplitude relative, dans l'un des cent canaux où elles sont "emmagasinées" dans un système de mémoire à matrice magnétique. Vu que l'amplitude des impulsions produites est en fonction directe de l'énergie de la radiation incidente, les données enregistrées dans la mémoire fournissent un histogramme indiquant l'abondance relative de radiation à chacun des 100 niveaux d'énergie à l'intérieur du spectre sous examen. Ces indications sont présentées sur un tube à rayons cathodiques de 152 mm, sous forme d'image continue, linéaire et analogique de la courbe spectrale, l'image étant projetée tant avant qu'après la période de comptage. En outre, des indicateurs numériques au néon, "en ligne", donnent une lecture directe des comptes enregistrés dans l'un des canaux. Cette lecture est non-destructive. La mémoire à matrice contient 100 × 16 noyaux Ferroxcube, donnant à chaque canal une capacité d'emmagasinage de 65 536 comptes. Le taux de comptage maximum n'est limité que par le pourcentage des comptes perdus par suite du temps mort, qui est constant à 25 microsecondes pour les impulsions admises.

La gamme d'entrée de base du spectromètre est de 0,5 à 10 volts, la largeur de chaque canal étant d'environ 0,1 volt. Une tension de polarisation arrière allant jusqu'à 30 volts peut être appliquée, donnant une étendue totale equivalant à 400 canaux. L'uniformité de la largeur de bande est supérieure à 1% et la stabilité des positions de canaux est supérieure à 10 mV sur la gamme d'entrée de base. Le vacillement occasionnel est inférieur à 3 mV.

Les spectromètre L.332 est un appareil parfaitement sûr, faisant un usage étendu de transistors. Il est basé sur un dessin des Etablissements de Recherche Atomique de Harwell, Angleterre.

EE 6771 *pour plus amples renseignements*

G. V. PLANER LTD

Windmill Road, Sunbury-on-Thames, Middlesex.

CIRCUITS IMPRIMES CUIVRE-SUR-VERRE

(Illustration à la page 112)

Dans ce type de circuits imprimés électroniques, les câblages de cuivre sont étroitement agglomérés à une base en verre. Grâce à ce système, les circuits se prêtent au fonctionnement à haute température et gardent leur inertie en raison de l'absence de substrates organiques ou de toutes matières agglomérantes. Les modèles exposés comprenaient des inducteurs faits en substrate de verre, coupés électroniquement et portant chacun un câblage de cuivre de chaque côté.

Les plaques peuvent être entassées de façon à former des inducteurs à couches multiples les interconnexions s'effectuant par les bords du verre. Un inducteur construit selon ce procédé est représenté dans la gravure ci-après, agrandie à environ × 2.

EE 6772 *pour plus amples renseignements*

W. G. PYE & CO. LTD

Granta Works, Cambridge.

POTENTIOMETRE DE DECADES DE HAUTE PRECISION

Ce potentiomètre perfectionné, de haute précision, a été étudié dans le but d'assurer un grand rendement sans entretien et de donner les indications voulues de la façon la plus appropriée. Il est entièrement à commutation par décades, éliminant ainsi les curseurs. La première décade est combinée avec le commutateur de changement de gamme, permettant donc la lecture directe pour toutes les gammes. La gamme de tensions complète s'étend de 1,99999 Volts à –0.1 mV. L'instrument comprende un dispositif de contrôle précis du courant de décade, donnant des variations de courant par plots de 1 partie sur 10, jusqu'à 1 partie sur 10^6. On peut aussi choisir par les commutateurs n'importe lequel des cinq circuits externes et transférer la pile-étalon, des circuits correcteurs au diviseur de tension, aux fins d'étalonnage. L'instrument est également muni d'un commutateur d'inversion de circuit, permettant de tenir compte de toute force électromotrice erronée (moins de 0,1 mV) qui pourrait se produire dans le galvanomètre ou le circuit de sortie.

Les trois gammes existantes sont comme suit:

10 μV à 1,99999 V en plots de 10 μV.
1 μV à 0,199999 V en plots de 1 μV.
0,1 μV à 0,0199999 V en plots de 0,1 μV.

La précision de la division de tension est de ±0,0001%. Cette précision peut être obtenue en comparant deux forces électromotrices. La précision absolue de mesure est de ±0,002%.

La correction de la pile étalon englobe une gamme de températures de 10°C à 37°C. Le plus petit plot est de 10 μV égalant 0,25°C.

EE 6773 *pour plus amples renseignements*

THE SOLARTRON ELECTRONIC GROUP LTD

Solartron Works, Thames Ditton, Surrey.

OSCILLOSCOPE A DOUBLE FAISCEAU

(Illustration à la page 112)

L'Oscilloscope CD711S.2 constitue un perfectionnement des oscilloscopes CD711 et CD711S. Tant la robustesse que la sécurité de fonctionnement ont été améliorées de manière à augmenter sa résistance aux conditions d'utilisation les plus rudes. Il a été agréé par les autorités militaires sous la désignation CT414. En outre, sa disposition intérieure a été modifiée de façon à permettre de mieux l'examiner.

Le CD711S.2 applique la technique des variations d'orientation du faisceau, et il est doté d'un dispositif d'écrêtage pour les vitesses de balayage à faible base de temps. Les deux canaux de déviation verticale ont des caractéristiques identiques, avec une largeur de bande de 7 MHz et une sensibilité maximum de 3 mV/cm.

L'appareil comporte une commande étalonnée de décalage Y qui peut être branchée sur l'un ou l'autre des canaux afin d'assurer l'étalonnage d'amplitude avec une précision de ±5%.

La base de temps a une gamme de vitesses de balayage s'étendant de 3 cm/μsec à 0,3 cm/sec.

On peut normaliser l'étalonnage de la base de temps sur un oscillateur à cristal intérieur et sur un système de division de fréquences.

EE 6774 *pour plus amples renseignements*

ANALYSEUR DE REPONSE DE PROCEDE

(Illustration à la page 113)

L'Analyseur de réponse de procédé, type JY743, facilite particulièrement certaines analyses dynamiques de très basse fréquence.

Sa plage de fréquences est de 0 0001 Hz à 100 Hz, avec une précision de fréquence de ±1,5% et une stabilité d'amplitude supérieure à 1% sur toute la plage de fréquences. Il assure des sorties d'ondes carrées et triangulaires, en plus des sorties d'ondes sinusoïdales à quatre phases et trois phases respectivement. La teneur en courant continu des sorties d'ondes sinusoïdales est inférieure à 0,1% de la sortie crête à crête de courant alternatif et la distorsion non linéaire est inférieure à 0,5%.

Les relais dont est muni l'analyseur assurent l'amorçage, le démarrage, le maintien et le réenclanchement du fonctionnement de l'oscillateur à des fréquences de moins de 10 Hz. Ces relais peuvent être actionnés à la main ou de l'extérieur, c'est à dire, par exemple, au moyen des commutateurs 'calcul', 'maintien' et 'réenclanchement' d'un calculateur ou d'un simulateur électroniques. On peut aussi effectuer le contrôle numéral de la tension momentanée de l'oscillateur.

Le bloc indicateur comprend deux enregistreurs de 127 mm (référence et quadrature) qui indiquent l'amplitude et le signe des composants réels et imaginaires. Un dispositif de lecture numérale permet également le traçage automatique de la fréquence et de l'amplitude. La sensibilité maximum est de 3 mV pour une déviation complète et le filtrage d'harmoniques est supérieur à 40 dB.

La technique spéciale appliquée dans l'image donne des indications stables, quelle que soit la fréquence des signaux fondamentaux.

Le processus d'intégration, pour toute mesure donnée, est donc déterminé par la teneur en bruit. Le filtrage est assuré à des constantes de temps de 0,1, 0,3, 1,3 et 10 secondes et l'intégration à cycle unique s'effectue à des fréquences de 1 cycle par seconde et 1 cycle par heure. On peut, enfin, régler à la main la période d'intégration.

EE 6775 *pour plus amples renseignements*

STANDARD TELEPHONES & CABLES LTD

Connaught House, Aldwych, London, W.C.2.

EQUIPEMENT D'ESSAI DE "BRUITS BLANCS"

(Illustration à la page 113)

Cet équipement a été étudié pour la mesure de bruits fondamentaux et de bruits d'intermodulation dans les systèmes téléphoniques multicanaux. Sa largeur de bande va jusqu'à 4 MHz (16 supergroupes) et il produit une gamme de bruits occasionnels ayant une distribution d'énergie à peu près uniforme dans la plage de fréquences choisie, pouvant correspondre, selon les besoins, à 2, 4, 10 ou 16 supergroupes. On fait passer les bruits par un jeu de filtres afin de déterminer la largeur de la bande de bruits et de former à l'intérieur de cette dernière une étroite bande silencieuse. Ce signal peut être appliqué à un système téléphonique multicanaux (hors des lignes de trafic) à un niveau de puissance analogue à la charge de trafic réel, ce qui permet d'effectuer des mesures de diaphonie et de bruit à l'extrémité éloignée du système. On peut mesurer des rapports bruit/puissance allant jusqu'à 70 dB, tandis que le fonctionnement dos à dos du générateur et du récepteur donnera des chiffres de rapport bruit/puissance de 75 dB ou davantage pour chaque bande de fréquences reçues. La densité minimum absolue de "bruit blanc" pouvant être mesurée est de -118 dBm/kHz, correspondant à $-114{,}5$ dBm/canal lorsqu'elle est psophométriquement atténuée par un réseau d'atténuation téléphonique C.C.I.F. (1951).

L'equipement comprend un générateur se composant de trois éléments portatifs, mesurant chacun 559 mm × 251 mm × 298 mm, et d'un récepteur composé de deux éléments de même grandeur que les précédents et d'un troisième élément mesurant 559 mm × 206 mm × 298 mm. Le générateur et le récepteur sont, tous deux, pourvus de leur propre bloc d'alimentation secteur c.a.

EE 6776 *pour plus amples renseignements*

SUNVIC CONTROLS LTD

Crown House, Aldwych, London, W.C.2.

ANALYSEUR DE HAUTEUR D'IMPULSIONS

(Illustration à la page 113)

Le nouvel analyseur d'impulsions SUNVIC est livré soit sur console soit monté sur bâti.

Transistorisé en grande partie, il comprend une matrice d'enregistrement à noyau de ferrite, fournissant une centaine de canaux donnant un comptage décimal de 65,535. Le comptage est indiqué sur un tube cathodique, sous forme de trace analogique où la déviation horizontale représente le numéro du canal et la déviation verticale est proportionnelle au comptage. L'analyseur fait usage de Dékatrons pour indiquer, sous forme décimale, le chiffre total d'impulsions dans l'un quelconque des canaux choisis, soit manuellement soit automatiquement en série. Les totaux peuvent aussi être imprimés automatiquement.

EE 6777 *pour plus amples renseignements*

VENNER ELECTRONICS LTD

Kingston-By-Pass, New Malden, Surrey.

ETAGES A TRANSISTORS

(Illustration à la page 113)

Les plus récentes réalisations de la gamme d'étages interchangeables Venner comprennent l'oscillateur à cristal type TS25, qui fonctionne à des fréquences entre 100 kHz et 1 MHz (selon le cristal employé). Cet instrument peut être utilisé pour la production d'ondes sinusoïdales ou de formes d'ondes d'impulsions. Il est de la même grandeur que l'oscillateur à cristal de 10 kHz, type TS5. La toute dernière réalisation dans la série des oscillateurs à cristal est le modèle TS30. C'est un oscillateur à basse fréquence qu'on emploie avec des cristaux à courbure XY. La gamme couverte par cet élément s'étend de 5k Hz à 100 kHz.

Il y a aussi le bloc binaire de 1 MHz, construit en deux modèles: le TS2A/HF, destiné aux chaînes de division exigeant un rapport de division qui ne soit pas en multiples de deux, et le TS2B/HF, qui est particulièrement indiqué pour être utilisé avec la vanne à mircrosecondes type TS36. Cette vanne peut être ouverte ou fermée en changeant le niveau de courant continu de la clavette de contrôle, de -9 V à -4 V, avec une alimentation de 10 volts, ces niveaux étant aisément fournis par le bloc binaire type TS2B/HF. Deux de ces blocs, reliés à la vanne à microsecondes, peuvent assurer un effet de couplage de manière que la première impulsion de contrôle reçue ouvre la vanne, la seconde la referme et les impulsions suivantes n'ont plus d'autre rôle dans ce sens.

L'instrument à former des impulsions de microsecondes, type TS27, a été étudié pour fournir une impulsion pouvant actionner des éléments binaires et de décades, d'une gamme étendue de formes d'ondes, à des fréquences allant jusqu'à 1 MHz. La sortie consiste en impulsions positives ayant une amplitude d'environ 8 volts, une largeur de 0,4 microsecondes et une durée d'établissement de moins de 0,2 microsecondes à 1 MHz. On peut, évidemment, l'employer également pour actionner la vanne de microsecondes.

Le bloc de décades à 1 MHz, type TS10/HF, qui fait partie du choix d'appareil exposés, est d'un fonctionnement analogue au bloc de décades courant, type TS10/5, en ce qu'il commande un compteur étalonné de 0 à 9, afin de donner une indication du comptage effectué. Ce bloc fonctionne à n'importe quelle fréquence usqu'à 1 MHz lorsqu'il est commandé, par exemple, par l'appareil d'impulsions type TS27.

Enfin, le réenclancheur électronique, type TS32, peut être utilisé avec n'importe quel étage binaire ou de décades, et il est capable de réenclancher ces derniers en moins de 4 microsecondes lorsqu'il est déclenché par une impulsion négative. Il n'est fait usage d'aucun contact, le fonctionnement étant entièrement électronique.

EE 6778 *pour plus amples renseignements*

Résumés des Principaux Articles

Dispositifs d'arrêt de guides d'ondes et réseaux connexes. par J. W. Sutherland.

Résumé de l'article aux pages 64 à 68

Cet article décrit et analyse les interrupteurs de guides d'ondes de type cylindrique ou actionnés par obturateur, ainsi que les interrupteurs de tubes à décharge et les interrupteurs de ferrite. L'auteur y compare les propriétés de chaque modèle et il examine en détail les caractéristiques de branchement en T. de même que plusieurs de ses applications. Il décrit également un réseau à connexions multiples faisant usage de différentiels.

L'emploi des Dékatrons pour la distribution d'impulsions. par G. H. Stearman.

Résumé de l'article aux pages 69 à 71

On emploie souvent des tubes sélecteurs à cathodes froides lorsqu'on a besoin d'une source pouvant produire des impulsions sur plusieurs lignes de sortie à intervalles réguliers. Le nombre des lignes de sortie est, cependant, limité à douze et il faut recourir à des méthodes spéciales pour augmenter ce nombre. Cet article traite d'une méthode éprouvée, c'est à dire celle de la matrice à diode, et en suggère deux autres pouvant, toutes deux, être fort utiles en certains cas.

Un diviseur de fréquence à transistors. par F. Butler.

Résumé de l'article aux pages 72 à 75

L'auteur décrit un générateur de fréquences étalonnées se composant d'un oscillateur de 100 kHz piloté par quartz et de deux diviseurs de fréquence donnant, à travers un amplificateur intermédiaire, des tensions de sortie de 1 et de 10 kHz. Les diviseurs de fréquence sont du genre des modulateurs à réaction, se caractérisant par le fait qu'une rupture de la tension motrice entraîne l'arrêt complet de la sortie, empêchant ainsi le fonctionnement libre normal.

L'appareil est entièrement muni de transistors et il ne comporte aucune commande préréglée ou réglable. Les tensions de sortie sinusoïdales sont d'un niveau permettant leur utilisation directe, sans qu'il soit besoin d'appareil supplémentaire.

L'article décrit, enfin, un circuit diviseur à oscillateur de blocage, fournissant une gamme de fréquences, dépendant de la source de 1 kHz, allant jusqu'à 25 Hz.

Compensation de fréquence pour sections de transformation de guides d'ondes étagés. par D. Wray.

Résumé de l'article aux pages 76 à 79

On tente souvent d'étendre la bande des sections de transformation de guides d'ondes étagés en faisant suivere aux taux d'impédance, à chaque étage, une certaine séquence prescrite (binomiale, logarithmique ou de Tschebyscheff). Le fonctionnement qui s'ensuit n'atteint généralement pas le résultat escompté car les taux d'impédance sont eux-mêmes sensibles à la fréquence, de sorte que la séquence voulue ne s'obtient qu'à la fréquence nominale. Cet article indique la manière de maintenir presque constants les taux d'impédance requis dans la bande de service, en changeant les dimensions, tant larges qu'étroites, à chaque étage. Il en résulte une amélioration sensible du fonctionnement.

Un circuit monostable à transistor. par H. Stinton.

Résumé de l'article aux pages 80 à 81

Cet article décrit un circuit générateur d'impulsions employant un simple transformateur d'impulsions et un transistor dont les caractéristiques sont des plus souples. Le circuit en question permet soit une augmentation quasi linéaire du courant d'émission et de captage, soit une impulsion de tension d'amplitude quasi constante. Une basse impédance de sortie rend le dispositif utile comme générateur d'impulsions intermédiaire.

Etude de grilles de polarisation de transistors d'après des caractéristiques thermiques différentielles. par L. M. Vallese

Résumé de l'article aux pages 88 à 93

Des formules d'étude précises pour la construction de réseaux de polarisation à un étage sont données dans cet article. L'emploi de ces formules exige la connaissance de deux paramètres thermiques différentiels, dont l'un se rapporte à dI_{co}/dT et l'autre au taux de changement du courant d'entrée. Des formules théoriques et des procédés expérimentaux pour déterminer ces caractéristiques sont discutées en détail dans cet article qui indique également des applications pratiques et des vérifications expérimentales de cette théorie.

Les modèles approximatifs pour lignes de transmission et les erreurs qui en résultent. par R. R. Vierhout.

Résumé de l'article aux pages 94 à 95

Il est souvent commode, dans les travaux expérimentaux et les calculs, de remplacer les lignes de transmission par des modèles simplifiés tels que les réseaux en T et en π. Cet article examine quelques formules simples donnant une indication immédiate de l'importance de toute erreur provoquée par l'emploi de ces modèles.

Un compteur de décades à transistors de 1 MHz. par C. G. Bradshaw.

Résumé de l'article aux pages 96 à 97

Cet article décrit un compteur de décades pouvant recevoir des impulsions d'entrée à une fréquence de répétition maximum de 1,3 MHz. Ce compteur fait usage de quatre étages binaires dont la rétroaction réduit les seize états stables possibles à dix. Ce procédé est d'ailleurs connu dans la technique des compteurs d'impulsions.

L'article représente également un circuit correcteur typique, fournissant les impulsions nécessaires au compteur.

Conçu pour une alimentation nominale de 12 volts, le compteur fonctionne régulièrement dans la gamme de 9 à 12 volts et à des températures s'élevant jusqu'à 55°C.

Des essais de fonctionnement en laboratoire. d'une durée d'anviron 150 heures, ont été exécutés avec plein succès.

Une introduction aux matrices et à leur emploi dans l'analyse des circuits à transistors. par J. S. Bell et K. Brewster.

Résumé de l'article aux pages 98 à 102

L'objet de Cet article est de présenter les méthodes dites "à matrices," pour l'analyse des circuits à transistors, à ceux qui abordent le sujet pour la première fois. Il est traité, du point de vue mathématique, de façon aussi pratique que possible, afin de ne pas voiler le but envisagé qui est celui de l'analyse de circuits pratiques.

Die Ausstellung der PHYSICAL SOCIETY

Beschreibungen, zusammengestellt auf Grund von Mitteilungen aus Herstellerkreisen, einer kleinen Auswahl elektronischer Geräte, die auf der 43. Ausstellung der Physikalischen Gesellschaft in London vom 19.-22. Januar zur Schau gelangten.

(Übersetzung de Seiten 108 bis 113)

ADVANCE COMPONENTS LTD

Roebuck Road, Hainault, Ilford, Essex.

F.M.-A.M. Signalgenerator

(Abbildungen auf Seite 108)

Der Signalgenerator Typ 60 liefert ein genau gesteuertes Signal in dem Bereich von 4 bis 230 MHz, frequenz- und/oder amplitudenmoduliert. Die Trägerfrequenzeichung lässt sich an Hand einer 5 MHz Innenkristalleichvorrichtung prüfen. Der Dämpfer besteht aus einem kontinuierlich regelbaren Potentiometer mit einem nachgeschalteten Vierstufen-Dekadendämpfer, die zusammen einen Bereich von 100 dB ergeben. Die Überwachung des Trägerausgangpegels geschieht mittels eines Kristallspannungsmessers.

Dieses Gerät verwendet einen Colpitts-Oszillator, und die Frequenzmodulation wird durch einen regelbaren Blindwiderstand bewirkt, der über den abgestimmten Kreis gespannt ist. Die Frequenzabweichung ist eine direkte Funktion der modulierenden Spannung, die den regelbaren Blindwiderstand steuert. Da bei unterschiedlichen Frequenzen unterschiedliche Amplituden zur Erzeugung derselben Frequenzabweichung erforderlich sind, ist eine Abweichungsausgleichsvorrichtung in dem Schaltkreis vorgesehen, die die Modulationsspannung auf einen von der Stellung des Hauptabgleichkondensators und des verwendeten Frequenzbereichs abhängigen Wert bringt. Die Ausgangsleistung des Trägeroszillators wird in einen HF-Verstärker gespeist und sein abgestimmter Kreis mit dem des Verstärkers gekoppelt. Die Amplitudenmodulation wird durch das Anlegen der Modulationsspannung an den Eingang des Verstärkers erzielt. Die Träger- und Modulationsspannung wird verschiedenen Elektroden der Verstärkerröhre zugeführt, um Frequenzschwankungen auf ein Mindestmass herabzusetzen.

Die Ausgangspegeleinstellung wird durch einen lose mit dem verstärkten Ausgang gekoppelten induktiven Potentiometer gesteuert, so dass die Einstellung nur geringe Wirkung auf die Signalfrequenz hat. Auf diese Weise wird die Signalfrequenz so gut als unabhängig sowohl von der Ausgangseinstellung als auch dem Amplitudenmodulationsgrad gestaltet. Die durch die Amplitudenmodulation verursachte Nebenfrequenzmodulation ist sehr gering. Ein sich bei 1 kHz betätigender Oszillator liefert eine Innenmodulationsquelle und wird wie erforderlich zum Bewirken der Amplituden- oder Frequenzmodulation geschaltet. Falls beide Modulationstypen gleichzeitig auf die Trägerwelle aufzutragen sind, ist eine Aussenquelle für die Modulationsspannung erforderlich.

EE 6751 *für weitere Einzelheiten*

AIRMEC LTD

High Wycombe, Buckinghamshire.

Annäherungsdetektor

(Abbildungen auf Seite 108)

Der Annäherungsdetektor Typ N 266 ist ein kapazitiver Schalter, der ein Relaissystem in Betrieb setzt, sobald irgendeine Substanz in einen vorbestimmten Abstand zu der Bestimmungssonde gebracht wird. Er hat einen weiten Anwendungsbereich und ist zum Einführen in Flüssigkeiten, Pulvermassen usw. geeignet. Gedruckte Schaltungen unter Verwendung von Transistoren werden im Verstärker benützt, der sowohl mit Netz- als auch Batterieantrieb arbeitet. Die Kraftquellen- und Verstärkeragregate sind verbolzt und durch eine gedruckten Schaltungsplatine und Steckdose verbunden, um je nach Wunsch leichtes Auswechseln von Kraft- und Batterieaggregaten zu gestatten. Die Standzeit bei Batteriebetrieb beträgt mehr als 200 Stunden.

Die Bestimmungssonde kann entweder auf dem Gerätekasten oder am Ende eines armierten Kabels montiert werden, das eine Länge bis zu 7·5 m haben kann. Sonden verschiedener Grössen sind in wasserdichten Überzügen erhältlich, und zwar zur Verwendung in Flüssigkeiten bei Temperaturen bis zu 100°C.

Die Schaltung verbleibt statisch, wenn die Kapazitanz zur Sonde bei einem vorbestimmten Wert konstant ist, und setzt sich in Betrieb, wenn die Kapazitanz grösser oder kleiner als der vorbestimmte Wert wird. Die Empfindlichkeit des Geräts entspricht einer Kapazitanzänderung von 0,02 pF, aber dies lässt sich durch Regelung jedem besonderen Anwendungszweck anpassen. Die Kapazitanz der Sonde beträgt im abgeglichenen Zustand 8 pF und lässt sich wie erforderlich durch Kapazitanzzusatz im Innern des Geräts erhöhen.

EE 6752 *für weitere Einzelheiten*

AVO LTD

Avocet House, 92-96 Vauxhall Bridge Road, London, S.W.1.

Transistoranalysator

(Abbildungen auf Seite 108)

Dies ist ein tragbarer, fester, batteriegespeister Transistortester zur Prüfung von kleinen und mittleren NPN oder PNP Transistoren, entweder an Ort und Stelle oder aber demontiert vom Muttergerät. Das Gerät entspricht den Vorschriften der U.K. Klima- und Haltbarkeitsspezifikation K.114.

Das Gerät ist zum Aufnehmen von I_b/I_c Kennzeichen mit wahlweiser V_c geeignet, mit dem Transistor in der geerdeten Emitter-Anordnung. α' kann an jedem beliebigen vorbestimmten Punkt auf dem I_b/I_c Kennzeichen gemessen werden, und I_{co}' wird direkt in mA angezeigt. Vorkehrungen für das Messen des Transistor-Rauschens wurden getroffen, und zwar durch Vergleichen des Spitzenrauschens über den annähernden Bereich 100 Hz bis 10 kHz, mit einem gleichwertigen Eingangssignal, das von dem inneren 1 kHz Oszillator geeicht wird. Das Gerät kann ausserdem als ein Mehrbereich-Prüfmesser für die Wartung von transistorisierten Geräten verwendet werden.

EE 6753 *für weitere Einzelheiten*

BRITISH PHYSICAL LABORATORIES

Houseboat Works, Radlett, Hertfordshire.

Bestandteilskomparator

(Abbildungen auf Seite 109)

Dieser Induktanz-, Widerstand- Kapazitanzbestandteilskomparator Typ CZ.457/1 wird mit 100 kHz betrieben und ist eine Weiterentwicklung des Modells 457, das mit 1 kHz betrieben wird.

Dieses Gerät wurde eigens für die Prüfung von Kondensatoren, Widerständen und Induktoren unter Massenfertigungsbedingungen entworfen. Ohne zusätzliche mechanische Hilfsmittel lassen sich mit Leichtigkeit Prüfgeschwindigkeiten von 1 500 bis 2 000 Bestandteilen pro Stunde erzielen.

Das Gerät vergleicht den Prüfling mit einer Normeinheit und zeigt die prozentuelle Abweichung direkt auf einer 125 mm Skala an. Plus- oder Minustoleranzen von 0,5% bis 25% werden deutlich angezeigt. Ausserdem können Kondensatoren und Induktoren auf Phasenwinkel geprüft werden, wobei die durch Vergleich mit dem Normbestandteil ermittelte Abweichung direkt auf der Skala des Messers angezeigt wird.

Ein besonderes Merkmal des Geräts besteht aus der automatischen Überlastschutzvorrichtung des Messers. Sowohl unter Arbeitsstrom als auch Kurzschlussbedingungen des Prüflings übersteigt der Ausschlag des Messers mehr als 20% des Höchstwerts des Messerbereichs.

Das Gerät ist vollkommen in sich abgeschlossen, und Regelorgane zur Prüfung der Eichungsgenauigkeit sind eingebaut, so dass keine Extraskalen oder Richtigstellungswerte erforderlich sind.

Es kann Kondensatoren innerhalb des Bereichs von 20 pF bis 0,1 μF, und Induktoren von 20 μH bis 100 μH vergleichen, wobei die Genauigkeit bis zu 3% des angezeigten Werts beträgt.

EE 6754 *für weitere Einzelheiten*

THE BRITISH THOMSON-HOUSTON CO. LTD

Rugby, Warwickshire.

ÜBERLAGERUNGS-(ZF) ZENTIMETER-SPEKTROMETER

Dieses Spektrometer ist in erster Linie für die Analyse der Spektren von impulsgespeisten Oszillatoren bestimmt, durch Bildanzeige der Frequenzverteilung der HF-Kraft auf einer Kathodenstrahlröhre. Der Bereich der Impulsbreite liegt zwischen 0,1 μsec und 5 μsec. Ebenfalls für das Messen von Frequenzverzerrung, und von stehenden Wellenverhältnissen von Signalen mit sehr niederen Amplitudenwerten geeignet, besteht das Spektrometer aus einem Schmalband- ZF-Empfänger, dessen Ortsoszillator frequenzmässig vermittels einer an den Spiegel angelegten Sägezahnwellenform geregelt wird. Diese Wellenform wird ebenfalls an die X-Platten der Kathodenstrahlröhre angelegt.

Das Eingangssignal wird mittels eines HF-Dämpfers in den Mischer eingespeist, und der Ortsoszillator wird mittels der Sägezahnwellenform moduliert und gleichfalls in den Mischer eingespeist. Die sich ergebende Zwischenfrequenz wird verstärkt, demoduliert, und der Ausgang wird vermittels eines Bildverstärkers an die Y-Platten der Kathodenstrahlröhre angelegt. Der Zwischenfrequenzverstärker arbeitet bei einer Frequenz von 48 MHz und hat eine Bandbreite von 50 Hz. Ein kristallgesteuerter Oszillator ist mit eingeschlossen, um eine Frequenzeichung der Kathodenstrahlröhrenleuchtspur in Abständen von 1 MHz zu ergeben. Die Zentimeterbestandteile sind auf einem abnehmbaren Teil der Vorderplatte montiert. Auswechselbare Platten machen das Spektrometer für die am meisten verwendeten Grössen von Zentimeterwellenbändern geeignet.

EE 6755 *für weitere Einzelheiten*

CAWKELL RESEARCH & ELECTRONICS LTD

Scotts Road, Southall, Middlesex.

DAMPFUNGSKONSTANTENMESSER

(Abbildungen auf Seite 109)

Dieses Gerät ist ein empfindliches, zerstörungsfreies Hilfsmittel zum Erlangen von Einzelheiten über den inneren Zustand jeder Art von Maschinenbauwerkstoffen. Das Grundprinzip beruht auf der genauen Messung der Abklingzeit, der Dämpfungskonstanten, der logarithmischen oder 'Q' Abnahme der mechanischen Schwingung in dem der Prüfung unterzogenen Werkstoff zum Zwecke der Erlangung von Einzelheiten in Bezug auf seinen inneren Zustand, Materialfehler, ungleichmässiges Gefüge, chemische Zusammensetzung usw. Der Prüfling wird zu mechanischer Schwingung bei seiner Eigenfrequenz oder harmonisch verwandten Art erregt. Sobald ständige Oszillationen erreicht sind, wird die Erregung gesperrt und das Nachklingen der Oszillation mittels eines geeigneten Umsetzers 'beobachtet' und innerhalb eines vorbestimmten Zeitabstands gezählt. Die Zählung wird in Form einer vierstelligen Zahl auf Dekatronen dargestellt. Die neueste Ausführung dieses Geräts erlaubt Messungen zwischen 50 Hz und 50 kHz.

EE 6756 *für weitere Einzelheiten*

DAWE INSTRUMENTS LTD

99-101 Uxbridge Road, Ealing, London, W.5.

TRAGBARER TONOSZILLATOR

(Abbildungen auf Seite 109)

Der Tonoszillator Typ 421 ist für Batteriebetrieb geeignet und verwendet durchweg Transistoren. Die Frequenz ist kontinuierlich regelbar innerhalb eines Bereichs von 20 Hz bis 20 kHz, der in drei Dekadenstufen gedeckt ist.

Diese Eigenschaften machen ihn besonders für den Betriebsstelleneinsatz geeignet, z.B., für die Eichung von Geräuschpegelmessgeräten. Für kontinuierliche Werkbanksverwendung kann eine Netzkraftquelle eingebaut werden.

Ein neuartiger Gegentakt-Oszillator mit Wienscher Brückenschaltung wird zum Überwinden der Schwierigkeiten verwendet die gewöhnlich bei den in Transistorschaltungen anzutreffenden niederen Impedanzen auftreten. Jeder Gegentaktteil des Oszillators verwendet zwei geerdete Kollektorstufen und eine stabilisierte geerdete Emitterstufe. Die Amplitudenkonstanz wird mittels Verwendung eines neu entwickelten Thermistors mit geringer Kraft aufrechterhalten. Das Signal vom Oszillator wird in zwei Gegentakt-Transistorausgangspaare gespeist, von denen ein jedes der 'Emitter - Rechtecksmitläufer' - Typ ist, mit einer Ausgangsleistung von bis zu 2,5 V in 600Ω. Ein Ausgangsspannungsmesser ist eingebaut.

Die Genauigkeit beträgt ±2% und die Verzerrung weniger als 1% Der Ausgang ist bis zu 2,5 V in 600Ω regelbar.

EE 6757 *für weitere Einzelheiten*

THE DISTILLERS CO. LTD

172/3 Tottenham Court Road, London, W.1.

PANORAMISCHES TEMPERATUR-SCHAUBILDGERAT

Dieses Gerät zeigt die Ausgänge von bis zu 24 Thermoelementen in Form eines Profils auf einer Kathodenstrahlröhre an. Die 24 Thermoelemente werden der Reihe nach abgetastet, wobei eine senkrechte Verschiebung des Flecks die Temperatur beim jeweiligen Einschalten eines Thermoelements darstellt. Das Bild besteht auf diese Art aus einer Anzahl von senkrechten Linien, die in waagrechter Richtung in gleichen Abständen verteilt sind. Der Umriss dieses Schaubilds stellt die Temperaturverteilung an den Punkten dar, an denen sich die Thermoelemente befinden. Das Abtasten von 24 Thermoelementen nimmt 1,5 Sek in Anspruch.

Eine Vorrichtung zum Abdrosseln aller Eingänge um denselben Wert ist eingeschlossen, so dass eine kleine Schwankung um eine hohe Durchschnittstemperatur mit hoher Empfindlichkeit gezeigt werden kann. Diese Vorrichtung gestattet ebenfalls das Messen von Einzeltemperaturen. Ausserdem kann jeder Eingang einzeln abgedrosselt werden, so dass, wenn eine gewünschte Temperaturverteilung erzielt wird, der Effektivwert aller Eingange gleich Null ist, und sich ein Schaubild in Form einer geraden Linie ergibt. Die Empfindlichkeit kann dann erhöht werden, so dass jede Abweichung von dem gewünschten Zustand sich leicht feststellen lässt. Falls erforderlich, kann eine Warnanlage betätigt werden, wenn die Abweichung vorgewählte Grenzwerte überschreitet, und ein Netz von Signallampen zeigt die den Alarmzustand verursachenden Punkte an.

Obwohl dieses Gerät in erster Linie zur Verwendung mit Thermoelementen entworfen ist, kann jede Umsetzertype, die ein elektrisches Ausgangssignal liefert, verwendet werden.

EE 6758 *für weitere Einzelheiten*

EDWARDS HIGH VACUUM LTD

Manor Royal, Crawley, Sussex.

VORRICHTUNG ZUM ANBRINGEN VON ALUMINIUMUBERZUGEN IN KATHODENSTRAHLROHREN

Das Erzielen eines gleichmässigen Aluminiumüberzugs an der Rückseite der Schirme von 110° Kathodenstrahlröhren bereitet auf Grund der kurzen Länge des Kolbens und des kleinen Halsdurchmessers ein schwieriges Problem.

Eine besondere Vorrichtung wurde geschaffen, um die Gleichmässigkeit eines von einem einzelnen Wolframheizfaden niedergeschlagenen Aluminiumüberzug zu verbessern. Der Heizfaden ist auf einem Arm angebracht, der zwecks leichteren Einführens in den Hals der Röhre mit einer Scharniervorrichtung versehen ist. Der Arm wird mittels einer Feder in seiner Stellung gehalten, und wird im Innern der Röhre gedreht, wodurch auf der Innenseite ein sehr gleichmässiger Niederschlag gebildet wird. Der radiale Abstand und die Stellung der Quelle sind zwecks Unterbringung verschiedener Röhrengrössen verstellbar.

EE 6759 *für weitere Einzelheiten*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex.

TRAGBARESHSTRAHLUNGSDOSIMETER

(Abbildungen auf Seite 110)

Der Ekco Beta-Gamma-Überwachungsmesser Typ N596 ist ein neuer Kofferstrahlungsdosimeter, der eigens dafur entworfen wurde, den Vorschriften des in Kürze erscheinenden Betriebsgesetzes für Beta- und Gammastrahlung zu entsprechen. Bereiche von 0 bis 3, 0 bis 30 und 0 bis 300 mRad/h sind vorgesehen, zusammen mit einer Integrationsvorrichtung bis zu 30 mRad. Die Luftausgleichs-Ionisationskammer lässt sich leicht gegen eine Ersatzkammer der Grössenordnung eines Zehntel zur Messung von sehr örtlich begrenzten Dosenintensitäten austauschen. Das Gerät ist ausserdem zur Messung von 'weichen' Röntgenstrahlen geeignet, und seine praktische Unempfindlichkeit HF-Störungen gegenüber macht es besonders in Verbindung mit Hochspannungsröhren, Kathodenstrahlröhren usw. nützlich.

Der Niederaktivitäts - Universaldosimeter Typ N645 ist ein Leichtgewicht-Koffergerät zum Nachweis von sehr kleinen Mengen jeder radioaktiven Quelle. Es umfasst einen transistorisierten, batteriegespeisten Dosenmesser, der an einem Schulterriemen getragen wird, und mit einer Szintillationszählersonde verbunden ist, die in der Hand gehalten wird. Auswechselbare Leuchtstoffe sind für Alpha- und Betateilchen, Gammastrahlen, und langsame und schnelle Neutronen vorgesehen. Die Zählbereiche am Dosimeter betragen 0 bis 10, 0 bis 100 und 0 bis 1 000 Zählungen/Sek.

EE 6760 *für weitere Einzelheiten*

ELLIOTT BROTHERS (LONDON) LTD

Century Works, Lewisham, London, S.E.13.

HOCHSTFREQUENZ-PRUFOSZILLATOR

(Abbildungen auf Seite 110)

Der Höchstfrequenz Prüfoszillator Typ B.689 umfasst einen Koaxialoszillatorenhohlraum konstruiert für Betrieb von 7 bis 12 kMHz unter Verwendung eines CV2346 Klystronoszillators, eine stabilisierte Kraftquelle und einen Quadratwellenmodulator für 400 Hz bis 4 MHz. Die HF-Kraft wird mittels eines Koaxialsteckers entnommen, und eine ausreichende Kraftausgangsleistung wird innerhalb des Breitbands für alle normalen Mikrowellen-Werktischmessungen erzielt. Eine Frequenzskala mit direkter Ablesung ist eingebaut, die klare Anzeigen der Betriebsfrequenz gibt. Vorrichtungen für externe Modulationen sind ebenfalls vorhanden.

EE 6761 *für weitere Einzelheiten*

ENGLISH ELECTRIC VALVE CO. LTD

Chelmsford, Essex.

SPEICHERROHREN FUR TAGESLICHTSICHTEN

(Abbildungen auf Seite 110)

Die E702 hat elektrostatische Ablenkung und ist in erster Linie zur Verwendung in Oszilloskopen gedacht, wo das Bildsignal gespeichert werden muss. Da das Bildsignal ohne Verschlechterung bis zu zwei Minuten nachleuchtet, ist die Röhre gleich gut zum Zeigen von sich langsam ändernden Wellenformen oder schnellen Sprüngen geeignet. Die Höchstschreibgeschwindigkeit beträgt etwa 8 mm/μsec. Die Schreibkanone arbeitet bei 2,2 kV, wodurch die Ablenkkraft klein gestaltet wird. Der Bildschirm wird bei 10 kV betrieben, wodurch sich die sehr hohe Helligkeit ergibt, die ein Kennzeichen dieser Röhren ist und die Verwendung von Schutzblenden unnötig macht, selbst bei vollem Tageslicht.

Die E703 ist ebenfalls für Oszilloskope bestimmt, und besitzt zwei Schreibkanonen, wodurch gleichzeitiges Registrieren von zwei unabhängigen Wellenformen ermöglicht wird.

Die E701 verwendet magnetische Ablenkung, und ist für Radaranlagen entworfen, wo ihre hohe Helligkeit vorteilhaft in Sichtvorrichtungen verwendet werden kann, die in einem Lotsensitz oder auf der Brücke angebracht sind. Alle diese Röhren haben wirksame Schirmdurchmesser von mindestens 102 mm.

EE 6762 *für weitere Einzelheiten*

E.M.I. ELECTRONICS LTD

Hayes, Middlesex.

MAGNETISCHE SPEICHERTROMMELN

(Abbildungen auf Seite 110)

Diese Speicher, die für die Erfordernisse des Digitalrechenmaschinen - Ingenieurs geschaffen wurden, sind mit hochempfindlichen Magnetkopfstücken einzigartiger Konstruktion ausgestattet. Die Schaltkopfaggregate der gedruckten Schaltungen, die einen wesentlichen Teil der Trommelmontage bilden, sind leicht zugänglich.

Die Speicherkapazität beträgt 8 000 oder 16 000 Worte in einer Länge von 36 Stellen, auf 128 + 8 Reservespuren oder 256 + 16 Reservespuren. Fünf Uhr- und Reservespuren sind vorhanden. Die ineinandergreifenden Kopfstücke bestehen aus 16 Ablese-Schreibköpfen und einem Reservekopf.

Alle Köpfe sind zum Ablesen und Schreiben geeignet. Die fünf Uhrköpfe sind für gleichzeitiges Ablesen und Schreiben geeignet. Alle Köpfe besitzen Hochleistungsabschirmung gegen Aussenstörung.

Der Durchmesser der Trommeln beträgt 203 mm. Die Länge der Trommeloberfläche auf der 8 000 Worttrommel beträgt ungefähr 178 mm, und auf der 16 000 Worttrommel 324 mm. Die Mindeststandzeit wird bei Dauerbetrieb als 20 000 Stunden bei einer Geschwindigkeit von 3 000 U/Min. unter normalen Bedingungen angegeben.

EE 6763 *für weitere Einzelheiten*

FERRANTI LTD

Hollinwood, Lancashire.

WATTLOSE TRANSISTORREGULATOREN

Diese Regulatoren verwenden einen oder mehrere Transistoren in Parallelschaltung, die in einem Schaltkreis mit einer Diode und einem Drosselkondensatorfilter als Schalter wirken. In dieser Schaltung ist die an den Eingang des Drosselkondensatorfilters angelegte Spannung abwechselnd die Speisespannung oder null, je nachdem der Transistor ein- oder ausgeschaltet wird. Der Gleichstromsausgang von dem Filter entspricht dem Durchschnittswert der Eingangswellenform. Eine Steuerschleife vergleicht die Ausgangsspannung mit einem Beziehungspotential, wie z.B. einer Zenerdiode, und gleicht automatisch das Spur/Raum-Verhältnis des Signals dem Schalter an, um auf diese Weise den Ausgang in Bezug auf Schwankungen der Eingangsspannung und Last zu stabilisieren. Leistungen einer Wirksamkeit von 90% werden erzielt, und in einem typischen Aggregat, in dem eine einfache Schaltung verwendet wird, wird die Eingangsspannungsschwankung einhundertfach vermindert.

EE 6764 *für weitere Einzelheiten*

THE GENERAL ELECTRIC CO. LTD

Magnet House, Kingsway, London, W.C.2.

INDUKTIONSDIGITALISATOR

Dies ist ein vollständig neuer Umsetzertyp zur Verwendung mit Auswertungs- und Rechensystemen für Digitalinformationen. Er liefert eine nummernmässige Darstellung der Stellung eines Schafts oder Schiene in Form von verschlüsselten elektrischen Signalen ohne die Verwendung von Schleifringen, Stromwendern oder Photozellen.

Er ist im Grunde genommen ein Transformator mit einer einzelnen Erregerwicklung und einer Anzahl von Sekundärwicklungen, die der Stellenanzahl entsprechen. Die Stellung einer beweglichen Brücke bestimmt die Kopplungspolarität zwischen der Erregerwicklung und den verschiedenen Stellenwindungen, die entsprechend den Anforderungen des Schlüssels angeordnet sind. Impuls- oder Sinuswellenerregung kann verwendet werden.

Schaftdigitalisatoren können in grob/fein Kombinationen betrieben werden, wobei eine Zusatzwicklung Ungenauigkeiten zwischen den Digitalisatoren verhindert. Geräte für 5-Stellenablesung in einem Standard - Synchro - Gehäuse Grösse 11 und für 7-Stellenablesung in einem Gehäuse Grösse 23 wurden hergestellt.

EE 6765 *für weitere Einzelheiten*

GRIFFIN & GEORGE LTD

Ealing Road, Alperton, Wembley, Middlesex.

RC-BRUCKE

Die 'Griffin Panmetron' Brücke ist ein Netzgerät, das nicht nur als eine Wechselstrombrücke zum Vergleichen von Widerstand und Reaktanz bei 50 Hz unter Verwendung einer externen Norm dient, sondern ausserdem als hochgradiger Detektorverstärker zur Verwendung mit externen Brückenschaltungen.

Zur Betätigung der letzteren ist eine 2 und 4 V Wechselstromquelle eingebaut. Der unsymmetrische Strom wird an die Eingangsseite eines hochgradigen Verstärkers angelegt, unter Verwendung eines EM84 Abgleichanzeigers. Mit dem Steuerschalter in der höchsten Stellung wird eine Abgleichsanzeige mit einem Eingang von lediglich 0,1 μV erzielt.

EE 6766 *für weitere Einzelheiten*

ISOTOPE DEVELOPMENTS LTD

Beenham Grange, Aldermaston Wharf, near Reading, Berkshire.

SKALIERGERÄT

(*Abbildungen auf Seite 111*)

Der hauptsächliche neue Zusatz zu der I.D.L. Reihe ist das Skaliergerät 1700. Dasselbe umfasst einen vollständigen und automatischen Zählapparat in einem einzigen Tischplattenaggregat. Es benötigt lediglich den Zusatz eines Strahlungsdetektors, um jede Art von Radioaktivitätsmessung zu ermöglichen.

Bei der Konstruktion des Skaliergeräts Typ 1700 wurden neue Wege zur Lösung von Wartungsproblemen beschritten. Mit dem empfohlenen Satz von Ersatzteilen können Ausfallzeiten weitgehend vermindert werden, und Instandhaltungsarbeiten können selbst von ungeschultem Personal in allen Teilen der Welt rasch ausgeführt werden.

Abgesehen von dem Kraftaggregat umfasst das Skaliergerät Typ 1700 ein einfaches gedrucktes Schaltungssystem mit Einzeleinsteck-Aggregaten. Auf diese Weise kann jeder beliebige Teil der Schaltung augenblicklich gegen einen Ersatzteil ausgetauscht werden, falls eine Störung auftritt.

Vollkommener Zugang zum Innern ist gleichfalls in wenigen Sekunden mittels Schnellspann-Abdeckplatten möglich, die an der oberen und unteren Seite des Geräts angebracht sind.

Das Kraftaggregat ist senkrecht montiert und von allen Hauptzählschaltungen getrennt, während seine Chassis zwecks raschen und leichten Zugangs mit Scharnieren versehen ist.

In dem Gerät ist eine Schnellskaliervorrichtung eingebaut, und ausserdem ein elektronischer Zeitgeber; Vorrichtungen für 'vorgewählte Zeit' oder 'vorgewähltes Zählen' Arbeitsvorgänge; ein Geschwindigkeitsmesser; ein Einkanal-Impulshöhenanalysator; stabilisierte Speisehochspannung (250 V bis 2 kV); und ein linearer Impulsverstärker. Das Gerät lässt sich mittels Fernsteuerung betätigen. Vorrichtungen sind ausserdem zum Anschluss eines externen automatischen Schreibgeräts und eines externen Registrierapparats für die Zählgeschwindigkeitvorgesehen.

EE 6767 *für weitere Einzelheiten*

KELVIN & HUGHES LTD

New North Road, Barkingside, Essex.

'SYNCRO' SYSTEM MIT GERINGER REIBUNG

Das 'Syncrotel' ist ein Schrägstellungs-Transmissionselement der 'Synchro' Familie, aber seine ausserordentlich geringe Reibung und Trägheit ermöglichen seine Verwendung in Fällen, wo die zur Verfügung stehende Antriebskraft sehr klein ist, z.B. in kapselbetriebenen Mechanismen.

Der Stator ist ein hergebrachter Dreiphasen-Synchrotyp, aber an Stelle des gewöhnlichen Läufers wird eine ortsfeste Einphasenwicklung auf einer Spule verwendet, mit ihrer Axis zusammentreffend mit der Spindel. Der Drehteil besteht aus einem schrägen Abschnitt eines Hohlzylinders aus Aluminium, und seine Welle läuft in Juwelenlagern.

'Syncrotels' können direkt in paarweiser Anordnung-Rückseite an Rückseite-für gewisse Anwendungszwecke verwendet werden, aber sie werden gewöhnlich in Servosystemen mit einem gewöhnlichen Synchro-Aufnahmelement benützt.

Das Läufergewicht beträgt 0,76 g und das Trägheitsmoment 0,11 g-cm².

EE 6768 *für weitere Einzelheiten*

MUIRHEAD & CO. LTD

Beckenham, Kent.

HF-ANALYSENOSZILLATOR

(*Abbildungen auf Seite 111*)

Mit einem Frequenzbereich von 200 Hz bis 650 kHz erstreckt dieses neue Doppelzweckgerät den durch Muirhead Oszillatoren und Analysatoren gedeckten Frequenzbereich auf das Trägerfrequenzband für Telefonsysteme.

Durch Betätigen eines Schalters kann das Gerät entweder als ein empfindlicher Frequenzanalysator mit konstanter Verstärkung (mit einem 'Q'-Faktor von nicht weniger als 70, einer Oktavengleichrichtung von 40 dB und einer Mindestanzeige von 50 μV), oder als ein Oszillator (mit einem Ausgangspegel von 2,5 V in 600Ω verwendet werden). Bei jeder dieser Anwendungen beträgt die Frequenzgenauigkeit 2%.

EE 6769 *für weitere Einzelheiten*

MULLARD EQUIPMENT LTD

Mullard House, Torrington Place, London, W.C.1.

SCHNELLVERSCHLUSSLER

Dieses Gerät ist für die Schnellumwandlung einer Analog-Eingangsspannung in einen binären Ausgang von 10 'Bits' entworfen, die in Parallel-oder Folgedarstellung, und in rein binärer oder binär-dekadisch verschlüsselter Form zur Verfügung stehen. Die Auslesezeit beträgt weniger als 1 mSek, wobei Auslesungen beim Eintreffen von Eingangsimpulsenvorgenommen werden, die bei jeder beliebige Frequenz bis zu 1 000 pro Sekunde betragen. Der Analog-Eingangsspannungbereich liegt zwischen 0 und 10 V (positiv.). Für den Folgeausgang beträgt die Impulsausgangshöchstzahl ungefähr 12 kHz/Sek, oder aber sie kann gegen eine beliebige externe Impulsquelle gesperrt werden, vorausgesetzt dass die Impulsfolgefrequenz 12 kHz nicht überschreitet. Für die Parallelausgangsdarstellung ist die Impulszahl übereinstimmend mit dem nachfolgenden Eingangsausleseimpuls. Der Verschlüssler ist vollständig transistorisiert unter Verwendung von gedruckten Schaltungsplatinen, und besitzt einen hohen Grad von Zuverlässigkeit. Er eignet sich besonders für informationsverarbeitende Verfahren, wo grosse Mengen von Informationseinzelheiten in sehr kurzer Zeit verarbeitet werden müssen, wie z.B. beim Vornehmen von dynamischen Messungen an Raketenmotoren, oder beim Prüfen von Flugzeugaufbauten in Überschallwindtunnellen.

EE 6770 *für weitre Einzelheiten*

MEHRFACHKANAL-IMPULSSPEKTROMETER

(*Abbildungen auf Seite 112*)

Der L.332 ist ein 100-Kanal-Impulsspektrometer, der schnelle, genaue Analysen der Energiespektren von nukleonischen Strahlungen liefert. Impulse von einem Proportionaldetektor können gemäss ihrer relativen Amplituden gezählt und sortiert werden, und zwar in jeden beliebigen der 100 Kanäle, wo sie in einem magnetischen Matrix-Gedächtnissystem gespeichert werden. Da die Amplituden der durch den Proportionaldetektor erzeugten Impulse in direkter Beziehung zu der Energie der Einfallsstrahlung stehen, ergibt die in der Gedächtnisvorrichtung gespeicherte Information ein Staffelbild, das die relative Fülle der Strahlung an jedem der 100 Energiepegel innerhalb des untersuchten Spektrums zeigt. Diese Information wird auf einer 152 mm Kathodenstrahlröhre in Form eines kontinuierlichen, linearen, Analogbilds der spektralen Kurve gezeigt, wobei das Bild sowohl während als auch nach der Zählperiode zur Verfügung steht. Eine direkte Ablesung der in einem besonderen Kanal registrierten Zählungen steht ebenfalls zur Verfügung, und zwar durch hintereinandergeschaltete Neonnummernanzeiger. Die Ablesung ist zerstörungsfrei. Die Matrix-Gedächtnisvorrichtung enthält 100 × 16 Ferroxblock-Kerne, die jedem Kanal eine Speicherfähigkeit von 65 536 Zählungen vermittelt. Die Höchstzählgeschwindigkeit ist lediglich durch den Prozentsatz der durch die Totzeit verursachten verlorenen Zählungen begrenzt, wobei die Totzeit bei 250 μSek für angenommene Impulse konstant ist.

Der Grundeingangsbereich des Spektrometer beträgt 0,5 bis 10 V, wobei jeder Kanal eine Breite von ungefähr 0,1 V besitzt, und eine Rückgitterspannung bis zu 30 V kann angelegt werden, was einer Gesamtdeckung von 400 Kanälen gleichkommt. Kanalbreitengleichmässigkeit ist besser als 1% und die Stabilität der Kanalstellungen ist besser als 10mV über den Grundeingangsbereich. Grundzittern beträgt weniger als 3 mV.

Der L.332 ist von hoher Zuverlässigkeit und macht ausgedehnten Gebrauch von Transistoren. Er beruht auf einem Entwurf der A.E.R.E., Harwell.

EE 6771 *für weitere Einzelheiten*

G. V. PLANER LTD

Windmill Road, Sunbury-on-Thames, Middlesex.

'KUPFER AUF GLAS' GEDRUCKTE SCHALTUNGEN

(*Abbildungen auf Seite 112*)

Bei diesem Typ der gedruckten elektronischen Schaltung werden Kupferleiterbilder fest auf eine Glasgrundplatte aufgebracht. Die Vorteile dieses Systems sind Eignung für Betrieb bei hohen Temperaturen und der Abwesenheit von organischen Schichten oder Bindemitteln zu verdankende Trägheit. Die gezeigten Beispiele umfassten aus ultrasonisch geschnittenen Glasoschichten hergestellte Induktoren, von denen jeder einzelne auf beiden Seiten ein Kupferleiterbild aufweist. Die Platten lassen sich zwecks Herstellung von Mehrschichteninduktoren aufschichten, wobei Zwischenverbindungen über die Glaskanten hergestellt werden. Ein mittels dieses Verfahrens hergestellter Induktor wird in der zugehörigen Abbildung in ungefähr X2 Vergrösserung gezeigt.

EE 6772 *für weitere Einzelheiten*

W. G. PYE & CO. LTD

Granta Works, Cambridge.

PRAZISIONS-DEKADENPOTENTIOMETER

Dieses verfeinerte und höchst genaue Potentiometer wurde zu dem Zweck entworfen, Hochleistung ohne Wartung zu bieten, und Information in bequemster Form wiederzugeben. Dekadenschalten kommt durchwegs zur Verwendung und Schleifdrähte erübrigen sich. Die erste Dekade wurde mit dem Bereichsumstellschalter verbunden, wodurch das Gerät in allen Bereichen direkte Ablesung ergibt. Ein Gesamtspannungsbereich mit einer Reichweite von 1·99999 V bis auf $-0,1$ μV steht zur Verfügung. Eine genaue Dekadenstrom-Feinsteuerung ist eingeschlossen, die Stromschwankungen in Stufen von 1 Teil in 10 bis herab zu 1 Teil in 10^5 ergibt. Schaltvorrichtungen sind vorhanden, die das Wählen einer jeder beliebigen der fünf externen Schaltungen, und die Übertragung der Normzelle von den Korrektionsschaltungen auf den Spannungsteiler für Eichzwecke ermöglicht. Ein Systemumkehrschalter ist angebracht, um etwaigen wilden Effektivströmen (weniger als 0,1 μV), die im Galvanometer oder der Ausgangsschaltung auftreten können, Rechnung zu tragen.

Die drei Bereiche sind wie folgt:

—10 μV bis 1·99999 V in Stufen von 10 μV

—1 μV bis 0,199999 V in Stufen von 1 μV

—0,1 μV bis 0,0199999 V in Stufen von 0,1 μV

Die Genauigkeit der Spannungsteilung beträgt bis zu $\pm$0,0001%, und diese Genauigkeit wird durch Vergleich zweier Effektivströme erzielt. Die absolute Messgenauigkeit beträgt bis zu $\pm$0,002%.

Die Normzellenkorrektion deckt einen Temperaturbereich von 10°C bis 37°C. Die kleinste Stufe beträgt 10 μV, was 0,25°C entspricht.

EE 6773 *für weitere Einzelheiten*

THE SOLARTRON ELECTRONIC GROUP LTD

Solartron Works, Thames Ditton, Surrey.

DOPPEL-SPUR OSZILLOGRAPH

(Abbildungen auf Seite 112)

Das CD711S. 2 ist eine Weiterentwicklung des CD711 und CD711S. Sowohl die kräftige Konstruktionsform als auch die Zuverlässigkeit wurden verbessert, um den Betriebsbedingungen zu genügen. Dem Gerät wurde die J.S.A.* Auszeichnung verliehen, und die ihm zugeordnete Kennummer ist CT414. Darüber hinaus wurde die interne Anordnung des Oszilloskops einer Neuordnung unterzogen, um verbesserte Zugänglichkeit zu ergeben.

Das Gerät verwendet eine Strahlschalt, mit 150 kHz Auslese-vorrichtlungen für niedere Zeitbasis-Abtastgeschwindigkeiten. Beide Senkrechtablenkkanäle haben dieselben Kennzeichen, mit einer Bandbreite von 7 MHz und einer Höchstempfindlichkeit von 3 mV/cm.

Eine geeichte 'Y'-Verschiebungssteuerung, die sich auf jeden der beiden Kanäle umschalten lässt, wird für die Amplitudeneichung mit einer Genauigkeit von $\pm$5% verwendet.

Die Zeitbasis hat einen Abtastgeschwindigkeitsbereich von 3 cm/μSek bis 0,3 cm/Sek.

Die Zeitbasiseichung kann gegen ein internes Kristallosillatoren und Spannungsteilersystem genormt werden.

* *J.S.A. = Joint Service Approval, eine Zustimmung der Regierung für den Geländeeinsatz dieser Geräte durch die Vereinigten Britischen Streitkräfte.*

EE 6774 *für weitere Einzelheiten*

FREQUENZGANG-MESS PLATZ

(Abbildungen auf Seite 113)

Der Frequenzgang-mess Platz Typ JY743 vermittelt eine Anzahl von besonderen Vorrichtungen für die dynamische Analyse sehr niederer Frequenzen.

Der Frequenzbereich umfasst 0,0001 H bis 100 Hz mit einer Frequenzgenauigkeit von $\pm$1,5% und einer Amplitudenstabilität besser als 1% über den gesamten Frequenzbereich. Quadratische- und Dreieckswellenausgänge sind zusätzlich zu Vier- und Dreiphasensinuswellenausgängen vorgesehen. Der Gleichstromgehalt der Sinuswellenausgänge beträgt weniger als 0,1% des Spitze-Spitze-Wechselstromausgangs und die nichtlineare Verzerrung beträgt weniger als 0,5%.

Für das Zünden, Anlaufen, Aufrechterhalten und Entladen des Oszillatorbetriebs bei Frequenzen unter 10 Hz sind Relais vorgesehen. Diese Relais sind für Handbetrieb eingerichtet, oder können durch externe Vorrichtungen, z.B. durch die 'Rechnen', 'Halten' und 'Umsetzen'-Schalter einer Rechen- oder Simulationsvorrichtung betrieben werden. Eine Vorrichtung für die digitale Überwachung der augenblicklichen Oszillatorspannung ist gleichfalls vorgesehen.

In dem Bildaggregat sind swei 127 mm Meter (Beziehungs und Quadratur) vorgesehen, die die Amplitude und das Vorzeichen der wahren und imaginären Komponenten anzeigen. Digitale Schreibvorrichtungen für das automatische Auftragen von Frequenz- und Amplitudenwerten sind ebenfalls vorhanden. Die Höchstempfindlichkeit beträgt 3 mV für Vollausschlag, und die harmonische Auslenkung ist grösser als 40 dB.

Die in dem Bildaggregat zur Verwendung kommende Sonderkonstruktion gewährleistet Anzeigenbeständigkeit ohne Rücksicht auf die Frequenz der Grundsignale. Die Verwendung von Integration für jede gegebene Messung hängt deshalb vom Geräuschgehalt ab. Glättungsvorrichtungen sind bei Zeitkonstanten von 0,1, 0,3 1, 3 und 10 Sekunden vorgesehen, und Einzelperiodenintegration ist für Frequenzen zwischen 1 Hz und 1 Periode pro Stunde vorgesehen. Ausserdem ist eine Vorrichtung für Handsteuerung der Integrationszeit vorhanden.

EE 6775 *für weitere Einzelheiten*

STANDARD TELEPHONES AND CABLES LTD

Connaught House, Aldwych, London, W.C.2.

'WEISSES RAUSCHEN' PRUFGERAT

(Abbildungen auf Seite 113)

Dieses Gerät ist zum Messen von Basisrauschen und Zwischenmodulationsrauschen in Mehrfachkanal-Fernsprechanlagen entworfen; mit einer Bandbreite von bis zu 4 MHz (16 Supergruppen) erzeugt es ein Grundrauschband mit einer etwa gleichmässigen Energieverteilung innerhalb des gewünschten Frequenzbands, das nach Wunsch 2, 4, 10 oder 16 Supergruppen entsprechen kann. Das Rauschen wird durch einen Satz von Filtern geführt, um die Rauschbandbreite festzustellen und innerhalb desselben ein schmales 'stilles' Band herzustellen, das ein belastetes System mit einem stillen Kanal simuliert. Dieses Signal kann einem Mehrfachkanal-Fernsprechsystem (unter Nichtbeanspruchungsbedingungen) zugeführt werden mit einem tatsächliche Beanspruchungsbelastung simulierendem Kraftpegel, der die Messung von Kreuzgesprächen und Rauschen in dem schmalen 'Stop'-Band am abgelegenen Ende des Systems ermöglicht. Rausch/Kraft-Verhältnisse bis zu 70 dB können gemessen werden, und der 'Rücken an Rücken' Betrieb des Generators und Empfängers erzeugt Rausch/Kraft-Verhältniszahlen von 75 dB oder mehr für jedes empfangene Frequenzband. Die absolute Mindestdichte des weissen Rauschens, die gemessen werden kann, beträgt -118 dBm/kHz, was dem $-114,5$ dBm/Kanal entspricht, wenn derselbe geräuschspannungsmässig mit einem C.C.I.F. (1951) Fernsprechbeschwerungsnetz beschwert wird. Das Gerät umfasst einen aus drei tragbaren Einheiten bestehenden Generator; die Abmessungen jeder Einheit sind 559 $\times$ 251 $\times$ 298 mm; und einem Empfänger, der aus zwei Einheiten derselben Grösse plus einer Einheit mit den Abmessungen 559 $\times$ 206 $\times$ 298 mm besteht. Sowohl der Generator als auch der Empfänger haben eigene eingebaute Netzwechselstromaggregate.

EE 6776 *für weitere Einzelheiten*

SUNVIC CONTROLS LTD

Crown House, Aldwych, London, W.C.2.

IMPULSHOHENANALYSATOR

(Abbildungen auf Seite 113)

Der neue Sunvic Impulshöhenanalysator kann sowohl in Konsolen als auch Ständerausführung geliefert werden.

Er ist weitgehend transistorisiert und

verwendet eine Ferritkernspeichervorrichtung mit 100 Kanälen, jeder mit einem Höchstfassungsvermögen von 65 535. Zählungen können in Form einer Kathodenstrahlröhren - Analogleuchtspur gezeigt werden, in welcher die waagrechte Ablenkung die Kanalnummer darstellt, und die senkrechte Ablenkung sich proportional zur Zählung verhält. Zur Anzeige in Dezimalform der in jedem beliebigen Kanal entweder durch Handbetrieb oder automatisch gewählten Gesamtimpulsfolge werden Dekatrone verwendet. Die Gesamtwerte lassen sich ausserdem automatisch drucken.

EE 6777 *für weitere Einzelheiten*

VENNER ELECTRONICS LTD

Kingston-By-Pass, New Malden, Surrey.

BAUSTEIN-TRANSISTORSTUFEN

(*Abbildungen auf Seite 113*)

Die neuesten Zusätze zu der Venner Reihe von Einsteckstufen umfassen den Kristalloszillatortyp TS25 der (gemäss dem verwendeten Kristall) bei Frequenzen zwischen 100 kHz und 1 MHz wirksam wird. Diese Einheit kann entweder zum Geben einer Sinuswelle oder eines Impulswellenformausgangs verwendet werden. Ein weiterer Zusatz zu der Kristalloszillatorenreihe ist der TS30, der ein NF-Oszillator zur Verwendung mit XY Biegungskristallen ist. Dieser Baustein deckt den Bereich von 5 kHz bis 100 kHz.

Ein weiterer neuer Artikel ist das binäre 1 MHz Aggregat, das in zwei Formen verfügbar ist, dem TS2A/HF, das für Untersetzungsteilungsketten geeignet ist, wo ein anderes Teilungsverhältnis als mehrfache Werte von 2 erforderlich ist, und dem TS2B/HF, das besonders zur Verwendung mit dem Mikrosekundentortyp TS36 entworfen wurde. Dieses Tor kann geöffnet oder geschlossen werden durch Regelung des Gleichstrompegels des Steuerstifts von −9 V bis −4 V vermittels einer 10 V Speisung, wobei diese Pegel bequem von dem binären Aggregat Typ TS2B/HF geliefert werden können. Zwei solche mit dem Mikrosekundentor verbundene Aggregate können zum Geben einer Verriegelungswirkung angeordnet werden, so dass der erste eintreffende Steuerimpuls das Tor öffnet, der zweite es schliesst, und sich ergebende Impulse keine weitere Wirkung ausüben.

Der Mikrosekunden-Impulsformer Typ TS27 ist zum Geben eines Impulses entworfen, der sich zum Triggern von binären und Dekaden-Einheiten aus einem weiten Wellenformbereich bei Frequenzen bis zu 1 MHz eignet. Der Ausgang besteht aus sich in positiver Richtung bewegenden Impulsen mit einer Amplitude von ungefähr 8 V, 0,4 µSek. Breite, und einer Anstiegzeit von weniger als 0,2 µSek bei 1 MHz. Es kann natürlich auch zum Betrieb des Mikrosekundentors verwendet werden.

Ebenfalls verfügbar ist das 1 MHz Dekadenaggregat Typ TS10/HF, das im Betrieb ähnlich dem Standarddekadenaggregat Typ TS10/5 ist, d.h. es betätigt einen Meter, der zwecks Anzeige der erzielten Zählung mit einer Eichung von 0-9 versehen ist. Es arbeitet bei jeder beliebigen Frequenz bis 1 MHz, wenn es beispielsweise von dem Mikrosekunden-Impulsformer Typ TS27 betrieben wird.

Das elektronische Umsetzaggregat Typ TS32 ist zur Verwendung mit jeder beliebigen binären oder Dekadenstufe bestimmt, und ist in der Lage, dieselben innerhalb von 4 µSek umzusetzen, wenn es durch einen sich in negativer Richtung bewegenden Impuls ausgelöst wird. Keine Kontakte werden benützt, da der Betrieb gänzlich elektronisch ist.

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Zusammenfassungen der Hauptartikel

Hohlleiterschalter und Abzweignetze. von J. W. Sutherland.

Zusammenfassung des Artikels auf Seite 64-68

Hohlleiterschalter, sowohl zylindrische als auch solche, die mittels Kreisblenden betätigt werden, ferner Gasentladungsschalter und Ferritschalter werden beschrieben und besprochen. Vergleiche der Eigenschaften jeder Type werden angestellt.

Die Eigenschaften der T-Kopplung im Hohlleiter, und einige ihrer Anwendungen werden ausführlich besprochen. Ein Vielfachabzweignetz, in dem gemischte Kopplungen (Hybriden) zur Verwendung kommen, wird beschrieben.

Die Verwendung von Dekatronen für die Impulsverteilung. von G. H. Stearman.

Zusammenfassung des Artikels auf Seite 69-71

Kaltkathodenselektorröhren werden häufig verwendet, wenn eine Quelle erforderlich ist, die an mehreren Ausgangsleitungen und in einer bestimmten Zeitfolge Impulse erzeugen kann. Die Anzahl der Ausgangsleitungen ist jedoch auf 12 beschränkt, und um diese Anzahl zu erhöhen, ist die Anwendung besonderer Methoden erforderlich. In diesem Artikel wird eine allgemein bekannte Lösung (die Diodenmatrix-Methode) erwähnt, und zwei weitere vorgeschlagen, die beide unter bestimmten Umständen von Vorteil sein können.

Transistorentzerrungsmodulator-Frequenzteiler. von F. Butler.

Zusammenfassung des Artikels auf Seite 72-75

Eine Beschreibung eines Standard-Frequenzgenerators wird hier gegeben, bestehend aus einem kristallgeregelten 100 kHz Oszillator mit zwei nachfolgenden Frequenzteilern, die durch Pufferverstärker Ausgangsleistungen von 10 kHz und 1 kHz ergeben. Die Frequenzteiler sind Entzerrungsmodulatortypen und zeichnen sich dadurch aus, dass jeder Betriebsspannungsausfall den Stillstand der Ausgangsleistung bewirkt, ohne jede Möglichkeit von freilaufendem ungesperrten Betrieb.

Transistoren kommen durchgehend zur Verwendung; es gibt keine vorgewählten oder verstellbaren Regelungen, und die sinusförmigen Ausgangsspannungen sind von geeignetem Niveau für direkte Verwendung ohne die Notwendigkeit von Zusatzgeräten.

Eine Sperroszillator-Teilerschaltung wird ebenfalls beschrieben. Sie ergibt eine Reihe von Frequenzen, die auf die 1 kHz—Quelle eingestellt sind, bis zu einer Mindestgrenze von 25 Hz.

Frequenzausgleich für abgestufte Hohlleiterumwandlungsabschnitte. von D. Wray.

Zusammenfassung des Artikels auf Seite 76-79

Die Banderweiterung von abgestuften Hohlleiterumwandlungsabschnitten wird häufig in der Weise unternommen, dass die Scheinwiderstandsverhältnisse bei jeder Stufe so angeordnet werden, dass sie eine vorbestimmte Folge einhalten (z.B. binomisch, nach der Tschebyscheff Methode, exponentiell). Die sich ergebende Leistung liegt gewöhnlich unter der zu erwartenden, da die eigentlichen Scheinwiderstandsverhältnisse selbst frequenzempfindlich sind, so dass die gewünschte Folge nur bei der Mittenfrequenz zutrifft. Dieser Artikel beschreibt, wie die erforderlichen Scheinwiderstandsverhältnisse durch Anderung sowohl der breiten als auch der engen Abmessungen bei jeder Stufe beinahe konstant für das ganze Betriebsband aufrecht erhalten werden können. Eine beachtliche Leistungsverbesserung wird erzielt.

Eine monostabile Schaltung unter Verwendung von Transistoren. von H. Stinton.

Zusammenfassung des Artikels auf Seite 80-81

Eine impulserzeugende Schaltung wird hier beschrieben, in der ein einfacher Impulswandler und ein Transistor zur Verwendung kommen, deren Kennzeichen beträchtlichen Spielraum haben können. Ein ungefähr lineares Ansteigen des Emitter und Kollektorstroms steht zur Verfügung, oder ein ungefähr konstanter Amplitudenspannungsimpuls. Die niedere Ausgangsimpedanz macht die Vorrichtung zu einem brauchbaren impulsbildenden Puffer.

Transistorengitterentwürfe unter zu Grunde Legung von wärmeinkrementalen Eigenschaften von L. M. Vallese

Zusammenfassung des Artikels auf Seite 88-93

Ausführliche Entwurfsformeln für die Konstruktion von Einstufengittervorspannungsnetzen sind hier gegeben. Die Anwendung dieser Formeln setzt die Kenntnis von Einzelheiten voraus, die mittels zweier wärmeinkrementaler Parameterr ermittelt werden, von denen der eine in Beziehung zu dI_{co}/dT, und der andere in Beziehung zum Anderungsgrad des Eingangsstroms steht. Theoretische Formeln und Versuchsverfahren zur Ermittlung dieser Eigenschaften werden ausführlich besprochen. Praktische Anwendungsbeispiele und Prüfversuche sind ebenfalls eingeschlossen.

Annäherungsmodelle für Ubertragungsleitungen und ihre Fehler. von. R. R. Vierhout.

Zusammenfassung des Artikels auf Seite 94-95

Für Versuchsarbeiten und Berechnungen ist es oft zweckmässiger, Übertragungsleitungen durch einfachere Modelle, wie z.B. T- und -Netzsysteme zu ersetzen. In diesem Artikel werden einfache Formeln entwickelt, die eine unmittelbare Anzeige der Grösse des Fehlers ergeben, der durch die Verwendung dieser Modelle bedingt ist.

Ein I MHz Transistordekadenzähler. von C. G. Bradshaw.

Zusammenfassung des Artikels auf Seite 96-97

Ein Transistordekadenzähler, der Eingangsimpulse bei einer Entzerrungshöchstfrequenz von 1.3 MHz aufnimmt, wird hier beschrieben. Er verwendet vier binäre Stufen mit einem Rückkoppler, der die möglichen sechzehn beständigen Stufen auf zehn vermindert. Diese Methode ist in der Impulszählertechnik weitgehend bekannt.

Eine charakteristische impulsbildende Eingangsschaltung, die die erforderlichen Impulse für den Zähler liefert, wird gezeigt.

Der für eine Nennleistung von 12 V entworfene Zähler arbeitet korrekt innerhalb des Bereichs von 9-15 V und bei einer Temperatur bis zu 55°C.

Bisher wurden etwa 150 zufriedenstellende Laboratoriumsbetriebsstunden erzielt.

Eine Einführung zu Matrixen und ihrer Verwendung in der Analyse von Transistorschaltungen. von J. S. Bell und K. Brewster.

Zusammenfassung des Artikels auf Seite 98-102

Dieser Artikel wurde für den Zweck geschrieben, Matrixmethoden in die Analyse von Transistorschaltungen einzuführen, und zwar in erster Linie für alle diejenigen, die sich zum ersten Mal mit diesem Gegenstand befassen. Die mathematische Annäherungsmethode wurde so praktisch als möglich gestaltet, da zu befürchten war, dass eine zu genaue Behandlung zur Verschleierung des Hauptpunktes—der Analyse von praktischen Schaltungen—führt.