ELECTRONIC ENGINEERING

VOL. 31 No. 371

JANUARY 1959

Commentary

THE European Common Market, as originally conceived by the Rome Treaty of March 1957, consisted of Western Germany, France, Italy, Holland, Belgium and Luxembourg, and it was proposed that these six countries should gradually establish themselves into a community or Common Market in a period of twelve to fifteen years, beginning on 1 January this year.

By 1970 to 1973 these countries, it was envisaged, would have completed the setting-up of a Customs Union when trade between them would be wholly free of import and export duties, quotas, subsidies and discriminations in international transport rates designed to favour the products of one country more than those of another.

The six countries forming the European Common Market would, in fact, be a single market—the second largest trading area in the world—surrounded by a single import applied equally to all imported goods, and acting as a single entity in their commercial dealings with the rest of the world in negotiating trade agreements, export policies, and so on.

It was obvious, of course, that Britain with her vital dependence on export trade could not afford to be cut off from this trading area and it was proposed that out of the European Common Market should be formed as soon as possible—and certainly before the end of the period set for the completion of the Common Market—a European Free Trade Area comprising, in addition to the Common Market countries, the United Kingdom, Austria, Denmark, Norway, Sweden and Switzerland.

Such a plan would offer the United Kingdom a closer association with Europe in the sphere of economies without abandoning her long-established trading and financial relationships with the Commonwealth.

This plan for a more closely integrated Europe, for all its boldness and imaginativeness, has not been easy to bring into being and the discussions which have been going on for so long, particularly between the United Kingdom and France, have yielded little in the way of positive results.

As far as this country is concerned, the plan was given a mixed reception at first, but further thought has led to the conclusion that the greater part of British industry would stand to gain by British participation in the Free Trade Area or, to put it in a less palatable way, the majority of Britain's industries would suffer if the Common Market only were set up.

The electronic industry of Britain is one which is likely to gain by the coming of the Free Trade Area, for it has attained a position of eminence, due in no small measure to the techniques which were evolved during the war. In the field of industrial electronics, as the recent Computer Exhibition has shown, the industry is more than holding its own and it should be able to meet with confidence the challenge of stiffer competition for European markets which the Free Trade Area imposes.

As a step in support of this, we are making one or two changes in the journal which we hope will commend themselves to our readers. Our increasing circulation in Europe leads us to believe that the electronic industries of Europe obtain much of their information on British activities from the pages of our journal.

That the journal is printed in the English language is no barrier.

The fact that the principal centres of development in electronics in the Free World have been, and continue to be, in Britain and America, have made a working knowledge of technical English a matter of necessity for the European electronic engineer.

We are proposing, however, starting with this issue, to print translations of the summaries of the main articles, and the 'New Electronic Equipment', in French and in German. For the convenience of the foreign reader, we are grouping the translations of the summaries and electronic equipment review and they will be found at the end of the English editorial section.

This arrangement has the additional advantage that the reader in English need not have his attention diverted from his main reading.

The translations will involve a matter of eight pages or so each month, which will be in addition to the usual number of pages given over to editorial matters and they will be so numbered and annotated that when the yearly volume is completed they can be bound in or omitted as desired.

The opportunity has been taken to introduce other changes in the appearance of the journal.

The contents page, which some readers have experienced difficulty in finding in its old position immediately preceding the main editorial section, has now been moved to the front of the journal.

At the same time, for reasons connected with the printing of the journal, the classified section of the advertisement pages has been moved to the rear.

These changes have involved us in a good deal of work but we think that they add considerably to the journal and we hope that they will meet with the approval of our readers.

A Microwave Frequency Standard

(Part 1)

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

This article describes an equipment which produces positively identified signals for calibration purposes in the frequency range 7kMc/s to 20kMc/s: Both limits may be extended.
The smallest intervals available are not greater than 0·03 per cent of the calibration frequency, the signals being synthesized from a low-frequency temperature-controlled crystal oscillator with a short term stability of better than 1 part in 10^7, which can be directly compared with an accurately known BBC frequency. Frequency division down to 50c/s gives several accurate low frequencies and enables a clock to be operated as a check on the long term stability of the oscillator.

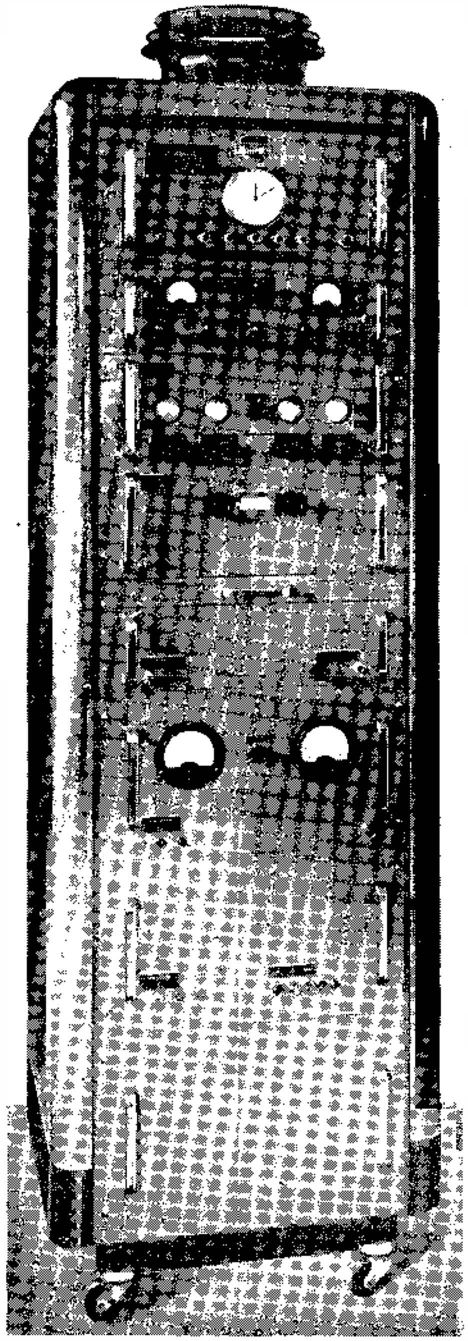
(Voir page 57 pour le résumé en Français: Zusammenfassung in deutscher Sprache auf Seite 61)

A MICROWAVE frequency standard is an essential requirement in any method for measuring unknown microwave frequencies and for accurate determination of cavity-frequency and Q; thus it forms part of the calibration service for wave-meters in frequency bands between 7kMc/s and 20kMc/s, in particular in X-band.

Most methods used to provide standard microwave signals can be considered as a mixture of two basic methods. In the first a stationary 'picket-fence' of frequencies all derived from a low frequency crystal by repeated harmonic multiplications and sideband injections is provided. In the other a few movable microwave frequencies are used; these are derived mainly by harmonic multiplication of the crystal frequency but in part from a stable low-frequency variable oscillator. Both methods employ a panoramic frequency display for comparison of standard frequency and unknown.

The advantage of the first method is that all frequencies obtained are known to the same accuracy as the crystal; the chief disadvantage is that unknown frequencies lying betwen those of the spectrum can be found only by interpolation, thus errors may be introduced. The second method eliminates this disadvantage but gives a signal of inherently slightly lower accuracy. It may be more useful however in cases where an isolated frequency is required rather than a group of harmonics.

The equipment described in this article uses a combination of the above methods. It provides a single output frequency variable in steps which are always harmonically related to the low frequency crystal, and hence the design difficulties associated with stable continuously variable oscillators are avoided.



The complete microwave frequency standard

The Principle of Operation

A block diagram of the equipment is shown in Fig. 1. A 100kc/s crystal oscillator feeds the frequency control system which generates a signal centred on 11Mc/s. The signal may be varied about this frequency, in small steps, by adjusting two calibrated controls, whose effects are additive. This signal is then multiplied to the region of 300Mc/s when the steps become 100kc/s. Further multiplication to the required microwave band is accomplished with crystal harmonic generators mounted in waveguide external to the apparatus. Thus a microwave harmonic will be produced approximately every 300Mc/s which can be moved by calibrated controls to overlap the range of movement of the adjacent harmonics.

The 100kc/s crystal oscillator also feeds a chain of dividers to produce accurately known low frequencies, Fig. 2, and can be visually checked against the carrier of the BBC Light Programme. The crystal is maintained in a commercially obtainable temperature-controlled oven.

Fig. 3 illustrates the frequency-control circuit which is based on several applications of the impulse governed oscillator (i.g.o.) principle in which an oscillator, harmonically or sub-harmonically related to some reference frequency, is locked to the correct harmonic by a frequency-controlling feedback loop. Frequency ratios of the order of 100 : 1 can be attained by this method, a tuning control enabling the ratio to be changed to 99 : 1, 98 : 1, etc. Two reference frequencies are used here, the basic 100kc/s itself, and 100/27kc/s derived from a subsidiary divider. Each is used to lock a higher frequency oscillator, and the sum of these two oscillators is taken as the output frequency at about 11Mc/s, it being possible to vary this frequency by variation of either of the components in their respective steps. Since the sum varies over

* *Mullard Ltd.*

a wider frequency range than the frequency of the lower of the two components, the use of fixed filters for the removal of unwanted products is impracticable, and to avoid using tunable filters for this purpose, a further locked oscillator system is employed.

To ensure that the i.g.o. multipliers are working at the multiplication ratios implied by the calibration, a system of calibration checking is employed whereby all the variable frequencies in the control circuit just described may be checked, some directly, others indirectly, against the crystal frequency. (See Fig. 4.) Visual frequency comparison is employed, the lower frequency providing a circular trace and the harmonic being applied to the grid of a small c.r.t., for brightening. The system also indicates the correct locking of the multiplier circuits during the operation of the equipment.

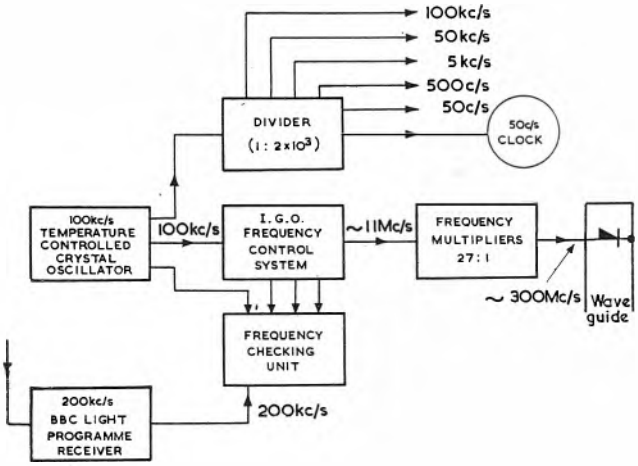


Fig. 1. Overall block diagram

Frequency Selection and Monitoring

The Impulse Governed Oscillator (I.G.O.) Principle

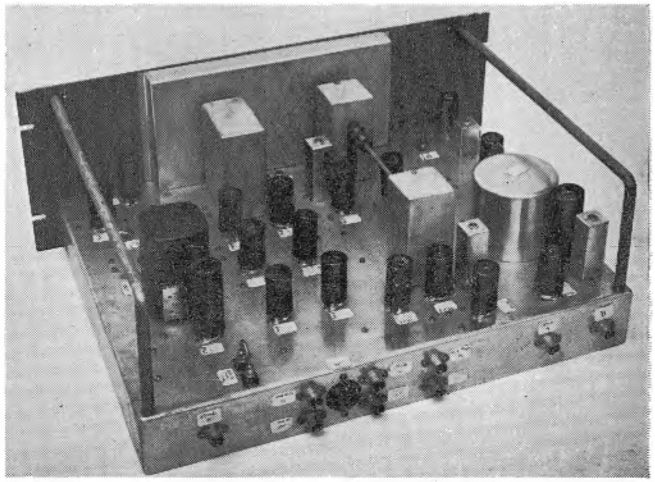
It is desired to lock a high frequency oscillator, frequency f_2, to a lower reference frequency, f_1. The input, f_1, is fed to a pulse generator which produces one pulse per cycle of f_1 with a pulse length of $1/2f_2$. The oscillator signal f_2 and this pulse are fed to an impulse mixer where every $(f_2/f_1)^{\text{th}}$ cycle of f_2 is compared in phase to one pulse. The mixer output is a pulse whose amplitude varies in accordance with the phase relationship. These pulses are passed to a peak reading detector and the resultant d.c. component and a.c. ripple fed through a low-pass filter, the output of which controls a reactance valve tuning the oscillator f_2, thus maintaining this oscillator locked to the reference frequency.

Care must be taken in the design of the low-pass filter in order to prevent modulation of f_2 by f_1 while keeping phase-shift small enough to prevent the control loop becoming unstable. If the condition obtains where f_2 is not locked to f_1, the signal from the mixer will have a varying amplitude at the beat rate between f_2 and the closest harmonic of f_1, and this will sweep f_2 in frequency until it approaches a locking position. The 'catching zone' is the frequency range within which f_2 will pull into a given lock, and the 'holding zone' the maximum manual tuning range possible without the oscillator pulling out of lock.

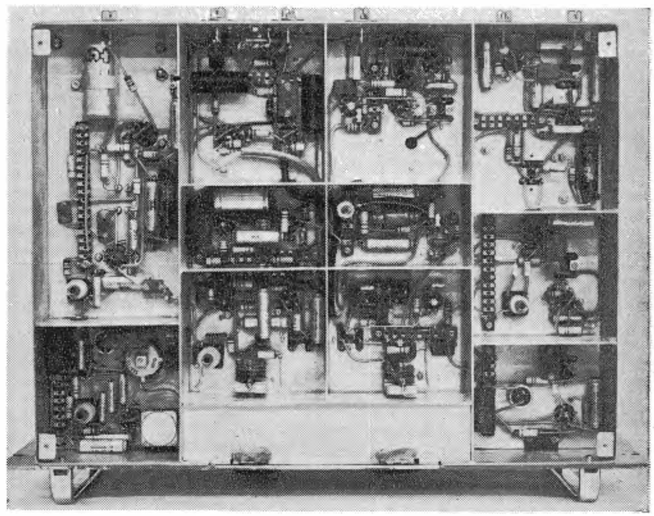
Method of Application

If intervals of approximately 3Mc/s are required at 10kMc/s then the intervals at the waveguide-crystal drive frequency of 300Mc/s must be 100kc/s. Hence at 11Mc/s, the input frequency to the chain of harmonic frequency multipliers, the intervals must be 100/27kc/s.

It is not feasible to use an oscillator at 11Mc/s controlled

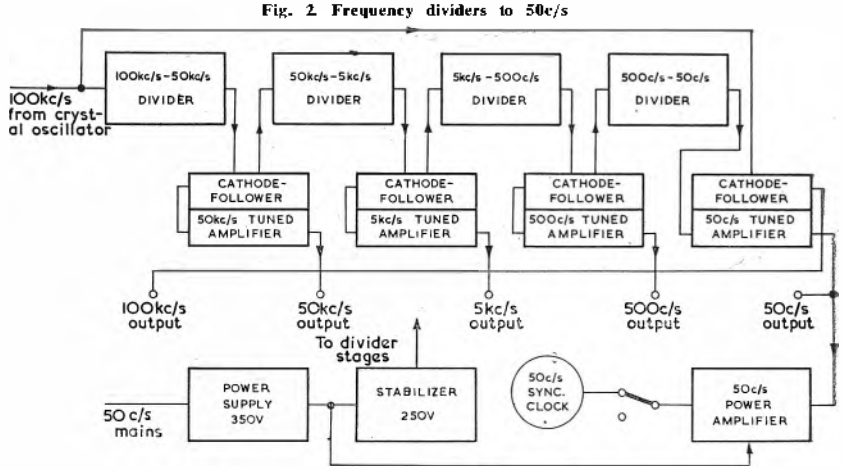


The frequency control chassis (top rear view)



The frequency control chassis (sub-chassis, with cover removed)

Fig. 2. Frequency dividers to 50c/s



by an i.g.o. circuit directly from a 100/27kc/s reference frequency to obtain such intervals, owing to the large multiplication ratio, about 3 000, which would be necessary. Instead two frequencies are added, one with the small intervals required and the other with 100kc/s steps. A 100/27kc/s oscillator is locked to the 100kc/s crystal oscillator by a simplified i.g.o. circuit, operating as a divider, in which the control loop corrects the lower frequency instead of the higher. An 11Mc/s oscillator tunable in steps of 100kc/s over a 400kc/s range is locked to the crystal frequency, and a 400kc/s oscillator tunable over 100kc/s in 100/27kc/s steps is locked to the i.g.o. divider.

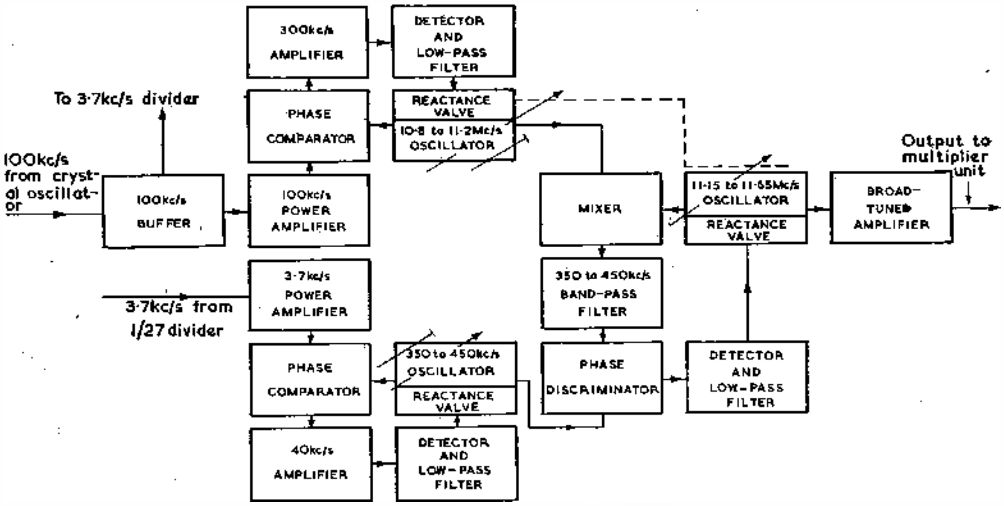


Fig. 3. Impulse governed oscillator frequency control chassis

The two frequencies are added in a slave oscillator circuit which eliminates the need for complex tunable filters. The principle of operation is illustrated in Fig. 3. The difference frequency between the slave oscillator and the 11Mc/s i.g.o. is compared in phase with the output of the 400kc/s i.g.o. and a correcting signal is applied to a reactance valve in the slave oscillator. The frequency of this oscillator is then variable, over the total frequency range required in 100kc/s steps, by variation of the 11Mc/s i.g.o. to which it is ganged, and in 100/27kc/s steps by tuning the 400kc/s i.g.o.

System of Checking

To ensure that during operation the i.g.o. circuits remain locked to the correct multiple or sub-multiple, a checking system is incorporated with monitor cathode-ray tubes using spot-wheel patterns (Fig. 4). The correct 1/27 divider ratio is permanently displayed and the 400kc/s and 11Mc/s i.g.o.'s can be checked against their individual reference frequencies, or the 400kc/s can be calibrated by checking it against 100kc/s, and then the 11Mc/s can be checked against the 400kc/s so obtained. Thus the output frequency can be checked at any time against the basic crystal reference frequency.

To compare two frequencies by the spot-wheel pattern method, the lower of the two is used to produce a circle by quadrature excitation of the X and Y plates of a c.r.t. while the higher is applied to the grid of the tube. Here the comparisons required are:

X and Y Plates	*Grid*	
3·7kc/s	100kc/s	to check divider
3·7kc/s	350 to 450kc/s	to check 400kc/s i.g.o.
100kc/s	10·8 to 11·2Mc/s	to check 11Mc/s i.g.o.
100kc/s	400kc/s	to check 400kc/s cal.
400kc/s	10·8Mc/s	to check 11Mc/s cal.

Crystal Frequency Check

The 100kc/s crystal is checked in the same way against the BBC's 200kc/s transmitter, a t.r.f. receiver amplifying the signal and producing a two-spot pattern with 100kc/s.

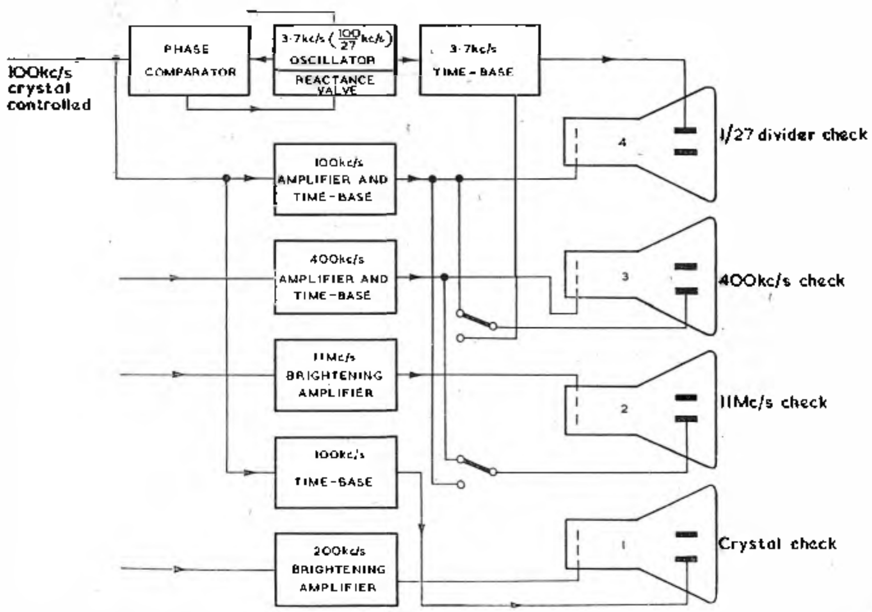


Fig. 4. 100/27kc/s divider and display unit

The direction and speed of rotation indicate the sign and amount of crystal frequency error; one cycle clockwise in 100sec is an error of 1 in 10^7 too high. Errors in the frequency of the 200kc/s transmitter are published monthly, enabling checks on the long term stability of the crystal oscillator to be made.

Circuit Design

Crystal Oscillator Unit

Oscillator and Buffer Stages

The master oscillator circuit is shown in Fig. 5. The crystal is a 5° X-cut bar by S.T. & C. maintained nominally at 47·5°C by a thermostatically-controlled oven.

The oscillator circuit is recommended by S.T. & C. for their crystal, and the capacitances in parallel with the crystal have been chosen to have as small a resultant temperature coefficient as possible. The available frequency change is approximately 2 parts in 10^5 with the 10pF trimmer provided. The cathode-follower buffer stage provides a r.m.s. output voltage of about 30V.

Oven and Control Unit (G.E.C. Type QC 242)

The oven contains a temperature-sensitive Wheatstone bridge which balances at 47·5°C, the supply being 50c/s. The bridge output is amplified by three triode stages, rectified, and used to control the oven heater supply via a relay in the anode of a fourth triode. An overheat protection device is incorporated should the control relay fail to operate. The stability claimed for this method of control is ±0·05°C.

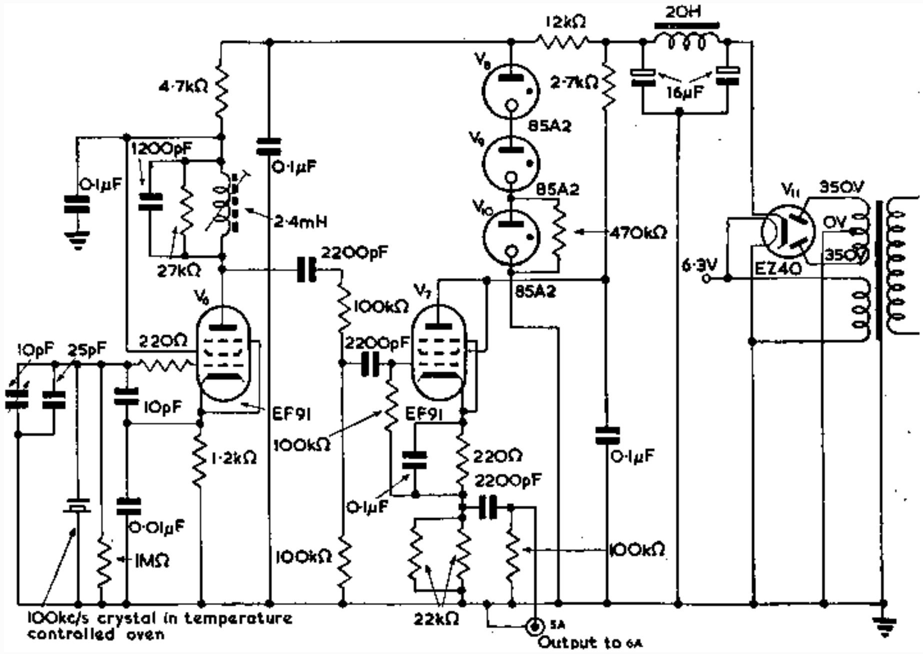


Fig. 5. 100kc/s master oscillator

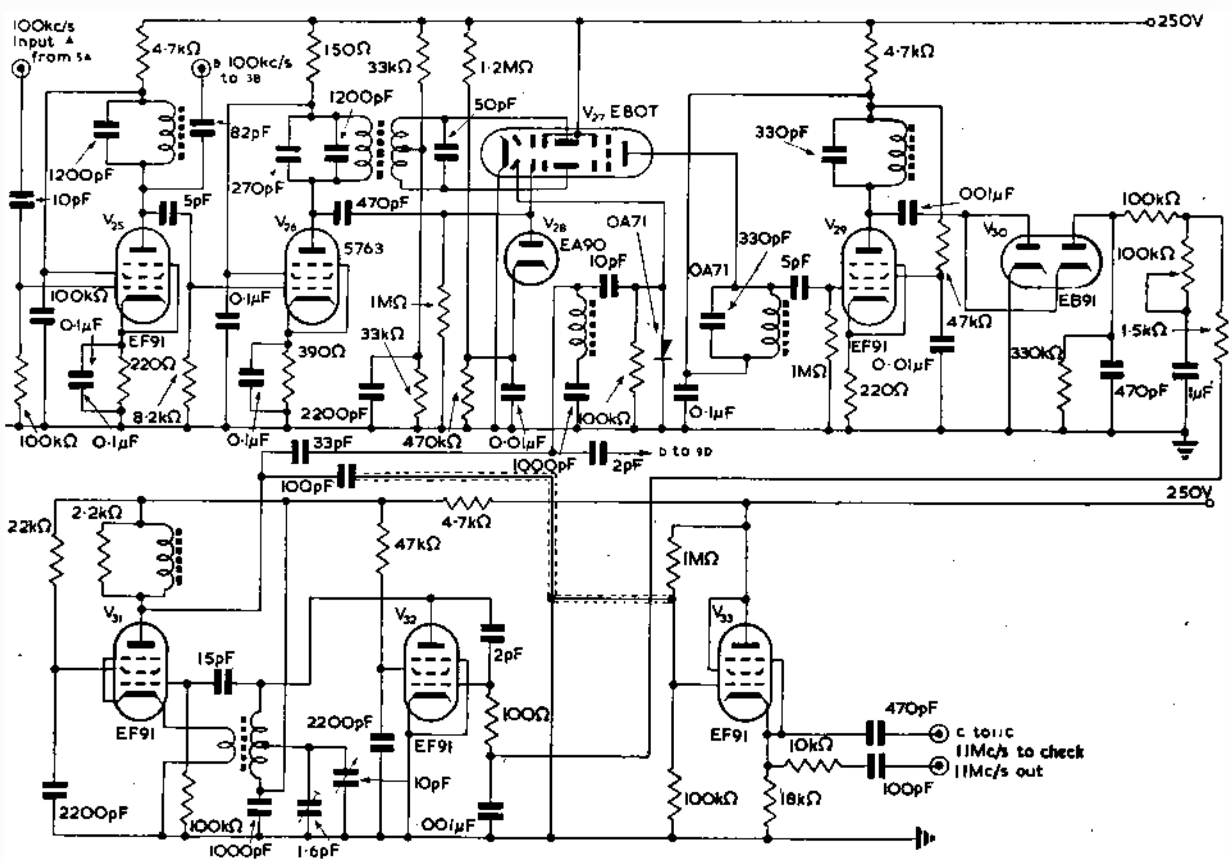
Frequency Selecting System

11Mc/s Multiplier

The 11Mc/s multiplier is located on the frequency control chassis; the circuit diagram is shown in Fig. 6.

V_{31} constitutes a cathode-coupled tuned-grid oscillator at 11Mc/s and V_{32} its associated reactance-valve. The oscillator frequency is adjusted by the left-hand front panel control and can be trimmed from inside the front panel compartment. V_{25} amplifies the signal arriving from the crystal oscillator unit and V_{26} is a further amplifier stage having a tuned transformer in the anode, the secondary of which produces 600V r.m.s. at 100kc/s, 90° out of phase with the primary voltage.

Fig. 6. 11Mc/s multiplier



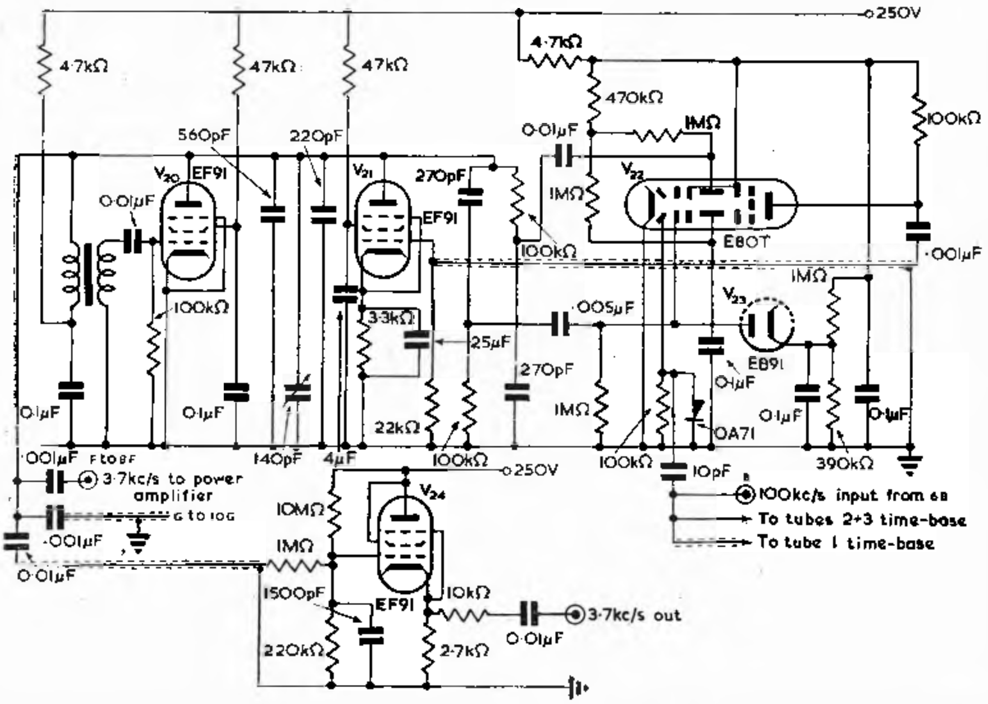


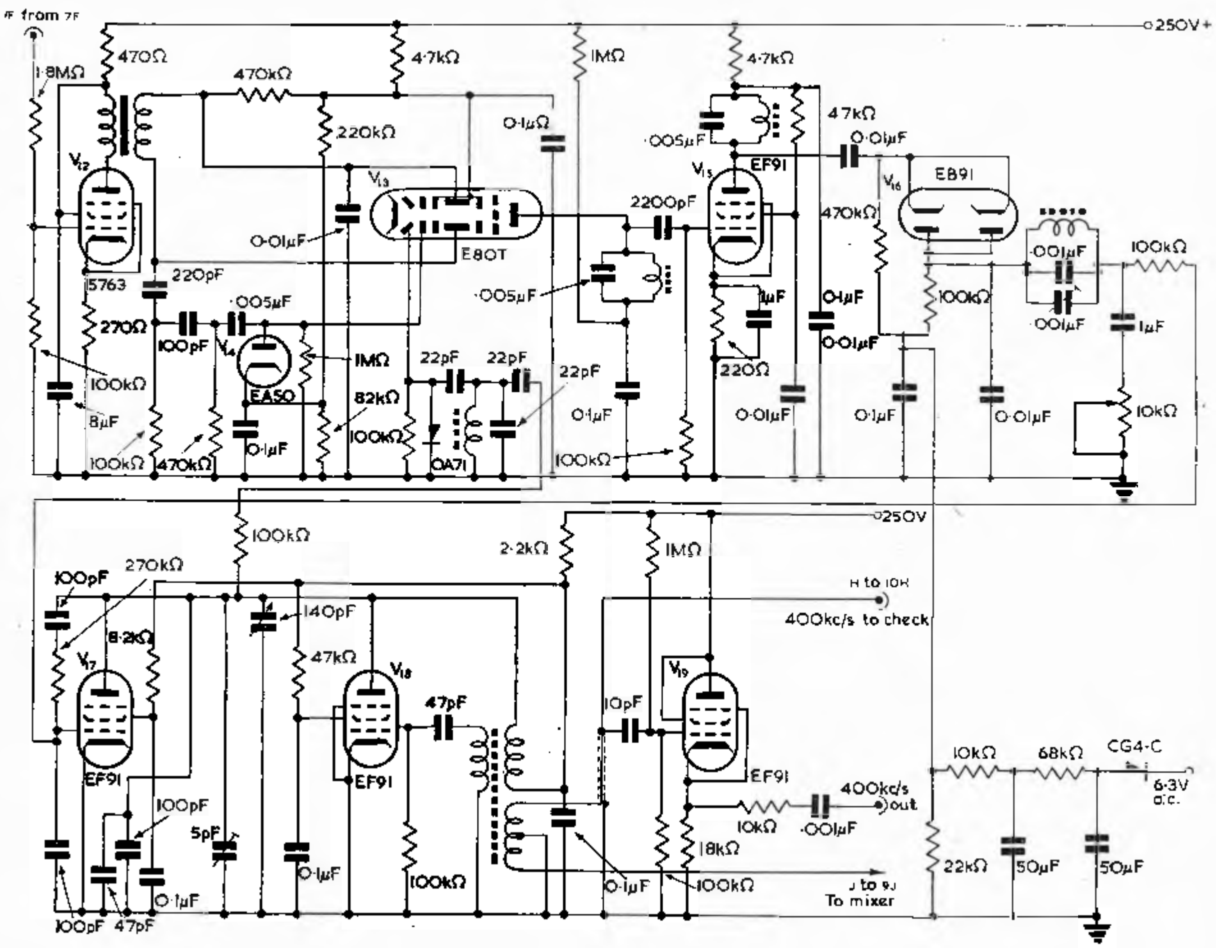
Fig. 7. 3·7kc/s divider

In V_{27}, type E80T, an electron beam is swept across a collector plate by the 600V, 100kc/s signal applied symmetrically to deflexion plates. One pulse of output current per cycle is produced as the beam passes an aperture and is collected by the final anode. The tube is gated by the quadrature anode voltage from V_{26} on g_3 to suppress the beam on the return sweep; this grid is held by the diode V_{28} to a maximum positive potential of 70V to focus the beam.

The beam current is also controlled by the 11Mc/s signal which is applied to g_1, so the output pulse amplitude from the collector will vary as the 11Mc/s moves in and out of phase with the 100kc/s sweep pulse.

V_{29} is a 300kc/s tuned amplifier which selects the third harmonic of the output from V_{27}; its output is rectified in V_{30} and fed via the low-pass filter to the reactance valve in the 11Mc/s oscillator. A band-stop filter is placed between the oscillator and g_1 of V_{27} to prevent any 100kc/s from

Fig. 8. 400kc/s multiplier



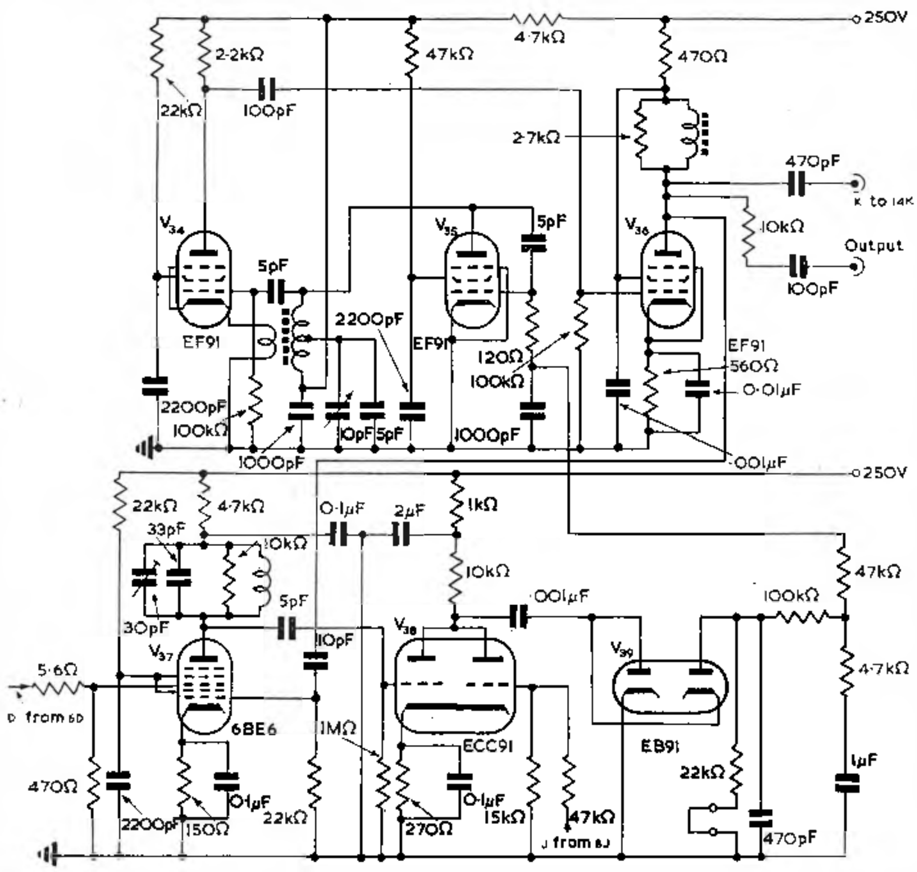


Fig. 9. 11Mc/s slave oscillator and mixer

modulating the oscillator by this path; a diode is connected between the same grid and earth to prevent this electrode taking any current, this being another possible source of modulation.

An output is taken via a cathode-follower V_{33} to a monitor socket.

1/27 Divider

The 1/27 divider is located on the frequency control chassis; the circuit diagram is shown in Fig. 7.

V_{20} and V_{21} comprise the 3·7kc/s oscillator and reactance valve; the frequency can be adjusted by a front panel trimmer over approximately 150c/s. From a *CR* network across the output quadrature signals are taken to V_{32}, an E80T; one is applied to the deflector plates and the other used to gate the return stroke of the beam on g_2. To g_1 is applied 100kc/s from V_{25} (Fig. 6), so the output from this beam-deflexion tube consists of pulses at a repetition rate of 3·7kc/s with a variation in amplitude suitable for controlling a reactance valve. By further retarding the phase of these pulses by the stray capacitance of the connecting lead they become of the correct phase to make V_{21} appear reactive when applied to its grid. Thus the oscillator frequency is controlled, and will lock when it is a sub-multiple of 100kc/s.

An output is taken via a cathode-follower V_{24} to a monitor socket.

400kc/s Multiplier

The 400kc/s multiplier is located on the frequency control chassis; the circuit diagram is shown in Fig. 8.

V_{18} and V_{17} comprise the oscillator and reactance-valve having a front-panel tuning range of 100kc/s from 350kc/s to 450kc/s. The calibration can be adjusted by a trimmer in the front panel compartment. Outputs are taken to the mixer and subsequent stages from a winding coupled to the tuned anode winding.

V_{12} is fed from the 3·7kc/s oscillator and is a power amplifier with a step-up output transformer producing 450V r.m.s. at 3·7kc/s sweep voltage for the E80T, V_{13}. This gives pulses of 1μsec duration at the anode of V_{13} every 270μsec, with amplitude variation due to the 400kc/s applied to g_1. A signal proportional to this amplitude is obtained by selecting the 11th harmonic of the repetition rate in a 40kc/s tuned circuit and amplifying it in V_{15}, the output of which is detected and passed via the low-pass filter to the grid of the reactance-valve V_{17}. The filter contains a parallel circuit resonant at 3·7kc/s to reduce modulation of the oscillator by this frequency, and an a.c. loop gain control for adjusting the loop to a stable operating condition. A small negative bias obtained by rectifying the 6·3V heater line is applied to V_{17} through the filter.

A high-pass filter is placed between the oscillator and g_1 of the E80T, and this grid is d.c. restored to earth potential; both measures are to prevent 3·7kc/s modulation of the oscillator.

An output is taken via a cathode-follower V_{19} to a monitor socket.

Slave Oscillator and Mixer

The slave oscillator and mixer are located on the frequency control chassis; the circuit diagram is shown in Fig. 9.

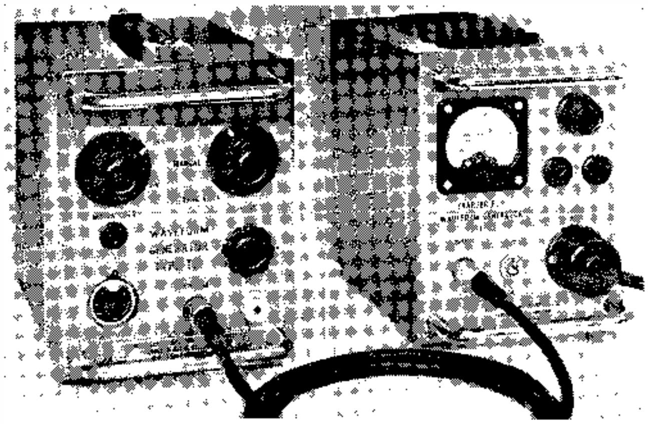
V_{34} and V_{35} constitute an oscillator and reactance valve similar to that of V_{31} and V_{32} in Fig. 6. The two oscillators have their tuning controls ganged and the slave oscillator is higher in frequency by an amount equal to the output from the 400kc/s multiplier. V_{36} is a broad-tuned amplifier centred on 11·4Mc/s, feeding the harmonic multiplier chain and a mixer stage V_{37}. The other signal to this mixer is from the 11Mc/s i.g.o. and their difference frequency is selected in the anode tuned circuit and applied to one grid of a double-triode, V_{38}. To the other grid is applied the 400kc/s i.g.o. signal at approximately the same level, thus beats between these two in the common anode load will have a maximum amplitude. This fluctuating signal is reproduced at the output of the detector V_{39} and applied as a control voltage to the reactance valve in the slave oscillator.

A link is provided to enable the load current of the detector, V_{39}, to be measured to ensure that the slave oscillator is in the centre of its electronic tuning range when the 400kc/s i.g.o. is in the centre of its manual tuning range.

(*To be continued*)

A Television Waveform Generator Using Transistors

By F. Rozner*, B.Sc.(Eng.)



This article describes a television waveform generator for use with transportable television broadcasting equipment. Very small size, weight and power consumption are achieved by the use of transistors and the complete elimination of thermionic devices throughout. The waveform generator uses digital techniques and, although this demands a large number of active elements, the resulting performance is, to a large degree, independent of the characteristics of individual transistors, h.t. variations, etc., and the stability and precision are adequate for broadcasting purposes.

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

THE instrument described in this article represents an attempt to reduce the size and weight of transportable television broadcasting equipment by transistorizing a significant part of it. The design was begun following the interest aroused when the experimental model of the Ferguson transistorized television pattern generator type T.1 was shown at the Television Society Exhibition in London, in March 1957.

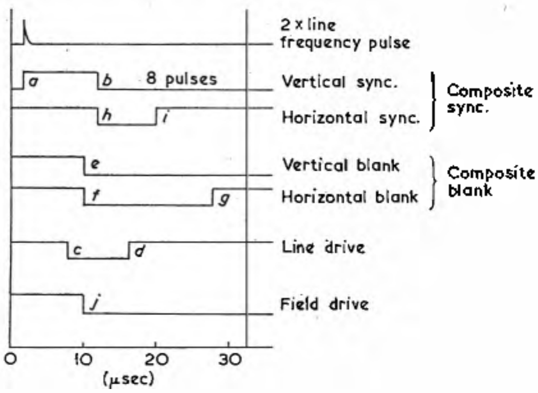


Fig. 1. Output waveforms based on a 32μsec timing cycle

The T.1 was conceived primarily for testing television receivers. Its output consisted of a grille picture pattern plus synchronizing waveform available either as a video signal or modulated r.f. carrier. A description of the instrument can be found elsewhere[1]. Although its performance conformed with the 405-line system specification[2] the precision of the waveforms was insufficient for broadcasting purposes. Consequently, when the requirement arose for a waveform generator for the British Broadcasting Corporation a fresh design approach was sought.

Logical Design

The BBC Specification required four outputs (Fig. 1):—

(a) Composite synchronizing waveform.

(b) Line drive.

(c) Field drive.

(d) Composite blanking waveform.

Each of these outputs is supplied by a separate output generator with two inputs for trigger pulses:—(a) ON-generator output negative, (b) OFF-generator output positive. The polarity of both trigger pulses was the same, viz., negative-going.

The implication here is that all the timing and mixing operations are carried out on narrow trigger pulses, not on the waveforms themselves.

The full cycle of waveforms (except the line drive) embraces two fields, i.e., about 40msec. It is obviously impracticable to provide a timing system with the required resolution over such a period, particularly as a great deal of redundancy would result. On the other hand the highest pulse repetition frequency occurs during the vertical synchronizing waveform. It is twice line frequency, i.e. 20·25kc/s.

Fig. 1 shows that, with the exception of the back edge of the field blanking pulses, it is possible to accommodate all the waveforms within 26μsec. It was decided, therefore, to provide a timing system with a period of about 50μsec (twice line frequency), and to use auxiliary circuits to gate the trigger pulses at appropriate times. The twice-line frequency was provided by a master oscillator running at 101·25kc/s, followed by a scale-of-five counter.

Turning again to Fig. 1 it will be seen that a resolution of about 2μsec is necessary. It is obtained from a 500kc/s timing oscillator which is keyed by the master oscillator. Fig. 2 shows the block arrangement. The timing oscillator drives four bistable circuits of the Eccles-Jordan type, connected in cascade. The chain of binaries forms a scale-of-sixteen counter the output pulse of which is fed back to the timing oscillator in order to stop it after 16 cycles of oscillation, i.e., when the binaries revert to their original state. The period covered is 32μsec of which 26μsec are useful.

Table 1 shows the state of the binaries during each of

* *Computer Developments, Ltd, formerly Ferguson Radio Corporation Ltd.*

The above illustration shows the waveform generator and the charging unit

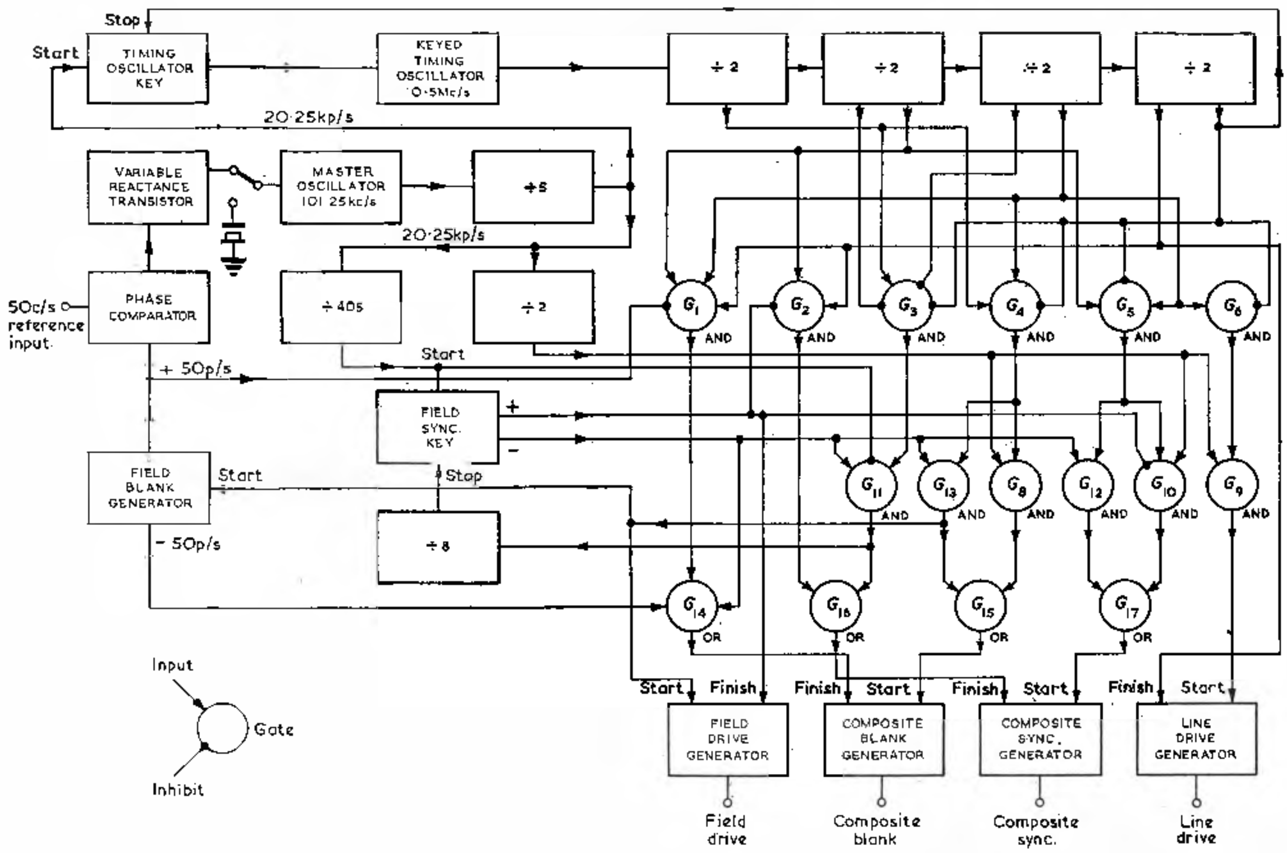


Fig. 2. Block diagram showing diode logic

the 16 cycles. In conjunction with Fig. 1 the state of the binaries for each of the triggering pulses can be determined. The actual selection is done in diode 'and' gates driven by the binaries (Fig. 2). Since the coincidence for a given wave edge, as specified in Table 1, can occur more than once during the cycle provision is made for the application of 'inhibit' pulses to the gates. In order to reduce the complexity and number of components it is not attempted to block all the unwanted pulses, because an 'on' trigger pulse following the wanted 'on' pulse will have no effect upon the output generator unless an 'off' pulse was applied in the meantime. For example, to obtain edge h (Fig. 1) the coincidence of binaries 2 and 3 is specified (Table 1). Such coincidence occurs at positions 6, 7, 14, 15 of which only number 6 is wanted. While number 7 cannot do any harm, it is necessary to block numbers 14 and 15 since they were preceded by the 'off' edge i. Hence an 'inhibit' pulse is applied from binary 4 to gate G_5 where the coincidence takes place (Fig. 2).

TABLE 1.

'And' gates for waveform edges shown in Fig. 1.

COUNT STATE	BINARY NO. 1	2	3	4	PULSE EDGE (FIG. 1)
0	0	0	0	0	
1	1	0	0	0	a
2	0	1	0	0	
3	1	1	0	0	
4	0	0	1	0	c
5	1	0	1	0	$e.f.j.$
6	0	1	1	0	$b.h.$
7	1	1	1	0	
8	0	0	0	1	d
9	1	0	0	1	
10	0	1	0	1	i
11	1	1	0	1	
12	0	0	1	1	
13	1	0	1	1	
14	0	1	1	1	g
15	1	1	1	1	
16	0	0	0	0	stop

The 'on' trigger pulses for the line drive, blank and synchronizing pulses are c, f and h (Fig. 1), selected in gates G_6, G_4 and G_5 respectively. As they occur at twice line frequency they are applied to a set of three coincidence gates G_9, G_8 and G_{10} which are also driven by a scale-of-two circuit dividing the master oscillator frequency. Coincidence occurs, therefore, at line frequency. For reasons already discussed no such operation is necessary upon the 'off' trigger pulses. It does not matter if two 'off' pulses follow each other so long as there is no intervening 'on' pulse. Hence, one output of binary 4 and the outputs of gates G_1 and G_2, which form the line drive, blank and synchronizing 'off' trigger pulses, d, g and i, are taken straight through to the mixing gates, or output generator in the case of line drive.

Composite Synchronizing Waveform

The vertical synchronizing signal consists of eight broad pulses[2] at twice line frequency. They follow the 50c/s output of the main divider chain. Referring to Fig. 1, b is the beginning of each broad pulse (coincident with h) and a is the end. The reason why a is shown to precede b is to accommodate both edges within the 32μsec. Edge a is generated in G_3 continuously.

In order to maintain the overall precision it is imperative that the beginning of the vertical synchronizing signal should be determined by the timing oscillator and not by the 50c/s pulse from the divider chain. The function of the latter is restricted to triggering the 'frame synchronizing key' which, in turn, opens certain gates and closes others. By opening G_{12}, b (coincident with h) is passed to the mixer G_{17} at twice line frequency. G_{11} is opened to pass on a from G_3 and at the same time G_2, normally emitting i, is closed.

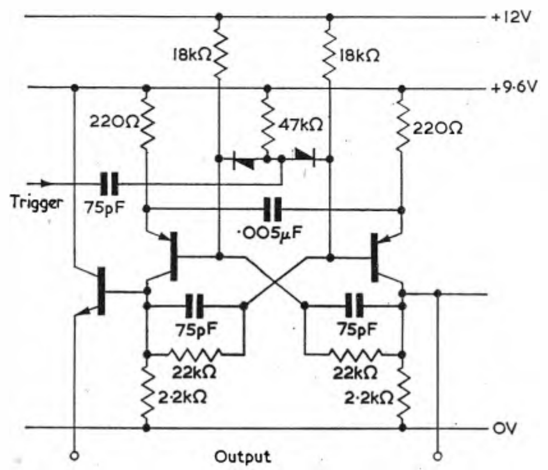


Fig. 3. Bistable circuit as used in the timing chain

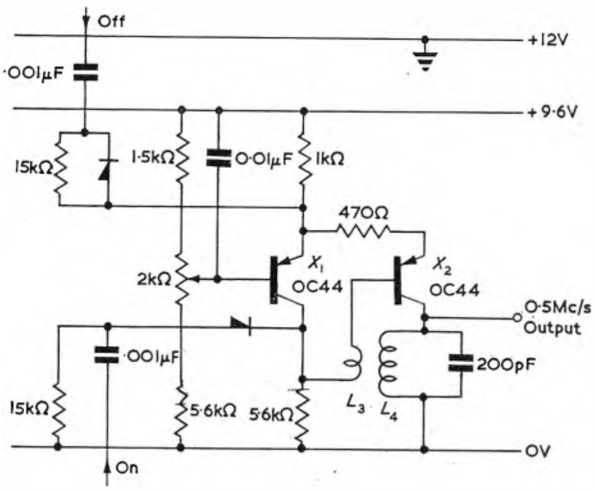


Fig. 4. Keyed timing oscillator

The output of G_{11} is also connected to a scale-of-eight circuit which counts off the eight broad pulses constituting the vertical synchronizing signal. The eighth pulse triggers the 'field synchronizing key' thereby terminating the vertical synchronizing signal. The mixing of the horizontal and vertical synchronizing waveforms is achieved in 'or' gates G_{17} and G_{16}.

The start of the vertical blanking pulse is also governed by the 'field synchronizing key' in a manner similar to the vertical synchronizing signal. By opening G_{13} the blanking pulse can begin at the appropriate time (in the middle of the line for the odd fields) since it is not inhibited by the scale-of-two circuit connected to G_9. The latter gate is not closed, the necessary inhibition being done on G_1 which normally emits the 'off' pulses for the horizontal blanking pulses.

The termination of the vertical blanking pulse is governed by the field blanking generator. The duration of the pulse is 14 lines and to define it digitally it would be necessary to count 28 half-line periods. Since the tolerance allowed is quite appreciable (± 1 line period) the incorporation of a counter was not considered justified. Consequently, the field blanking generator is the only timing circuit in the equipment which is based upon an *RC* network.

The Bistable Circuit

The cross-coupled bistable circuit of the Eccles-Jordan type is used throughout the equipment for counting, gate-setting, timing, etc. In Fig. 3 it is shown in the form used in the timing chain which is driven by the timing oscillator (Fig. 2). The conventional circuit is modified by the

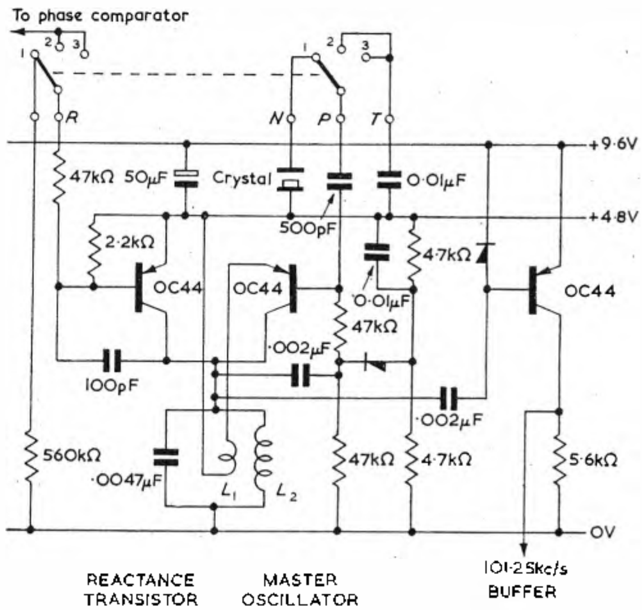


Fig. 5. Master oscillator with alternative crystal or reactance control

insertion of small resistors in series with the emitters which are cross-coupled by means of a capacitor. The aim of this modification is to improve the output waveform and trigger sensitivity. The drawback of transistorized Eccles-Jordan circuits is the integration of the 'off' edge of the collector waveform (negative-going for pnp transistors) due to the collector-base cross-coupling capacitor. Hence it is desirable to keep the value of these capacitors as low as possible and to compensate for the loss of positive feedback by cross-coupling the emitters.

The most sensitive trigger points are the bases of the transistors. However, it is usual to apply the trigger pulses to the collectors in order to provide firm routing for the former by utilizing the potential difference between the two collectors. By inserting resistors in the emitter circuits a potential difference of about 2V (for values given in Fig. 3) is introduced between the base of the saturated transistor and that of the cut-off one. Hence, trigger pulses can be applied to the bases resulting in high trigger sensitivity without abandoning the advantage of routing.

Timing Oscillator

It follows from the discussion earlier on that the keyed timing oscillator must be switched by a bistable device separately triggered for the 'on' and 'off' states. The arrangement is shown in Fig. 4, where both the oscillator and switch are incorporated in a single two-transistor circuit. The switch is a Schmidt type binary in which the two states are defined as follows:—

ON: X_1 saturated, X_2 conducting
OFF: X_1 cut off, X_2 saturated

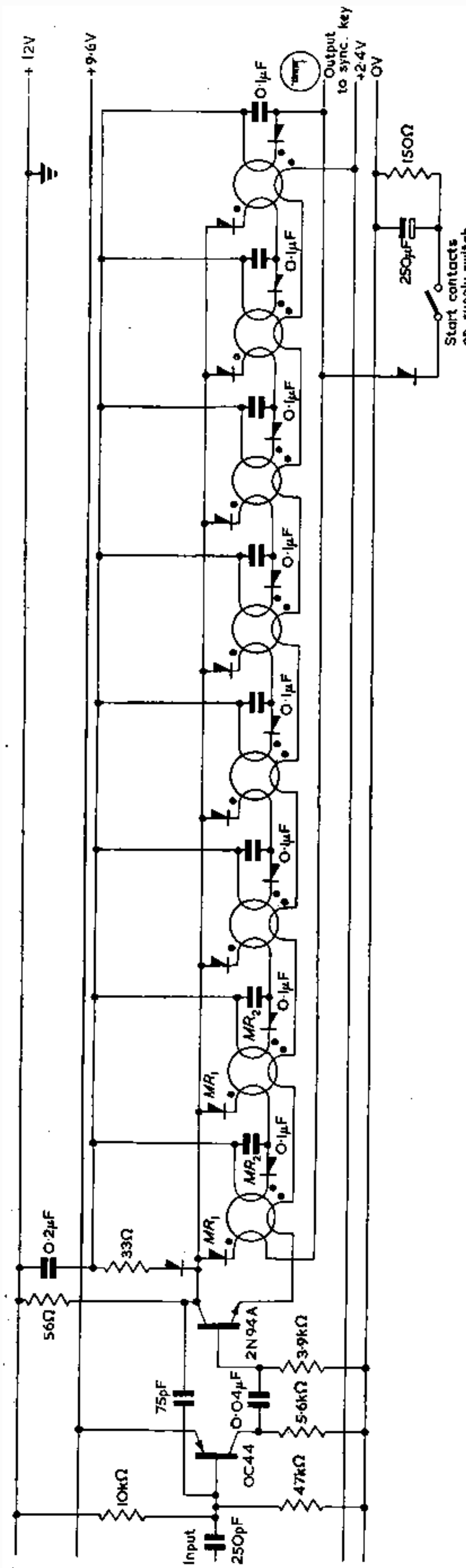


Fig. 6. Scale-of-eight counter using square loop ferrites

In the 'on' state X_2 oscillates at a frequency determined by the tank circuit in series with the collector. One end of the feedback winding in series with the base is effectively grounded because the collector of X_1 presents a low impedance when X_1 is saturated.

In the 'off' state, when X_2 is saturated, obviously no oscillation will take place.

The trigger requirements are rather heavy for two reasons. Firstly, in order to prevent the oscillation from changing the state of the binary, it is necessary to saturate X_1 rather heavily. Secondly, the changeover from the 'on' to the 'off' state has to be accomplished quickly enough to prevent unwanted cycles of oscillations from triggering the binaries that follow the oscillator.

Master Oscillator and Phase Comparator

The master oscillator (Fig. 5) is basically a single transistor, grounded base oscillator. The tank circuit is in series with the collector and feedback is taken to the emitter. For crystal control (switch position 1) the crystal is placed in series with the base. There is another transistor connected across the tuning capacitance, which acts as a reactance control. The operation is similar to the well-known a.f.c. arrangement employing thermionic valves. That is, the transistor takes a leading collector current the magnitude of which depends on the gain which can be controlled by base bias setting. With the crystal in circuit the reactance transistor has a constant bias.

When it is required to synchronize the master oscillator with mains (switch position 3) the bias of the reactance transistor is governed by the smoothed output of a conventional four-diode phase comparator bridge. The two inputs to the bridge are an amplified clipped mains waveform, which serves as the reference waveform and a pulse derived from the main divider output, which, fed to the bridge in push-pull, serves to open it at 20msec intervals. The control range of the reactance transistor is ± 5 per cent.

The switch has also an intermediate position 2. It is identical with position 3 except that the mains reference waveform is removed. In conjunction with the 'manual' control provided on the front panel this position can be used for setting up the d.c. conditions in the a.f.c. loop so that the master oscillator frequency is about nominal.

Scale-of-Eight Counter

This circuit is used to count off the eight pulses which form the vertical synchronizing signal. It deserves a mention, because the counting is done using small ferrite cores instead of better established methods. The circuit (Fig. 6) consists of a shift register connected as a ring counter using one core per bit of a type described elsewhere[3]. However, modifications had to be introduced to compensate for certain limitations arising out of the use of transistor drive. The core is a Mullard type FX.1508 2mm o.d. wound with 3 windings of 30 turns each.

There are two major modifications. Firstly, the discharge paths of the temporary storage capacitors lead through diodes MR_1 to the shift pulse generator. The diodes MR_1 are cut off during the shift pulse and conductive in the latter's absence.

It was found that the voltage transferred to the temporary store when a core switched, was insufficient to ensure reliable switching of the succeeding core in the register. Hence, the $0{\cdot}1\mu F$ capacitors were connected in series with a $0{\cdot}2\mu F$ capacitor which was charged by the shift pulse generator with a polarity which aided the temporary storage capacitors. The resultant output waveform is shown in Fig. 6.

For a shift register to act as a ring counter the contents of the register must be a '1', the rest being zeros. It was found that whenever the instrument was switched on or off, using a simple switching arrangement, the gradual build-up or decay of the shift pulses tended to eliminate this condition. In order to ensure the correct functioning of the counter it was found necessary, wherever the instrument was switched on, to apply automatically to one core a long and powerful 'start' pulse. In this starting arrangement use is made of the fact that transistors have no 'warm-up' time, and all generators and amplifiers reach full output as soon as circuit time-constants will allow it.

When not active the counter consumes virtually no current. Since it is counting for 0·4msec on a 20msec cycle, the average current consumption is negligible.

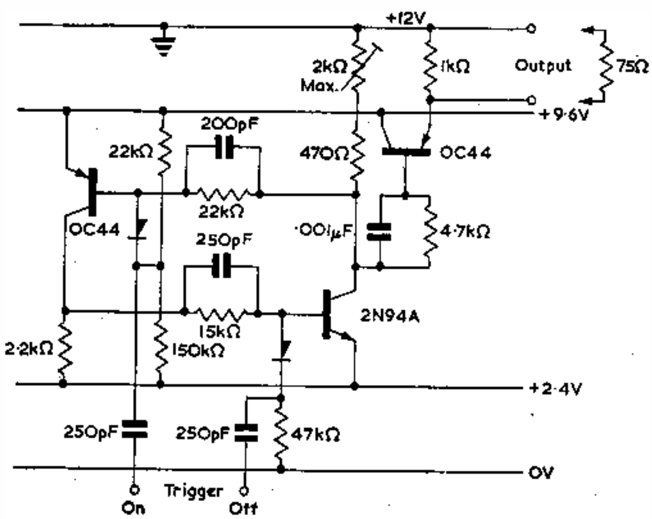


Fig. 7. Typical output generator with emitter-follower

Output Pulse Generators

The four output pulse generators are basically similar. The output requirements are 2V peak into 75Ω. The maximum mark-to-space ratio is that of the composite blank generator which is about 1:4. Even allowing for the fact that the output is taken from emitter-followers, the power per pulse delivered by the generators is still considerable. The pnp-npn circuit[4] was, therefore, judged superior to the more usual pnp-pnp (or npn-npn) particularly, as in view of the small mark-to-space ratio the average current consumption is greatly reduced. It was about 20mA for the four generators (plus the emitter-followers) as against 3·5 times as much for experimental pnp-pnp circuits giving the same performance.

It will be seen from the circuit shown in Fig. 7 that the generated pulse is clipped at the base of the emitter-follower. During the 'on' (negative-going) stroke the emitter-follower saturates and the coupling capacitor restores via the base collector diode. During the 'off' stroke the base potential rises considerably above that of the emitter. In these circumstances a certain amount of overshoot is inevitable but it was generally less than 5 per cent.

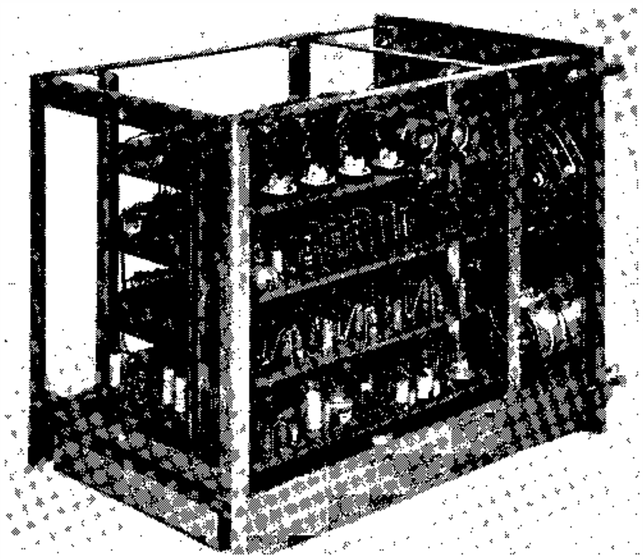
Power Supply

Ten rechargeable DEAC cells, 1·2V nominal each, form the power supply. The total power consumption of the instrument was 1W, but by fitting load equalizing resistors across some cells to ensure an even current drain the consumption went up to 100mA at 12V, i.e., 1·2W. The capacity of the cells is 900mAh giving 9 hours' operation. The cells can be recharged via a socket on the front panel, a small charger being supplied for the purpose. No h.t. stabilization is used.

The cells are held in a tray which plugs in at the bottom of the instrument. If an operating time longer than 9 hours away from base is envisaged a spare tray with cells can be used to replace the discharged cells.

Conclusions

The experience of computers shows that the use of digital techniques demands a large number of active elements, but the resulting performance is to a high degree independent of the characteristics of individual transistors, h.t. variation, etc. In this instrument a total of 73 transistors is used, of which 63 are Mullard type OC44 and 10 are Sylvania npn type 2N94A. The employment of *RC* networks for timing would reduce substantially the number of transistors required but it is doubtful if a similar stability and precision could be reached. As it is the performance is adequate for broadcasting. A prototype undergoing tests with the BBC has been used successfully for live transmissions from helicopters.



An internal view of the waveform generator

The weight of the instrument, including batteries, is 8lb and the size is 5¼in × 6¾in × 10in. The small dimensions are the result of using a light framework carrying four printed circuit boards. The small power consumption and high circuit efficiency enable close packing of components and eliminate all cooling problems. Access to components is gained by sliding out the boards.

Acknowledgments

A substantial part of the circuit design and development work was carried out by Mr. J. V. Corney of the Ferguson Radio Corporation Laboratories. Valuable assistance was rendered by other members of the staff of the laboratories, and thanks are due to the directors of the corporation for permission to publish this article.

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A Low Frequency Phasemeter

By N. Hambley, B.Sc.

The unit described in this article provided a quick and accurate measurement of the phase difference between two waveforms. The waveforms operate gates which allow electronic digital counters to display the phase difference in degrees. The phase measuring equipment requires an input level between 1V and 10V r.m.s. The frequency range is 1c/s to 100c/s but the upper limit can be extended, by a simple modification, to 3kc/s.

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

THE equipment described in this article has been developed particularly for use in the study of the vibration characteristics of aircraft structures. In this application the airframe is excited electromagnetically at several points. The resulting movements at other points are studied for differences in phase and amplitude. The same kind of study is also applied between the pilot's control column and the aircraft control surfaces. The mechanical vibrations are converted into electrical form by strain gauge transducers. These waveforms are basically sinusoidal, but square waves and pulses are equally acceptable inputs to the phasemeter. While the phasemeter is capable of working from 1c/s to 3kc/s, the application to aircraft structures is usually limited to an upper figure of 40c/s. At these very low frequencies oscilloscope displays of phase difference present some difficulties, and it is felt that this phasemeter offers one solution to the problem of phase measurement.

Basic Operation

The method of measurement is based upon a digital counting process which is controlled by electronic gates. These gates are, in turn, operated by the reference and phase delayed waveforms. With reference to Fig. 1 the reference input, at A, is fed into the logic gate and at a time T_1, opens the gate. This occurs when the waveform crosses the time axis in a positive going sense. The phase delayed waveform f', at B, also travels through the logic gate and closes it at a time T_2. This means that the logic and counting gates are open only during the period corresponding to the phase difference between the inputs at A and B. The input at E is arranged to be the frequency of the fundamental f multiplied by 360. If the phase difference between the waves at A and B is a quarter of a cycle, then the gates will only be open for this time period. During this period the counters will count up to 90 and then stop at that figure, indicating a phase difference of 90°.

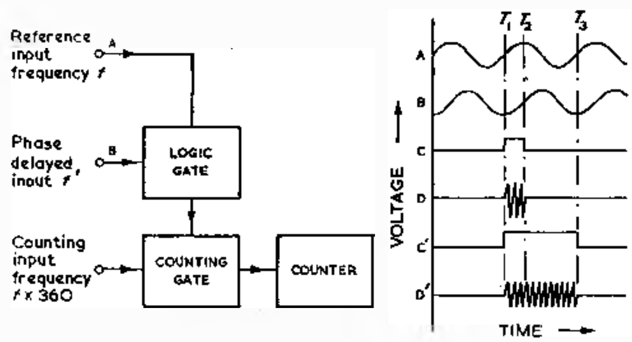
In order to check that the counting frequency is 360 times the reference f a switch is incorporated to open-circuit the input f' at B. The gates are then only dependent upon f which, at a time T_3, closes both gates after one complete cycle. The number then shown on the counters should be 360. The associated waveforms are shown at C′ and D′ in Fig. 1(b).

In order to achieve the required accuracy it is essential that the frequencies f and $f \times 360$ should be highly stable. It has been found that an Ediswan low frequency oscillator, type R666, has a short term stability of better than 0·5 per cent. This is suitable for energizing the electromagnetic exciters which cause the aircraft structure to vibrate at a frequency f. The reference and phase delayed waveforms are derived from strain gauges and amplified before being applied to the phasemeter at A and B respectively. The counting frequency $f \times 360$ is obtained from a Muirhead oscillator, type D650-B, which is accurate, in frequency to 0·1 per cent. The counters used are Racal type CU542A.

Gating System

With reference to Figs. 2 and 3, the input at A is squared before being applied to the logic gate at G. Also the phase delayed input at B is squared, limited and differentiated before being applied to the logic gate at F. The count input $f \times 360$ at E is squared before going to the counting gate at I. The output of the counting gate is taken to electronic counters at D. The action of the counting gate is controlled by the output from the logic gate at C. The logic gate must allow the counting gate to operate as follows:—

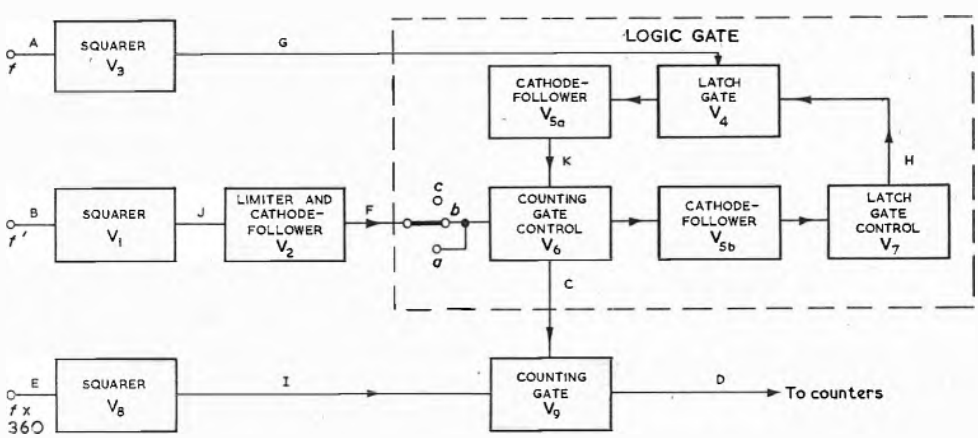
(1) To be first of all in a closed position.



(a) (b)

Fig. 1. (a) Simplified block diagram (b) Waveforms

Fig. 2. Detailed block diagram



(2) Then to be ready for being switched by the input waveforms.

(3) To be opened then by the first positive excursion of f, the input reference waveform.

(4) To be closed by the next positive excursion of f', the phase delayed input.

(5) To have one condition where the f' input is open-circuited, so that the gate is closed by the next positive excursion of f. This enables the setting $f \times 360$ to be checked.

(6) After (4) or (5) above, to remain closed until reset, so that the phase delays of successive cycles are not added together.

The combination of a three pole switch and two multivibrators satisfy the above requirements. The switch is

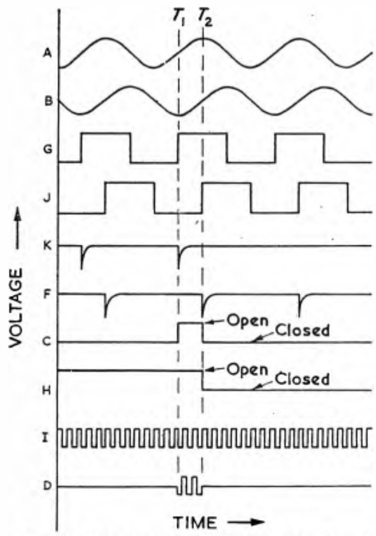


Fig. 3. Waveforms associated with Fig. 2

shown in position b in Fig. 2. The multivibrators are then connected to the reset voltage which opens the latch gate (H in Fig. 3), and closes the counting gate. The reset voltage holds the multivibrators in this position and they are not affected by the switching pulses at K and F. With the switch in position a the reset voltage is removed and the multivibrators are ready for being switched. The first following positive excursion by the reference input at A causes the pulse at K to switch over the counting gate multivibrator and the count gate opens. When this has occurred the next pulse at F switches the multivibrator back again to its original state and closes the counting gate. The switching action is also passed to the second multivibrator which closes the latch gate at time T_2. This prevents further counts occurring due to successive cycles of the input waveforms. When the switch is returned to b the count is returned to zero and the multivibrators are returned to their original states. In position c the phase delayed input at F is open-circuited. Now the first multivibrator is switched by two successive pulses at K. In this way the counting gate is open for exactly one cycle of the reference input, which enables the count to be checked at 360.

Circuit Details

Limiter Circuit

Fig. 4(a) shows the type of limiter circuit and Fig. 4(b) the associated waveforms. The positive half cycle of V_g is grid limited and the negative half cycle is limited by anode bottoming. In the complete circuit diagram, Fig. 7, valves V_2 and V_4 act as limiters, while V_4 is also the latch gate.

Multivibrator Circuit

A general form of the multivibrator[1] used is shown in Fig. 5(a), and the waveform in Fig. 5(b). When the grid resistor switch is open V_{g2} rises towards h.t. and V_{a2} is low.

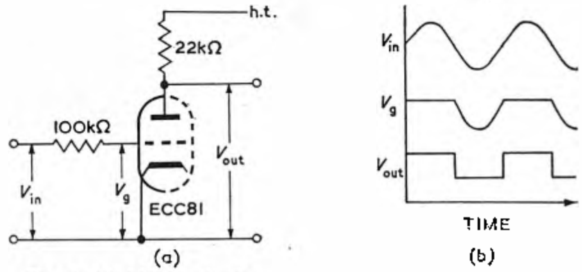


Fig. 4. (a) Limiter circuit (b) Limiter waveforms

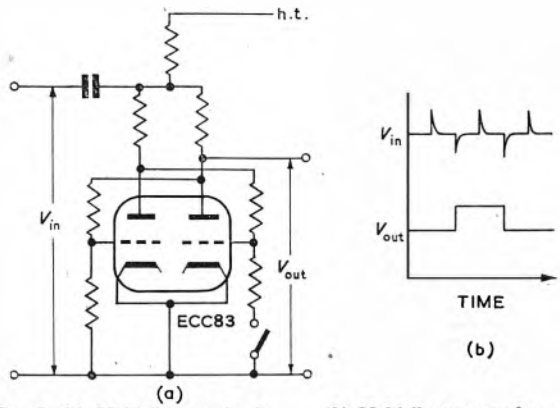


Fig. 5. (a) Multivibrator circuit (b) Multivibrator waveforms

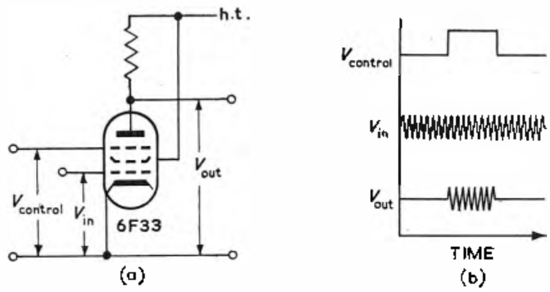


Fig. 6. (a) Gate circuit (b) Gate waveforms

While in this condition the circuit is insensitive to the switching pulses V_{in}. When the switch is closed, however, the multivibrator is switched by the first negative pulse injected into the anode resistor network. The phase delayed negative pulses are injected into g_1 and switch the multivibrator back into its original state. In this way the voltage V_{a2} is high only for a period corresponding to the phase difference between the two input waveforms. This voltage is used to control the counting gate. The multivibrator which operates the latch gate has only one input and that is into the anode nework. It is the negative going pulse from the first multivibrator which triggers the second one and closes the latch gate. This cuts off the reference input pulses and prevents any further action by the multivibrators until the circuit has been reset by the grid resistor switches.

The detailed circuit arrangements are shown in Fig. 7 where the first and second multivibrators are V_6 and V_7 respectively. The counting gate control voltage is taken from the second anode of V_6 and is preset to the correct level by a potentiometer. The latch gate control voltage is taken from the second anode of V_7 and is preset by its potentiometer. The reset switch for V_6 and V_7 is common with the reset line for the electronic counters.

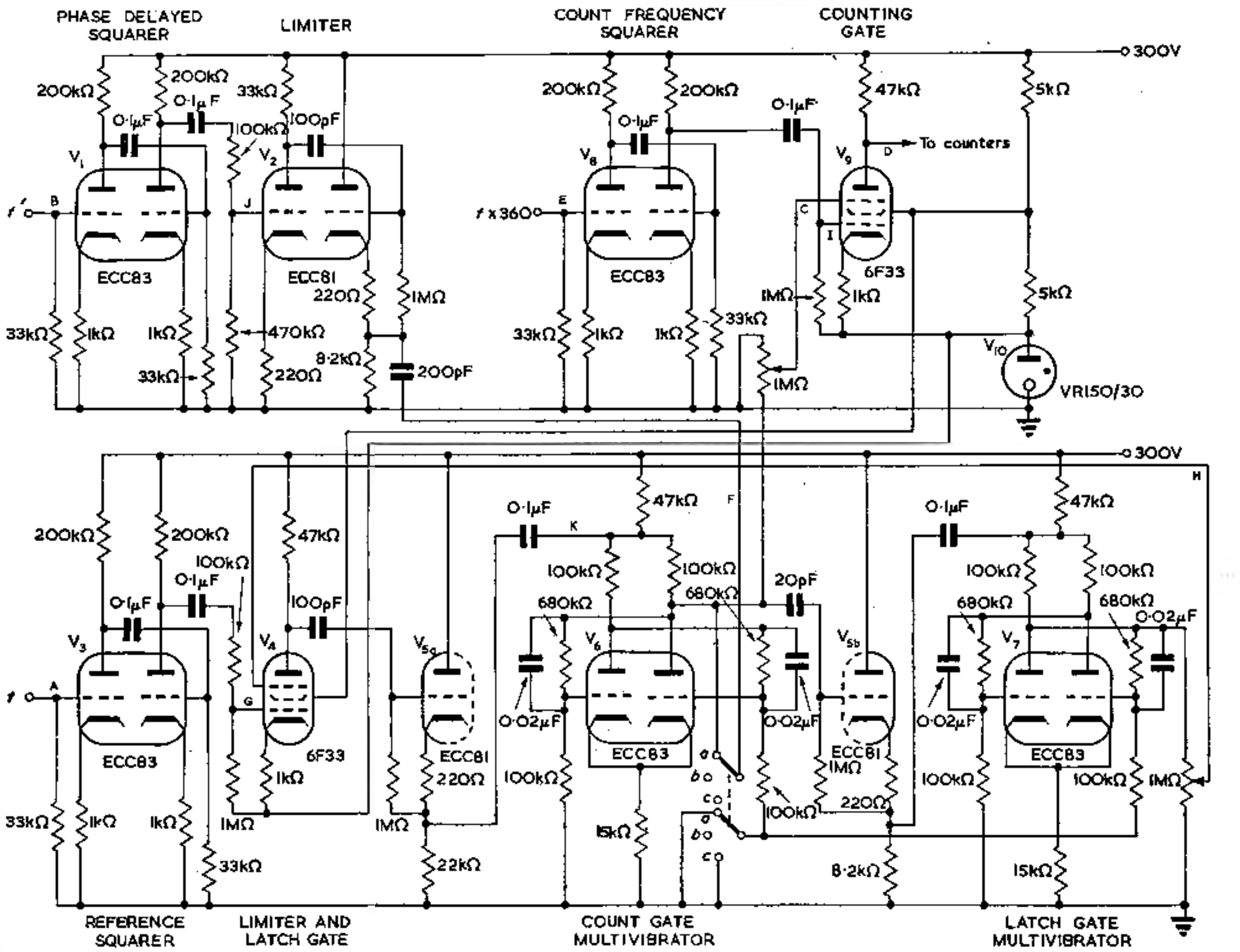


Fig. 7. Detailed circuit

GATE CIRCUIT

The form of gating circuit is shown in Fig. 6. A 6F33 pentode valve with a short suppressor grid base is used to switch the anode output on and off. When a positive step function is applied to the suppressor, the valve amplifies the control grid input in the normal way. When a negative step function is applied to the suppressor the valve is rapidly cut off and no output signal appears at the anode. The control voltages for the suppressor are derived from the multivibrators shown at H and C in Figs. 2, 3 and 7. In Fig. 7 the gating valves V_4 and V_9 are the latch gate and counting gate respectively.

POWER SUPPLIES

The phasing unit requires a stabilized h.t. supply of 300V at 40mA, and the electronic counters consume a further 45mA. The need for a negative supply line is avoided by the use of a neon valve to obtain the cathode and screen voltages for the gates.

Operating Procedure and Performance

The procedure for an operator using the phasemeter is firstly to set the low frequency oscillator to the required frequency *f*. The higher counting frequency is then set to $f \times 360$. This is checked by raising a key-switch to its 'check' position, and noting that the counters read 360°. This action corresponds to switch position *c* in Figs. 2 and 7. The gates are then re-set and the counters re-set to zero by returning the key-switch to its centre position (corresponding to *b* in Figs. 2 and 7). If the key-switch is pressed into its 'on' position (corresponding to *a*) the phase delay between the two input waveforms is shown on the counters in degrees.

The phase measuring equipment requires an input level between 1V and 10V r.m.s. The frequency of the input waveform may be between 1c/s and 100c/s. A simple modification, however, would extend this upper limit to 3kc/s. The present limit to accuracy (± 2°) is caused by variation of the oscillator frequency *f*. If a more stable oscillator is used, the phase-angle can be measured to within 1°. One possible improvement to the apparatus would be the combining of both oscillators into a single unit, having one output with a frequency 360 times the frequency of the other output.

Since the construction of this phasemeter, the problem of measuring low frequency phase, associated with vibrating structures, has been considerably eased. The use of this simple counting technique has also improved the accuracy of measurement, without the introduction of complicated operating controls.

Acknowledgments

The apparatus described has been developed as part of the programme of the Engineering Research Division of Messrs. A. V. Roe & Co., Ltd.

The author would like to thank his colleagues for helpful discussions and Mr. H. R. Ashley, Chief of Engineering Research Division, for permission to publish this article.

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A Voltage Controlled Continuously Variable Low-Pass Filter

By A. J. Seyler*, Dipl.Ing., M.E.E., M.I.R.E.(Aust) and A. Korpel*, Ir.(Delft)

A continuously variable low-pass filter is described which has been designed and constructed by using the time series method and delay line techniques. The cut-off characteristic of this filter approximates a 6dB/octave slope and its 3dB loss frequency is varied by applying a control voltage. The developmental model permits a cut-off frequency variation of 1:1000 (4·5kc/s to 5Mc/s) for the control voltage varying from −85V to 0V.

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

THE subsequently described low-pass filter was designed and constructed to serve in experimental investigations concerned with the statistics of the power spectrum of television picture signals. Its particular performance requirements were:

(a) Continuous variation of cut-off frequency by means of a control voltage.

(b) Constant amplitude of step function response irrespective of rise-time (bandwidth).

(c) Independent input and output impedances.

Although the design of this filter was governed by the requirements of the particular experiment, its principle and method of practical realization are general enough to be applicable to other frequency ranges and purposes.

For the practical realization of the filter, delay line techniques were considered most suitable, because of their inherently more convenient variability. For this reason the network response function was expressed in the form of a time series with real coefficients. The coefficients of this series were then determined from the sampled impulse response which was obtained from the inverse Laplace transform of the filter transfer function specified in the frequency domain. For the generation of the time series coefficients a rather simple method was found using repeated reflections in a delay line having a delay of half the sample spacing. This delay line is terminated in an open-circuit on one end and in a variable impedance on the other, the latter determining the time series coefficients and thus the filter response.

Filter Specification and Synthesis

The response function of a network $G(t)$, when excited by the input function $F(t)$ may be approximated by a time series of the form[1]

$$G(t) = c_0F(t) + c_1F(t-T) + c_2F(t-2T) + c_3F(t-3T) + \dots$$

$$= \sum_{r=0}^{\infty} c_rF(t-rT) \quad \dots\dots \quad (1)$$

where c_r are real coefficients which in the general case are given by $-1 < c_r < 1$, and T is the 'sampling' interval[2] defined as $T = 1/2 f_{max}$ with f_{max} being the maximum significant frequency component of $F(t)$. In this application $F(t)$ is a television signal of C.C.I.R. standard with $f_{max} =$ 5Mc/s, hence $T = 0{\cdot}1 \times 10^{-6}$sec. It is then only necessary to define the coefficients c_r to obtain the time series for the desired filter.

The frequency transfer function $h(p)$ was required to approach a 6dB/octave cut-off characteristic and the response can thus be written in complex notation.

$$g(p) = f(p)h(p) = f(p)\frac{\omega_0}{p+\omega_0} \quad \dots\dots \quad (2)$$

where ω_0 is the cut-off frequency (−3dB) of the filter. Since in this equation $h(p)$ is the transfer function of the network, its inverse Laplace transform is the response of the network to a unit impulse excitation $F(t) = \delta(t)$ (e.g. ref. 3). Consequently

$$H(t) = \mathcal{L}^{-1}h(p) = \omega_0 \exp(-\omega_0 t) \quad \dots\dots \quad (3)$$

From this the coefficients of the time series (1) are obtained by letting $t = rT$ and

$$c_r = H_r(rT) = \omega_0 \exp(-\omega_0 rT) \quad \dots\dots \quad (4)$$

Substituting $b = \exp(-\omega_0 T)$, equation (4) may be written

$$c_r = \omega_0 b^r \quad \dots\dots \quad (5)$$

Consequently for the specified low-pass filter the time-series coefficients are positive and represent a geometrical series of b, with b for any value of $\omega_0 > 0$ being $0 < b < 1$.

For $F(t) = u(t)$, the unit step function at $t = 0$, the amplitude of the output for $t \gg T$, equation (1) reduces to

$$\sum_0^{\infty} c_r = \omega_0 \sum_0^{\infty} b^r = \frac{\omega_0}{1-b} \quad \dots\dots \quad (6)$$

In order for this output to be constant independent of the cut-off frequency, the right-hand side of equation (1) must be multiplied by $(1-b)/\omega_0$. Hence in its final form, on which the physical realization is to be based, equation (1) becomes

$$G(rT) = (1-b)\{F(t) + bF(t-T) + b^2F(t-2T) + b^3F(t-3T) + \dots\} \quad \dots\dots \quad (7)$$

$$= (1-b)\sum_{r=0}^{\infty} b^rF(t-rT) \quad \dots\dots \quad (7a)$$

From $b = \exp(-\omega_0 T)$ follows

$$\omega_0 = -(1/T)\ln b$$

and the cut-off (−3dB) frequency

$$f_0 = -(1/2\pi T)\ln b = -0{\cdot}366(1/T)\log b \quad \dots \quad (8)$$

Finally with $T = 0{\cdot}1 \times 10^{-6}$sec the cut-off frequency in Mc/s becomes

$$f_0 = 3{\cdot}66 \log 1/b \quad \dots\dots \quad (8a)$$

The rise-time of the output waveform for a step function input is now to be derived. This is conventionally defined as the time between 10 per cent and 90 per cent amplitude. The amplitude of $G(rT)$ for $F(t) = u(t)$ and a limited number of coefficients n becomes from equation (7a)

$$\alpha = (1-b)\sum_{r=0}^{n} b^r$$

$$\alpha = 1 - b^n$$

Hence the number of coefficients for any amplitude α obtains from

$$n \log b = \log(1-\alpha)$$

* *Postmaster General's Department, Commonwealth of Australia.*

as

$$n = \frac{\log(1-\alpha)}{\log b} \qquad \ldots\ldots\ldots\ldots (9)$$

For 10 per cent amplitude, i.e. $\alpha = 0{\cdot}1$,

$$n_{0.1} = \frac{\log 0{\cdot}9}{\log b}$$

The rise-time for a delay T between each sample is then obviously

$$\tau = (n_{0.9} - n_{0.1})\, T \text{ sec}$$

and with numerical values for log 0·9 and log 0·1 and $T = 0{\cdot}1\mu$sec.

$$\tau = \frac{0{\cdot}0954}{\log 1/b}\ \mu\text{sec} \qquad \ldots\ldots\ldots (10)$$

and with equation (8)

$$\tau = 0{\cdot}35\ 1/f_o\ \mu\text{sec}$$

for f_o in Mc/s.

Practical Circuit Design

Time Series Generation

A possible practical realization of equation (7) would be a delay line with taps at $0{\cdot}1\mu$sec intervals and facilities to adjust the output of each tap in accordance with equation (7), then to add all outputs and multiply the sum with the correction factor $(1 - b)$. In order to vary the cut-off frequency of the filter all tap outputs would need to be varied simultaneously, maintaining the relationship as given by equation (7). This obviously would be a very elaborate, though not complicated arrangement, and for a good approximation a great number of coefficients, i.e. delay line sections would be required. However, the particular configuration of the time series for this network offered a much simpler solution. Consider a delay line of length $T/2$ having a characteristic impedance R_o. This may be terminated at the far end in an open-circuit and at the input end in a resistance R_1, e.g. the internal resistance of a signal generator. A signal V_1 injected into the line at $t = 0$ will be reflected at the far end and (neglecting losses in the line) will arrive back at the input after T seconds. Owing to the termination R_1 it will be reflected there with a reflection factor

$$b = \frac{R_1 - R_o}{R_1 + R_o} \qquad \ldots\ldots\ldots\ldots (11)$$

This signal bV_1 will then travel to the far end and return to the input again after T seconds, where it will be reflected multiplied with b and thus will start down the line as b^2V_1 and so forth. Hence the successively reflected signal, bouncing back and forth in the line will be characterized by the series

$$V_1(0) + bV_1(T) + b^2V_1(2T) + b^3V_1(3T) + \ldots\ b^rV_1(rT) \qquad \ldots\ldots\ldots (12)$$

This series fulfills the requirements for the generation of the time series coefficients of the network as given by equation (7). The actual circuit problem is then to devise input and output circuits for such a delay line and a means of controlling the value of the terminating resistance to satisfy the requirement of variability of the coefficients.

Fig. 1 shows a circuit configuration which, as will be shown, satisfies the operating requirements with a good approximation. A delay line characterized by its delay $T/2$ and characteristic impedance R_o is fed from a cathode-follower V_1, the dynamic impedance of which in series with a resistor R represents the termination of the delay line at the input end. A signal (e.g. pulse) of amplitude V_o applied to the input grid of the cathode-follower results in a signal V_1 at the cathode equal to

$$V_1 = \frac{g_m R_c}{g_m R_c + 1} V_o \qquad \ldots\ldots\ldots\ldots (13)$$

where g_m is the mutual transconductance of V_1 and R_c the cathode load impedance.

$$R_c = R + R_o \qquad \ldots\ldots\ldots\ldots (14)$$

The cathode impedance $1/g_m$ may be assumed to contain a constant component $1/g_o$ and a variable one $1/g_v$, hence

$$1/g_m = 1/g_o + 1/g_v \qquad \ldots\ldots\ldots\ldots (15)$$

Then equation (13) with equation (14) and (15) becomes

$$V_1 = \frac{R + R_o}{R_c + 1/g_o + 1/g_v}$$

The signal at the input of the delay line, V_2, obtains from

$$V_2 = \frac{R_o}{R + R_o} V_1 \qquad \ldots\ldots\ldots\ldots (16)$$

and

$$V_2 = \frac{R_o}{R_o + R + 1/g_o + 1/g_v} V_o \qquad \ldots\ldots\ldots (17)$$

At the open (far) end of the line, owing to the total reflec-

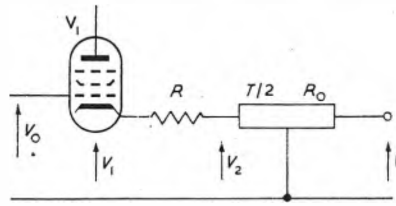


Fig. 1. Fundamental circuit

tion, the signal V_3 obtains as

$$V_3 = 2V_2 \qquad \ldots\ldots\ldots\ldots (18)$$

One half of this signal travels back to the origin of the line. There it 'sees' the impedance R_1, the input termination of the line, given by

$$R_1 = R + 1/g_o + 1/g_v \qquad \ldots\ldots\ldots (19)$$

from which the reflection coefficient b results as

$$b = \frac{R_1 - R_o}{R_1 + R_o} = \frac{R + 1/g_o + 1/g_v - R_o}{R + 1/g_o + 1/g_v + R_o} \qquad \ldots\ldots\ldots (20)$$

Let $1/g_o + R$ be chosen so that

$$1/g_o + R = R_o \qquad \ldots\ldots\ldots\ldots (21)$$

in other words, let the delay line be terminated by its characteristic impedance, when the variable component of the cathode impedance is zero, then from equations (20) and (21)

$$b = \frac{1/g_v}{2R_o + 1/g_v} = \frac{1}{1 + 2R_o g_v} \qquad \ldots\ldots\ldots (22)$$

If the same condition (21) is applied to the output voltage of the line V_3, (equation (18)) this becomes

$$V_3 = \frac{2R_o}{2R_o + 1/g_v} V_o = (1 - b)\, V_o \qquad \ldots\ldots\ldots (23)$$

There will be a constant delay of $T/2$ between the input signal to the cathode-follower V_o and the output signal of the line V_3 which is, however, insignificant for the mechanism of the time series generation. Consequently this constant delay may be eliminated by writing for V_o the time function $V_o(t + T/2)$ for which then with equation

(23) the delay line output becomes

$$V_3(t) = (1 - b)\left\{V_0(t) + bV_0(t-T) + b^2V_0(t-2T) + \ldots\right\} \qquad \ldots\ldots\ldots\ldots (24)$$

$$= (1 - b)\sum_{r=0}^{\infty} b^r V_0(t - rT) \qquad \ldots\ldots\ldots (24a)$$

Equations (24) and (24a) are thus identical with the required time series as given by equations (7) and (7a). In order for the coefficients b to be positive as postulated by equation (4) the condition for the magnitude of the terminating impedance R_1 of the line at the cathode of V_1 is

$$R_1 \geqslant R_0 \qquad \ldots\ldots\ldots\ldots\ldots\ldots (25)$$

or with equation (19)

$$R + 1/g_0 + 1/g_v \geqslant R_0 \qquad \ldots\ldots\ldots\ldots (26)$$

The condition (21) i.e. $R_0 = R + 1/g_0$ applied when deriving equation (24) governs, together with equation (26), the choice of the valve and the characteristic impedance of the line. g_0 is the maximum operating transconductance of the valve and R_0 could be chosen to be equal to $1/g_0$ thus eliminating the series resistor R. However, since g_0 may vary considerably from valve to valve, it is desirable to make provision for an easy adjustment of the matching between valve and delay line impedance. For this reason a.o.t. ('adjust on test') resistor R is provided. The existence of this resistor makes it also possible to specify the delay line impedance with less rigid tolerances. In this matched condition the cathode-follower delay line circuit is an amplifier having gain one between input grid and delay line output (apart from the constant delay $T/2$). For this to be valid for the maximum frequency range $B = 1/2T$, the delay line should be free of amplitude and phase distortion over this range. Under these conditions the equation (8) for the filter cut-off frequency

$$f_0 = -(1/2\pi T)\ln b$$

becomes with

$$b = \frac{1}{1 + 2R_0 g_v}$$

from equation (22)

$$f_0 = (1/2\pi T)\ln(1 + 2R_0 g_v) \qquad \ldots\ldots (27)$$

For decreasing cut-off frequency the cathode impedance must increase so that

$$1/g_m \gg 1/g_0$$

and from equation (15)

$$1/g_v = 1/g_m - 1/g_0 \simeq 1/g_m \qquad \ldots\ldots\ldots (29)$$

with which equation (27) can be written

$$f_0 \simeq 1/2\pi T \;\ln(1 + 2R_0 g_m)$$

With this approximation, however,

$$2R_0 g_m \ll 1$$

and

$$\ln(1 + 2R_0 g_m) \simeq 2R_0 g_m$$

hence

$$f_0 \simeq (1/2\pi T)\,2R_0 g_m \qquad \ldots\ldots\ldots\ldots (30)$$

i.e. the cut-off frequency becomes a linear function of the valve transconductance. Obviously a remote cut-off type

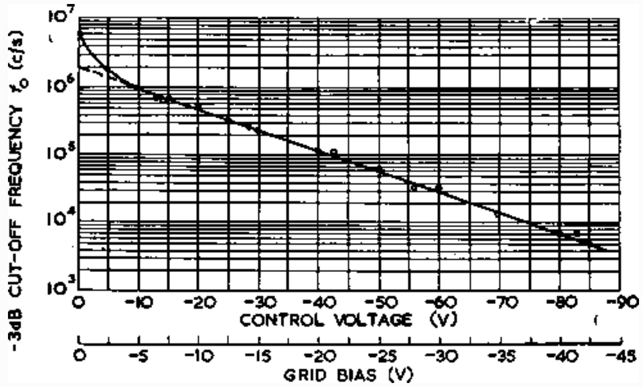


Fig. 2. Cut-off frequency control voltage

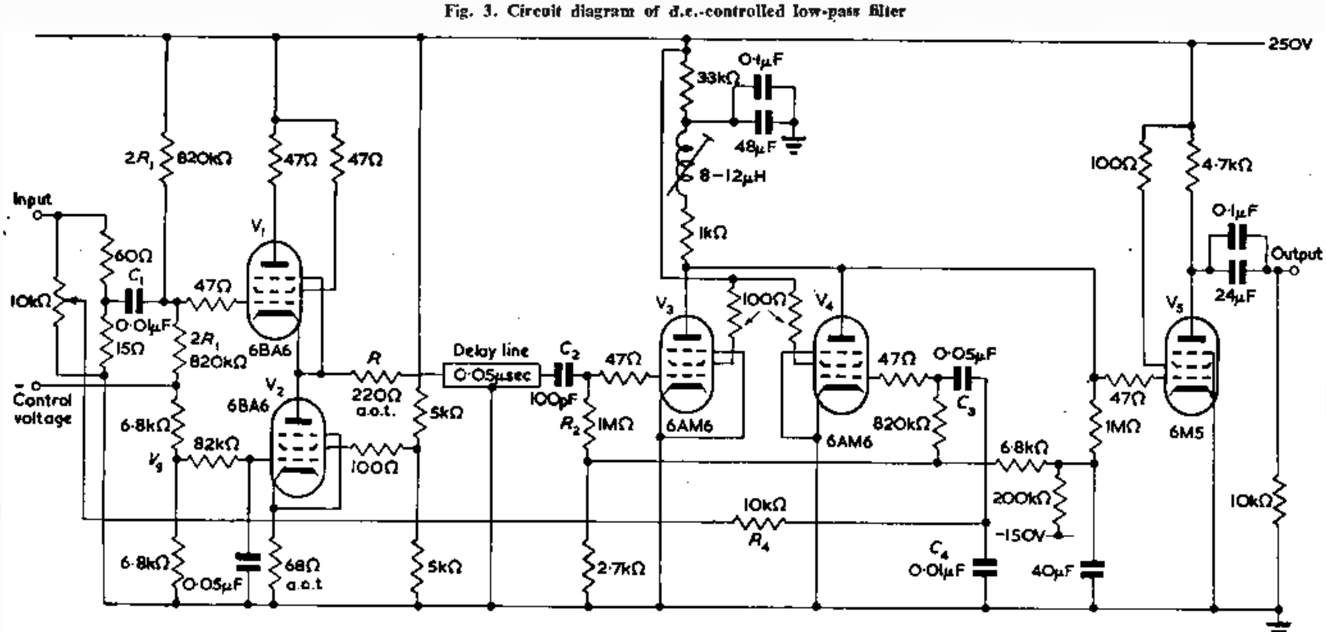
of valve will give a large control range of the cut-off frequency, owing to the large variations of transconductance obtainable.

For such a valve, e.g. the type 6BA6, the transconductance g_m (for small g_m) is approximately an exponential function of grid bias and may be expressed by

$$g_m = k\,\exp(\gamma V_g) \qquad \ldots\ldots\ldots\ldots (31)$$

where V_g is the grid bias voltage (which is negative) and k and γ are valve constants.

Fig. 3. Circuit diagram of d.c.-controlled low-pass filter



Hence, from equations (30) and (31), it follows that the cut-off frequency can be controlled by varying the grid of the valve feeding the delay line and that for low cut-off frequencies it will approach the function

$$f_0 \simeq (kR_0/\pi T)\ \exp(\nu V_g) = K\ \exp(\nu V_g) \dots . \quad (32)$$

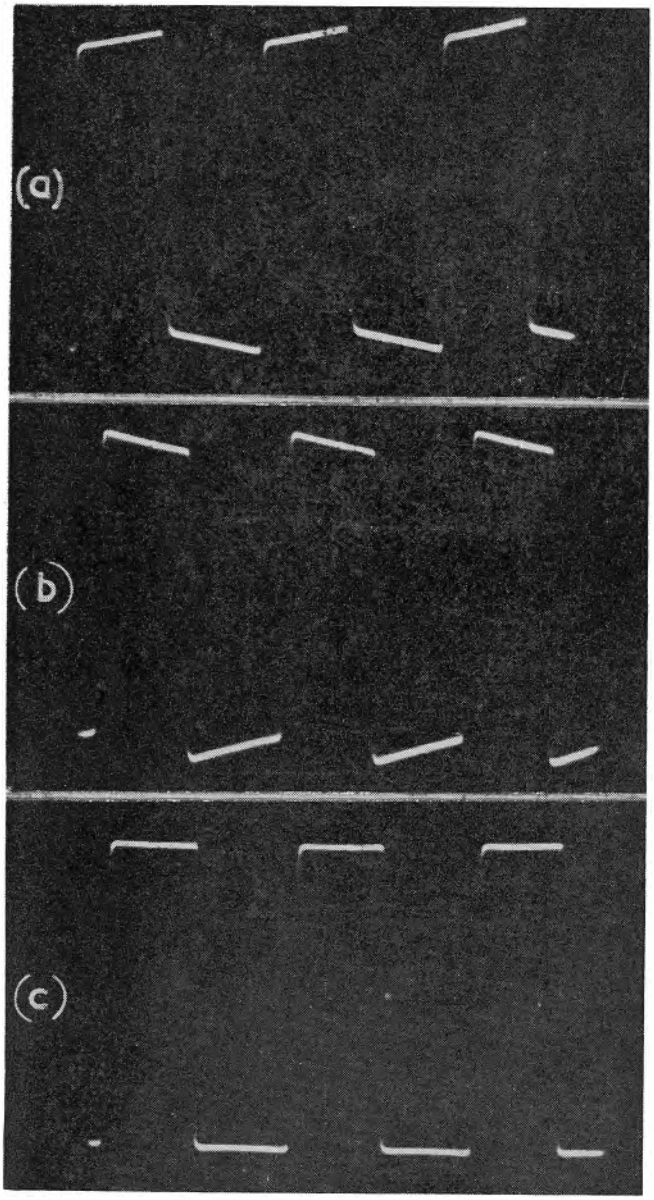


Fig. 4. Waveforms showing balancing of l.f. and h.f. paths
(a) l.f. > h.f. (b) h.f. > l.f. (c) l.f. = h.f.

Fig. 2 shows the measured values of cut-off frequency against control voltage of an actual filter. The valve used was a 6BA6, the delay line had a delay of $T/2 = 0{\cdot}5\mu$sec and a characteristic impedance $R_0 = 400\Omega$. Using a logarithmic scale for f_0 a function of the form of equation (32) will represent a straight line and this is in close agreement with the experimental results for $f_0 < 1$Mc/s. The numerical values for k and γ are found to be approximately $k = 1{\cdot}8$Mc/s and $\gamma = 0{\cdot}14\mathrm{V}^{-1}$ for this valve and circuit configuration. The complete circuit of an actually constructed filter will now be described.

Circuit of the Filter

As stated in the introduction, this filter was constructed for experimental investigations on the power spectrum of television signals. This determined the maximum frequency component and thus the maximum cut-off frequency of the filter to be 5Mc/s. Instead of requiring the filter to pass all frequencies below this value including d.c. with the inherent complications of d.c. amplifiers, the usual specifications for video amplifiers of permitting a 50c/s square wave to have a sag of not more than 5 per cent was adopted. In order to fulfil this condition the time-constants of anode-grid coupling and anode decoupling networks must be correspondingly large.

It was found that these time-constants restricted the speed with which the cut-off frequency could be controlled. The signal path through the filter was therefore split into a low frequency path which remained constant and a high frequency path which was bandwidth-controlled. The amplitudes are adjusted by means of the 10kΩ potentiometer in the l.f. path (see Fig. 3). This is readily accomplished by feeding a low frequency square wave into the filter and adjusting for optimum flatness of the horizontal portions of the displayed waveforms (Fig. 4). The fact that there is a constant delay between the h.f. and l.f. path of $0{\cdot}05\mu$sec, is not important in this application, since this represents a very small relative phase shift of the frequencies passed by the l.f. path and in the overlapping ranges of the two paths, in view of the relatively low crossover frequency. By addition in the common anode load of the valves V_3 and V_4 the two paths are joined again. The complete signal is then fed to the output valve V_5 through a conventional RC coupling network of a sufficiently high time-constant. The transfer characteristics of the l.f. and h.f. paths are made complementary in the vicinity of the crossover frequency f_c.

This frequency is determined by the upper cut-off of the l.f. path and the lower cut-off of the h.f. path. By letting the time-constant T_2 of the network consisting of C_2R_2 be small compared with that of C_1R_1 and large compared with that of the variable filter circuit $T_0 = 1/\omega_0$ at minimum cut-off frequency, T_2 alone determines the l.f. cut-off characteristic of the h.f. path. Similarly, $T_4 = R_4C_4$ is made small compared with $T_3 = C_3R_3$ and thus governs the upper cut-off characteristic of the l.f. path. Then, if $T_2 = T_4$ and because of the reciprocity of the two RC networks, the transfer functions of the l.f. and h.f. path are complimentary in the vicinity of the crossover frequency $f_c = 1/2\pi T_2 = 1/2\pi T_4$. T_2 and T_4 were chosen to be 10^{-4}sec giving a crossover frequency of $f_c = 1{\cdot}5$kc/s approximately. The sag in the l.f. square wave response introduced by the time-constant of the network C_3R_3 is compensated in the conventional way to the required tolerance by the anode decoupling network in the anodes of V_3 and V_4.

It was shown in the description of the filter principle that the cut-off frequency is controlled by the ratio of the cathode impedance (of the valve feeding the delay line) to the characteristic impedance of the line. To achieve a large range of control a remote cut-off valve type 6BA6 is used here. The d.c. return path to earth could be carried either through the delay line and a terminating resistor at its end or through a resistor parallel to the input of the line. In both cases a large resistor would be necessary in order not to interfere with the terminating conditions of the delay line and thus with the operation of the circuit. But this would make the establishing of the proper d.c. conditions for V_1 somewhat difficult. Instead, the cathode of the input valve is returned to ground through another valve of the same type, the anode impedance of which is large enough to be negligible compared with the maximum cathode impedance of the input valve. The cathode current and thus the cathode impedance is then controlled by a voltage applied to the grid of the bottom valve (V_2). The configuration of the voltage divider from +250V to ground, supplying the two grids with the required d.c potentials, ensures that for any voltage applied at the control voltage input the grids of V_1 and V_2 move by the same amount. This prevents any spurious transients from being generated and passed through the filter as a result of rapid changes in control voltage. This effect is still further suppressed by introducing an *RC* time-constant into the grid of V_2 which is equal to the time-constant of R_1C_1 at the grid of V_1.

The delay line is a laboratory-made model consisting of 5 *m*-derived T-sections and an *m*-derived terminating half-section. An $m = 1{\cdot}27$ is used for the T-sections and $m =$ 0·7 for the terminating half-section. This method of construction ensures constant amplitude and delay response and matching of the impedance to a resistive termination well beyond the required frequency range[4]. As required, the total delay is 0·05μsec and the characteristic impedance was chosen to approximate 400Ω to facilitate matching with the cathode impedance of V_1 as described earlier. Any other delay line will, of course, be usable as long as the delay and impedance values are compatible with the filter design and the frequency response of the line is uniform over the range of frequencies the filter is required to pass.

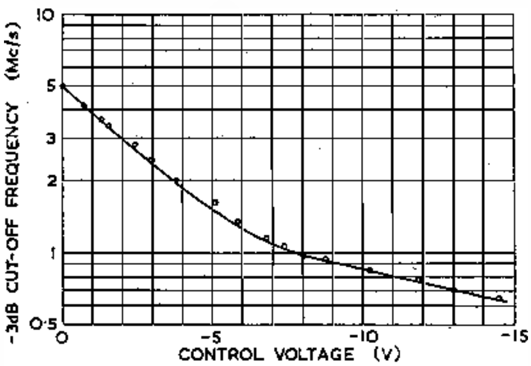


Fig. 5. Cut-off frequency control voltage

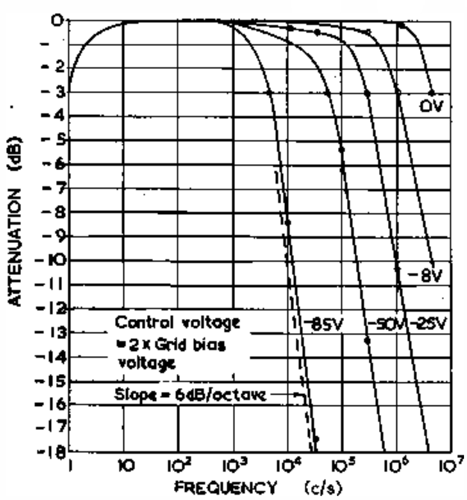


Fig. 6. Amplitude-frequency response for various control voltages

Performance Measurements of the Filter

Typical steady state measurements of the amplitude response of the filter are shown in Fig. 2, Fig. 5 and Fig. 6. As discussed above, Fig. 2 depicts the operating range of the filter where the control voltage is high and thus the valve transconductance follows essentially an exponential law of the grid bias. This range extends from about −10V to −85V control voltage (which is twice the grid bias on the control valve) with a variation in cut-off frequency from approximately 1Mc/s to 4·5kc/s respectively. When the control voltage falls below about −10V the simplifying assumptions made to derive equation (32) are no longer valid and also the valve transconductance deviates from the exponential characteristic. This operating condition is shown in Fig. 5. It indicates that the variation in cut-off frequency with control voltage is much more rapid there and that only about 8V are required to shift the filter from its maximum design bandwidth of 5Mc/s to 1Mc/s. For the evaluation of these measurements it must be kept in mind that the cut-off frequency versus control voltage characteristic of this filter is a function of valve constants; these not only vary from one valve to the other, but also for the same valve during its useful life and with variations in supply voltages such as filament and high tension. These voltages must consequently all be stabilized. Recalibration is necessary, when valves are replaced and valves should preferably be pre-aged for some time. Since the filter described here was to be used in laboratory experiments, no special efforts were made to achieve a long term stability of these operating conditions and routine checks of the characteristic were carried out, when it appeared desirable. Some cut-off characteristics of the filter, i.e. amplitude-frequency responses for fixed control voltages are shown in Fig. 6. This shows the approximation to the theoretical 6dB/octave slope quite clearly. Owing to the split path arrangement for low and high frequencies, the response below 1kc/s is practically constant for all settings of the variable high frequency path.

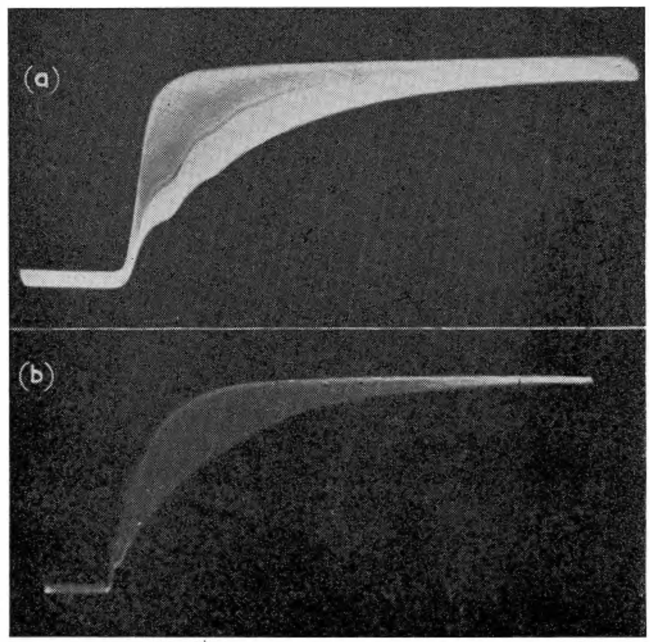


Fig. 7. Step function response. Control voltage d.c. and 50c/s

(*a*) *Control voltage −9V d.c. ±9V a.c. cut-off frequency 5Mc/s > f_o > 0·55Mc/s*

(*b*) *Control voltage −30V d.c. ±9V a.c. cut-off frequency 450kc/s > f_o > 110kc/s*

In Fig. 7 two photographs are shown of the rising flank of a 100kc/s square wave, when a 50c/s sine wave of 6·3V r.m.s. was superimposed on the d.c. control voltage. Fig. 7(a) was taken with −9V and Fig. 7(b) with −30V d.c. control voltage (the time-base is different in the two figures). The envelope of the modulated step corresponds to minimum and maximum rise-time at the positive and negative peak of the modulating sine wave. In the first case the filter swings through a frequency range of from 5Mc/s to about 500kc/s and in the second case from about 400kc/s to 120kc/s. The change in slope of the cut-off frequency versus control voltage characteristic, when the control voltage goes through the 'knee' of the characteristic at about −10V is quite distinctly visible in Fig. 7(a) from the lighter shading of the pattern towards the steeper rising edge. In both waveforms a kind of staircase pattern is visible near the beginning of the step function except in the one with the shortest rise-time. This is firstly due to the fact that the test equipment used has a much wider frequency spectrum (20Mc/s) than the waveforms for which the filter was designed (5Mc/s), and secondly, that these higher frequencies are still contributing significantly to the filter output because of its slow roll-off cut-off characteristic. Hence, the individual coefficients of the time series of which the output waveform consists are distinctly visible. However, the more often the excitation function has passed the delay line, i.e. with increasing order of the coefficient, the more these high frequencies are attenuated and the waveform becomes smooth. Of course, since for the fully open filter only the zero order coefficient appears at the output ($b = 0$), the fastest visible edge in Fig. 7(a) is free of this staircase pattern. Fig. 7 also illustrates the independence of the step amplitude of the cut-off frequency of the filter as postulated by equations (6) and (7).

Discussion and Outlook

The filter described in this article is of a very simple nature, in fact, it could have been realized by means of an *RC* network, since it contains only one simple pole on the real p-axis. However, the delay line technique employed for its synthesis, demonstrates the direct translation of the time series concept of network synthesis in such a convincing simplicity and presents such negligible complications that this form of realization of the filter may be accepted without hesitation. It can as such be considered as a basic building block by means of which other transfer functions may be synthesized.

Since each such low-pass filter produces one pole on the real p-axis in the transfer function, any number of poles on the real p-axis may be generated by cascading a number of these units. The location of each of these poles may be controlled individually or in a functional relationship to the others by manual d.c. voltage settings or by varying voltages derived from quantities to be controlled by the filter.

A high-pass filter may be synthesized with equal control facilities by subtracting the low-pass filter output from its input in a differential amplifier, leading to the configuration of Fig. 8(a). Since there is a constant delay of $T/2$ present in the low-pass filter an identical delay line must be inserted into the unfiltered signal path to obtain correct subtraction. This then gives the transfer function.

$$g(p) = 1 - \frac{\omega_0}{p + \omega_0} = \frac{p}{p + \omega_0} \quad \ldots\ldots\ldots (33)$$

which is that of a high-pass with a single pole on the real p-axis and a zero in the origin. Similarly, bandpass configurations can be synthesized by cascading low-pass and high-pass units (Fig. 8(b)) whereby bandwidth and centre frequency can be controlled by the applied voltages.

While the transfer functions of these configurations contain a simple pole on the negative real axis for each basic low-pass unit used,* the same circuit configuration as shown in Fig. 1. can be used to generate complex conjugate pole pairs. For the low-pass filter it was postulated that the time series coefficients be positive (equations (1) and (7)), leading to the further postulate that the reflection factors at the delay line input be always positive and hence the dynamic cathode impedance be always larger than the characteristic impedance of the delay line.

If the reflection factor at this point is made negative, by letting the characteristic impedance of the delay line be always larger than the dynamic cathode impedance, and the open-circuit condition is maintained at the output of the delay line, the time series describing the response of this circuit to an excitation $F(t)$ becomes, from equation (7):

$$G(rT) = (1+b)\left\{F(t) - bF(t-T) + b^2F(t-2T) - b^3F(t-3T) + \ldots\right\} \ldots\ldots\ldots (34)$$

$$= (1+b)\sum_{r=0}^{\infty}\left\{(-1)^r b^r F(t-rT)\right\}$$

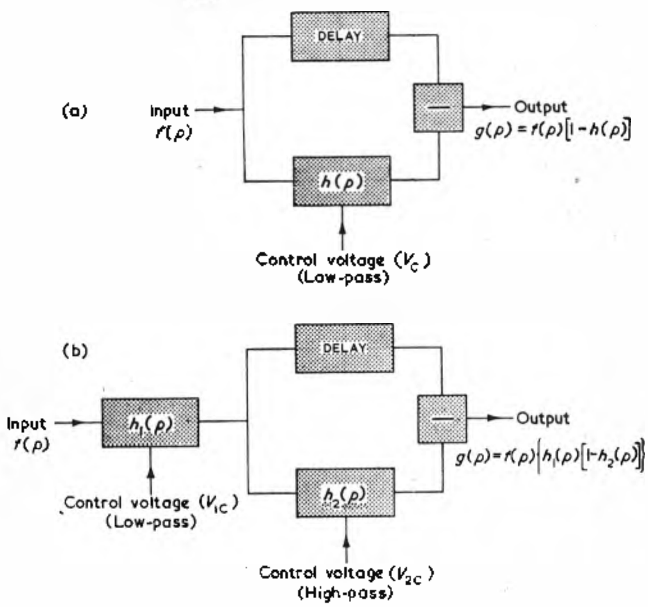


Fig. 8. Synthesized filter configurations
(a) High-pass
(b) Band-pass

With the same definitions as used for the development of equation (7), equation (34) can be written in the continuous form $H(t)$, representing the unit impulse response of the circuit, as*:

$$H(t) = (1+b)\cos(\pi t/T)\ \exp(-\omega_0 t) \quad \ldots.(35)$$

After transformation this then yields the transfer function[3]

$$h(p) = (1+b)\frac{p+\omega_0}{(p+\omega_0)^2 + (\pi/T)^2}$$

$$= (1+b)\ \frac{p+\omega_0}{(p+\omega_0+\mathrm{j}\pi/T)(p+\omega_0-\mathrm{j}\pi/T)} \quad \ldots\ldots\ldots\ldots\ldots\ldots (36)$$

This transfer function obviously contains a conjugate complex pole pair with the co-ordinates

$$p_{00} = -\omega_0 \pm \mathrm{j}\pi/T$$

and a zero at

$$p_0 = \omega_0$$

* *It should be realized that this is due to the condition that the maximum signal frequency be $1/2T$ applied here. If this restriction is removed the transfer function of such a delay line filter becomes periodic, having poles at $p_\infty = -\omega_0 \pm 2jk\pi/T$ $(0 < k < \infty)$, thus assumes the character of a 'comb' filter.*

* *$(-1)^r$ are the sample amplitudes at intervals T seconds of the function $\cos \pi t/T$. By definition $b = \exp(-\omega_0 T)$ and hence $b^r = \exp(-\omega_0 rT)$ which with $rT = t$ becomes $\exp(-\omega_0 t)$.*

Again the inherent possibility of voltage control of the circuit transfer function applies to the 'damping' co-ordinate ω_0 of poles and of the zero while the 'resonance frequency' $\omega = \pi/T$ is determined by the length of the delay line.

Although the delay line method of network synthesis offers a still wider range of possibilities, this brief outlook was intended to describe some possible variations of the one particular circuit configuration operating under the conditions specified.

Conclusion

A circuit has been developed which generates a theoretically infinite time series with exponentially decreasing coefficients. It has been shown that this approximates a low-pass filter with a 6dB octave cut-off characteristic. The particular feature of the circuit is the possibility of varying the time series coefficients by means of a control voltage which results in the continuous variation of the cut-off frequency (−3dB) of the filter. In the particular model of the circuit the cut-off frequency may be varied from 4·5kc/s to 5Mc/s (a ratio 1:1000) by a control voltage variation of 85V.

This facility provides a convenient means of automatic and remote control of variable transfer functions and makes this type of filter circuit a suitable component for non-linear filter applications. Furthermore, as is indicated in the discussion, the basic circuit configuration can also be applied to the realization of other transfer functions, e.g. high-pass and band-pass characteristics.

Acknowledgment

The authors are indebted to the Engineer-in-Chief of the Postmaster-General's Department of Australia for his permission to publish this article.

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An Electronic Wheel Truing Machine

Wheels that are completely true and free from any irregularities are essential if a bicycle is to run smoothly and efficiently. Building wheels to these close limits has always provided difficulties for the manufacturers, as the rims are never in themselves completely free from irregularities.

The wheel builder has to correct these defects by applying exactly the right tension to each spoke so that the finished wheel is completely round and concentric. This is a task which has always proved extremely difficult to mechanize, being normally done by skilled operators who need a long period of training before becoming proficient.

Now, for the first time, The British Cycle Corporation Limited, in conjunction with Tube Investments Technological Centre, has designed and developed an electronically controlled machine which will automatically carry out this difficult operation.

The wheel, roughly laced but with the spokes still loose, is placed into the machine by the operator, who then merely presses a button and takes no further part in the operation until the machine has finished truing and testing the wheel one minute later.

When the wheel is placed in the machine it is automatically lowered horizontally into a bed incorporating a ring of 40 electric motors, each of which is individually controlled, each one tightening one spoke only. When the spoke nipples have been gripped by the spanners driven by these electric motors, a magnetic sensing head is lowered on the end of an arm which is then rotated at high speed just above the rim of the wheel—the head scanning the rim all the time to detect any divergence from the true circle.

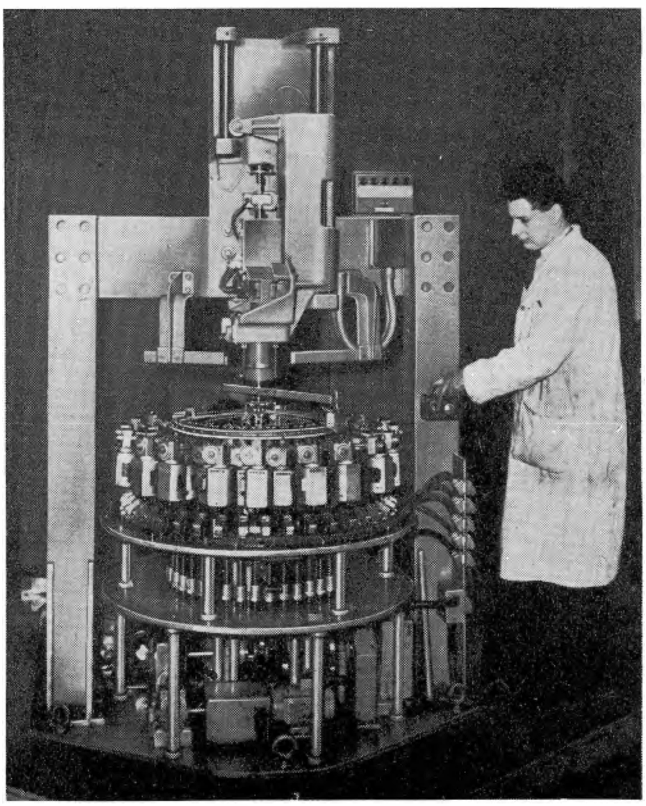
The information from this inductive magnetic head is fed into two control cubicles at the rear of the machine, where it is translated by electronic devices into individual 'instructions' for each one of the 40 electric motors tightening the spokes.

The machine is so designed that the only forces acting on the rim are those imposed by the spokes themselves, so that when the wheel is true in the machine it will stay true when removed from it.

The spokes are tightened until there is the correct overall tension and when the truing is completed the magnetic inductive head continues to rotate as a final check that the operation is completed. When it is satisfied that each spoke is under the right tension the rotation stops and the sensing head is withdrawn. The spanners then rotate slightly to free the spokes, and the wheel is released and ready for the operator to remove it.

The operations are in fact rather more complex than has been suggested in this brief outline, as the variations in dimensions of the rims are such that the machine has to determine its own tensioning limits before truing can start. Nevertheless, it still trues and tests at the rate of one wheel per minute, the wheel being true to within ·001in. This is a much faster rate of production and a greater degree of accuracy than could be achieved by even the most skilled wheel builder.

The wheel truing machine



The New ACE Computer

One of the world's largest and fastest electronic computing machines has begun to operate at the National Physical Laboratory. Known as ACE, it can carry out any calculation process for which exact rules are known.

Very fast computers of this type are required for many problems of which design calculations for high-speed aircraft and nuclear reactors are typical examples. The emphasis in the N.P.L's use of ACE will be, however, into the general methods by which computations such as these can be carried out.

ACE is the end-product of research, which began at the National Physical Laboratory at the end of the war. In 1945 the late Dr. A. M. Turing, F.R.S., joined the staff of the Mathematics Division at N.P.L. and in the early post-war years he produced, with his collaborators, a logical design for a computer intended to carry some of the load of numerical calculation. He called it the Automatic Computing Engine—or ACE[1].

In 1947 a special section—now the Control Mechanisms and Electronics Division—was set up to carry out the development of the machine. It was thought, however, unwise to attempt to build so large and complex a machine in one step—so it was decided to make first a pilot model of considerably smaller power and capacity.

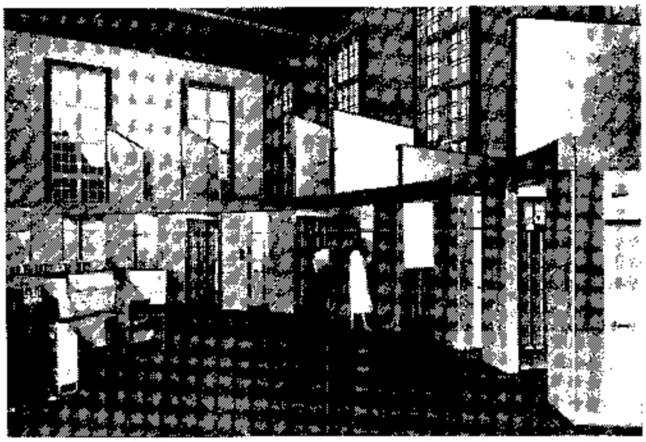
Thus, a pilot-model ACE[2], only one-sixth of the size of the present machine, was built to investigate the engineering problems which would arise in the full-scale machine. It was completed in 1951 and during its five-year life at N.P.L.—it was presented to the Science Museum in 1956 —it carried out a considerable amount of work, including extensive calculations to discover the cause of failure in the early Comet. It was, in fact, the pilot-model which led to the commercial production of the English Electric DEUCE computer. Moreover, it provided the team of mathematicians, logical designers, electronic engineers and mechanical engineers with the essential experience and 'know-how' to undertake the more ambitious full-scale ACE project.

Short Description of Machine

ACE is a serial computer in which numbers and instructions have 48 binary digits: the digit rate is $1{\cdot}5 \times 10^6$/sec and therefore its word time or minor cycle is 32μsec.

The main working store consists of 24 mercury delay lines each containing 32 words, circulating in a major cycle of 1 024μsec. Rapid access storage is obtained by using mercury delay lines of one, two and four words capacity. The backing store is four magnetic drums containing a total of 32 768 words.

The ACE instruction specifies a three address operation of the form '*A* function *B* to *D*' where *A* and *B* are the store addresses of the operands and *D* is the store address to which the result of the operation is to be sent. Instructions also specify a further store address *N* from which the next instruction is to be extracted. The time needed to execute this operation (for single length numbers) is 32μsec and since each operation may be followed immediately by another, a maximum rate of operation of 30 000 per second is possible.



A general view of the ACE computer showing the rising doors.

The drum store, the multiplier and the divider are independent units. These are put into operation by the appropriate instructions and then work independently until they have completed their operation. This feature provides a parallelism of operation, since a multiplication, division, drum transfer and a series of ordinary operations could be in progress at the same time.

The three address operation of the machine allows the multiplier and multiplicand to be selected and the multiplication process to be initiated by a single instruction. The product is formed in 14 minor cycles, i.e. 430μsec. The division process is similarly specified and the quotient (rounded or unrounded) is formed in approximately 1·5msec. The divider also contains an automatic standardizing process for use in floating point arithmetic operations.

The magnetic drum store has a total capacity of 1 024 long delay lines, i.e. about $1{\cdot}5 \times 10^6$ digits. There are four drums each having 256 tracks, each track storing the contents of one delay line. Each drum has 16 read heads and 16 write heads, the 256 tracks being obtained by moving the heads as one unit into one of 16 discrete positions. Each drum is 6·75in long by 5in diameter, allowing a linear digit packing of 100 per inch. The drum is driven by a hysteresis motor running synchronously at 12 000rev/min and is phase corrected so as to rotate exactly once in five major cycles of the machine.

Punched card and magnetic tape equipment will be provided for input and output. The installation will initially consist of two machines.

(1) A broadside card reader, running at 450 cards per minute.

(2) A broadside card punch running at 100 cards per minute.

In order to read or punch 80 columns of card the computer has an 80 digit input dynamicizer and an 80 digit output staticizer.

The ACE has been designed and constructed to allow easy maintenance. Extensive marginal checking facilities are provided and the chassis units are designed to give complete accessibility to all components, valve connexions, etc.

The machine is housed in 10 cabinets, each having a cooled air circulation system. Each cabinet is fitted with a rising door which permits immediate access to all the 24 chassis units contained therein. The number of valves is about 6 000.

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A Microwave Reflectometer Display System

for 7 500 to 11 000Mc/s

By J. C. Dix*, B.Sc., and M. Sherry*, M.Sc.

An X-band reflectometer display system which has recently been developed, is described. This instrument gives a continuous display of the match of an X-band waveguide component over a frequency band of 7 500 to 11 000Mc/s.

The basic principle is that when the frequency is changed the variation of the modulus of reflection coefficient of a test load is indicated by the change in level of the reflected wave if the amplitude of the incident wave is maintained constant and independent of frequency.

The X-band r.f. source is a backward wave oscillator electronically scanned at 50c/s over any frequency band within 7 500 to 11 000Mc/s. Output of the b.w.o. is fed into a waveguide system consisting of two directional couplers arranged back to back, with the device under test as the termination. Balanced crystal detectors in the coupled arms of the directional couplers detect respectively the incident wave and the reflected wave set up by the termination. A feedback circuit from the incident crystal to the b.w.o. ensures that the incident power level is held sensibly constant. The level of the detected reflection signal after amplification is displayed on a c.r.t. whose time-base is synchronized to the scanning voltage, thus giving a display of variation of (reflection coefficient)^2 with frequency.

The instrument is primarily intended for the display and comparison of standing wave ratios of the order of 1·5 or more. In normal use the match of a test component is compared against that of a known mismatch of roughly the same order of magnitude. With an additional circuit component the instrument can also be used to display the variation of transmission coefficient with frequency.

Some discussion of the sources of error to be expected in the system is given together with details of tests made to ascertain operation accuracy.

The article deals only with an X-band reflectometer, but it is pointed out in conclusion that since a range of b.w.o.'s is available to cover the microwave frequency range of 1 600 M/cs to 11 500Mc/s, the system can easily be extended to any other frequency band within these limits.

(Voir page 58 pour le résumé en Français: Zusammenfassung in deutscher Sprache auf Seite 62)

Up to a comparatively short time ago work at microwave frequencies has been done at certain well defined narrow frequency bands. This has been largely because of the limited tuning range of available types of microwave oscillator. Measurements of the impedance of circuit components in these narrow bands has usually been performed by patient exploration of the frequency band in small steps, using a slotted line standing wave meter and manually tuned oscillator.

With the advent of the backward wave oscillator (b.w.o.) and travelling wave tube (t.w.t.) the microwave field is no longer restricted to such narrow frequency ranges. These devices can have tuning ranges of an octave or more and both in the development of the tubes themselves and in the subsequent development of microwave systems making full use of them, the engineer is faced with the experimental evaluation of the performance of microwave circuit components over much wider frequency bands than hitherto. It is obvious that impedance measurements using a manually operated standing wave meter become very tedious and time consuming when applied to these wide bands and some other method enabling a relatively rapid check of impedance is desirable.

It was to meet this requirement that the present reflectometer system was devised. It gives an oscilloscope display of the magnitude of the reflection coefficient of a microwave component plotted against frequency. Matching information over a wide band is thus presented visually and continuously, although not to the accuracy obtainable by a standing wave indicator.

* *Communication No. 782 from the staff of the Research Laboratories of The General Electric Co. Ltd, Wembley, England.*

The system permits a rapid comparison of the match of a test component against that of a known standard. It has been used to assess the performance of new designs of delay lines for use in b.w.o.'s and t.w.t.'s and components associated with them. In addition since the display shows immediately the effect of an adjustment of any variable parameter, it has provided a rapid means of setting the match of a component to a desired level. The reflectometer does not give information about the phase of the reflection coefficient whereas from a standing wave measurement both the magnitude and phase can be ascertained. However in many cases knowledge of the phase is not essential. This is particularly so in wide band devices where the phase must in general vary between wide limits through the frequency band.

The initial development of this reflectometer was undertaken at S-band but more recently it has been applied to higher frequencies. The particular version to be described here is an X-band one covering a frequency range of 7 500Mc/s to 11 000Mc/s.

Basic Principle

In a microwave system consisting of a generator feeding through a transmission line into a terminating impedance there are normally present the incident wave from the generator and the reflected wave set up by the termination. The amplitude of the reflected wave is proportional to the modulus of the reflection coefficient and the amplitude of the incident wave.

Hence if over a range of frequency the amplitude of the incident wave is maintained constant the amplitude of the reflected wave varies in proportion to the reflection coefficient of the termination. Thus a crystal detector coupled to the reflected wave through a directional coupler gives an output signal approximately proportional to the square of the voltage reflection coefficient (assuming the crystal operates in its square law region). This detected signal is displayed as the Y deflexion on a cathode-ray tube against an X deflexion approximately linearly proportional to the frequency of the microwave source to give a visual display of the square of reflection coefficient against frequency.

The essential requirement for a system such as this is a microwave generator which can be easily frequency modulated over a wide band together with some form of output amplitude control. The b.w.o. lends itself admirably to this requirement as it can be tuned electronically on the line electrode with its output controlled by a suitable voltage applied to its grid. This latter voltage is obtained from a crystal detector coupled to the incident wave through a directional coupler. The system then functions as a negative feedback loop tending to stabilize the amplitude of the incident wave.

These are the basic features of the system which is described in detail in the following section.

Detailed Description of the System

A block schematic diagram of the complete reflectometer system is shown in Fig. 1.

The microwave generator (1) which supplies the variable frequency microwave input to the system is a 'Carcinotron' type CO43 made by the Compagnie Générale de T.S.F. This is a low power O type backward wave oscillator with a nominal electronic tuning range of 7 000 to 11 000Mc/s.

As used here the b.w.o. frequency scanning is done at 50c/s with a sinusoidal voltage derived from the mains. The scan voltage supply (2) incorporates an isolating transformer and a variable auto-transformer, permitting adjustment of the scanned frequency range. The d.c. supply (3) for the line is also adjustable. The two adjustments give a frequency scan of 7 500 to 11 000Mc/s or any smaller scan about any mean frequency within this band.

The r.f. output of the b.w.o. is taken through a coaxial cable and coaxial waveguide transition (4) to a circuit in waveguide No. 16. From the input section of this the r.f. signal passes through the preset attenuator (5) and the -10dB directional couplers (6) and (7) to the test load (8). Wideband balanced crystal detectors (10) and (11) with preset pads (9) and (12), are connected in the appropriate auxiliary arms of the couplers to sample the incident and reflected waves respectively. The unused arms are terminated internally in well matched loads. An absorption wavemeter (13) is connected in series with the reflection crystal to provide a check of microwave frequency.

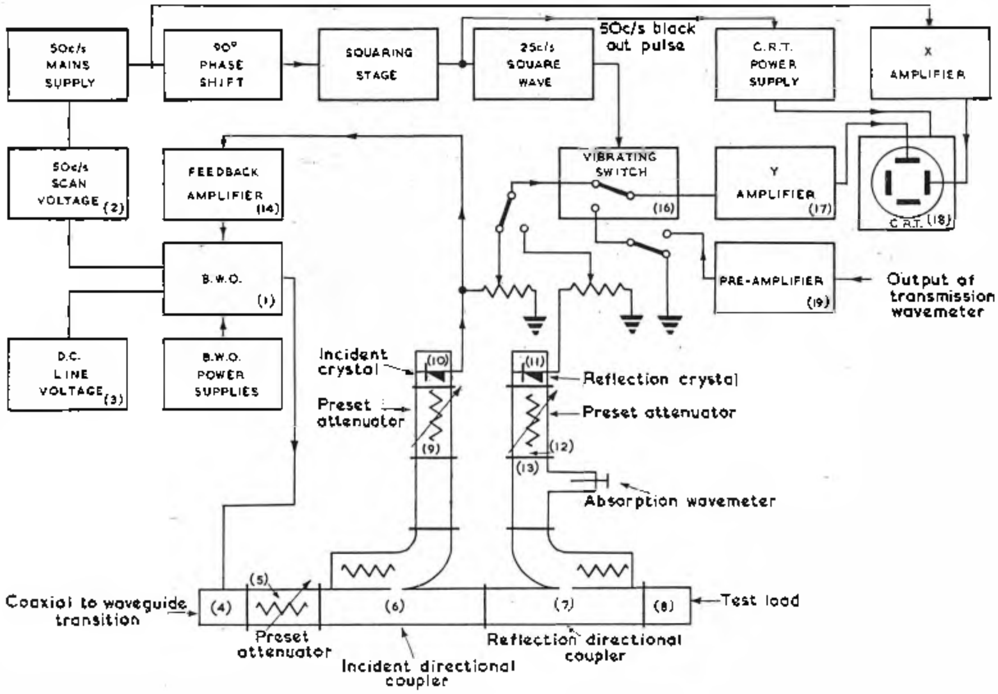


Fig. 1. Block schematic diagram of reflectometer

Part of the incident detected signal from the detector (10) is amplified in the amplifier (14) and fed back in reverse phase to the grid of the b.w.o. This negative feedback approximately stabilizes the level of the incident wave in the main guide to a constant value independent of the microwave frequency.

The output from the reflection crystal (11) is taken through the vibrating switch (16), Y amplifier (17) to the Y plates of a 6in c.r.t. (18). The function of the vibrating switch is to chop the input to the Y amplifier periodically to zero voltage. This ensures that the true level of the detected signal is transmitted through the a.c. coupled Y amplifier which otherwise only transmits the signal variations about a mean level. The functioning of the associated time-base, black out, and vibrating switch drive circuits is such that a continuous presentation is obtained on the c.r.t. of the detected reflection signal above a base line defining the zero level, the X displacement being approximately linearly proportional to microwave frequency.

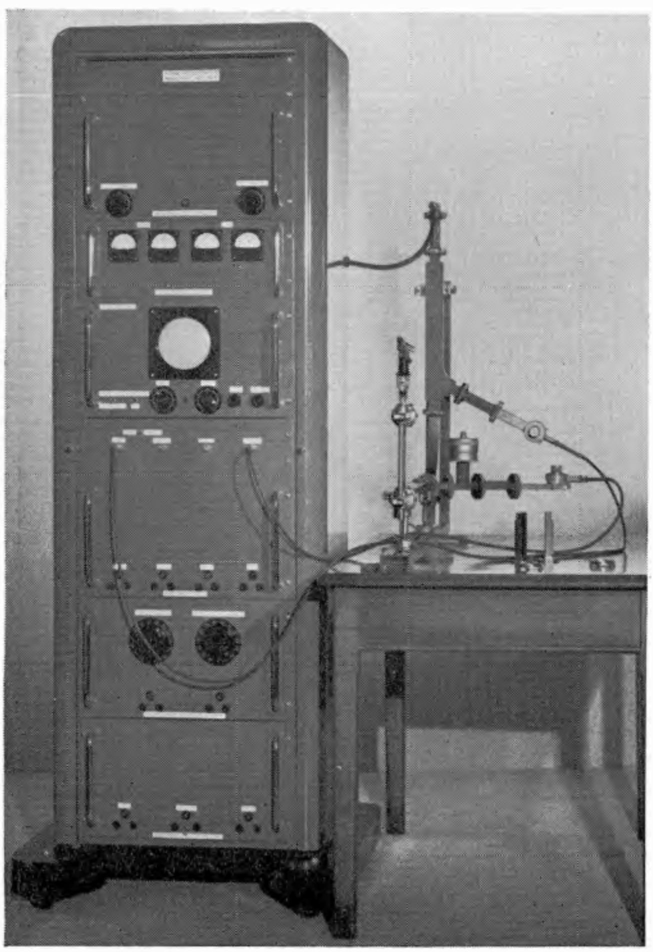
The wavemeter gives, superimposed on the signal trace, an absorption pip by means of which the frequency corresponding to any point on the trace is ascertained. When greater precision is required the output of a high Q transmission type wavemeter may be fed into the pre-amplifier

(19) its response appearing on the base line of the display. When it is necessary to adjust the stabilization, e.g. when changing the range of scan, the switch (15) permits the display of the incident signal in place of the reflection signal.

The electronic equipment of the reflectometer which includes the b.w.o., c.r.t., associated circuits and power supplies is housed in a 6ft rack cabinet. The waveguide circuit is external to the cabinet and is arranged for setting up on an adjacent bench. Photographs of the complete equipment and of the waveguide circuit only are shown in Figs. 2 and 3. The most important components of the waveguide circuit are the two directional couplers since they determine to a large extent the accuracy obtained from the system. They should have as near as possible identical coupling characteristics and high directivities over the operative bandwidth. The couplers used are of the multi slot type[1] with twenty longitudinal and twenty transverse slots. The typical performance curves in Fig. 4 show the performance achieved. The coupling factors of two couplers constructed to the same design do not differ by more than 0·4dB at any point in the band.

Crystal holders having as good a match as possible over the operative band are used in order to avoid appreciable reflection back into the main guide. A tapered ridge to coaxial design is used. A v.s.w.r. of 1·5 has been achieved over the major part of the band but at the ends of the band the v.s.w.r. rises to about 3·0 and therefore, to reduce the reflection into the main guide at these frequencies, it is necessary to insert a matched 3dB attenuator in front of each detector. For the detectors, balanced pairs of

Fig. 2. The complete equipment



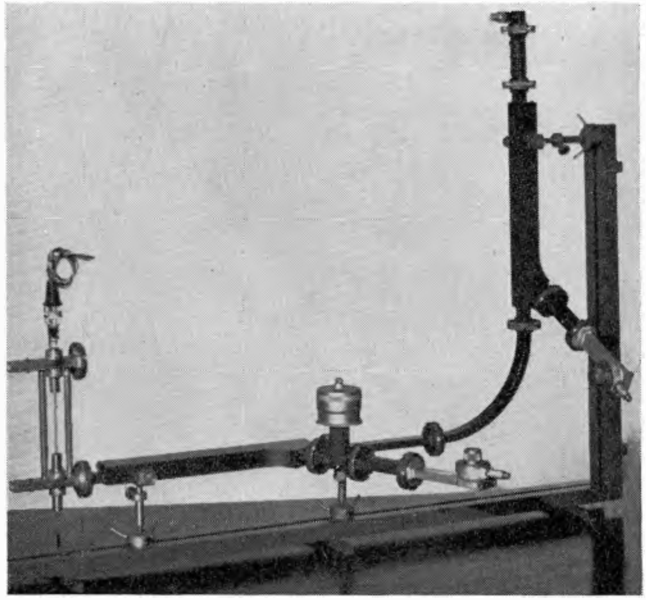


Fig. 3. The waveguide circuit

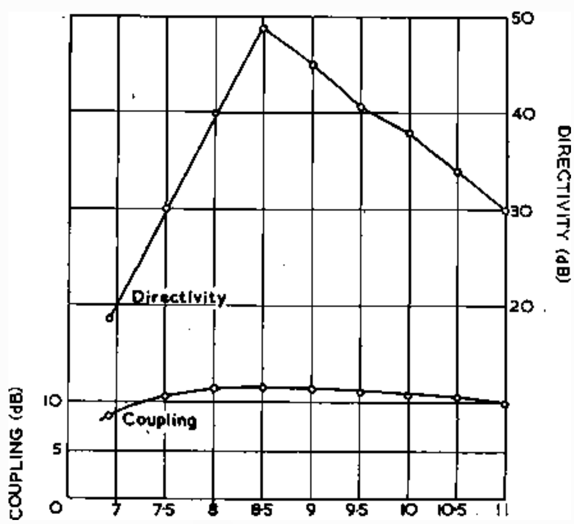


Fig. 4. Performance of X-band directional coupler

crystals must be chosen giving as near as possible identical outputs over the operative bandwidth when subjected to the same r.f. input level. This should be approximately that at which they normally operate in the reflectometer circuit. The b.w.o. output is adequately restricted by the preset attenuator in the input to ensure that the crystals function in their square law region.

Testing Procedures

The most accurate method of using this reflectometer is in conjunction with a calibrated adjustable mismatcher. A record is made of the height of the signal trace from the component under test at desired frequency points in the band under investigation. The test component is replaced by a variable calibrated mismatcher which is then successively adjusted to give corresponding trace heights at these frequencies. The values of the reflection coefficients can be read off from the calibration of the mismatcher. This method gives actual values of reflection coefficient at definite frequencies but it requires considerable time for the performance to be explored in detail.

In fact for most applications the main interest is in knowing that a component has a v.s.w.r. better than a given value, in which case the component may be compared with a preset mismatcher having a mismatch which does not vary appreciably with frequency. Mismatchers having v.s.w.r.'s of 1·5 : 1 and 3 : 1 (reflection coefficients of 0·2 and 0·5) have been constructed by mounting tapered 'lossy' cards in short-circuited sections of guide. Plots of v.s.w.r., reflection coefficient and (reflection coefficient)² against frequency for these devices are shown in Figs. 5 and 6. It can be seen that very nearly constant values of reflection coefficient can be achieved over the frequency range of the reflectometer. The reflection signals are of the form shown in the photographs of Fig. 7 which includes also photographs of the incident signal both stabilized and unstabilized. Due to the various errors in the system these reflection signals exhibit variations above and below the level corresponding to the true value of reflection coefficient. This point is further discussed in a later section.

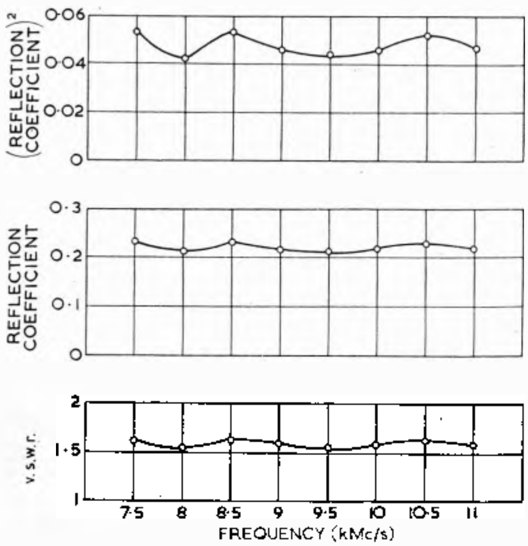


Fig. 5. Performance of standard mismatch—nominal v.s.w.r. = 1·5

The test procedure is to set the mean height of the signal reflected from the appropriate mismatcher to a convenient level by adjustment of the Y amplifier gain. This height is recorded and the mismatcher is then replaced by the load to be tested. A direct comparison of the signal reflected from the load and the calibration level provided by the mismatcher can then be made, and the frequency band over which the load is substantially better or worse than the mismatcher ascertained from the wavemeter.

When the test load has variable circuit parameters it is possible to observe the effect of changes of these on the reflected signal. Suitable adjustments may be effected to optimize the performance. Matching of the input and output circuits of t.w.t.'s and adjustment of the structures of b.w.o.'s has been carried out in this way.

Although primarily designed as a reflectometer, the equipment may be used to measure the magnitude of the transmission coefficient of a four terminal device. To do this an additional crystal detector is connected to the output termination of the test load and the detected signal is fed into the display system in place of the output from the reflection crystal. Alternatively the reflection crystal detector itself may be used. The display on the c.r.t. then shows the transmission characteristic of the test load. Detailed measurements of the transmission coefficient may be made by replacing the test load with a calibrated attenuator (or alternatively comparisons against a preset attenuator can be effected). This facility has been used to measure the attenuation of t.w.t. helices and transmission coefficients of waveguide windows. In addition to these specific applications the reflectometer can be used in an inspection test to provide a rapid check on the general

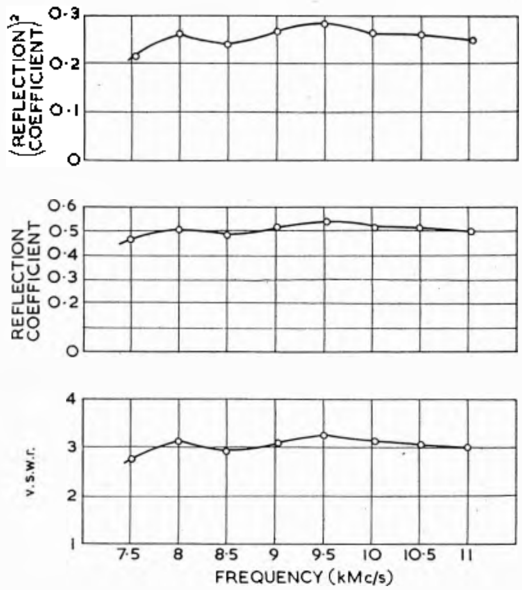
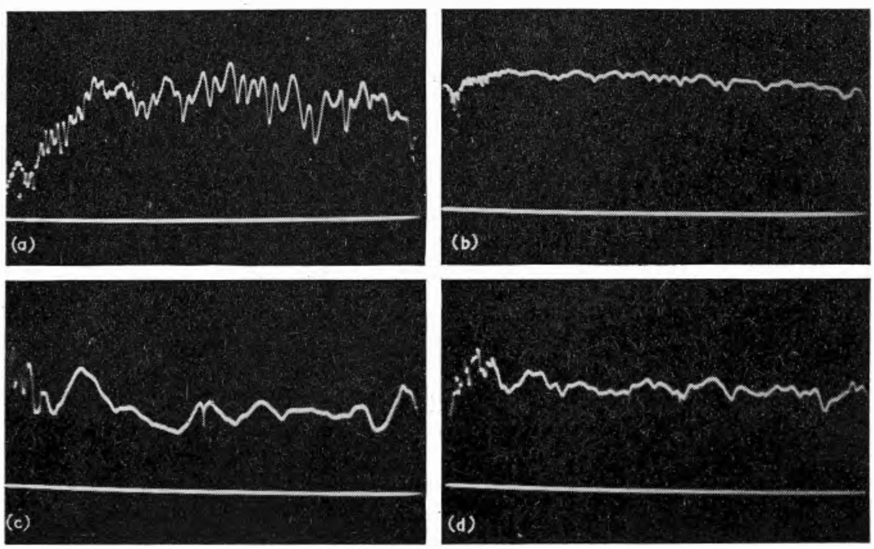


Fig. 6. Performance of standard mismatch—nominal v.s.w.r. = 3·0

Fig. 7. Signal traces on the c.r.t.

(*a*) *Incident signal unstabilized*
(*b*) *Incident signal stabilized*
(*c*) *Reflected signal from* 1·5 : 1 *standard mismatch*
(*d*) *Reflected signal from* 3 : 1 *standard mismatch*



performance of a test component. This is of considerable service in revealing relatively narrow band imperfections such as unwanted resonances or stray couplings which would otherwise quite easily go undetected. A more detailed examination of any narrow band effect that comes to light in this way is possible by reducing the displayed frequency scanning range. In the X-band instrument the minimum range to which the instrument may be set varies between 10 and 15Mc/s over the full tuning range.

Accuracy

The principal factors limiting the accuracy of the display are:—

(1) Finite directivity of the directional couplers

(2) Imperfect balance of the crystal detectors

(3) Incomplete stabilization of the incident power.

In addition there are small errors due to differences in the coupling factors of the directional couplers, internal reflections in the r.f. circuit and variations in frequency characteristics of the attenuators.

On the display from a standard mismatch having a substantially constant reflection coefficient over a wide frequency band the effect of these errors is manifested as an undulation of the signal trace around the mean level which corresponds to the true magnitude of the reflection coefficient. The maximum and minimum values of the undulation indicate points in the frequency band where the various errors combine to give the condition of maximum overall error.

The magnitude of the error introduced by the finite directivity of the couplers can be calculated as shown in the Appendix given the constants of the couplers and the reflection coefficient of the termination. Table 1 shows the calculated errors for two values of reflection coefficient and directivities of 35 and 40dB, these being the measured minimum directivities of the couplers over the bands 8 to 11kMc/s and 8 to 9·5kMc/s respectively.

These results show that substantially increased accuracy is obtained if the display is restricted in bandwidth to the region 8 to 9·5kMc/s where the performance of the couplers is an optimum.

The error introduced by the crystal detectors is mainly due to the unbalance produced in a balanced pair of crystals when the r.f. input level to one of them is changed. Pairs of crystals can be chosen to give substantially identical output-frequency characteristics when operated at the same r.f. level but when this changes, as it must necessarily do at the reflection crystal, there is a change of crystal law and of its dependence on frequency. Thus the crystals no longer have similar characteristics and the accuracy of the system is impaired. This error can be minimized when measuring reflections covering only a small range of values by adjusting the crystal padding attenuators to give approximately equal r.f. inputs to the two crystals for the mean value of reflection coefficient in the range under test. Thus when this procedure is carried out with a mismatch of 3:1 (reflection coefficient of 0·5) the error in the display of reflection coefficients of this order due to crystal unbalance has been shown to be about ± 3 per cent over the 8 to 11Mc/s band.

TABLE 1.

DIRECTIVITY	TRUE		INDICATED		APPROXIMATE ERROR IN REFLECTION COEFFICIENT (PER CENT)
	V.S.W.R.	REFLECTION COEFFICIENT	V.S.W.R.	REFLECTION COEFFICIENT	
35dB	3·0	0·5	2·82-3·19	0·477-0·522	± 4·5
	1·5	0·2	1·44-1·55	0·181-0·218	± 9
40dB	3·0	0·5	2·9-3·1	0·488-0·512	± 2·5
	1·5	0·2	1·47-1·53	0·190-0·210	± 5

Any variation of incident power level over the frequency band degrades the accuracy since the method assumes a constant level of incident power. This cannot be completely stabilized because the magnitude of the feedback which can be applied is limited by phase change in the feedback amplifier. After stabilization therefore there still remains some variation of output with frequency, the amount being roughly proportional to that present in the unstabilized output from the backward wave oscillator. For one b.w.o. over the 8 to 11kMc/s band the variation of stabilized incident power was of the order of 0·6dB while for another it was only of the order of 0·3dB. The resultant error in reflection coefficient therefore ranges between ± 3·5 per cent to ± 1·6 per cent, depending on the particular b.w.o. in use.

Having regard to the variety of these errors it is obvious that the overall error obtained will depend very much on the test procedure adopted. For instance when making measurements with a calibrated mismatcher which is adjusted to give a reflection signal equal to that from the test load the error due to variation of signal level at the reflection crystal is avoided. On the other hand when the reflectometer is used to compare a test load with a preset mismatcher there is some error due to variation of level at this crystal. Tests have shown that in this case the overall maximum error which may be obtained, when reasonable care is taken in the initial adjustment, does not exceed ± 6 per cent on a reflection coefficient of 0·5 rising to ± 12 per cent on 0·2. This is for the case when the crystal pads are preset to give equal detected signals at the two crystals for a mismatch reflection of 0·5 and the display is over the full band of 7 500 to 11 000Mc/s.

Obviously when only a limited band is displayed the accuracy is appreciably increased since not only the errors in the directional couplers but all the other errors tend to decrease with reduction of frequency scan.

At present the accuracy of the system is adequate for existing requirements in the t.w.t. and b.w.o. development field and while higher standards are sometimes required these are usually only over a restricted frequency range. Higher accuracies will however be achieved with the development of broader band directive couplers of higher directivity and these will be useful particularly for measuring low values of reflection coefficient.

Conclusion

This reflectometer has proved to be of great service for work in connexion with microwave valves by reducing enormously the time required to make an assessment of the matching and transmission properties of slow wave structures and associated microwave circuit components. Following its successful use at S- and X-band a model has recently been built for the 6 000 to 8 000Mc/s band. Since the range of backward wave oscillators now available from C.S.F. covers a total frequency band of 1 600 to 11 000Mc/s the system can in fact be applied to any part of this band provided that suitable r.f. components are available. Moreover since the power supply requirements for the various frequency bands are all very similar one equipment can be adapted to cover additional wavebands by the provision of additional backward wave oscillators and the appropriate external r.f. circuits.

Acknowledgment

The authors are indebted to the Admiralty for permission to publish this article.

Appendix

Calculation of the Errors Due to Imperfections of the Directional Couplers

Referring to Fig. 8(a) assume a wave having voltage amplitude V_1 incident at port 1 and waves having voltage amplitudes V_2 V_3 V_4 leaving the ports 2, 3 and 4 which are terminated in perfect matched loads. The coupling K and directivity D can be defined as:—

$$K = V_3/V_1 \qquad (1)$$

$$D = V_2/V_3 \qquad (2)$$

(Both K and D are complex quantities).

By conservation of energy

$$V_1^2 = V_2^2 + V_3^2 + V_4^2$$

$$\text{i.e. } V_4^2 = V_1^2\,[1 - K^2 - K^2D^2]$$

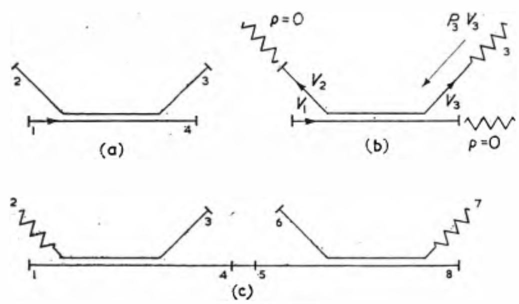


Fig. 8. Directional coupler circuits

(a) Four-arm directional coupler
(b) Three-arm directional coupler
(c) Two three-arm directional couplers as used in the reflectometer

Thus

$$\left.\begin{aligned} V_2 &= KDV_1 \\ V_3 &= KV_1 \\ V_4 &= (1 - K^2 - K^2D^2)^{\frac{1}{2}}V_1 \end{aligned}\right\} \qquad (3)$$

In a three-arm directional coupler the arm designated 3 is terminated in an internally matched load which in practice is imperfect. Let this load have a reflection coefficient $\rho_3 = |\rho_3|\,e^{j\theta}$. A measurement of the effective directivity D_1 made on a three-arm coupler must include the effect of this load.

Referring to Fig. 8(b) and using the definition of directivity (2) $D_1 = V_2/V_3$, where V_2 is made up of two voltages

(a) $= KDV_1$ due to V_1

(b) $= \rho_3 V_3 = \rho_3 KV_1$

due to reflection from load terminating arm 3 (ignoring power coupled to arms 1 and 4)

$$\text{i.e. } D_1 = \frac{KDV_1 + \rho_3 KV_1}{KV_1} = D + \rho_3$$

Thus since these are vector quantities D_1 will have maximum and minimum values given by

$$|D_1| = |D| \pm |\rho_3| \qquad (4)$$

The reflectometer arrangement consists of two such couplers arranged back to back as shown in Fig. 8(c) with detectors on arms 3 and 6 which are assumed to have perfect match for the purpose of the analysis. Let V_2 to V_8 be the voltages at the terminations of arms 2 to 8 due only to the wave V_1 in arm 1, if all arms are perfectly matched and both the couplers have the same values for K and D as defined in equation (1) and (2).

Then the values of V_2 to V_8 are as follows, neglecting small terms proportional to K^2 and D^2

$$\begin{aligned} V_2 &= DV_3 = DKV_1 \\ V_3 &= KV_1 \\ V_4 &= (1 - K^2)^{\frac{1}{2}}V_1 \\ V_5 &= V_4 = (1 - K^2)^{\frac{1}{2}}V_1 \\ V_6 &= DV_7 = DK(1 - K^2)^{\frac{1}{2}}V_1 \\ V_7 &= KV_5 = K(1 - K^2)^{\frac{1}{2}}V_1 \\ V_8 &= (1 - K^2)V_1 \end{aligned} \qquad (5)$$

If now arm 8 is terminated in a load having a reflection coefficient $\rho = |\rho|\,e^{j\phi}$ and arms 2 and 7 are terminated in loads having small reflection coefficients

$$\rho_2 = |\rho_2|\,e^{j\eta} \text{ and } \rho_7 = |\rho_7|\,e^{j\lambda}$$

Then ignoring small terms the voltage at arm 3 is made up of three terms

(a) $= KV_1$ due to forward wave in arm 1.

(b) $= DK(1 - K^2)^{3/2}\rho V_1$ due to coupling (by directivity to backward wave at 4).

(c) $= K(1 - K^2)^{3/2}\rho\rho_2 V_1$ due to reflection from termination at 2.

$$\text{i.e. } V_3 = KV_1[1 + \rho(1 - K^2)^{3/2}(D + \rho_2)]$$

thus from equation (4)

$$V_3 = KV_1[1 + \rho(1 - K^2)^{3/2}D_1] \qquad (6)$$

i.e. where D_1 is the measured directivity.

Similarly at 6 there are three waves.

(a) $= K(1 - K^2)\rho V_1$ due to coupling to backward wave.

(b) $= DK(1 - K^2)^{\frac{1}{2}}V_1$ due to coupling by directivity to forward wave at 5.

(c) $= K(1 - K^2)^{\frac{1}{2}}\rho_7 V_1$ due to reflection from load 7 which is coupled to forward wave.

$$\text{i.e. } V_6 = KV_1[(1 - K^2)\,\rho + (1 - K^2)^{\frac{1}{2}}(D + \rho_7)]$$

$$V_6 = KV_1[(1 - K^2)\,\rho + (1 - K^2)^{\frac{1}{2}}D_1] \qquad (7)$$

This assumes $\rho_7 = \rho_2$

The indicated reflection coefficient is proportional to

$$V_6/V_3 = \frac{(1 - K^2)\,\rho + (1 - K^2)^{\frac{1}{2}}D_1}{1 + \rho(1 - K^2)^{3/2}D_1} = \rho(1 - K^2)\left[\frac{1 + (1 - K^2)^{-\frac{1}{2}}D_1/\rho}{1 + (1 - K^2)^{3/2}D_1\rho}\right] \qquad (8)$$

In practice provided that K does not vary with frequency the factor $(1 - K^2)$ outside the square brackets does not constitute an error and since the test procedure allows for the adjustment of the overall proportionality factor the indicated value of $\rho = \rho'$ may be taken as

$$\rho' = \rho\left[\frac{1 + (1 - K^2)^{-\frac{1}{2}}D_1/\rho}{1 + (1 - K^2)^{3/2}D_1\rho}\right]$$

Furthermore, since in this case $K^2 = 1/10$ the factors $(1 - K^2)^{-\frac{1}{2}}$ and $(1 - K^2)^{3/2}$ approximate to unity

$$\text{i.e. } \rho' = \rho\left[\frac{1 + D_1/\rho}{1 + D_1\rho}\right]$$

since these are vector quantities ρ' will have maximum and minimum values given by

$$|\rho'| = |\rho|\left[\frac{1 \pm |D_1|/|\rho|}{1 \mp |D_1|\,|\rho|}\right] \qquad (9)$$

Reference

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A Vibrating-Head Scanner for Inspection and Indexing of Magnetic Recordings

By J. S. Gill*, B.Sc.(Eng.), A.M.I.E.E.

When ti is necessary to read a stationary or very slowly moving magnetic tape a conventional playback system cannot be used because the output is proportional to speed and falls to zero when the tape is stationary. In this article a vibrating playback head is described which can be used for reading stationary magnetic recordings. The principle has been successfully applied to the extraction of larynx excitation from recorded speech by direct visual interpretation of the waveform.

(Voir page 58 pour le résumé en Français: Zusammenfassung in deutscher Sprache auf Seite 62)

MAGNETIC tape is now used for the recording of many types of signals. The usual method of playback may be explained by consideration of an 'idealized' system in which the longitudinal variation of flux on a recorded tape is proportional to the voltage waveform of the recording signal. During normal playback the recorded tape is drawn past a head at the same constant speed as used during the recording process. The voltage generated within the head is proportional to the rate of change of flux linking the windings and consequently the signal from the head is proportional to the rate of change of the original signal. An equalizer with an 'integrating' characteristic, that is to say a network whose loss increases by 6dB/octave throughout the range of frequencies recorded, must be included in the playback system if a replica of the original signal is required. In practical systems it is usually necessary to provide additional equalization to compensate for losses[1].

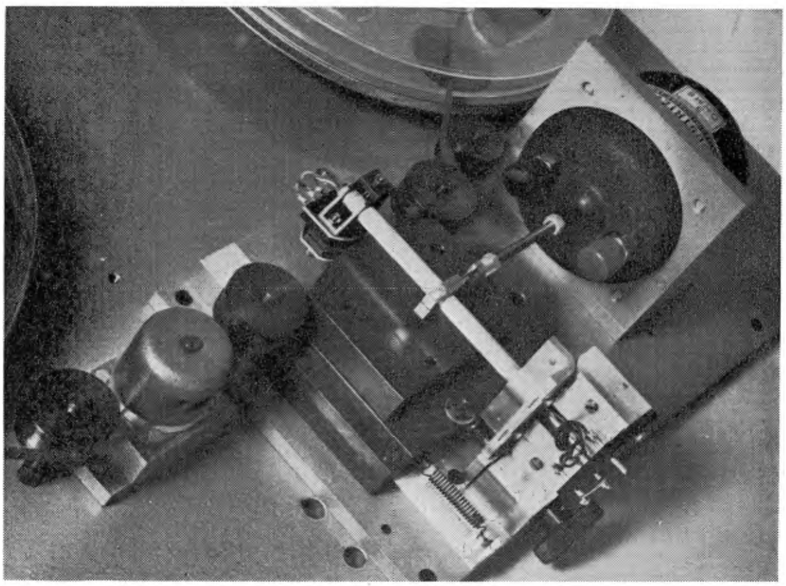

Fig. 1. The vibrating head mechanism

When it is necessary to read a stationary or very slowly moving tape a conventional playback system cannot be used because the output is directly proportional to playback speed and falls to zero when the tape is stationary. Flux sensitive playback systems have been developed to overcome this difficulty[2].

However, if a persistent oscillographic display of the waveform recorded over a short length of stationary tape is required the head must repeatedly scan the tape. Consequently a flux sensitive head is not essential for this application. A scanned display of this type may be used for detailed examination of the recorded signal. Events of particular interest may be indexed by synchronously recording pulses on another track of the same tape. The technique described in this article was developed for this purpose.

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Method

Stationary recorded tape is scanned cyclically by a playback head and the output is displayed on an oscilloscope with a time-base synchronized to the head motion. A monostable pulse generator, which can be triggered at any point on the display, is used for indexing.

A rotating playback head system could probably have been successfully developed for the scanner, but at the time the most convenient solution was that shown in Figs. 1 and 2. A conventional twin-track head, driven by a small vibration generator, vibrates just out of contact with the tape. With a system of this type head wear is completely absent and there is no possibility of varying frictional drag between tape and head causing 'jitter' on the display. Since both head and time-base execute the same simple harmonic motion the display is free from any distortion of the time scale. An 'integrating' playback characteristic is employed. This compensates for the effect of the variable scanning velocity on the amplitude scale of the display at all points except those near the extremities of the scan. In practice forward and reverse traces are not exactly superimposed and so for this reason one direction of scan only is displayed.

Distortion at the extremities of the scan could be reduced by replacing the conventional head and equalizer with a flux sensitive head[2]. In neither of the applications discussed later in this article has such a modification been considered to be necessary.

At the time of development of this apparatus (mid 1956) the use of an oscillating head for tape readout was

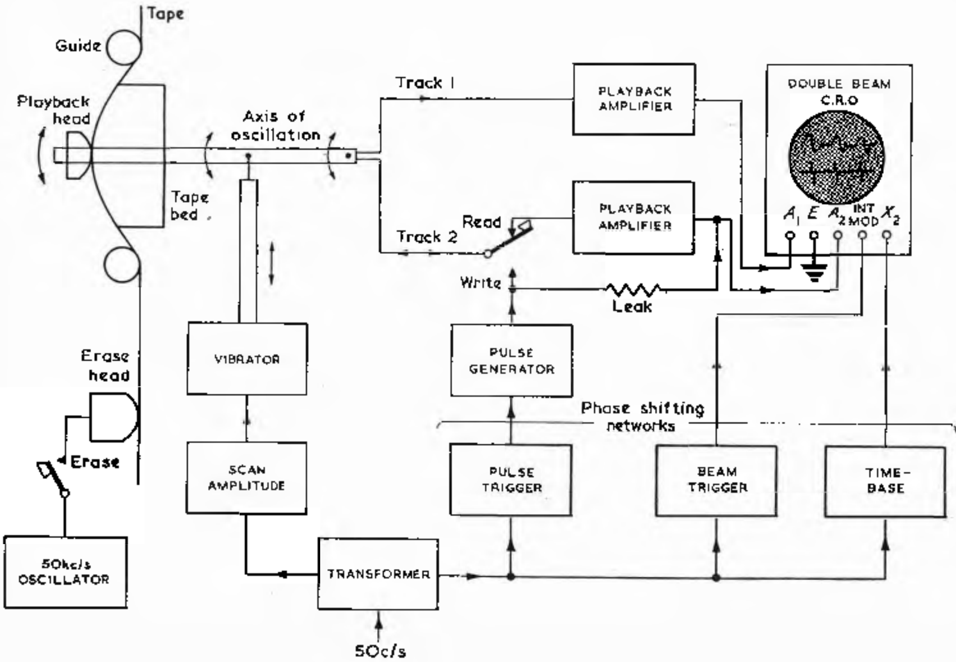

Fig. 2. General arrangement of the apparatus

believed to be novel but references[3,4] to similar schemes have since been noted.

Design Details

The vibration system is as shown in Fig. 1. A conventional 0·00025in gap twin track head is clamped to a light alloy tube attached to a hinge. Essentially this hinge consists of two spring steel suspension units. Each unit comprises two strips (0·006in thick) mounted at right-angles and clamped in slots milled in the two halves of the hinge. By this method bending of the strips is confined substantially to the axis of oscillation. Head movement is confined to an arc of radius 3in in the horizontal plane. The system is driven by a small vibration generator from the 50c/s supply. A total head excursion of ¼in can easily be obtained.

The recorded tape passes between the vibrating head and a cylindrical surface of radius of curvature 3in against which it is held by its own tension. Tape to head clearance must be kept to an absolute minimum. This requirement has been met by mounting the cylindrical tape bed on a dovetail slide system actuated by a differential screw of effective pitch of 0·001in. Two helical springs load the differential screw and thus eliminate backlash. The vibrating system sub-assembly is bolted to the base plate on which are fixed the tape guides, spool mountings and an erase head.

The associated electrical equipment is shown on the block schematic (Fig. 2). Scan amplitude is controlled by a tapped potentiometer and the phase shifters for time-base and beam trigger inputs to the oscilloscope are preset. Phase adjustment of the pulse generator triggering circuit is provided for use when indexing. Track 1 output is displayed on beam 1 of the oscilloscope and the position of the indexing pulse is shown on beam 2. A 50kc/s oscillator and erase head are provided for erasure of any errors which may be recorded on track 2 of the tape.

Operation

Initially the time-base and beam trigger phase shifters are set to synchronize head and spot motions and the pulse generator is triggered at approximately mid-scan. Head to tape clearance is set to the minimum possible consistent with freedom from 'jitter' and the tape is drawn past the head until the required waveform is approximately at the centre of the scan. Any point of interest on the waveform can be indexed by first triggering the pulse generator at the same point on the scan and then using it to mark track 2 synchronously with the event.

Applications

The tape scanner was designed primarily for the extraction of larynx excitation from recorded speech. The larynx, when vibrating, may be considered to excite the vocal tract with a quasi-periodic pulse train. The pulse shapes are modified by the transfer function of the tract but it is still possible to make a reasonable estimate of their times of occurrence by examination of the recorded speech waveform. In this application the indexing facility is used to mark a second track of the tape at times corresponding to the 'larynx pulses'. Recordings of this type have been of considerable use during investigations of methods of analysis-synthesis telephony.

Stationary digital recordings have been read with the same apparatus. Binary signals recorded at 1 000 digits per inch on tracks 0·1in wide have been correctly resolved. At these high packing densities it is convenient to display only the central portion of the scan since ¼in would correspond to 250 digits. In these circumstances the scanning velocity is substantially constant over the displayed length and more sophisticated playback equalization can be employed if required.

The apparatus described in this article has been in service since mid 1956 and its performance has been wholly satisfactory.

Acknowledgments

The spring steel suspension unit used in the vibration assembly was suggested by Mr. R. J. Jury and Mr. S. H. F. Parrott of the Post Office Research Station, Dollis Hill.

Mr. J. Lennan constructed the mechanism shown in Fig. 1 and also made several most helpful suggestions concerning the mechanical design.

This apparatus was developed for use in research at the Joint Speech Research Unit of the Post Office, Eastcote, and permission to make use of the information is acknowledged.

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AN ELECTRONIC RATIO CALCULATOR

By A. D. Boronkay, M.Sc. (Eng.), Grad. I.E.E.

In electronic computers all-electronic ratio circuits are seldom used due to the difficulty of finding a strict analogue of a quotient. The system described in this article is based not on a strict physical analogue but on an approximation valid within certain limits; the principle being that an approximately linear relation exists between the phase-angle of the sum of the two signals and the amplitude ratio. The ratio measuring device consists of an adding circuit which sums the numerator and denominator, a phase sensitive circuit to provide an electrical quantity proportional to the sum vector and a phase sensitive controlled rectifier.

(Voir page 58 pour le résumé en Français: Zusammenfassung in deutscher Sprache auf Seite 62)

ALL-ELECTRONIC ratio circuits are rarely used in computers because of the difficulty of finding a strict analogue of a quotient. Even the system examined in this article is not based upon a strict physical analogue, but on an approximation valid within certain limits. It was developed originally for use in spectrophotometry[1,2,3] and was later suggested by the author for use in computers[4].

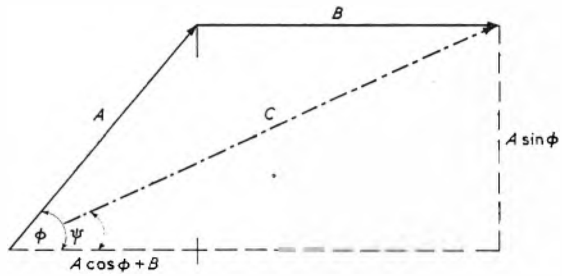


Fig. 1. Vectorial diagram of the addition of numerator A and denominator B

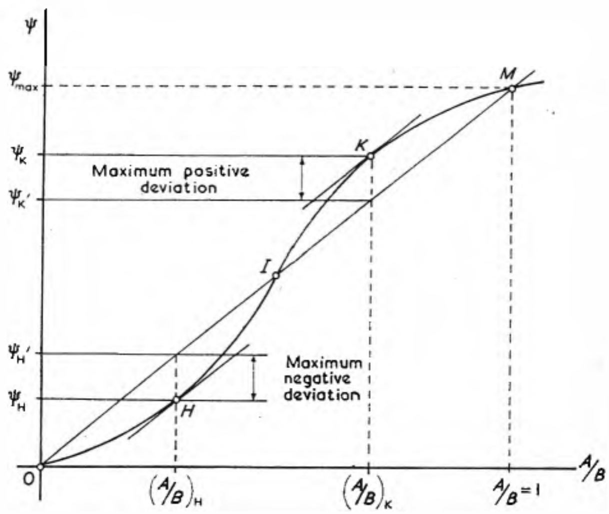


Fig. 2. Phase angle ψ of the sum vector as a function of the A/B ratio (heavily distorted)

Let the components of the ratio to be calculated be converted into two sine wave signals of the same frequency (ω) and a predetermined phase-angle (ϕ) between them. The principle of the present system is that an approximately linear relation exists between the phase-angle of the sum $\mathbf{c} = \mathbf{a} + \mathbf{b}$ of the two signals

$$\mathbf{a} = A\sin(\omega t + \phi)$$
$$\mathbf{b} = B\sin\omega t$$

and the amplitude ratio A/B. It is easy to see from the vectorial diagram (Fig. 1) that the relationship between ψ and A/B can be represented by the function

$$\psi = \tan^{-1}\frac{\sin\phi}{\cos\phi + B/A} \quad \ldots\ldots\ldots\ldots \quad (1)$$

as is proved in Appendix 1, equation (1) does not represent a linear relationship but approximates to one under certain conditions. It is desirable that the value of the ratio to be measured be positive and less than unity:

$$0 < A/B < 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

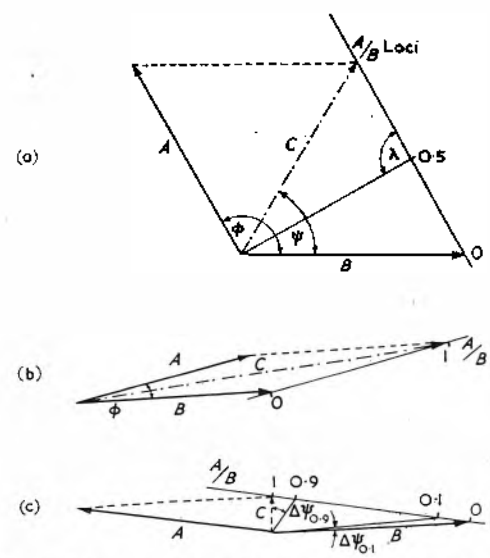


Fig. 3. Effect of variation of the angle ϕ between A and B
(a) optimum (b) too small (c) too large angle

since the value $A/B = 0$ corresponds to $\psi = 0$, and the curve of the function $\psi(A/B)$ becomes rapidly bent above $A/B = 1$. To determine the condition of linearity it is necessary to examine the derivatives of equation (1) from which it can be established that the curve has an inflexion at $A/B = -\cos\phi$. The average deviation from a straight line is smaller if this inflexion is within the range mentioned in condition (2), (see Fig. 2 and Appendix 2).

To fulfil the requirement (2) the phase angle ϕ must be obtuse. In Fig. 3(a) the line A/B is the loci of vectors representing different A/B values. The diagram shows the sum vector at a right-angle to the A/B line in the ψ value of inflexion. If ϕ is small (Fig. 3(b)) there is no inflexion in the measuring range; and the change of ψ corresponding to A/B will be almost linear, as a small section of any curve may be substituted by a straight line. If ϕ is too large (Fig. 3(c)) the linearity becomes poor. The diagram shows the obvious difference between the two angle proportions belonging to the same ratio change. On the other hand the system of better linearity (Fig. 3(b)) is less sensitive than the non-linear one (Fig. 3(c)) if the sensitivity is defined by the value of $\psi_{max} = \phi/2$.

$$\left(\text{Since the sensitivity is } \frac{\Delta\psi}{\Delta(A/B)} \simeq \frac{\psi}{A/B} \simeq \frac{\psi_{max}}{1}\right)$$

Nevertheless, a compromise can be made between the demands of sensitivity and linearity, as for instance the

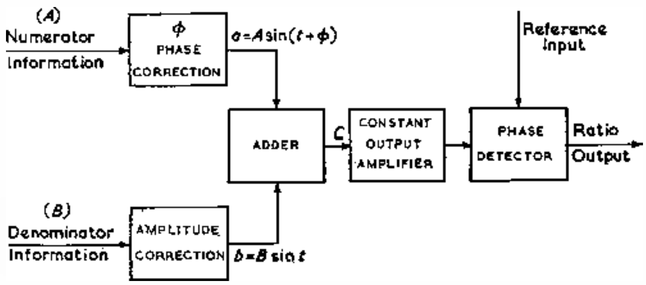


Fig. 4. Block diagram of the ratio calculator

arrangement in Fig. 3(a) which takes $\phi = 120°$, as this is the only solution to put the inflexion in the middle of the range. The deviation of this curve from the straight line is 0 in three points: $A/B = 0$, $A/B = 0{\cdot}5$ and $A/B = 1$ and less than $\pm$ 2 per cent below and above the inflexion point. (For calculation of error see Appendix 3).

Principle of Operation

The ratio measuring device thus reminds one of conventional phase measuring equipment[5,6]. It consists of an adding circuit, which sums the numerator with the denominator, and a phase sensitive circuit to produce an electrical quantity proportional to the phase-angle of the sum vector. The block diagram (Fig. 4) shows this function. The adder produces the $\mathbf{c} = \mathbf{a} + \mathbf{b}$ signal, the amplitude of which has to be made independent of the amplitudes A and B by a constant output amplifier. After this the signal is rectified by a phase sensitive controlled detector.

When using sine wave signals most types of synchronous rectifiers produce a current proportional to the cosine of phase-angle between signal and reference inputs[7,8,9]. (See Appendix 4.) If square wave signals are used (or squaring units to transform sine waves into square waves) the current obtained from the synchronous rectifier will be a linear function of the phase-angle, that is to say it will truly represent the ratio function given in equation (1). This arrangement is sometimes useful, in particular if the

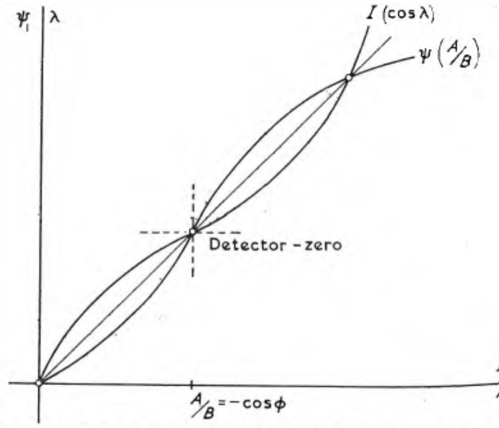
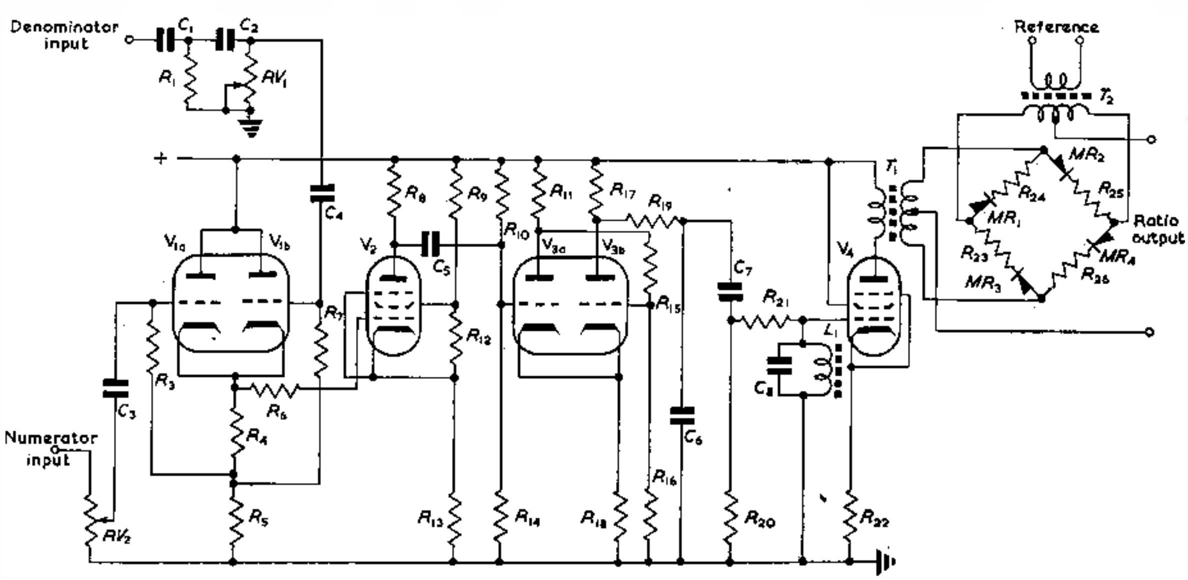


Fig. 5. Cancellation of the ψ (A/B) curve by the cosine characteristic of the rectifier

current is supposed to drive immediately a meter or a recorder[1] since the curvature of function (1) may be eliminated by calibration.

The non-linearity is of the order of some few per cent, which can be reduced. A handy method is given by the cosine characteristic of the phase detector. The $\psi(A/B)$ curve is reminiscent of a small part of an inverted cosine curve with its zero point corresponding to the detector zero to $A/B = -\cos\,\phi$ along the A/B axis, and to the corresponding value along the ψ axis (Fig. 5). If it is presumed that the reference signal is in phase with the numerator signal, $\mathbf{c}$, the sum vector, consequently becomes perpendicular to $A/B = 0{\cdot}5$. At 0 and 1 the angle

Fig. 6. The circuit diagram



between **c** and the reference will be $-30°$ and $+30°$ respectively. Since the curvature has an opposite tendency to the cosine curve of the phase detector, the two deviations from linearity cancel each other out. The maximum deviations of the cosine curve from the line running through $-30° - 0 - +30°$ are about $+$ and $-17°$ and their values are approximately $-$ and $+$ 1·9 per cent, which values correspond very conveniently to the above mentioned data given to the $\psi(A/B)$ curve (Fig. 5).

The Circuit

The circuit shown in Fig. 6 represents a simplified modification of the suggested system. The control grids of V_1, a double triode adder, are fed by the denominator and numerator information. The input circuits are provided with an amplitude control RV_2 and with a phase shift network producing the accurate 120° setting: $C_1C_2R_1RV_1$.

The output of the adder is connected to the amplifier V_2, the square wave output of which is differentiated by $C_5R_{10}R_{14}$ to drive the multivibrator, V_3. The uniform signal leaving the band-pass filter $R_{19}C_6C_7R_{20}$ and the tank-circuit C_8L_1 controls the ring-demodulator by the power-amplifier V_4. All values are conventional, according to the chosen frequency. The reference generator which drives the ring demodulator has a highly stabilized output. The same generator provides the rest of the computer with its carrier frequency.

Performance

The circuit shown in Fig. 6 operates with an acceptable linearity and accuracy if the amplitude of the denominator signal is between 3 and 30V r.m.s. The represented ratio between 0 and 1 is within $\pm$ 0·002 (0·2 per cent). The long term stability of the zero depends on the properties of the rectifier diodes; germanium diodes giving the best performance.

Appendix

1. The sum of vectors **a** and **b**:

$$\mathbf{c} = \mathbf{a} + \mathbf{b} = A\,.\,\sin(\omega t + \phi) + B\,.\,\sin\omega t$$
$$= (B + A\cos\phi)\sin\omega t + (A\sin\phi)\cos\omega t$$

If the angle between the vector **c** and the sine-co-ordinate is ψ it follows that:

$$\tan\psi = \frac{c\,.\,\sin\psi}{c\,.\,\cos\psi}$$
$$= \frac{A\,.\,\sin\phi}{B + \cos\phi}$$

thus

$$\psi = \tan^{-1}\frac{\sin\phi}{B/A + \cos\phi} \quad \ldots\ldots\ldots\ldots (1)$$

2. The derivatives of the above functions are:

$$\frac{d\psi}{d(A/B)} = \frac{\sin\phi}{(A/B)^2 + 2A/B\cos\phi + 1} \quad \ldots\ldots\ldots (3)$$

$$\frac{d^2\psi}{d(A/B)^2} = \frac{2\sin\phi\,(\cos\phi + A/B)}{[(A/B)^2 + 2(A/B)\,.\,\cos\phi + 1]^2} \quad \ldots (4)$$

The second derivative is zero if $-\cos\phi = A/B$, from which the ϕ co-ordinate of the inflexion points can be obtained.

3. A straight line may be put through the points $A/B = 0$, $A/B = -\cos\phi$ and $A/B = 1$. The maximum deviation from this line occurs at those points where the tangent of the curve is parallel to the line, so

$$d\psi/d(A/B) = \psi_{max}/1 \quad \ldots\ldots\ldots\ldots (5)$$

From equation (3), one obtains $(B/A)_H$ and $(B/A)_K$ (Fig. 2) by calculating the roots of the equation:

$$(A/B)^2 + (A/B)\,.\,2\cos\phi + (1 - (\sin\phi/\psi_{max})) = 0 \quad \ldots (6)$$

The actual error for example in H is:

$$\frac{\psi_H - \psi_H'}{\psi_{max}} = \frac{\dfrac{\sin\phi}{(B/A)_H + \cos\phi} - \psi_{max}\,.\,(A/B)_H}{\psi_{max}} \quad \ldots (7)$$

In the case of $\phi = 120°$:

$$\psi_H = 12{\cdot}25°$$

and

$$\psi_H' = 13{\cdot}38°$$

The error:

$$-1{\cdot}13/60 = -1{\cdot}9 \text{ per cent}$$

4. The output of an ideal phase detector is determined by integration over the time of one cycle of the multiplied signals.

(a) If the two signals are $S\,.\,\sin\omega t$ and $R\,.\,\sin(\omega t - \lambda)$, the output current of the detector becomes:

$$I = k_1/T\int_0^T S\,.\,\sin\omega t\,.\,R\,.\,\sin(\omega t - \lambda)\,dt =$$
$$= k_2/T\int_0^T \sin\omega t\,.\,\sin(\omega t - \lambda)\,dt =$$
$$= k_3\,.\,\cos\lambda \quad \ldots\ldots\ldots\ldots\ldots\ldots (8)$$

(The factors k represent electrical parameters of the circuit).

(b) When the two signals are supposed to be square waves:

$$s = S\Big|_{t=0}^{t=T/2}\;;\; s = -S\Big|_{t=T/2}^{t=T}$$

and

$$r = -R\Big|_{t=0}^{t=\tau}\;;\; r = R\Big|_{t=\tau}^{t=T/2}\;;$$
$$r = +R\Big|_{t=T/2}^{t=T/2+\tau}\;;\; r = -R\Big|_{t=T/2+\tau}^{t=T}$$

where $\tau = (\lambda/2\pi)$ and λ is the phase-angle (in 1st quadrant only). $(\lambda = \pi - \phi + \psi)$.

The output current:

$$I = k_4/T\Big[\int_0^{\tau} -R\,Sdt + \int_{\tau}^{T/2} R\,Sdt + \int_{T/2}^{T/2+\tau} -R\,Sdt + \int_{T/2+\tau}^{T} R\,Sdt\Big] =$$

$$k_4\,.\,R.S.\,\frac{T - 4\tau}{T} = k_5\,(1 - (2\lambda/\pi)) \quad \ldots\ldots\ldots\ldots (9)$$

Thus I is a linear function of λ, $I = 0$ if $\lambda = \pi/2$ and $I = 1$ if $\lambda = 0$.

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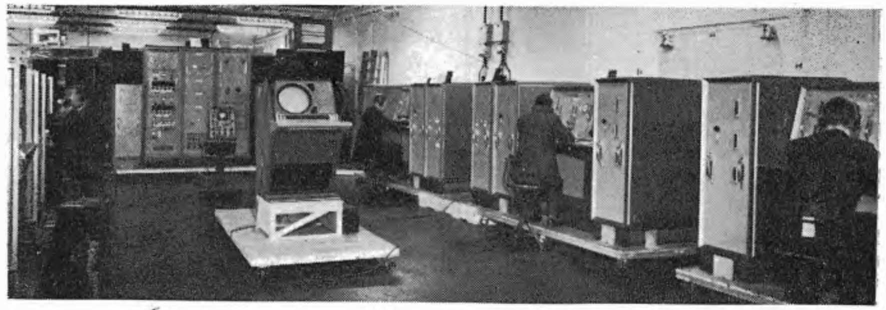
A Radar Simulator System for the Italian Army



A RADAR simulator system has been developed by Solartron, Farnborough (Hants), and the first model, valued at £125 000, has recently been despatched to the Italian Army authorities. It will enable fully realistic training and exercises to be performed without having to use operations staff, airfields, aircraft and radar stations and without interfering with civil air routes.

The Solartron system uses analogue computers to simulate targets, radar stations, attacking and defensive aircraft and electronic counter-measures, in the Italian scheme. Surface vessels or submarines, wholly or partly submerged and using Asdic, shipborne radar, tidal motions, height-finding radars and air traffic control systems could also be simulated if required in other Solartron radar simulator systems.

The system consists of a number of desk consoles with control panels, a central rack system and two control equipments each fitted with radar plan position indicators on which aircraft and electronic radar defence measures are seen as blips and marks.

The two control panels at the back of the desk are identical and represent two 'aircraft'. On the right- and left-hand sides from the floor to shoulder height are two computer units immediately concerned with the 'aircraft'.

The 'pilot' sits opposite his panel. If necessary one 'pilot' may control two 'aircraft'.

The complete system for the Italian Army consists of seven consoles, making fourteen 'aircraft', which may be allocated to be 'belligerent' or 'interceptor' aircraft as required. In addition, four targets are 'window' jammers to simulate the dropping of reflecting metal foil and two are parachuted noise jammers. The dropping of noise jammers is done by a keyswitch.

The simulators for these effects are in the side computers and their effects fed into the main system. The signals from a flying-spot scanner are processed by suitable noise and function generators to simulate video response from metal strips of varying characteristics and fed to the p.p.i's. of the main controls. The X and Y co-ordinates of wind velocity are fed-in electronically. In other words, the drift of metal strips by the wind in the sky is simulated.

The central control rack is largely composed of the main supplies and the main switch controls.

Two large consoles stand by themselves behind the computers or may be installed away from the 'aircraft' console room. They have plan position indicators on which the 'aircraft' may be seen in their courses, and the effect of 'window' and noise jammers seen on the location of the 'aircraft' by the simulated radar system.

One console is fitted for 'fighter and bomber' control. The other can be 'fighter' or 'bomber'. The controller of each console is in communication with his 'pilots' through the usual headphone-throat microphone arrangement. The equipment plugs into a socket under the desks of the desk-consoles.

Naturally, when simulating an air battle, one controller at one console would be in communication solely with his type of flight—'fighters' or 'bombers', and the other controller at the other console similarly in control. They speak to their respective flights or to individual aircraft through the switching arrangement and their hand-microphones.

When a pupil is being instructed he is connected only to the 'aircraft' allocated to him. The instructor can work his own section of 'aircraft' or radar system and talk to and supervise his pupil in the ordinary way.

The 'aircraft' are grouped in pairs so that one operator may control two aircraft. After the initial starting position, speed, rate of turn, dive and climb have been set at their maximum and minimum limits, the aircraft, on the command of the controller, go into action. Indication of the aircraft position on a cartesian grid, heading angle and height are given to the operator by means of the edge-on dials. Aircraft slant range, bearing and elevation angle are indicated on similar dials. The flank computers enable the true shape of both horizontal and vertical antenna lobes, including side-lobes, to be reproduced. Certain 'aircraft' can drop 'window' or parachuted noise jammers, which gradually sink to the ground as in real life.

The exercise is conducted by two controllers, one in charge of the offensive aircraft and the other of the interceptor aircraft. On operation, the controller of the offensive aircraft would be the tactical control instructor, and his pupils would man the interception control displays as ground controllers or radar operators according to the training being carried out.

The radar simulator system described is that designed to meet the customer's requirements. There are many additional facilities.

Air targets may be calibrated in Mach numbers and they may be fitted with 'fuel expended' and 'fuel left' indicators, to which are tied engine boost and aircraft height values.

The delay in getting off the ground after receiving the signal to attack at the air station—'scramble and take-off'—may be simulated.

Track computers as described are used when surface vessels or submarines are simulated for marine radar problems and exercises.

The surface vessels and fully or part-submerged submarines are simulated with their various limiting factors on the desk panels as are aircraft. They give their radar echoes or indication of position, except submarines, when submerged. However, Asdic can then be simulated and the underwater position indicated.

In most marine applications the radar would be mounted on 'ships' and the information displayed can be given as unstabilized or stabilized relative motion or as true motion and the effect of the tides also included.

The above illustration shows a general view of the simulator

A Temperature-Stabilized Photo-Transistor Relay Circuit

By J. C. Anderson*, M.Sc., A.M.I.E.E., A.M.Brit.I.R.E., and T. Winer*

A junction photo-transistor can be used to operate a relay directly but the performance obtained varies widely with ambient temperature. In this article a circuit is described in which a photo-transistor and an ordinary junction transistor are used in a 'long-tailed pair' arrangement. In this way the variation in dark current with change in ambient temperature is much reduced, while high sensitivity is achieved; the transient response is also shown to be good.

(Voir page 58 pour le résumé en Français: Zusammenfassung in deutscher Sprache auf Seite 62)

THE advent of the commercially produced junction photo-transistor has brought with it the possibility of operating a relay directly from the light-sensitive element. This has already been put into commercial use in a number of ways, but the common failing of all circuits relying on the photo-transistor alone is variation of performance with ambient temperature. This is particularly troublesome in climates where the ambient temperature varies between very wide limits, as is the case on the South African high veld. A simple and cheap design in which this effect was minimized, at the same time maximizing sensitivity, was sought.

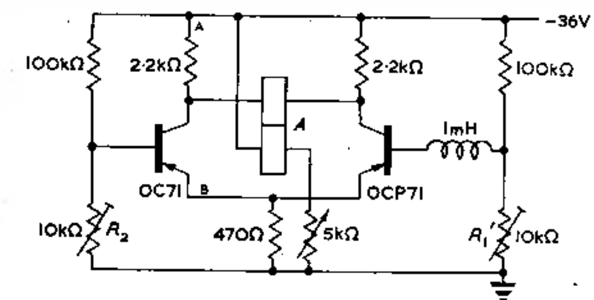


Fig. 1. Practical transistor circuit

The Circuit

The well-known 'long-tailed pair' valve circuit has been translated directly into a transistor circuit and is shown in Fig. 1. A basic feature is the discrimination against in-phase changes, i.e. changes affecting both transistors equally, such as a rise in ambient temperature. Increase in current through both transistors simultaneously does not affect the current through the relay coil connected between the collectors.

When the photo-transistor is illuminated the current through it increases the p.d. across the common emitter resistor so applying an effective voltage to the emitter of the ordinary transistor. This gives a base bias in such a direction as to decrease the current through the transistor. Thus, at the collectors, one end of the relay coil goes down in voltage while the other end goes up, applying a p.d. across the coil which is sufficient to drive the operating current through it.

The steps in empirical design of the circuit are as follows: The circuit is broken at points A and B, and a milliammeter is connected in series with the collector of the OCP71. The base bias resistor R_1 is then adjusted for maximum change of current between dark and light, which was found to be about 5mA using illumination from a 40W bulb at a distance of about four inches. The connexions at A and B are then made and R_1 is re-adjusted to bring the dark current back to the same value as when the connexions were broken. The meter is then transferred to be in series with the collector of the OC71 and its base bias resistor R_2 is adjusted for maximum change in current between dark and full illumination of the OCP71 The second stage of adjustment of R_1 is then repeated followed by re-adjustment of R_2, and the alternate re-setting of the two resistors is repeated until the current changes in both the transistors are the same on illumination. The meter is transferred to between the collectors and the total current change on illumination is measured. With perfectly matched transistors it should, of course, be possible to get the current changes in each transistor to be identical, providing high-precision components are used.

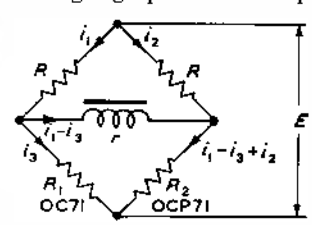


Fig. 2. Equivalent bridge circuit

In practice, using normal 20 per cent tolerance resistors and a pair of transistors chosen at random, the current changes were within 10 per cent of each other.

Theoretical Design

From the above description it will be appreciated that, in a correctly adjusted and perfectly matched circuit, the p.d. across the common emitter resistor will remain substantially constant. Also, due to the low resistance of the relay coil, the p.d. across it when the operating current flows will likewise be small; it may be assumed therefore that the transistor currents are not affected by these small changes in collector voltage. On this basis the circuit may be treated as a bridge, as shown in Fig. 2.

Using the well-known condition for maximum bridge sensitivity, the value of the collector load should clearly be such as to equal the effective resistance of the transistor under non-illuminated conditions. The bridge is then initially balanced, with no current flowing through the relay coil, when the photo-transistor is dark. Referring to Fig. 2 the initial conditions are then $(i_1 - i_3) = 0$, and $R_1 = R_2 = R$.

Applying Kirchoff's Laws to the unbalanced bridge it can be shown that

$$\frac{i_1 - i_3}{i_3} = \frac{R_1 - R_2}{2R_2 + r + rR_2/R}$$

where i_3 is the current in the OC71 and $i_1 - i_3$ is the current through the relay coil.

In the actual circuit used, $r \simeq R/4$. When the photo-transistor is illuminated, assuming that its current increases

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by as much as the current, i_3, in the OC71 decreases, then one may graph the above expression as shown in Fig. 3. Here it is assumed that when R_1 increases to a value $2R$, R_2 decreases to $\frac{1}{2}R$, and so on. Using this graph, if it is assumed that the current in the OC71 is initially 5mA, and it decreases to 1mA on illumination of the photo-transistor, then the current in the relay will be $1 \times 6{\cdot}9 = 6{\cdot}9$mA. The current in the OCP71 will have increased from 5mA to 9mA, so that the sum of the changes in current in the two transistors is 8mA. Since the maximum current for the OCP71 and OC71 is 10mA in each case, this was considered to be the maximum sensitivity which could safely be used.

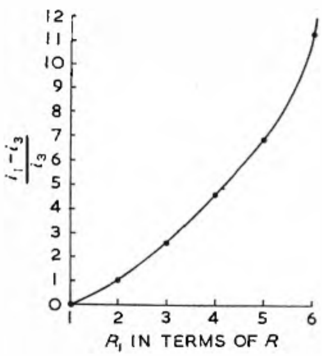
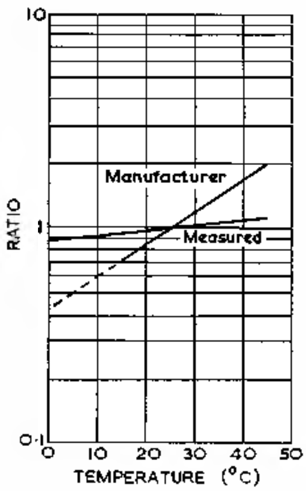


Fig. 3. Ratio of relay current to transistor current as a function of effective transistor resistance



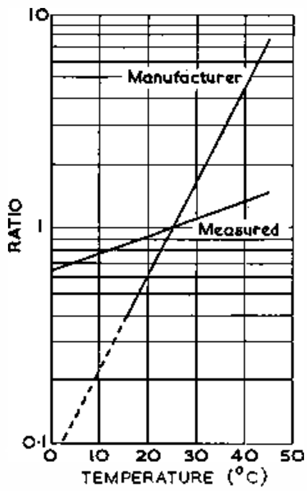


Fig. 4 (left). A graph of the ratio light current at t° C : light current at 25° C against temperature

Fig. 5 (right). A graph of the ratio dark current at t° C : dark current at 25° C against temperature

There is a danger of thermal run-away of the transistors implicit in the circuit. If the current in the OC71 falls by more than the current in the OCP71 initially rises, there is a net decrease in p.d. across the common emitter resistor which will bias the OCP71 such as to increase the current through it. This will, in turn, cause further diminution of the current through the OC71, and if the operating point on the transistor characteristics is suitable, this positive feedback mechanism, may drive the photo-transistor current well beyond its permissible maximum, leading to thermal breakdown. The empirical design procedure suggested will generally enable this to be avoided.

Circuit Performance

To test temperature stability, the whole circuit, as constructed, was placed in a temperature-controlled enclosure, and dark and light currents were measured as the temperature was raised. To check the low end of the scale, the equipment was placed in a refrigerator. The results are shown in Figs. 4 and 5 together with the manufacturer's published curves for the OCP71. It will be seen that considerable improvement in temperature stability is obtained.

The transient response of the circuit was tested by illuminating with a stroboflash. The pulse from the circuit, being the voltage across a resistor of value equal to that of the relay coil placed between the collectors, was displayed on an oscilloscope. A pulse derived directly from the stroboflash was also displayed, and the result is shown in Fig. 6. The rise-time of the pulse is of the order of 10μsec, showing that the circuit will respond satisfactorily to frequencies up to 100kc/s.

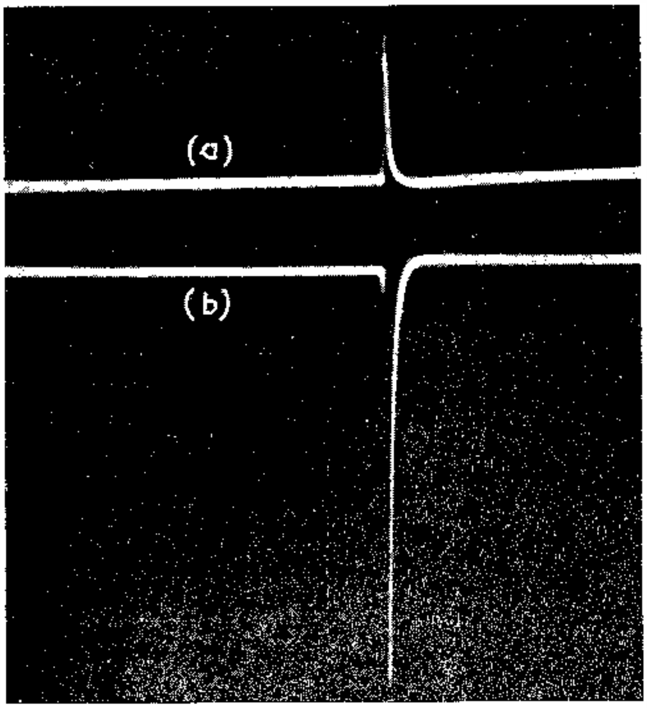


Fig. 6. (a) Transistor response pulse
(b) Stroboflash pulse

To make use of the maximum sensitivity of the OCP71, which occurs at a wavelength in the infra-red region, illumination from an infra-red source was used. This was constructed from an old car headlamp, in which the light from a portion of a 500W electric fire element mounted in the headlamp was concentrated on the transistor by means of a spherical mirror, silvered on its front surface. With the polarized relay used in the circuit, it was found that the minimum change in relay current for reliable operation was 1mA. Sensitivity was accordingly measured on this basis. With the source running at 30W, at which level it emitted no visible light at all, the change of 1mA through the relay was obtained at a distance of 2 metres, on interrupting the beam.

The circuit has been built up in portable form, using hearing-aid batteries, with the relay contacts brought out to terminals which may be connected to an alarm bell system. The whole chassis measured 4in by 3in by 3in and was used as a concealed alarm, in which role it proves most effective.

Acknowledgment

The authors wish to thank Messrs. S. A. Philips (Pty.) Ltd for supplying the transistors.

Incremental Frequency Control of RC Oscillators

By D. L. A. Smith*, B.Sc.(Eng.), A.M.Brit.I.R.E.

The provision of a continuously-variable, incremental frequency control which is accurately calibrated has not hitherto been available with an oscillator of the resistance-capacitance type. It is shown in this article that, by introducing a phase-shifting circuit auxiliary to the conventional phase-shift circuit, the generated frequency can be varied incrementally over a small range. It is further shown that, where a comparatively large increment is required, an error in the increment exists which is, to a first order, proportional to the time-constant of the main phase-shift circuit. A practical means of systematically correcting this error is also included.

(Voir page 58 pour le résumé en Français: Zusammenfassung in deutscher Sprache auf Seite 62)

A CONTINUOUSLY-VARIABLE oscillator for the generation of audio and low radio frequencies is almost always one of two kinds: it is either of a type using *RC* networks, or it is a heterodyne or beat-frequency oscillator. Of these, the resistance-capacitance oscillator has known advantages over the beat-frequency type and it will be sufficient to enumerate them briefly: an intrinsically higher stability of frequency, a more pure output waveform, especially at low frequencies, and a complete absence of spurious output frequencies.

However, the beat-frequency oscillator possesses one property which makes its application necessary where other considerations, such as those referred to above, would seem to dictate the use of an *RC* oscillator. By varying separately that heterodyne oscillator which is normally fixed, it is simple to provide an incremental control calibrated in frequency, so giving an effective expansion of the main frequency control at any point in the range. This is valuable for the investigation of any sort of high-Q resonant system, or indeed of any network or filter where response changes rapidly with frequency. It also leads to accurate and convenient assessment of small changes in any externally-generated frequency. It will be realized that in order fully to exploit the advantages of such an incremental control, it is essential that the frequency stability of the instrument as a whole be of a high order. Reference has already been made to the superior stability of the *RC* oscillator; the following account is a description of a way in which incremental control of the frequency of an *RC* oscillator can be achieved.

The normal *RC* oscillator is of the general configuration shown in Fig. 1, where the output of a maintaining amplifier is connected to its input through a phase-shifting network. At a frequency where the loop phase shift is 0° or $n.360°$, oscillation will occur, given that the corresponding loop gain is not less than unity. In order that the oscillation frequency shall be controlled solely by the constants of the network, every effort is usually made to reduce the phase-shift in the maintaining amplifier to proportions which are negligibly small over the whole of the working frequency range. However, it is profitable to consider the effect of phase-shift which is not negligible.

The essentials of a Wien-bridge type of circuit are shown in Fig. 2 (the resistive arms, which are used for amplitude control, are omitted). The series and parallel arms are shown as having equal components, C_1 and R_1. Although this is not essential for the analysis that follows, such equality has been assumed as the expressions thus obtained are somewhat less cumbersome and, in any case, continuously-variable-frequency oscillators always incorporate equal C and R arms for practical reasons. Assuming that the amplifier stages have high input and low output resistances compared with R_1 and R_2, the loop phase-shift is

$$\tan^{-1}\frac{1}{\omega C_1R_1} - \tan^{-1}\omega C_1R_1 + \tan^{-1}\frac{1}{\omega C_2R_2}$$

Equating this to zero and solving for ω gives

$$\omega = \frac{1}{C_1R_1}\left(1 + \frac{2C_1R_1}{C_2R_2}\right)^{\frac{1}{2}} \qquad \ldots\ldots\ldots\ldots (1)$$

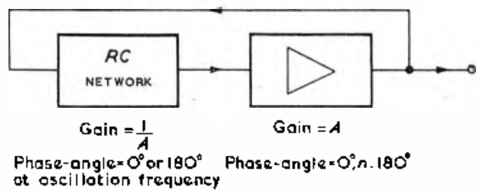


Fig. 1. General configuration of *RC* oscillator

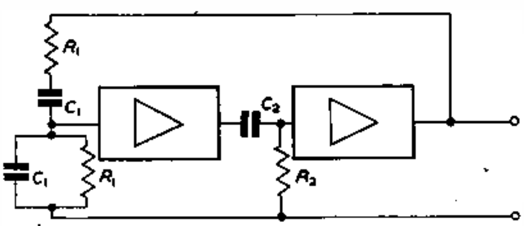


Fig. 2. Oscillator in which amplifier phase-shift is not negligible

If $C_2R_2 > 2C_1R_1$, this may be expanded to

$$\omega = \frac{1}{C_1R_1}\left\{1 + \frac{C_1R_1}{C_2R_2} - \frac{1}{2}\left(\frac{C_1R_1}{C_2R_2}\right)^2 + \frac{1}{2}\left(\frac{C_1R_1}{C_2R_2}\right)^3 - \frac{5}{8}\left(\frac{C_1R_1}{C_2R_2}\right)^4 + \ldots\ldots\right\} \ldots (2)$$

Thus to a first approximation

$$\omega \simeq \frac{1}{C_1R_1} + \frac{1}{C_2R_2} \qquad \ldots\ldots\ldots\ldots (3)$$

which would be the desired result, viz. that the frequency can be varied partly by changing C_1R_1 and partly by changing C_2R_2, each acting independently. The available error-free increment is quite small, however. For example, it cannot exceed about 2 per cent of the main frequency if the error in incremental calibration is limited to, say, 1 per cent. For increments which correspond to larger proportional changes the series should be examined further.

Suppose C_1R_1 to be controlled by a variable capacitor to form the main frequency control and C_2R_2 to be controlled by a second variable capacitor, the maximum capacitance of which is n times its minimum capacitance,

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C_{2min}. Then from equation (2) it follows that the increment, $\Delta\omega$, is given by

$$\Delta\omega = \omega_2 - \omega_1 = \frac{1}{C_{2min}R_2}\left(\frac{n-1}{n}\right) - \frac{C_1R_1}{2(C_{2min}R_2)^2}\left(\frac{n^2-1}{n^2}\right) + \frac{(C_1R_1)^2}{2(C_{2min}R_2)^3}\left(\frac{n^3-1}{n^3}\right) - \dots\dots$$

$$= A\left\{1 - \tfrac{1}{2}\frac{C_1R_1}{C_{2min}R_2}\left(\frac{n+1}{n}\right) + \tfrac{1}{2}\left(\frac{C_1R_1}{C_{2min}R_2}\right)^2\left(\frac{n^2+1}{n^2}\right) - \dots\right\} \quad \dots\dots\dots\dots\dots\dots (4)$$

where $A = \dfrac{1}{C_{2min}R_2}\left(\dfrac{n-1}{n}\right)$

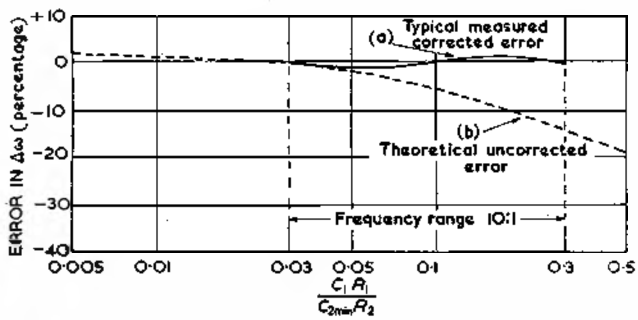


Fig. 3. Variation of $\Delta\omega$ with $C_1R_1/C_{2(min)}R_2$ showing effect of correction capacitance ($n = 2{\cdot}5$)

This is a general expression for the increment which may be applied to any part of the tuning range of the oscillator. It will be seen, however, that due to the presence of terms in powers of $(C_1R_1/C_{2min}R_2)$, the increment will become smaller as C_1R_1 is increased (i.e. in changing the main frequency from the high to the low end of the range). This is illustrated in the curve (a) of Fig. 3 which shows, for a typical value of n, the percentage reduction in $\Delta\omega$ as a function of $C_1R_1/C_{2min}R_2$.

Consider now the low-frequency end of the range: the series (4) is an alternating one and if terms in $(C_1R_1/C_{2min}R_2)^2$ and higher be neglected the error in doing so does not exceed $50(C_1R_1/C_{2min}R_2)^2((n^2+1)/n^2)$ per cent. In the practical case it is possible to make $(C_1R_1/C_{2min}R_2)^2$ sufficiently small to justify the approximation. At the high-frequency end of the range, the term in $(C_1R_1/C_{2min}R_2)$ is also sufficiently small to be neglected (assuming that the capacitance of C_1 will swing to about 1/10 of the low-frequency value, which is usual). Then the error in the increment at the lowest frequency, expressed as a fraction of the increment at the highest frequency, is

$$\frac{\Delta\omega_H - \Delta\omega_L}{\Delta\omega_H} = \tfrac{1}{2}\,\frac{mC_1R_1}{C_{2min}R_2}\left(\frac{n+1}{n}\right) \dots\dots\dots (5)$$

where m is ratio of maximum to minimum capacitance of C_1.

Clearly the error is proportional to C_1R_1, so that if $\Delta\omega$ can be increased as C_1R_1 increases, to the degree indicated by equation (5), the error can be eliminated. Reference to the expression for $\Delta\omega$ in equation (4) shows that $\Delta\omega$ is to a first order inversely proportional to C_{2min}.

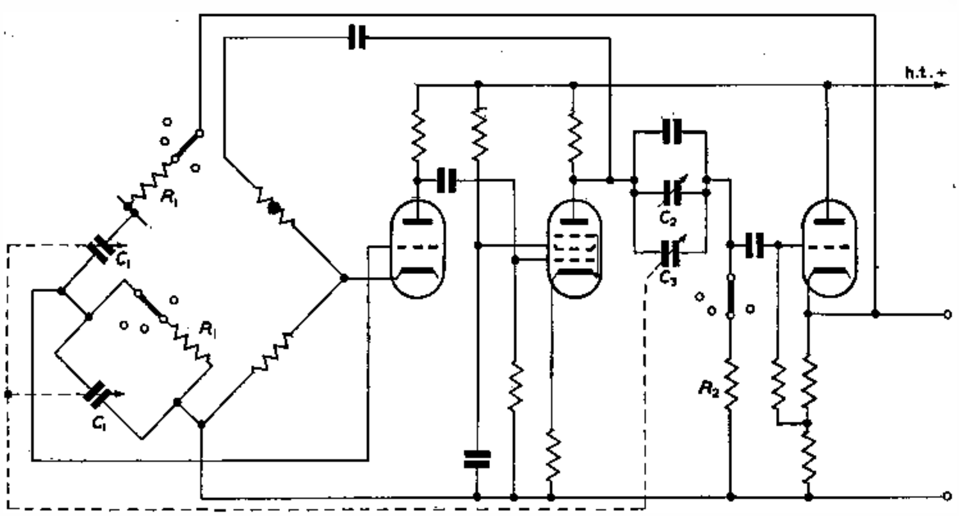
This suggests a convenient method of correcting the error: a small variable capacitor, C_3, is connected in parallel with C_2 and coupled mechanically to C_1 (its capacitance/rotation law should be similar to that of C_1). The coupling is arranged so that C_3 increases as C_1 decreases and vice-versa. The maximum value of C_3 is

$$C_3 = C_{2min}\left\{\frac{1 + \frac{1}{2}\,.\,C_1R_1/C_{2min}R_2\,.\,(n+1)/n}{1 - \frac{1}{2}\,.\,C_1R_1/C_{2min}R_2\,.\,(n+1)/n}\right\} \dots\dots (6)$$

The basic circuit of an oscillator incorporating this correction is given in Fig. 4. It will be seen that loading on the auxiliary phase-shift circuit C_2R_2 is minimized by the use of a cathode-follower and that R_2 is switched with R_1 to preserve the ratio of increment to main frequency throughout the ranges of the instrument. A typical curve of the measured error in $\Delta\omega$ between the extremes of a (10 : 1) frequency change is also shown in Fig. 3, for comparison with that for the intrinsic error previously estimated. The value of n is similar in the two cases and the nominal $\Delta\omega$ is 2π.50, the main frequency range being 250 to 2 500c/s. It will be seen that the correction is sufficiently close to maintain an accuracy of incremental calibration within $\pm 1{\cdot}5$ per cent limits, even though the increment amounts to 20 per cent of the main frequency at the low-frequency end.

An instrument embodying this circuit has been built. It covers the frequency band 25c/s to 250kc/s in four decade ranges, with an increment of 5c/s to 5kc/s, also in four decade ranges. The calibration accuracy of the main dial is ± 1 per cent and that of the incremental dial is ± 2 per cent. The frequency stability against changes of supply voltage and of temperature is good; sufficiently so that a tuning-fork-controlled calibrator forms part of the instrument. Using this, the main, or reference frequency can be set to any of a large number of frequencies defined to an accuracy better than 0·01 per cent, leaving the incremental dial as a directly calibrated fine frequency control.

Fig. 4. *RC* oscillator with incremental frequency control



Acknowledgments

The author wishes to express his thanks to Furzehill Laboratories Ltd. for permission to publish this article. Thanks are also due to Mr. E. S. Freeman, who conceived the need for this development, for his helpful criticism.

REFERENCE

Developments in and Relating to Variable-Frequency Oscillators. Brit.Prov. Patent Spec. 14,264/57.

Surface Structure of Saturated-Diode Filaments

By F. A. Benson*, D.Eng., Ph.D., A.M.I.E.E., M.I.R.E., and M. S. Seaman†, B.Eng.

The life of a tungsten-filament diode when operated from a d.c. filament supply is not necessarily the same as when supplied with a.c. Filaments which have been run on d.c. develop a step-like surface structure whereas a.c. heated filaments become smooth except near the supports. A number of diode filaments have been examined and the results are briefly discussed in this article.

(Voir page 58 pour le résumé en Français: Zusammenfassung in deutscher Sprache auf Seite 62)

SATURATED diodes are commonly used as controlling elements in certain voltage-stabilizer circuits[1,2,3] and for measurement purposes[4]. The relative merits of four different types of valve for this purpose have been determined by the authors[5,6] in recent years, and some characteristics of diodes with a.c. heating have been presented and discussed[7].

Attree[8] has commented on some of the work by the authors[5] and has drawn attention to the fact that in most practical applications of the saturated diode the filament is heated from an a.c. supply. The tests he referred to were made with the filaments run from a d.c. supply and the life on d.c. is not necessarily the same as on a.c., since tungsten filaments which have been run on d.c. exhibit a step-like surface structure, whereas a.c. heated filaments are uniformly smooth except for a restricted region near the supports where, in any case, the filament is too cool to give significant emission. The effect has been fully discussed by Johnson[9] in a paper which reports results of a microscopic study of filaments from several hundred burnt-out incandescent lamps of various sizes and from various manufacturers. He shows that the surface structure of tungsten-lamp filaments varies with age and that the variations depend on whether the filament was heated by d.c. or a.c. and also whether the lamp was a gas-filled or vacuum type.

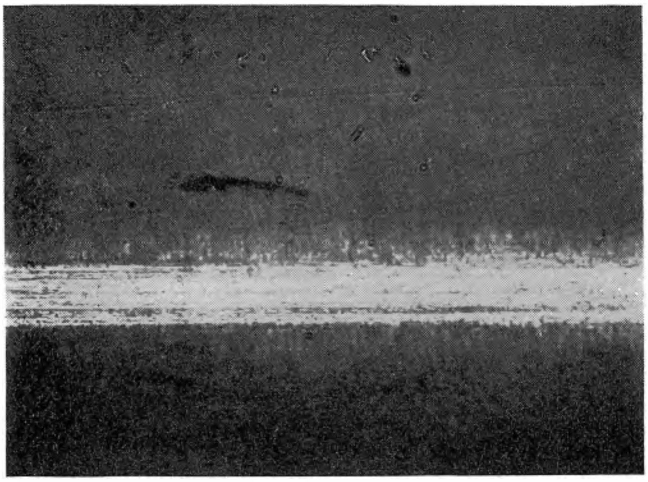


Fig. 1. Filament of a 29C1 diode after operating for only 25 hours
Magnification 450

The filaments of a few diodes which had been life tested, one which was used in development work on an a.c. stabilizer and another which was accidentally burnt out after a very short life, have been examined under a microscope. The results of these investigations, which are given here, confirm many of Johnson's findings for lamps.

Results and Discussion

With a new valve the die marks are easily seen running along the wire as shown in Fig. 1 which is a photograph of the filament of a 29C1 valve. This particular valve had run for about 25 hours only before it was burnt out. As a valve ages these die lines gradually disappear and the filament develops a surface determined by the nature of the heating current. It is somewhat surprising that variations in the surface structure of these valves, similar to those reported by Johnson, have been observed, when the accelerating field at the cathode of a saturated diode is considered. Johnson examined a few tungsten-filament rectifier valves, but these had been operated either under normal conditions or with the anode negative with respect to the filament, the results being similar to lamp filaments.

Fig. 2. Filament of a 29C1 diode which has developed a crystalline structure after operating on d.c.
Magnification 450. Total operating time 305 days

Fig. 3. Filament of an AV33 diode which has developed a crystalline structure after operating on d.c.
Magnification 450. Total operating time 95 days

When operated on d.c. the filaments develop a crystalline structure as is shown by the photographs of Figs. 2 and 3. There is a herring-bone appearance about this structure which is directed towards the negative end of the filament, except near the supports, similar to that observed by Johnson with lamps, who attributes the effect to the migration of tungsten ions along the filament under the action of the longitudinal electric field.

When run on a.c. the filaments gradually develop fairly smooth surfaces. In the case of one valve which had an appreciable running time (of the order of 100 days) on a.c. at a filament voltage of about 3·7V which is less than the makers recommended rating, the original die-marks are still visible as can be seen from Fig. 4. This result is similar to that obtained by Johnson for a lamp which had been operated on a.c. for a fairly short time. The effect is not visible near the supports. In the vicinity of

* *University of Sheffield.* † *Pye Ltd., formerly University of Sheffield.*

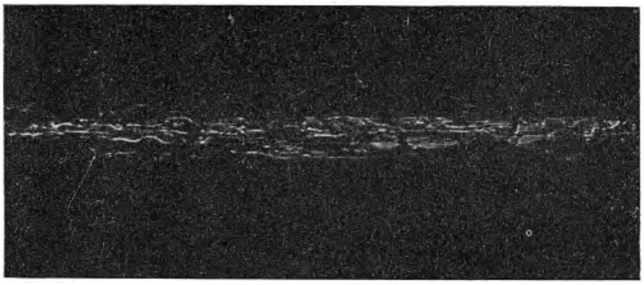


Fig. 4. Filament of a 29C1 valve after running for an appreciable time on a.c.
Magnification 450

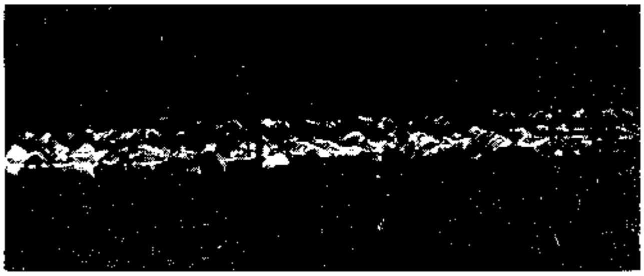


Fig. 5. Filament of a 29C1 diode near a support after running on a.c.
Magnification 450

the supports, where there is a temperature gradient on the wire, a crystalline structure has been observed, again with a herring-bone appearance. In this case, the herring-bone structure was directed towards the supports and Johnson has concluded that this is due to the migration of tungsten ions along the filament down the temperature gradient. The structure is the same as that observed on filaments heated by d.c. and occurred on all the filaments examined whether operated on a.c. or d.c. The structure shown in Fig. 5 was photographed from a 29C1 filament which had been heated by a.c. The herring-bone effect is not clear on this photograph, but it is evident that the structure shown is quite different from that illustrated in Fig. 4.

It is interesting to note that some preliminary studies by Johnson of tantalum, molybdenum, platinum, iron and nickel filaments in vacuum, indicated that these metals also develop a roughened surface on d.c. heating but remain smooth on a.c.

Acknowledgments

The authors wish to thank The University of Sheffield for facilties afforded in the Department of Electrical Engineering, where the work described was carried out. They also acknowledge the kindness of The Amalgamated Wireless Valve Co. Ltd (Pty), Australia, in supplying a number of AV33 diodes for examination.

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A CORDLESS TELEPHONE SWITCHBOARD

The country's first cordless switchboard to receive Post Office sanction recently came into operation.

The switchboard controls a private automatic branch exchange, manufactured by Automatic Telephone and Electric Ltd and installed at the European headquarters of the Canadian Pacific Railway Company in Trafalgar Square, London.

In place of the normal jacks, plugs and cords are neat rows of push-buttons and keys, mounted on a small desk, smartly styled in light oak with grey panelling. Even the familiar dial is only there for stand-by use, since 'dialling' has been replaced by the much faster system known as 'key-sending', which the operator (not the extensions) uses for calling both internal and external numbers.

The exchange has 110 extensions, 29 lines to the public exchange and four private wires to other Canadian Pacific offices. Any extension can dial any other extension and can also dial outside numbers direct, from the same telephone. Direct access to an outside line can be barred to certain extensions if desired.

The two operators—four were needed to operate the original equipment—are almost entirely occupied with incoming calls. A system of lamps tells the operator how many calls are waiting. To accept a call, she simply moves a key—the waiting calls are automatically brought up in rotation—and sets up the number of the required extension by pressing the appropriate push-buttons. She then presses the 'send' key, the extension is rung automatically and a lamp indicates 'free' or 'busy'. If the line is free, the outside call is put through automatically.

If busy, a 'ring when free' facility can be used. This device rings the extension when it becomes free and automatically connects the outside call. Sometimes the operator will wish to advise a 'busy' extension that an important outside call is waiting; she can do this by transmitting a warning tone. When the required extension hangs up he is automatically rung and, on answering, is connected to the incoming calls.

An extension can 'hold' an outside call while he makes an internal call; he can automatically transfer an outside call to another extension, without help from the operator; and if perhaps he is not sure to which extension the call should be transferred, he can return it to the operator for action.

A large number of telephone calls coming in to Canadian Pacific naturally consists of enquiries about the company's ship, 'plane and train services. To cut delay to a minimum, a special system has been devised which diverts these calls to the appropriate department, where they 'queue up' for attention. Nine enquiry lines go to each department, any one of which can be answered by any of five clerks, so no one is kept waiting for long. Each clerk has a control panel fitted with keys and indicator lamps to show which lines are 'busy' or 'waiting'. A clerk can transfer a call from the queue to an ordinary extension, or vice versa. He can also transfer a call from the queue to the queue in another department. All these transfer operations are carried out automatically, without assistance from the switchboard operator.

To the operators, with all controls at their fingertips and many operations that formerly had to be done manually now taken over by automation, the new cordless switchboard means smoother and faster operation and considerably less fatigue. To Canadian Pacific, it means much improved service for their customers and an all-round increase in efficiency.

A Reversible Dekatron Counter

By D. L. A. Barber*, B.Sc.(Eng.)

A decade counter capable of both addition and subtraction is described. The counter comprises a transistor circuit combined with valves and a Dekatron; any required number of decade circuits may be connected in cascade

(Voir page 58 pour le résumé en Français: Zusammenfassung in deutscher Sprache auf Seite 62)

THE circuit to be described provides a means of coupling successive decades of a Dekatron counter so that the counter will both add and subtract. A transistor trigger is used as a memory and valves are employed for pulse shaping and gating purposes.

In a multi-stage decimal counter any stage changing from 0 to 9 must add a 'carry' to the next higher stage, while a 'borrow' must be subtracted for changes from 9 to 0. A circuit capable of satisfying these requirements is shown schematically in Fig. 1.

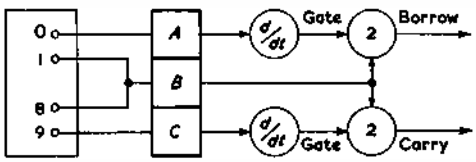


Fig. 1. Basic arrangement of reversible counter

The trigger *ABC* has three stable states *A* off, *B* off and *C* off. If any one input is present the trigger is set with the corresponding output absent, and with no inputs present the trigger remains in the condition set up by the last input. The three inputs are connected to Dekatron cathodes 0, 1 and 8 in parallel, and 9 respectively so that if, for example, the Dekatron glow rests on cathode 9 output signals will be present only at *A* and *B*. A direct change from 9 to 0 will transfer an output of the trigger from *A* to *C*, while *B* remains unchanged. The signal appearing at *C* is 'differentiated' and, since with *B* present the gates are open, a carry is generated. A signal will also appear at the output of *C* for a change from 9 to 8, but the simultaneous removal of the signal from *B* closes the gates to this pulse. A 'borrow' is produced, in a similar manner, by a change from 0 to 9. A trigger circuit with the properties required, shown in Fig. 2, consists of three transistors of which any one transistor is non-conducting. A conducting transistor is cut-off by applying a positive pulse to its base.

In order to transfer the glow in a Dekatron from one cathode to an adjacent cathode, two guide electrodes *a* and *b* are provided. A pulse applied to *a*, followed immediately by a pulse to *b*, causes the glow to step forward, while successive pulses to *b* and then *a* cause the

* *National Physical Laboratory.*

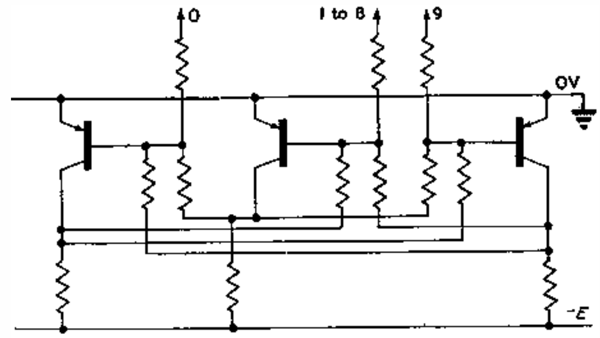


Fig. 2. Transistor trigger circuit

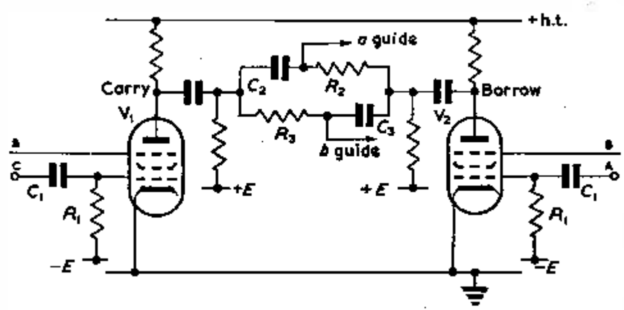
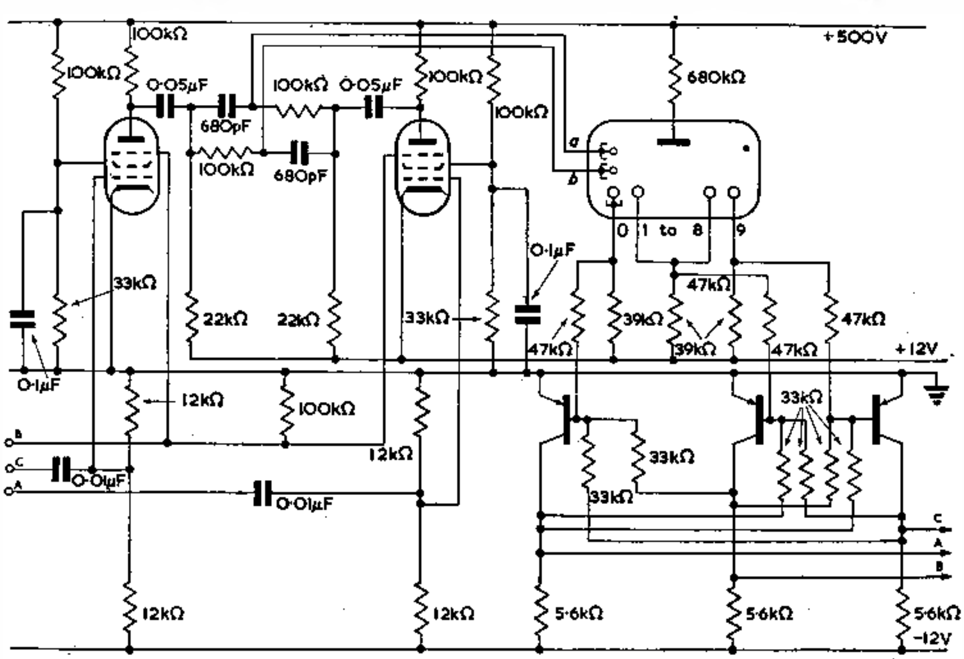


Fig. 3. Pulse circuit

Fig. 4. Complete circuit for one decade



glow to step backward. These two pulses are obtained from a carry or borrow pulse by means of the circuit shown in Fig. 3.

Valves V_1 and V_2 act as gates and amplifiers. A signal from C is differentiated by C_1R_1 and, if signal B is present, the resulting pulse is amplified by V_1 to form a carry pulse. Each carry pulse appears differentiated by C_2R_2 at the a guide and integrated by R_3C_3 at the b guide. Guide signals produced in this way cause the Dekatron to step forward. Borrow pulses, which appear at the anode of V_2, are integrated by R_2C_2 and applied to a and differentiated by C_3R_3 and applied to b. These guide signals cause the Dekatron to step backward.

The complete circuit for one decade, of which any number may be connected in cascade, is shown in Fig. 4.

Acknowledgments

The work described was carried out as part of the research programme of the National Physical Laboratory and is published by permission of the Director.

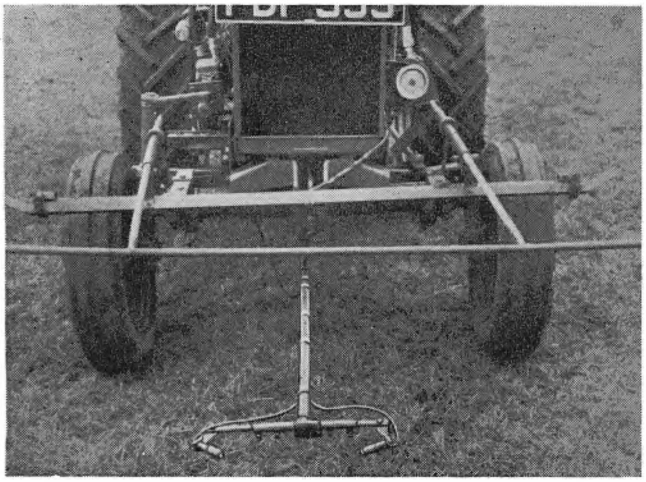
The author wishes to acknowledge the assistance of Mr. J. D. Gilbert in constructing and testing the circuit.

An Automatically-Controlled Farm Tractor

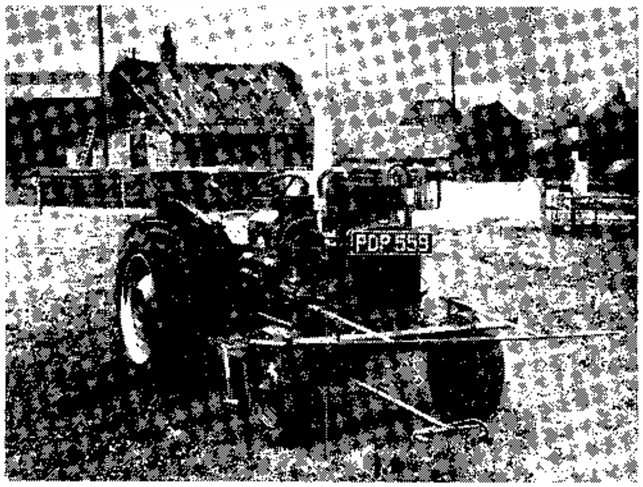
A driverless automatically-controlled tractor has recently been demonstrated jointly by the Farm Mechanization Department of Reading University and the Rural Electrification Section of the Electrical Research Association.

During the demonstration the tractor was made to leave a farmyard, pass through two gateways into a field and then perform several different functions and, finally, to return to the farmyard where it automatically came to a halt. The tractor was also made to obey a set of traffic lights on entering and leaving the field and sound its horn at points along the route.

For a tractor to operate without human supervision, the main functions to be controlled are its steering, clutch, power take off and hydraulic lift mechanisms. If these can be independently and remotely controlled, a tractor can be made automatically to carry out a number of tasks in any desired sequence.



The sensing coils at the front of the tractor



The complete tractor showing sensing coils at the front and side

The control system in use is a relatively simple one and, although not entirely novel, it is believed that this is the first time that a control system of this type has been fitted to an agricultural tractor.

It is designed primarily to replace a tractor driver on operations where a tractor can be given a task to perform in an allotted sequence, the details of which have been prescribed. Examples of such tasks are the routine carting over regular routes to and from the farmyard, and loading and unloading in the field.

In essence, the control system consists of a cable fed with an 800c/s supply and buried to a depth of about 9in along the regular routes the tractor is desired to work.

The steering mechanism, which consists of a double acting ram and conventional hydraulic steering booster, is controlled by two pick-up coils mounted in front of the tractor and a short distance above the ground.

Any unbalance in the coils caused by a deviation of the tractor from its route is detected by a transistor amplifier carried on the tractor, and the unbalance is made to operate the hydraulic steering mechanism and restore the tractor to its original path.

At each side of the tractor is mounted a further pick-up coil with its associated transistor amplifier, the output of which is connected to a hydraulic servo system. The two pick-up coils, either separately or in combination, can then be made to operate the hydraulic systems for controlling the clutch, power take off and hydraulic lift mechanisms as desired. These pick-up coils receive their operating signals from loops laid out from the main cable at predetermined points.

The present control system is cheap and robust, and it has negligible running costs but, before a more flexible and comprehensive system of automatic control for field work is evolved, considerable development will be required. But it is stated that studies of the existing system are sufficiently encouraging to warrant development of a complementary control system which would include such tasks as ploughing, cultivating, seeding, hoeing, spraying and harvesting to be done with great precision and without human intervention.

A Cardiac Ratemeter Using Transistors

By L. D. Kitchen*

A simple transistorized instrument for the continuous direct reading of the heart rate during surgery is described.

Of small physical dimensions, it has an input sensitivity of 250μV and is relatively insensitive to 50c/s interference. A calibrating pulse generator is incorporated.

(Voir page 58 pour le résumé en Français: Zusammenfassung in deutscher Sprache auf Seite 62)

IT is necessary to monitor a patient's pulse rate during surgical operations, but often it is neither easy nor convenient for the anæsthetist to do this by the usual means of placing a finger over an artery. Simple pulse indicators for use during prolonged operations and as an aid in hypotensive anæsthesia have been described[1,2] but these do not enable a quantitative indication of pulse rate to be shown. A valve operated instrument for the measurement of pulse rates in beats per minute has been described[3] and has proved a useful aid to the anæsthetist. Franklin and James[4] have discussed the application of transistors to ratemeter circuits for radiation monitors and it was thought that it might be possible to use a similar circuit for the measurement of pulse rates. Difficulties were experienced however because of the low frequencies involved and the following instrument was finally developed. It is simple in design and gives a direct reading of pulse rate in beats per minute on a 3½in moving-coil meter. It has the advantage of being small, self-contained and easily operated. In use, a clear indication of pulse rate is obtained both in anæsthetized and in normally relaxed patients.

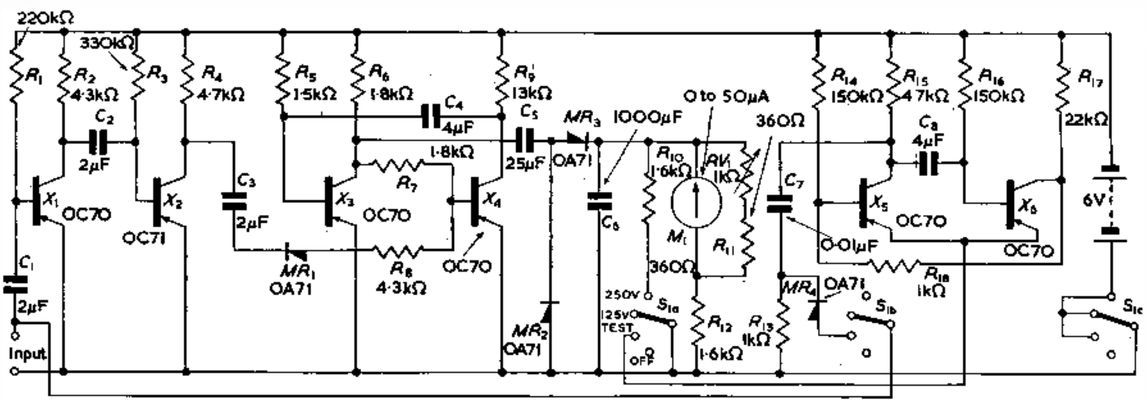


Fig. 1. Circuit diagram of cardiac ratemeter

Basic Considerations

The main factor influencing the design of any instrument for displaying electrophysiological phenomena is the means by which the initial signals are picked up. In the case of pulse rates there are three possible methods, namely an electro-mechanical transducer on the surface of the skin and sited directly over an artery, a pressure transducer inserted in a blood vessel, or the use of the electrical potentials generated by the heart muscle itself. Simplicity of design and operation suggested the latter method as being the most suitable, and it was decided to operate the counting rate circuit by means of the Q.R.S. complex of the electrocardiograph waveform after suitable amplification. This complex consists of a voltage pulse of amplitude approximately 1mV and duration between 50 and 100msec. Its sense can be made either positive or negative, depending upon the placing of the electrodes on the patient. If the wrists are used as electrode sites the complex is normally positive going over the right wrist, although this can vary in certain cardiac conditions.

The second major consideration is the reduction of artefacts which would be liable to give a false reading of pulse rates. In an instrument of this type artefacts would consist largely of (1) 50c/s pick-up induced in the patient by the proximity of unscreened electrical apparatus, and (2) muscle potentials of sufficient amplitude to trigger the ratemeter. Since the ratemeter is designed primarily for use on anæsthetized patients the latter consideration is not so important. The fact that occasional heavy diathermy fields would be present was also considered, but it was decided that to attempt to provide for their rejection would unnecessarily complicate a design which was inherently simple. The artefacts mentioned above would, in any case, send the meter completely off scale and would obviously not be confused with any pulse rate reading.

Circuit Description

The input pulse from the right and left wrists (Cardiac Standard Lead I) is fed in the negative sense to the two stage RC coupled amplifier X_1 and X_2. This amplifier has an overall voltage gain of 4 000 which is reduced to 3 000 by the coupling to the pulse forming circuit which follows. The negative going portion of the output signal is fed through MR_1 and R_8 to initiate the monostable trigger pair X_3 and X_4. C_5 charges through MR_2 between pulses and discharges into the meter circuit through MR_3 and the

* *Physics Dept., King's College Hospital, London.*

emitter-collector path of X_3 during each pulse. For each heart beat, therefore, a current due to the charge Q on C_5 is 'pumped' through M_1 and if the average recurrence rate is N pulses per minute, the average output current is given by $I_{av} \propto NQ$. I_{av} is thus proportional to N and is indicated on the meter. The scale is graduated in two ranges, 0 to 125 and 0 to 250 beat/min. C_6 controls the time-constant of the meter circuit and this has been fixed at 5sec in order to allow the meter to follow changes in pulse rate within 3 or 4 beats, while keeping meter fluctuations due to individual beats to a minimum. In practice during each beat the pointer moves plus and minus one beat, thus giving excellent indication that the instrument is functioning correctly but not interfering with ease of reading. The movement also allows a visual check on calibration if required.

For calibration purposes a modified form of transistor multivibrator is used to produce a pulse repetition frequency of 100 pulse/min. The output of this can be switched, by means of S_{1b}, into the amplifier input. The meter can then be matched to read appropriately by the preset control RV_1.

With the component values shown the instrument will trigger on an input pulse of 250μV and preselected control of trigger sensitivity of the pulse forming circuit can be achieved by varying the value of R_8. This therefore provides a method of reducing interference caused by low amplitude muscle activity and minor 50c/s pick-up by setting the trigger threshold above their level. In practice this has produced satisfactory results.

The ratemeter is powered by a 6V Leclanché type dry battery, mercury batteries being found unsuitable for this particular application because their relatively high internal resistance caused interaction between amplifier and pulse forming circuits. The complete instrument is housed in a diecast aluminium box of dimensions 7in by 4½in by 3in.

Acknowledgements

The author wishes to acknowledge the helpful interest of Dr. J. F. Fowler and the co-operation of Dr. A. H. Galley who sponsored the project. Acknowledgment must also be made of the support of the Research Fund of King's College Hospital and Medical School.

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'FRED'—A Figure Reading Electronic Device

A private exhibition of industrial electronic equipment and systems was recently staged by E.M.I. Electronics Ltd at their Hayes factory.

Among the exhibits was a new Figure Reading Electronic Device (FRED), the outstanding feature of which is its simplicity.

FRED can read either optically, by means of a photocell, or magnetically with the aid of specially prepared inks at speeds of many thousand characters a second.

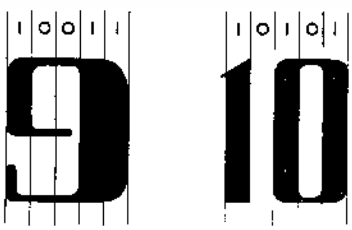

Fig. 1. Illustrating the method of reading

Both methods have widespread uses, but where documents to be read have to be passed through several hands and are likely to accumulate overprints or defacements, the magnetic system has definite advantages.

A special typeface that varies little in appearance from standard numerals, and is therefore easy to read, has been designed to allow maximum tolerances. The accuracy of size and shape is well within the capabilities of all the normal printing processes.

The possible applications of FRED are only just beginning to become apparent. In its magnetic form it is ideally suited for reading cheques and invoices for the purpose of sorting or feeding information into accounting machines.

Cash register entries can also be scanned and the information fed into a central accounting system, thus keeping an up-to-the-minute check on cash takings, throughout a large store or chain of businesses.

As a computer input system the equipment offers great advantages. Much of the information which is now fed into computers by means of punched cards or perforated tape can be prepared 'in clear' and read automatically into the computer.

FRED can also be used for reading route information or serial numbers of objects on conveyors or railway trucks. The logic of the circuits in FRED enable fifteen characters of which the Arabic numerals 0 to 9 will normally be included with alternative options for using the other five. The figures 10 and 11 in the pence column of an accounts entry can be read as single digits, and a wide range of conventional symbols can also be provided.

The characters are read by moving them smoothly past the reading head. There is no need to bring them to rest or centralize them. This relative motion can be achieved either by moving the head, or the characters themselves.

The maximum speed of operation of the equipment is determined in general by mechanical limitations rather than those concerned with electronics.

The extreme simplicity and reliability of the E.M.I. system have been achieved by studying the philosophy and circumstances of the applications to which it can be applied.

The basic principle of operation can be seen by reference to Fig. 1. Each character is looked at, from right to left, in five vertical strips and the circuits will first decide whether the strip is predominantly black or white. The first black strip will act as a trigger and, reading black as 1 and white as 0, the figure 9 will be coded as 'trigger' 1001 and the figure 10 as 'trigger' 0101 and so on. The typeface itself is so designed that the transition from black to white will always be sharp so that no ambiguity can arise.

LETTERS TO THE EDITOR

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

A Simple Improvement to a Three-Electrode Gap

DEAR SIR,—Three electrode gaps are commonly used for triggering time-base circuits for transient oscillographs. While high voltage valves have largely replaced these, many gaps are still in use and provide consistent operation when a sufficiently large trigger voltage is available.

Large trigger voltages may be obtained electrostatically from an aerial when the transient oscillograph is used with an impulse generator, but for recording the required calibrating traces, only conventional power supply voltages are available. Because this trigger voltage is generally less than 750V, the operation of the gap is critically dependent on the setting of the gap and the state of the discharge surfaces. This results in varying time lags necessitating constant cleaning and adjustment of the gap.

A sufficiently large trigger voltage may be obtained from a normal power supply with a pulse transformer. Following a design procedure described by Smith[1], a pulse transformer with a 10:1 turns ratio was wound on a Mullard Ferroxcube core. Optimum rise-time was achieved by varying the thickness of insulation between the primary windings and the core.

Thus with a supply voltage of 500V, a nominal 5kV tripping pulse is available. This gave consistent operation of the three electrode gap and was sensibly independent of the gap setting and the state of the discharge surface. The average time delay for 50 operations of the gap after the arrival of the trigger pulse was 0·2μsecs with a maximum jitter of 0·05μsecs.

A pulse transformer wound over a Ferroxcube core may be used to provide a large trigger pulse from conventional power supplies, thus ensuring consistent operation of a three electrode gap for transient oscillograph time-bases with a consistent delay of less than ¼μsec.

Yours faithfully,

M. DARVENIZA

Department of Electrical Engineering,
Queen Mary College,
University of London.

REFERENCE

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Response of a Capacitance-resistance Divider to a Step, Exponential, and a Ramp Function

DEAR SIR,—Mr. S. Turk's article in the October issue is of interest because it is an attempt to generalize results. When faced with this problem the writer has to decide on the best way of presenting his results, and Mr. Turk has selected

$$t/v\left(\frac{\text{variable time}}{\text{overall time-constant of network}}\right)$$

as his variable. This has the advantage of providing an open scale, but it does tend to obscure important properties, thus if he had used

$$t/\tau\left(\frac{\text{variable time}}{\text{time-constant of the input wave where appropriate}}\right)$$

he would have produced voltage time curves which can be quickly compared with the input wave shape.

An example of this principle in action is seen in particular in Figs. 6(d) (exponential function) and 7(d) (ramp function), which become one single curve when plotted against t/τ, thus showing what is well known, that when $R_1/R_2=C_2/C_1$ the wave shape of the output is an attenuated (but otherwise exact) copy of the original wave shape. Similarly the peaks of the ramp function for $a < 1$, all occur at $t/\tau = 1$, and the modification in the rate of rise brought about by the circuit is seen immediately.

Incidentally, there appears to be an error in the numbering of the curves in Fig. 7(d). The product of τ and t/v at an ordinate value of unity should be unity.

Yours truly,

K. R. STURLEY

The British Broadcasting Corporation,
Wood Norton,
Evesham.

The Author replies:

DEAR SIR,—I thank Dr. K. R. Sturley for his observations and suggestions and I apologize for the faultily drawn Fig. 7(d). The correct version is shown in Fig. A.

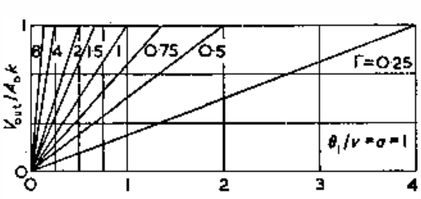

Fig. A. Corrected version of Fig. 7(d)

Dr. Sturley has well appreciated the problem of choosing the proper time variable. Using one or the other gives the same results but they point out different qualities and the choice depends on the individual problem. With respect to the input waveform a quicker understanding of the general properties of the divider is obtained using t/τ while t/v gives a better insight to some properties of the divider with respect to its constants. Since the chosen variable does not depend on input waveform properties, it allows the representation of the graphs for all three input waveforms on the same scale. These graphs are, in fact, divider response, as seen on an oscilloscope, using a constant sweep speed.

Yours faithfully,

S. TURK

Institute Ruder Boskovic,
Zagreb,
Jugoslavia.

Noise Limitation of the Maximum Gain-Bandwidth Product of Travelling Wave Amplifiers

DEAR SIR,—In designing a helix type travelling wave tube, one is usually aiming at a required output power and a given gain, which has not usually exceeded 40dB. If greater gains than this have been required, these have been obtained by the use of a second appropriately designed valve used as a driver for the output valve. If the output valve is operating at a high power level, it necessarily has a high-power beam and probably requires a large focusing field provided by a solenoid. The driver valve, on the other hand, can be designed so that its magnetic field requirement is considerably lower and being a low power valve it will often be possible to use periodic magnetic focusing. The advantage, therefore, of using two valves instead of one is that the high power solenoid is kept to a minimum length, which will be fixed by the fact that the minimum gain for an efficient tube is about 25dB. The price paid for this advantage is that the overall amplifier chain is considerably larger than it would be if a single valve were used, and additional power supplies are required for the driver valve and its solenoid if it has one.

The length of a travelling wave tube helix is arrived at by the length required to establish a growing wave, the length to accommodate the attenuation and its matching sections, and the length required to give the designed gain and to restore the loss due to the attenuator. Typically, the first two items account for about half the total length, and for a given electron beam the total length of the attenuator is fairly constant, since most of this is occupied by matching. Consequently, if it is desired, for example, to double the gain of the valve, keeping the same output power, the total helix length is not doubled, but is increased by about 30 per cent. The total length of the electron beam is comprised also of the lengths occupied by coupling antennae, the collector and the electron gun, hence the increase in tube and solenoid length would be, for example, only about 25 per cent or less.

The question which naturally occurs when considering any very high gain device is whether it will be stable. Provided that the attenuation is adequate and well matched, instability of a travelling wave tube usually arises in the loss-free output helix. In order to achieve efficient operation, this region must provide about 30dB gain. For short-circuit stability, this means that the attenuation must have a power reflection coefficient of less than 1/1 000, i.e. a match better than 0·94 over the whole frequency range of the interaction. In fact, the output helix is not quite loss free, but it can be seen that increasing the gain of this part of the tube will rapidly lead to an impossible requirement of match due to the attenuator or helix imperfections. There is, however, no reason why the additional gain should not be obtained on the input side of the attenuator, and here the only apparent limit is that there must not be any loss-free region giving a greater gain than about 30dB, to avoid this region oscillating on its own. Thus any valves having a greater total gain than about 60 to 70dB will have to have multiple or extended attenuators. If the valve in question is a high-frequency high power valve, this would also be necessary to prevent backward wave oscillation. With this proviso it would appear that the minimum stable gain for which a travelling wave tube can be designed can increase indefinitely.

There is however a practical limit when one considers the noise powers involved. The noise figure which can be expected in practice from a normal medium-power travelling wave tube is about 27dB, hence the equivalent noise power at its input is $-$ 177dBW/c/s (taking kTB at 290°K $= -$ 204dBW/c/s). Hence for a tube giving, say, 150db gain, there will be a noise power at the output of $-$ 27dBW/c/s. A C-band tube, for example, would have a bandwidth of several thousand megacycles per second. If we take the bandwidth as 3 000 Mc/s, this becomes a total noise power of 68dBW, i.e. 6MW. In order to avoid gain degradation due to cross-modulation, such a tube would have to have a beam power of about 60MW. Clearly this is not practical. This problem would not necessarily arise in a chain of tubes, since the bandwidth could be restricted say by the coupling circuits at a lower level than that of the final output. The limit is thus fixed by the noise power, and it is clear that for any given beam power there is a maximum gain-bandwidth product which can be achieved, and further that this is directly proportional to the beam power. Approximately it could be said that: $G\ B = KW$ where G is the gain expressed as a factor, B is the bandwidth, W is the beam power and K is a constant depending on the dynamic range required and the cross-modulation tolerable.

More generally, it may be said that

$$\frac{\text{Saturated Output}}{\text{Noise Output}} = \text{Dynamic Range}$$

i.e.

$$\frac{\eta W}{GF\ kTB} = R \text{ where } R = \text{dynamic range}$$

W = Beam power

η = Maximum efficiency

G = Gain factor

F = Noise factor

T = Temperature (°K)

B = Bandwidth (c/s)

k = Boltzmann's constant

hence $(G+R+F)_{\text{dB}} = 144 +$

$$10 \log \frac{\eta W}{B} \begin{matrix}\text{(watts)}\\ \text{(Mc/s).}\end{matrix}$$

The efficiency η is often written as αC where α usually has a numerical value about 2 and C is the Pierce gain parameter. Then

$$(G+R+F)_{\text{dB}} = 144 + 10 \log \frac{\alpha K I^3}{4B}$$

where K = Circuit interaction impedance

I = Beam current

It is interesting but not surprising that the quantity which varies with the power per megacycle is the sum of the gain plus noise figure plus dynamic range. In some applications, e.g. in transmitters, the dynamic range can be made very small, since the input power can be maintained within close limits, making possible a higher gain per watt/megacycle than could be used in some receiver applications.

Yours faithfully,

C. H. Dix

Research Laboratories,
The General Electric Co. Ltd.
Wembley, England.

Power Supply Measurements : Temperature Measurements

Dear Sir,—The two simple pieces of test equipment described below, have been found of considerable use in this laboratory and may be of interest to readers of your journal.

Long Term Stability of Stabilized Power Supplies

The recurrent problem of measuring the stability and output impedance at d.c., of a stabilized h.t. supply is usually solved by balancing the output voltage against another voltage source and measuring changes on a sensitive meter between the two. These results always require interpretation in the light of the (presumably known) characteristics of the reference supply. The circuit of Fig. 1 shows a method of checking the h.t. line variations without the provision of an external source.

It depends, in this case of a 250V h.t. line, on the fact that the voltage dropped across three 83A1 reference tubes is within a few volts of 250V. The low incremented impedances of these tubes means that virtually the whole voltage movement of the h.t. line is impressed across the voltmeter which may be switched to a range with a full scale deflexion of a few volts. The stability and temperature

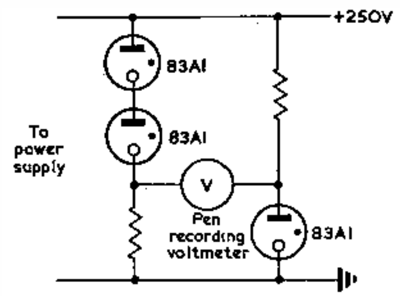

Fig. 1. Method of checking h.t. line variations

coefficent of this tube are at least as good as those of a Weston Standard cell. (Temperature co-efficient $\simeq$ 10 parts in 10^6/°C.)

Although Fig. 1 shows the circuit applied to a 250V h.t. line, other combinations of reference tubes can obviously be used for other supply voltages.

Ambient Temperature Variations

A constant-current fed, copper or platinum, resistance thermometer ar-

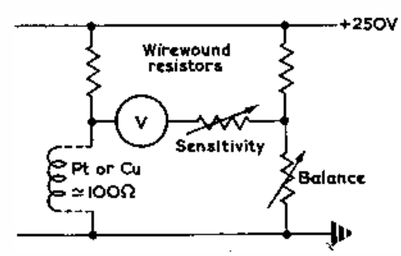

Fig. 2. Circuit for temperature measurement

ranged to run from the equipment h.t. supply lines is shown in Fig. 2 and is capable of an accuracy of about ± 1°C over the range 0 to 100°C.

In several investigations into circuit stability it was found convenient to use the same recording voltmeter, automatically switched between the circuits of Fig. 1 and Fig. 2 to give a continuous record of the supply and temperature variations to which the circuit under test was subjected.

Yours faithfully,

J. S. Johnston

Decca Radar Ltd.

BOOK REVIEWS

Les Ondes Centimetriques (Centimetric Waves)

By G. Raoult. 401 pp. 330 figs. Demy 8vo. Masson et Cie, Paris. 1958. Price 6.500fr.

THIS book is well filled with material intended for students making a fundamental study of centimetric waves as a basis for engineering application. The author is Professor of Physics at Clermont-Ferrand, where a course on centimetric waves is given to students with a wide range of qualifications. The author explains that for this reason mathematics has been included only where it is essential, and in many cases mathematical results are quoted without reasoned development. At the same time, a groundwork of mathematical physics is essential for any reader wishing to acquire a sound knowledge of the subject.

After an introductory chapter detailing the main properties of centimetric waves and their relationship to other parts of the spectrum, the idea of propagation in waveguides is developed from the concept of waves in free-space. A good knowledge of mathematics is essential to benefit from this material, in spite of the inclusion of physical representations wherever possible. The treatment of propagation in lines is more straightforward and is a useful fundamental study of the subject; adequate space is given to the uses of the Smith chart, with many useful examples of its application. The section on slotted lines summarizes the difficulties of construction; the mechanical accuracy required for satisfactory measurements is well stressed.

The practical application of waveguide is covered in three chapters which describe the action of various obstacles in the guide, and the bending elements, twists, junctions, couplings and transformers required. The numerous illustrations add value to the work; numerical quantities could be added in some cases, e.g., to the illustration on page 111.

More space is devoted to dielectric constant measurement than to either power or frequency measurement; the former subject is treated in a way which should appeal to the physicist, but the engineer might wish for greater stress on the practical problem of dielectric loss and an expansion of other aspects of centimetric wave measurement.

Klystron and magnetron oscillators are dealt with in some detail, and the more recent tubes are mentioned. The problem of amplification is briefly discussed, including a short section on "Maser" devices.

The properties of semiconductors are discussed in the chapter on detectors. The fundamental problem of noise factor is somewhat abbreviated but the essentials are covered. The ideas of available power and thermal agitation noise are introduced, but their importance would justify a more detailed treatment, particularly in a book dealing with fundamentals.

All types of aerials used in the centimetric band appear to be covered, with the necessary basic physics and the problems of polar diagram and impedance measurement are indicated. Reflector type aerials are given a separate chapter with many illustrations of the different types; lens aerials are given perhaps too much space, considering their limited economic application.

The mathematical appendices deal with imaginary notation, the Maxwell equations and units. Throughout the book references are given as footnotes.

This book is most suitable for students of microwave fundamentals who also wish to improve their French. It is evidently written in connexion with the author's course, and anyone following this will find the book invaluable. As a reference book for the English speaking engineer it is obviously of less value than the excellent British and American books on the subject. Despite what is said in the introduction, a high standard of mathematics is required to understand some parts of the material.

G. L. GRISDALE

Introduction to Non-Linear Analysis

By W. J. Cunningham. 349 pp. 80 figs. Demy 8vo. McGraw Hill Publishing Co. Ltd., New York, London. 1958. Price 74s.

THERE is an ever growing number of text books on mathematical subjects which are written for engineers who are supposed to be non-mathematical. As a consequence the 'niceties of the mathematics' are deliberately avoided in most cases. The work under review is no exception to this. However, there is a difference between Prof. Cunningham's book and many that have appeared since the war, for too often freedom from rigour has become licence for slap-dash mathematics, and in the reviewer's opinion this leads to slap-dash science. Prof. Cunningham has been at some pains to avoid this sort of thing with the result that we have a thoroughly useful and instructive text book. The field covered is a wide one, but the contents could be summed up by saying that the book is a systematic collection of techniques for obtaining solutions from non-linear differential equations. After the introduction the next two chapters are devoted to numerical and graphical methods respectively, then comes a chapter on known exact solutions, in which many of the more common functions met with in physics and engineering are discussed: the fifth chapter discusses the types of singularities which occur, and then follows a chapter on analytical methods. The next three chapters cover forced oscillating systems, systems described by differential-difference equations and linear differential equations with varying coefficients. The latter chapter (Chapter 9) would be beneficial to anyone beginning a study of so-called parametric amplification, in which there has been a revival of interest lately. The final chapter is on the stability of non-linear systems, a subject of paramount importance in non-linear work. Prof. Cunningham points out the difficulty of defining stability in non-linear systems, and then goes on to apply the known definitions to various types of system. There are many worked examples throughout the book, and also a set of problems for each chapter. No answers to the problems are given, and although many of the questions are so worded that the student has a fair idea of what the answer should be, nevertheless it might improve the value of a valuable book to supply a list of solutions. There is an extensive bibliography and an index at the end: there are also many figures and illustrations, and the book is well printed.

R. H. C. NEWTON

Telecommunications

By A. T. Starr. 459 pp. 80 figs. Demy 8vo. 2nd Edition. Sir Isaac Pitman & Sons Ltd. 1958. Price 37s. 6d.

THE preface to the second edition states that it differs from the first only in minor corrections and additions, and the comments made on the first edition (ELECTRONIC ENGINEERING, April, 1955, page 148) are still applicable. These were, in brief, that the book is a useful potted encyclopaedia, especially as it covers the construction of telegraph apparatus as well as telephone exchange equipment in the sections on line communication, but that it lacks references to original papers and more detailed works to supplement the necessarily condensed account of the whole field of line and radio communication which is given in this single book. It still defines the *neper* in terms of a power ratio, whereas the standard definition of the neper (unlike the decibel) is in terms of a current ratio.

The principle additions are: a brief account of pulse-code modulation in Chapter I; a discussion of iterative parameters of networks and a brief account of quartz crystals in Chapter V; and the addition to the Exercises throughout the book of a substantial number of questions from recent University of London examinations. The book does not yet contain any material on trunk dialling, cross-bar switches or electronic switching for telephone exchanges.

D. A. BELL

Principles of Noise

By J. J. Freeman. 299 pp. 167 figs. Demy 8vo. John Wiley & Sons, Inc., New York. Chapman & Hall Ltd., London. 1958. Price 74s.

DURING the past twenty years the subject of noise has moved from the domain of the physicist to that of the electronic engineer. Its importance, as judged by the outflow of technical literature, particularly in the field of information theory, has not, in the past, been matched by the appearance of text books suitable for studying the principles of this difficult subject. Dr. Freeman's book has been written to help fill this gap and, according to the preface, it should enable the student to 'read the literature with enough ease to use it as a professional tool'.

The first third of the book is concerned with the mathematical background, and this is covered in three chapters on Fourier series and integrals, probability, and stationary random processes. Following this, chapters four to six deal with matters more familiar to electronic engineers, viz. physical sources of noise, i.e. resistances and various types of valve, equivalent noise generators, and the determination and use of noise factors. An interesting discussion is then given on the minimum detectable signal in a linear system. Before passing to the corresponding subject of the detection of a.c. signals by a non-linear system, the nature of the Gaussian random process is considered. The book terminates with a short discussion of 'target' noise as an indication of the application of the theory to a practical problem.

In a reasonably sized book on a subject which is based on two broad mathematical techniques, it is inevitable that a compromise must be made in the selection and presentation of the basic material. The compromise in the present case is a fair one, though criticism can be made of the lack of attempt to highlight the more important definitions and formulae. Although such omission can be rectified by the lecturer when the book is used as a class text, a little more thought could have been given to the individual student who, in this country, is much more likely to be the reader.

The problems and worked examples form a useful addition to each chapter, but it is a pity that the author has relied for his bibliography on that published by Chessin some three years ago, which therefore misses much recent work. It also has some surprising omissions, particularly that of Dr. Norman Campbell's work on discontinuous phenomena (1910), which gave rise to the theorems which are used in the present book for much of the final analysis.

The book should provide a sound basis for a self-contained course on noise theory, and also serve as a source book on its fundamentals. As in other fields, the principles would need to be supplemented by further reading of the literature to make the student properly conversant with the subject.

W. H. ALDOUS

Halbleiter und Phosphore (Semiconductors and Phosphors)

684 pp. 200 figs. Demy 8vo. Friedr. Vieweg & Sohn, Brunswick. 1958. Price DM.68.

THE International Union of Pure and Applied Physics has been responsible for instigating a number of conferences on the subject of Semiconductors since 1950. The latest of this series was held at Rochester, U.S.A., in August 1958 and the preceding one was held at Garmisch-Partenkirchen, South Germany, in August 1956. It has been a general rule that the subjects dealt with at these conferences have been those relating to fundamental aspects of solid state physics. Any one who has been fortunate enough to attend one or more of these will know how valuable they have been in affording the opportunity to meet fellow workers in the same field. Both for these, and the less fortunate, the published proceedings of the conferences provide a valuable record of the general state of knowledge and ignorance at the time.

The present volume is a collection of nearly 100 papers which were presented at the Garmisch conference in 1956. In such a rapidly expanding field it is unfortunate that there has been a delay of nearly two years before the appearance of these proceedings, although it is apparent that some of the review papers have been expanded and brought up to date. The first half of the book contains a number of general reviews by leading authorities. The subjects covered include crystal growth, crystal defects, foreign atoms in semiconductors, statistics, surfaces, irradiation effects, magnetic susceptibility, thermoelectric effects, thermal conductivity and luminescence. Shorter contributions on specialized aspects are present in the second half. Eight of these short papers are in French and the other contributions are almost equally divided between English and German.

Specialists in the field of solid state research will certainly wish to have this book and it is appropriate to record here, on their behalf, their gratitude in particular to the editors and to all who have collaborated in its production for having successfully accomplished what must have been a formidable task.

J. R. DRABBLE

Atomic Energy

By Egon Larsen. 189 pp. 40 figs. Demy 8vo. George G. Harrap & Co. Ltd. 1958. Price 12s. 6d.

THIS book, written for those who know little about science but are looking for intelligible information, traces the story of atomic energy back to the end of our century when quite fantastic changes will have taken place. The book has been checked for scientific accuracy by the United Kingdom Atomic Energy Authority. It is illustrated with photographs and drawings; there is an appendix on careers in atomic science and engineering, and a glossary which explains things at a glance, together with a full index which makes the book a useful work of reference.

CHAPMAN & HALL

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37 ESSEX STREET, LONDON, W.C.2

Tables for Solving the Laplace Equation Inside an Ellipse

By A. I. Vzorova. 256 pp. Medium 4to. Cleaver-Hume Press Ltd. 1958. For Infosearch Ltd. Price 50s.

MR. Vzorova is at the U.S.S.R. Academy of Sciences Computing Centre and this book has been translated from the Russian by L. Herdan.

The tables are intended for the approximate solution of the first boundary value problem of the Laplace equation for the domain interior to an ellipse. All work connected with the computing, control, and preparation of the tables for printing has been carried out on electronic computers under the supervision of the scientific assistant, M. G. Rappoport.

Electronic Components Handbook Volume 2

Edited by Keith Henney and Craig Walsh. 357 pp. 150 figs. McGraw Hill Book Co. Inc., New York, Toronto, London. 1958. Price 97s.

THE components covered in this handbook are power sources and convertors, fuses and circuit-breakers, electrical indicating instruments, printed wiring boards, solder and fluxes, choppers, blowers, r.f. transmission lines and waveguides. For each component, this volume offers a general description of the kinds available. This includes information on both their advantages and disadvantages under various conditions.

This book was sponsored by the Electronic Components Laboratory, Wright Air Development Center, U.S. Air Force.

ELECTRONIC EQUIPMENT

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

(Voir page 55 pour la traduction Française: Deutsche Ubersetzung auf Seite 59)

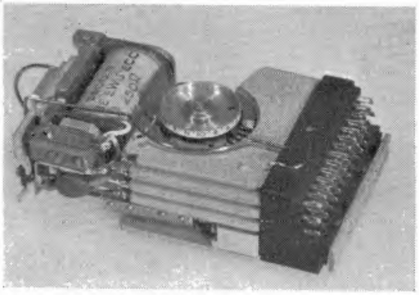
MINIATURE UNISELECTOR

(Illustrated below)

Siemens Edison Swann Ltd, 155 Charing Cross Road, London, W.C.2.

THIS small plug-in uniselector switch has been introduced for use in automatic systems: it has 36 outlets and measures 3¾in by 2 3/16in by 1 5/32in (95 by 55 by 30mm) and weighs 12oz (341g). The uniselector is designed for use in 22 or 50V circuits and incorporates its own spark quench device. It is claimed that the unit will operate for two million operations without need for adjustment and that it has a working life of at least eight million revolutions before replacement of any part will be necessary.

There are three main elements: a driving mechanism and wiper assembly;



a bank of contacts over which the wipers search at 60 steps/sec; and a jack which is pre-wired and assembled to the mounting enabling the uniselector mechanism to be plugged in and out without disturbing any wiring.

The driving mechanism is mounted on a steel side plate and has all parts readily accessible. It is operated by a ratchet and pawl system of the reverse drive type having a single driving magnet and a single break action interrupter spring set with tungsten contacts. The wiper assembly comprises three pairs of nickel silver treble-ended wipers of either bridging or non-bridging type. There is a number disk.

The bank comprises 36 contact segments arranged in three levels. The design of the bank is such that one of the arms of the wiper assembly is always touching a specially fitted solid contact segment, and this arrangement enables feeder springs to be dispensed with.

EE 5751 *for further details*

POWER SUPPLY UNIT

E.I.R. Instruments Ltd, 329 Kilburn Lane, London, W.9.

THIS unit provides a convenient continuously variable source of alternating or direct voltage and alternating and direct current.

The unit is mains driven and comprises two separate supplies built into a common panel and chassis assembly, these supplies may be switched to give either 'voltage' or 'current' output.

When switched for d.c. output metal rectifier and smoothing circuits come into operation.

The same continuously variable control is used on both the 'voltage' and 'current' outputs and is switched from one to the other as required.

Variable transformers are not used for this purpose but control is effected by means of 'coarse' and 'fine' heavy duty potentiometers connected across the input to the unit. This method enables the output of either voltage or current to be varied from virtually zero up to the maximum.

The outputs are: Alternating or direct current from 0 to 10A. Alternating or direct voltage from 0 to 1kV at 60mA. Outputs are floating with respect to panel chassis and case, which enables a more flexible use to be made of the equipment especially when voltages not connected to 'earth' are required. All controls are mounted together with the mains fuses on the front panel, their use is simple and obvious.

EE 5752 *for further details*

DIRECTIONAL COUPLER

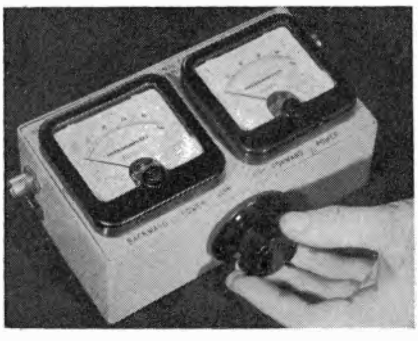
(Illustrated below)

Sir W. G. Armstrong Whitworth Aircraft Ltd, Baginton Aerodrome, Nr. Coventry, Warwickshire.

THIS coupler is of the 'loop' type and is suitable for measurements of r.f. power and s.w.r. (standing wave ratio) in small to medium power coaxial cables and aerial systems.

The directional coupler may be inserted at any point in series with the line in which power flow is to be measured. However, for measurement of power above 3W it is recommended that the point of insertion should be as far from the generator as possible.

Although this is a wide band device the meter readings are dependent on the frequency of transmission, the calibration curve provided with each instrument is accurate only at the frequency indicated on the calibration chart. The directional properties are, however, largely unaffected by frequency changes, so that the coupler may be used as an aid to obtaining optimum termination



of a 52Ω coaxial system, for frequencies up to 600Mc/s.

EE 5753 *for further details*

CRYSTAL CLOCK

(Illustrated below)

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey.

THIS instrument (type TSA) has been designed as a compact and portable frequency source of high accuracy. The equipment uses transistors throughout and incorporates a 50c/s synchronous clock with a seconds hand.

The internal clock can be omitted if desired and a 12V 50c/s synchronous motor can be supplied for external use.

Switched standard output frequencies of 10kc/s, 1kc/s, 100c/s and 50c/s are available, via a coaxial socket mounted on the front panel. The emitter-follower output circuit incorporates series resistance to prevent damage to the final transistor due to accidental short-circuiting of the output. Provision can be made for mains, accumulator, or dry battery operation.

The accuracy is within 1sec per week operating in ambient temperature.

EE 5754 *for further details*

NON-DESTRUCTIVE INSULATION TESTER

(Illustrated above right)

A/S Danbridge, Frederikssundsvej 60C, Copenhagen, Denmark. Distributed by: B & K Laboratories Ltd, 4 Tilney Street, London, W.1.

THE insulation tester type JP1 comprises a high frequency generator supplying the test voltage. This may be varied continuously by a potentiometer up to a maximum of 12kV d.c.

The output voltage is indicated on a meter with two ranges in order to allow accurate measurements at lower voltages.

The test voltage is fed through a screened coaxial cable to an insulated test prod. A safety switch is incorporated in the handle of the test prod. On closing this switch the high tension is switched on through a relay thus providing a maximum of safety for the operator.

A socket is provided on the instrument so that if required the relay may be operated remotely. The 'low' terminal of the test voltage is earthed directly so that earthed objects may easily be tested and in addition hum disturbances are largely eliminated.

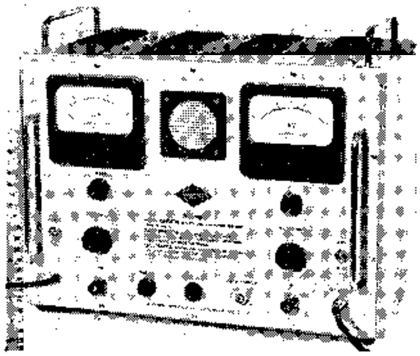
If ionization occurs in the test object the resulting current develops small voltage pulses in the primary of a transformer connected in series with the test voltage.

The secondary voltage is fed into a three-stage amplifier driving an electrostatic loudspeaker. Thus even very minute discharges provide an audible noise in the loudspeaker.

The amplifier gain may be adjusted by a potentiometer to suit the local noise level.

Any leakage currents in the test object may be checked by a microammeter on the front panel. Resistance may be calculated from the voltmeter and microammeter readings if required.

The instrument is of robust construction and designed for continuous use.



All components belonging to the h.t. circuit are mounted in an oil filled tank to ensure reliable working under all conditions.

The instrument is inherently safe to operate, as the output current is limited to a safe value by the high generator impedance, the safety switch also reducing the risk of accidental contact with points at high tension.

EE 5755 *for further details*

TWO-PHASE LOW-FREQUENCY DECADE OSCILLATOR

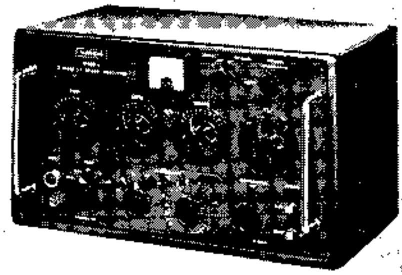
(Illustrated above right)

Muirhead & Co. Ltd, Beckenham, Kent.

THE frequency range of the D-880-A two-phase decade oscillator is 0·01c/s to 11·2kc/s. Over the whole of this range the output is constant within ±1dB.

The circuit of this oscillator, which represents a new approach to the problem of providing very low frequency selective circuits, is based on a development of a circuit covered by British Patent No. 781374 granted to the National Research Development Corporation; the instrument is made under licence agreement and incorporates several improvements.

The principle of operation is that of solving a second order differential equation with damping by using two integrator stages and a sign changer.



More simply, the overall phase shift of 360° which is required for oscillation to occur is achieved by two electronic integrating circuits—each having a phase shift of 90°—and a sign changer circuit having a phase shift of 180°. Two outputs—90° apart—are taken from the integrator circuits and are available at a level of 10V. Both outputs are directly coupled and the d.c. content is adjusted to a very low value by preset controls. The '0°' output' is suitable for feeding into a load of 600Ω and is adjustable by an output attenuator down to −62dB ref 10V; the 90° output is suitable for a load of 10kΩ and is adjustable down to zero by means of a potentiometer.

In addition to forming part of the Muirhead Transfer Function Analyser, this oscillator finds many applications in the servo-engineering field where the study of low frequencies is of increasing importance.

Other features of the instrument include its frequency accuracy, its very low harmonic content and the equality, or proportionality, of the two output levels.

EE 5756 *for further details*

WIDE RANGE OSCILLATOR

(Illustrated below)

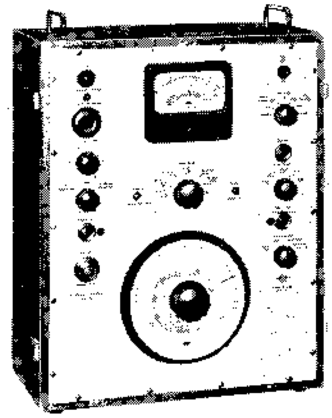
Dawe Instruments Ltd, 99 Uxbridge Road, London, W.5.

DAWE type 441 wide-range oscillator is a resistance-capacitance tuned design covering all frequencies from 5c/s to 600kc/s in five ranges with either balanced or unbalanced output. The same instrument can thus be used to obtain complete coverage from sub-audio to low radio frequencies. The wide range, coupled with the low distortion and constant output associated with this type of tuning, offers considerable advantages for vibration investigations, testing of wide-band amplifiers, transducers, medical equipment and carrier communications systems, as well as for the majority of routine measurements within the audio range.

This unit also comprises a modified Wien bridge acting as a frequency controlling network, the output from which feeds a balanced push-pull amplifier with cathode-follower output arranged for either balanced or unbalanced operation. The output is 0·1W (7·75V) into 600Ω, the open-circuit voltage being 15·5V. The source impedance is 600Ω ±20 per cent.

The good stabilization arrangements together with the large amount of negative feedback at harmonic frequencies, ensures a very pure sine wave output. A typical distortion figure is below ½ per cent at all frequencies, except below 20c/s where the distortion increases to about ½ per cent at 5c/s. Hum and noise voltages are below 0·1 per cent except near the frequency of the supply mains, at which the hum may increase up to about 0·2 per cent.

EE 5757 *for further details*



AUDIO FREQUENCY SPECTROMETER

(Illustrated above)

Brüel & Kjær, Nærum, Denmark. Distributed by: B & K Laboratories Ltd, 4 Tilney Street, London, W.1.

THIS instrument has been developed basically to facilitate rapid and accurate frequency analysis of complex signals such as the ones encountered in noise and vibration measurements. It may, however, also be used as a true r.m.s. reading valve-voltmeter and for harmonic analysis of electrical signals.

It is basically a frequency analyser which divides the frequency band 35c/s to 35kc/s into ⅓ octave ranges by means of 30 band-pass filters with very high selectivity (40dB rejection at 2 times pass-band centre frequency) and zero pass-band attenuation.

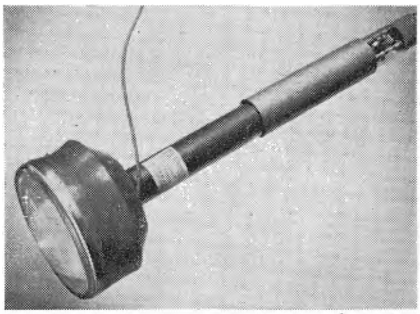
The inputs to all filters are paralleled and frequency band selection is carried out by means of a switch in the filter output circuit. The switch can be operated manually or be driven automatically with the aid of an external motor drive. When the motor drive of the graphic Level Recorder type 2304 is used for this purpose the switching operation can be synchronized with the paper drive of the Level Recorder in such a way that spectrograms (frequency-amplitude diagrams) are recorded automatically on pre-printed, calibrated recording paper. Due to pre-loading of the filters, ringing time is kept to a minimum, and a complete frequency amplitude diagram can be recorded in less than 20sec.

An important feature of the Spectrometer is its ability to measure not only the arithmetic average value of the input signal, but also the true r.m.s. value (for signals with crest factors up to 5) as well as the peak value. The desired type of indication is selected by means of a switch.

The instrument can also be switched to operate as a linear amplifier and valve-voltmeter in the frequency range 2c/s to 35kc/s. Two different meter damping characteristics are provided so that accurate measurements and easy meter reading are obtained also at the lower frequencies.

Furthermore, the four internationally standardized weighting networks for sound level measurements are included, and can be switched in whenever desired.

EE 5758 *for further details*



MICRO-SPOT CATHODE-RAY TUBE

(Illustrated above)

Ferranti Ltd, Hollinwood, Manchester, Lancashire.

CLAIMED to be the only one of its type in the world, a new micro-spot cathode-ray tube capable of resolving 5 000 lines, is now available from the Electronics Department of Ferranti.

Measuring 5in (127mm) across, the 5/71 CM tube, as it is known, has been developed for airborne applications such as aerial mapping. The spot size is considerably less than ·001in (0·025mm) in diameter.

The high resolution of the tube has been made possible by the use of an exceedingly fine screen and a novel design of electron gun using two focusing elements, one of which is electromagnetic and external to the tube, while the other is electrostatic and of fixed focal length.

The tube has an optical flat face with a non-darkening glass and a short cylindrical bulb, coated—except over the screen surface—with a thick layer of plastic resin, enabling it to be operated under adverse atmospheric conditions, viz. 30kV at 75 000ft (24kM) without danger of e.h.t. breakdown. Similarly, no tube socket is required, the leads to the rear end of the tube being encapsulated in a way which does not hinder the tube from being mounted easily in its focus and scanning assembly.

While the new tube is primarily intended for the display or photography of repetitive information, single transients may be photographed at writing speeds which are limited by the scan coil requirements.

The standard phosphor used in the tube has green fluorescence, decaying to 1/e level within 10μsec after removal of excitation. It is also suitable for use with many types of photographic material. Tubes with a blue fluorescence can also be supplied.

Great care must be taken in considering the quality of e.h.t., scanning and focus supplies for the tube, since multiple effects due to e.h.t. ripple, imperfect focusing and poor scanning fields can cause such an enlargement of the spot that no advantage is apparent when using these tubes.

Ferranti Limited can provide scanning and focus coils specially designed for use with the type 5/71CM tube, together with a small alignment magnet which is clamped to the neck of the tube in the region of the cathode.

EE 5759 *for further details*

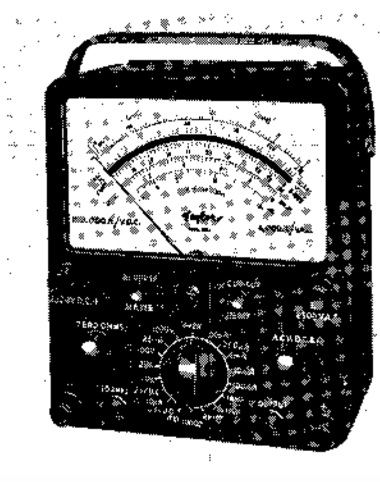
MULTI-RANGE METER

(Illustrated below)

Taylor Electrical Instruments Ltd, Monstrose Avenue, Slough, Buckinghamshire.

THE main feature of this multi-range meter (type 100A) is that it has a sensitivity of 100kΩ/V. The instrument incorporates an 8μA centre-pole moving-coil movement and has a mechanical overload protection device which is operative on all ranges, and also a reverse polarity facility.

The d.c. voltage range is from 0·5V to 2·5V and the a.c. range 10V to 2·5kV. The current ranges are 10μA to 10A d.c. and the resistance ranges are 2kΩ to 200MΩ f.s.d. An adaptor is available to increase the voltage range up to 25kV d.c.



EE 5760 *for further details*

SUB-MINIATURE RADIO COMPASS

(Illustrated below)

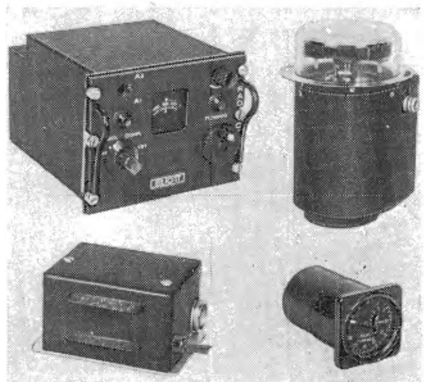
Elliott Brothers (London) Ltd, Elstree Way, Borehamwood, Hertfordshire.

A NEW lightweight sub-miniature aircraft radio compass is being built by Elliotts under licence from its designers, Air-Equipment of Asnieres, France, and is primarily intended for applications in which weight, size and price are all-limiting factors. Nevertheless, it is highly accurate, sensitive and particularly suited for fitting to all types of light military and civil aircraft, medium sized commercial aircraft and helicopters.

The complete system consists of three units—the receiver, loop aerial and bearing indicator—which when installed in the aircraft provide the pilot with continuous automatic indication of relative bearing and communication reception. These facilities are used for m.c.w., and r.t. operation within the frequency range of 200 to 800kc/s.

The loop aerial, which may be installed above or below the fuselage, is wound on a ferrite core and fixed in an isolating mount which protrudes through the fuselage by only 1½in (38mm).

The loop drive and indicator transmission system are housed in this unit,



which measures 4½in by 3½in by 5in (113 by 89 by 127mm) long, the whole assembly being hermetically sealed and completely enclosed.

The receiver, housed in the control box itself and measuring 4½in by 5½in by 6½in (113 by 140 by 165mm), consists of the following sub-units:

(a) R.F. unit, using sub-miniature valves.

(b) I.F. unit with detector, local oscillator and a.g.c. circuits (the i.f. amplifier and a.g.c. use sub-miniature valves, the local oscillator and detector circuits are transistorized).

(c) Servo amplifier unit—fully transistorized.

(d) A.F. unit—fully transistorized.

The indicator is supplied in two alternative sizes having diameters of 3¼in (82·5mm) or 2¼in (57mm), with scales marked in 2° graduations.

The power unit supplies all necessary power for the system and uses a transistor oscillator, enabling it to operate direct from the aircraft 28V d.c. supply.

EE 5761 *for further details*

Short News Items

The Seventh Annual Convention of the Scientific Instrument Manufacturers' Association was held at the Hotel Majestic, Harrogate, from 6-9 November. Among more than 200 participants were many delegates from the big instrument user industries. Groups represented were iron and steel; oil and chemicals; textiles and fibres; fuel and power. The privileges of SIMA membership for this special occasion were extended to representatives of Commonwealth Dominions, among them Canada, Australia, New Zealand, South Africa, India and Pakistan: foreign Embassies in London including the U.S.A., U.S.S.R., France and Germany; and to the press. At this year's convention members of the Scientific Instrument Manufacturers' Association of Great Britain for the first time invited the big users of instrumentation equipment to meet them in force to talk of current trends and future needs. His Worship the Mayor of Harrogate, Councillor Bernard H. Wood, was the principal guest of honour at the convention. As a result of SIMA talks with some of the big instrument users plans are revealed to develop, through the Engineering Equipment Users' Association, regular liaison among British manufacturers concerned with instruments to aid production. Details have now to be approved. It can, however, be said that the Distillers Company and Imperial Chemical Industries are among those expected to be taking part in developing the scheme. The president, Mr. R. B. Brock, stressed that the majority of the larger industries well know the value of scientific industrial instrumentation. 'The instrument industry will have to listen more to what the big users have to say about their needs.' He was convinced, however, that there is room for more British instruments among some small firms backward in realizing that scientific industrial measurement and control instrumentation can save time and money. The outcome of technical discussions was summed up in plenary session by the four panel chairmen: Mr. A. H. Campbell, Hilger & Watts Ltd; Mr. R. B. Abraham, C. F. Casella & Co. Ltd; Mr. J. Steward, R. B. Pullin & Co. Ltd; Mr. G. M. Sisson, Sir Howard Grubb Parsons & Co. Ltd. A main basic requirement for the future in all industries is for accurate 'ruggedized' instruments to monitor continuously the end product and to provide fast 'feedback' controlling input of materials and adjustments of machines along the processing line.

International Convention and Exhibition on Transistors. The Radio and Telecommunication Section of the Institution of Electrical Engineers is holding an International Convention on Transistors and Associated Semiconductor Devices at Earls Court from 21-27 May 1959. Under the same roof, the I.E.E. is taking the initiative in promoting the first international scientific exhibition devoted entirely to these devices. Acceptances of invitations to deliver the opening lectures at the Convention have already been received from the American inventors of the transistor, Dr. W. B. Shockley, Dr. W. H. Brittain and Professor J. Bardeen. The Convention will be divided into 23 sessions, 12 of them general and 11 specialist. Apart from the opening session, which will be concerned with world trends in transistors and associated semiconductor devices, the sessions will cover: Materials; Basic Theory; Characteristics and their Measurements, and Transistors as a Circuit element; Technology and Design and Manufacture; and Applications of Transistors. Including over 30 lectures from invited speakers of international repute, the total number of papers to be submitted at the Convention is expected to exceed 200. The text of the papers presented, together with the discussions, will be subsequently published in a special Supplement to Part B of the Proceedings of the Institution of Electrical Engineers. Further particulars, including details regarding registration, can be obtained from the Secretary, the Institution of Electrical Engineers, Savoy Place, London, W.C.2.

The exhibition will be essentially scientific in character and will cover all aspects of transistors and semiconductor devices and their applications. The International Transistor Exhibition is being organized on behalf of the Institution of Electrical Engineers by Industrial and Trade Fairs Ltd, Drury House, Russell Street, London, W.C.2.

BINDING OF VOLUMES

Arrangements are now in hand for binding the 1958 volume at an inclusive charge of 25s. Copies will be bound, complete with index and with advertising pages removed, in a good quality red cloth covered case blocked in gold on the spine.

Home and Overseas readers who wish to have their copies bound are asked to comply with the following instructions:—

(1) Tie the twelve issues (January to December, 1958) securely together before parcelling.

(2) Enclose a remittance for 25s. and a gummed label bearing the sender's name and address. (A cheque or Postal Order should be made payable to Morgan Bros. (Publishers) Ltd.).

(3) Enclose the copies, remittance and label in a closed parcel and address to:—
The Circulation Dept. (E.E. Binding), 28, Essex Street, Strand, London, W.C.2.

(No other correspondence is necessary.)

★ ★ ★ ★

The following will also be available from our Circulation Dept.:—

Complete Bound Volumes for 1958. Price £3 3s. 0d., post free.

Binding Cases for twelve issues. Price 7s. 6d. postage 9d.

The Index for Volume 30 (1958) free.

The Sixteenth Annual Radio and Electronic Component Show, organized by the Radio and Electronic Component Manufacturers' Federation, is to be held at Grosvenor House and Park Lane House, London, W.1, from 6-9 April. Admission (from 10 a.m. to 6 p.m. daily) is by invitation only, applications for tickets to be made to the Secretary, R.E.C.M.F., 21 Tothill Street, London, S.W.1.

The Council of the British Institution of Radio Engineers announces that the president-elect for 1958-59 is Professor E. E. Zepler, who occupies the Chair of Electronics at the University of Southampton.

Blackburn & General Aircraft Ltd, makers of the Navy's N.A.39 bomber, have decided on a new venture—to design and manufacture electronic equipment commercially. A new organization, with its own design and development departments and production and sales organizations has been formed at Brough under Dr. Harry Fuchs. This new addition to the Group's varied range of activities which embrace aircraft, gas turbines, building and civil engineering, mowers, rollers and agricultural equipment and motor engineering, will concentrate on electronic equipment for industrial applications, such as data handling, control and instrumentation.

Chimel S.A. of Geneva announce the development of a broad new range of solid electrolyte tantalum capacitors. They are one of the pioneers in the development of sintered tantalum capacitors. The new series include capacitors of up to 400μF at 6V, 220μF at 10V, 150μF at 15V, 100μF at 20V, and 22μF at 35V. These units supplement the wide range of existing Chimel solid electrolyte tantalum capacitors.

Errata. In the article, 'A Three-Band Aerial Combining Network' by S. L. Fife which appeared in the December issue, the following amendments should be made. On page 721 the blocks for Fig. 3(a)(b) and Fig. 4(a)(b) should be transposed. On page 722, line 2, left-hand column, should read
'B = phase constant of feeder $(2\pi/\lambda)$.'

In the write-up on the Benson-Lehner Oscillograph Trace Reader and Electro-Plotter on pages 728-9 of the December issue the illustrative blocks should be transposed.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 28 January. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: Speech Recognition and the Phonetic Typewriter.
By: D. B. Fry.

South Wales Section

Date: 14 January. Time: 6.30 p.m.
Held at: Cardiff College of Advanced Technology.
Lecture: Computer Control of Machine Tools.
By: H. Ogden.

South Western Section

Date: 29 January. Time: 7 p.m.
Held at: The School of Management Studies, Unity Street, Bristol 1.
Lecture: Recent Developments in Printed and Potted Circuits.
By: H. G. Manfield.

South Midlands Section

Date: 29 January. Time: 7 p.m.
Held at: The Winter Gardens, Malvern.
Lecture: Industrial and Underwater Television.
By: B. V. Somes-Charlton.

West Midlands Section

Date: 21 January. Time: 7.15 p.m.
Held at: The Wolverhampton and Staffordshire College of Technology, Wulfruna Street, Wolverhampton
Lecture: Learning Machines.
By: P. Huggins.

North Eastern Section

Date: 12 January. Time: 6 p.m.
Held at: The Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle upon Tyne.
Lecture: The Design and Construction of an Electronic Digital Computer.
By: R. W. Walker.

BRITISH SOUND RECORDING ASSOCIATION

Date: 16 January. Time: 7.15 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.
Lecture: Some Further Developments in Loudspeakers.
By: A. R. Neve.

THE INSTITUTE OF NAVIGATION

Date: 16 January. Time: 5.15 p.m.
Held at: The Royal Geographical Society, 1 Kensington Gore, London, S.W.7.
Lecture: Blind Landing Problems.
By: W. J. Charnley.

North Lancashire Sub-Centre

Date: 14 January. Time: 7.15 p.m.
Held at: The North-Western Electricity Board Demonstration Theatre, Friargate, Preston.
Lecture: British Columbia-Vancouver Island 138kV Submarine Power Cable.
By: T. Ingledow, R. M Fairfield, E. L. Davey, K. S. Brazier and J. N. Gibson.

South-East Scotland Sub-Centre

Date: 6 January. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: Electrical Installation at Calder Hall Nuclear Power Station.
By: N. J. Mackay and E. Hardwick.
Date: 7 January. (Time and Place as above.)
Lecture: A New Cathode-Ray Tube for Monochrome and Colour Television.
By: D. Gabor, P. R. Stuart and P. G. Kalman.

South-West Scotland Sub-Centre

Date: 6 January. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: A New Cathode-Ray Tube for Monochrome and Colour Television.
By: D. Gabor, P. R. Stuart and P. G. Kalman.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Ordinary Meetings

Date: 8 January.
Lecture: First Hunter Memorial Lecture.
By: L. G. Brazier.
Date: 22 January
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

Measurement and Utilization Sections

Date: 6 January.
Lecture: Silicone Electrical Insulation.
By: J. H. Davis.

Radio and Telecommunication Section

Date: 5 January.
Lecture: The Application of Transistors to Line Communication Equipment.
By: H. T. Prior, D. J. R. Chapman and A. A. M. Whitehead
Date: 19 January.
Lecture: High-Current-Density Thermionic Emitters.
By: A. H. W. Beck.
Date: 21 January.
Lecture: Dielectric Materials—Trends and Prospects.
By: C. G. Garton.
Date: 29-30 January.
Convention on Long-Distance Transmission by Waveguide.
(All wishing to attend must register; forms available on application).

Faraday Lecture

Date: 26 January. Time: 6 p.m
Held at: The Royal Festival Hall.
Lecture: Automation.
(Admission by ticket available from the Institution).

North-Eastern Radio and Measurement Group

Date: 19 January. Time: 5.30 p.m.
Held at: Rutherford College of Technology, Northumberland Road, Newcastle upon Tyne.
Lecture: Power-System Automatic Frequency Control Techniques.
By: F. Moran.

Sheffield Sub-Centre

Date: 22 January. Time: 6.30 p.m.
Held at: The Grand Hotel, Sheffield.
Informal Lecture.
By: Sir Christopher Hinton.

North-Western Measurement and Control Group

Date: 20 January. Time: 6.15 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: The Atomic Clock.
By: L. Essen.

South Midland Centre

Date: 5 January. Time: 6 p.m.
Held at: The College of Technology, Gosta Green.
Lecture: Design of the 330kV Transmission System for Rhodesia.
By: F. C. Winfield, T. W. Wilcox and G. Lyon.
Date: 12 January. Time: 7.30 p.m.
Held at: Malvern Winter Gardens.
Lecture: Electrostatic Loudspeakers.
By: P. J. Walker.

South Midland Radio and Measurement Group

Date: 26 January. Time: 6 p.m.
Held at: James Watt Memorial Institute, Birmingham.
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

North Staffordshire Sub-Centre

Date: 16 January. Time: 7 p.m.
Held at: The Technical College, Stoke-on-Trent.
Lecture: Operational Experience at Calder Hall.
By: K. L. Stretch.

Southern Centre

Date: 7 January. Time: 6.30 p.m.
Held at: The C.E.G.B. Offices, 111 High Street, Portsmouth.
Informal Evening on Ultrasonics in Industry.
Talk by: C. F. Brockelsby.
Date: 28 January. Time. 6.30 p.m.
Held at: The College of Technology, Anglesea Road, Portsmouth.
Lecture: Geo-electric Survey.
By: L. S. Palmer.
Date: 30 January. Time: 6.30 p.m.
Held at: South Dorset Technical College, Weymouth.
Lecture: Radio Observations on the Artificial Satellites.
By: R. L. F. Boyd.

Western Centre

Date: 12 January. Time: 6 p.m.
Held at: The South Wales Institute of Engineers, Park Place, Cardiff.
Lecture: The Design of the 330kV Transmission System for Rhodesia.
By: F. C. Winfield, T. W. Wilcox and G. Lyon.

Western Supply Group

Date: 19 January. Time: 6 p.m.
Held at: The South Wales Institute of Engineers, Park Place, Cardiff.
Lecture: The Future Pattern of Nuclear Electricity Power Stations.
By: I. Everson.

Western Utilization Group

Date: 26 January. Time: 6 p.m
Held at: The South Wales Institute of Engineers, Park Place, Cardiff.
Lecture: A Transatlantic Telephone Cable.
By: Sir Gordon Radley, K.C.B., C.B.E.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

Date: 5 January. Time: 6 p.m.
Held at: Manson House, Portland Place, London, W.1.
Lecture: Control Systems as Applied to Railways Signalling.
By: J. C. Kubale.
Date: 14 January (Time and place as above).
Symposium on Storage Media.
Date: 27 January. (Time and place as above).
Symposium on Flow Measurement.

THE TELEVISION SOCIETY

Date: 9 January. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Discussion: A Television Link with America?
Date: 22 January.
The Fleming Memorial Lecture: Modern Optics in Relation to Television.
By: G. H. Cook.
(This meeting is not held at Shaftesbury Avenue, and admission is by ticket only).

PUBLICATIONS RECEIVED

IMPERIAL COLLEGE OF SCIENCE AND TECHNOLOGY (UNIVERSITY OF LONDON) CITY AND GUILDS COLLEGE RESEARCH REPORT 1955-58. As is no doubt known, the City and Guilds College represent engineering within the Imperial College. The object of the present report is to describe some of the research in engineering which has been carried out by the staff and students of the College during the last three years. The items selected have been chosen for their general interest to industry and outside bodies. Imperial College of Science and Technology (University of London), City and Guilds College, South Kensington, London, S.W.7.

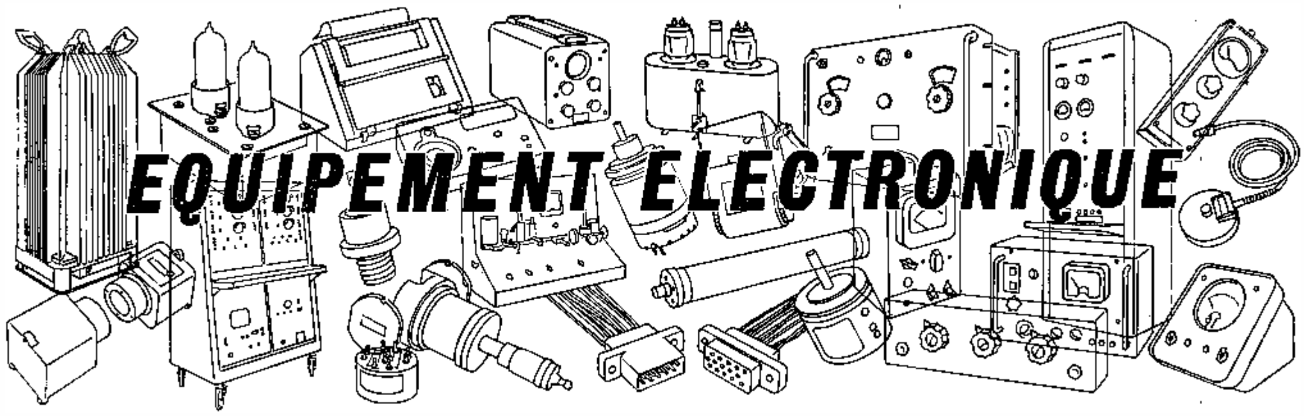
ROYAL MILITARY COLLEGE OF SCIENCE ANNUAL REPORT 1956-7, which covers the year to 31 July, 1957, has recently been published. It is now eleven years since the College was installed at Shrivenham. The programme of improvements to instructional and living accommodation, estimated to cost £100 000, was launched and is proceeding according to plan.

PROGRAMMES AVAILABLE IN THE INTERCHANGE SCHEME is the title of a document issued by Ferranti Ltd. to publicize the computer programmes of which they have been informed to date. They seek to encourage all users of a Mercury computer to advise them of the titles of all programmes written, or being written, subject to considerations. Ferranti issue lists of these titles from time to time, giving also (unless requested not to) the name of the author and his organization. The aim is to enable any other worker to get into direct touch with the originator of a programme if further details would be of interest. Ferranti Ltd., Computer Department, West Gorton, Manchester, 12, and 21 Portland Place, London, W.1.

BUSINESS AND SPECIALIZED PUBLICATIONS OF GREAT BRITAIN is the title of a booklet published by the Council of the Trade and Technical Press to meet what it believes to be a long-felt want for concise information about the business and specialized press of Great Britain. It contains brief details of journals listed in alphabetical order, together with similar details of year books and other annual publications in a separate section. At the end will be found an index, classified under subject matter, where those concerned with a particular sphere of activity can find grouped together the publications likely to be of most interest to them. All the journals listed in this booklet are in membership of the Council of the Trade and Technical Press, which is a section of the Periodical Proprietors Association. They fall under a variety of headings, technical, trade, professional, specialized and export, and range in frequency of publication from weekly to quarterly. The Council of The Trade & Technical Press, Imperial House, Kingsway, London, W.C.2.

THIS IS MAGNESIUM is a booklet dealing with the properties and uses of magnesium and its alloys. It shows a little of what magnesium is doing in a broad cross-section of modern industry. Magnesium Industry Council, Dickens House, 15 Took's Court, London, E.C.4.

(Pages 55 to 62 French and German section)



EQUIPEMENT ELECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai. *(Traduction des pages 50 aux pages 52)*

SÉLECTEUR UNIDIRECTIONNEL MINIATURE

(Illustration à la page 50)

Siemens Edison Swan Ltd, 155 Charing Cross Road, W.C.2.

CE petit sélecteur unidirectionnel à prises a été conçu à l'usage des systèmes automatiques: il est doté de 36 prises, mesure 95 × 55 × 30mm et son poids est de 341 grammes. Destiné à l'usage des circuits de 22 ou 50V, il comprend son propre circuit éliminateur d'étincelles et peut remplir jusqu'à deux millions d'opérations sans qu'il soit besoin de réglage. Sa durée de fonctionnement est d'au moins huit millions de révolutions avant qu'il n'y ait nécessité de remplacer une pièce quelconque.

Il comporte trois éléments principaux: un mécanisme d'entraînement et un ensemble de frotteurs; une batterie de contacts sur laquelle les frotteurs avancent à 60 plots par seconde; et un jack dont les fils sont pré-reliés à la monture, permettant ainsi de brancher et de débrancher le mécanisme unidirectionnel sans déranger l'enroulement.

La batterie de contacts comprend 36 segments de contact disposés sur trois étages. Elle a été conçue de manière à ce qu'un des bras de l'ensemble de frotteurs touche constamment un robuste segment de contact spécialement fixé à cet effet. Grâce à ce dispositif, l'emploi de ressorts d'alimentation est rendu inutile.

EE 5751 *pour plus amples renseignements*

BLOC D'ALIMENTATION

E.I.R. Instruments Ltd, 329 Kilburn Lane, London, W.9.

CE Bloc constitue une source à variation continue de tension alternative ou continue et de courant alternatif ou continu.

Il fonctionne sur secteur et comporte deux éléments d'alimentation encastrés dans un panneau et un ensemble-châssis communs. Ces éléments peuvent être branchés pour débit de tension ou de courant.

Lorsque le courant continu est branché, cela actionne un redresseur métallique et des filtres éliminateurs d'ondulations de courant.

La même commande à variation continue est employée tant pour le débit de tension que pour celui de courant et elle est branchée sur l'un ou l'autre, selon les besoins.

On n'emploie pas de transformateurs à rapport variable à cet effet et le contrôle s'effectue au moyen de potentiomètres de grande puissance, "approximatifs" et "précis", reliés au bloc par les contacts d'entrée. Cette méthode permet de varier le débit de tension ou de courant, virtuellement de zéro au maximum.

Les valeurs de débit sont: 0 à 10A pour le courant alternatif ou continu; 0 à 1kV—à 60mA—pour la tension alternative ou continue.

EE 5752 *pour plus amples renseignements*

COUPLEUR DIRECTIF

(Illustration à la page 50)

Sir W. G. Armstrong Whitworth Aircraft Ltd. Baginton Aerodrome, near Coventry, Warwickshire.

CE coupleur est du type "à boucle" et convient aux mesures de hautes fréquences et de t.o.s. (taux d'ondes stationnaires) dans les câbles coaxiaux de faible et moyenne puissance et les systèmes d'antennes.

Le coupleur directif peut être intercalé à n'importe quel point, en série avec la ligne dont le débit de puissance doit être mesuré. Cependant, pour des mesures de puissance supérieures à 3W, il est conseillé d'intercaler le coupleur aussi loin que possible du générateur.

Quoiqu'il s'agisse d'un appareil à large bande, les indications de l'enregistreur dépendent de la fréquence de transmission, la courbe d'étalonnage fournie avec chaque appareil n'étant précise qu'à la fréquence indiquée sur le diagramme d'étalonnage. Les propriétés directionnelles ne sont, cependant, presque pas affectées par les changements de fréquence, de sorte que le coupleur peut être utilisé en vue d'obtenir une bonne terminaison de système coaxial de 52Ω, pour des fréquences allant jusqu'à 600MHz.

EE 5753 *pour plus amples renseignements*

UNE MONTRE À CRYSTAL

(Illustration à la page 50)

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey.

CET appareil (Modèle TSA) constitue un étalon de fréquences compact et portatif de grande précision. Il est exclusivement muni de transistors et il comprend une montre synchrone de 50Hz avec une aiguille de secondes.

La montre intérieure est un organe facultatif, et un moteur synchrone de 50Hz est fourni sur demande pour être employé à l'extérieur.

Les fréquences de sortie normales, à savoir 10KHz, 1KHz, 100Hz et 50Hz, s'obtiennent au moyen d'une prise de courant coaxiale fixée sur le tableau avant.

Le degré de précision est de moins d'une seconde par semaine de fonctionnement dans la température ambiante.

EE 5754 *pour plus amples renseignements*

CONTRÔLEUR D'ISOLATION NON-DESTRUCTIF

(Illustration à la page 51)

A/S Danbridge, Frederikssundsvej 60C., Copenhague, Danemark. Distributeurs: B. & K. Laboratories Ltd, 4 Tilney Street, London, W.1.

LE Contrôleur d'isolation Type JP1 comprend un générateur haute fréquence fournissant la tension de contrôle. Cette dernière peut être variée continûment au moyen d'un potentiomètre jusqu'à un maximum de 12V c.c.

La tension de sortie est indiquée par un indicateur à deux gammes, de façon à permettre des mesures rapides aux basses tensions.

La tension de contrôle est fournie par un cable coaxial blindé à une sonde isolante. La poignée de la sonde comprend un interrupteur de sécurité. Lorsque cet interrupteur est fermé, la haute tension est branchée par un relais, assurant ainsi le maximum de sécurité à l'opérateur.

L'instrument est muni d'un raccord afin de permettre le fonctionnement du relais à distance. La borne 'basse' de la tension de contrôle est directement à la terre afin que les objets à la terre puissent être examinés facilement. En outre, les bourdonnements sont éliminés dans une grande mesure.

En cas d'ionisation dans l'objet sous examen, le courant qui en résulte produit de petites impulsions dans la tension primaire d'un transformateur relié en série à la tension de contrôle.

La tension secondaire alimente un amplificateur à trois étages actionnant un haut-parleur électrostatique. Ainsi, même les décharges les plus faibles provoquent un bruit audible dans le haut-parleur.

Le gain de l'amplificateur est réglé par un potentiomètre en fonction du niveau de bruit local.

Tout courant de fuite dans l'objet contrôlé est vérifié à l'aide d'un microampèremètre sur le panneau avant. La résistance est calculée d'après les indications du voltmètre et du microampèremètre.

L'instrument est de construction robuste et destiné à l'usage permanent.

Tous les accessoires du circuit haute tension sont fixés dans une cuve remplie d'huile afin d'assurer la sûreté de leur fonctionnement dans toutes les circonstances.

L'instrument présente une parfaite sécurité d'emploi car le courant de sortie est limité à une valeur sûre par l'impédance de générateur élevée, l'interrupteur de sécurité réduisant d'autre part le risque de contact accidentel avec des points de haute tension.

EE 5755 *pour plus amples renseignements*

OSCILLATEUR DE DÉCADES DEUX FILS BASSE FRÉQUENCE

(Illustration à la page 51)

Muirhead & Co. Ltd, Beckenham, Kent.

LA gamme de fréquences de l'oscillateur de décades deux fils D-880-A est de 0,01Hz à 11,2KHz. Sur toute l'étendue de cette gamme, la puissance de sortie est constante à ±1dB près.

Le circuit de cet oscillateur, qui représente une conception nouvelle du problème de la fourniture de circuits sélectifs de très basse fréquence, est basé sur le développement du circuit faisant l'objet du Brevet britannique N° 781,374, accordé à la National Research Development Corporation; cet appareil est construit sous licence et comprend de nombreuses améliorations

Le principe de fonctionnement consiste à résoudre une équation différentielle de second ordre avec amortissement en employant deux étages d'intégrateurs et un changeur de signes. Plus simplement: le déphasage général de 360°, qui seul permet des oscillations, est obtenu au moyen de deux circuits intégrateurs électroniques ayant chacun un déphasage de 90°, et d'un circuit changeur de signes ayant un déphasage de 180°. Deux sorties, à 90° l'une de l'autre, sont prises des circuits intégrateurs et sont utilisables à un niveau de 10V. Les deux sorties sont directement accouplées et la teneur en courant continu est ajustée à une très faible valeur par des commandes pré-réglées. La sortie '0°' convient à l'alimentation d'une charge de 600Ω et elle peut être abaissée par un atténuateur de sortie jusqu'à −62dB Réf. 10V. La sortie '90°' convient à une charge de 10kΩ et elle peut être abaissée à zéro au moyen d'un potentiomètre.

EE 5756 *pour plus amples renseignements*

OSCILLATEUR À GAMME ÉTENDUE

(Illustration à la page 51)

Dawe Instruments Ltd, 99 Uxbridge Road, London, W.5.

L'OSCILLATEUR à gamme étendue DAWE, Type 441, est un appareil à résonance résistance-capacité comprenant toutes les fréquences de 5Hz à 6000KHz sur cinq gammes, avec choix de sortie équilibrée ou non-équilibrée. Le même appareil peut donc être utilisé pour obtenir la gamme complète allant des fréquences infra-audibles aux basses fréquences de radio. Cette gamme étendue, ajoutée au taux réduit de distorsion et à la tension constante caractérisant ce genre d'accord, offre des avantages considérables pour les recherches de vibrations, l'essai d'amplificateurs à large bande, de transducteurs, d'appareils médicaux et de transmissions par courant porteur, ainsi que pour la plupart des mesures courantes dans la gamme des fréquences acoustiques

Cet appareil comprend aussi un pont de Wien modifié faisant fonction de réseau stabilisateur de fréquence, dont la sortie alimente un amplificateur symétrique à charge cathodique pour fonctionnement équilibré ou non-équilibré, selon les besoins. La puissance de sortie est de 0,1W (à 7,75V) dans une charge de 600Ω, la tension à vide étant de 15,5V. L'impédance de source est de 600Ω ±20%.

EE 5757 *pour plus amples renseignements*

SPECTROMÈTRE BASSE FRÉQUENCE

(Illustration à la page 51)

Bruel & Kjær, Noerum, Danemark, Distributeurs: B. & K. Laboratories Ltd, 4 Tilney Street, London, W.1.

CET instrument a été développé essentiellement afin de faciliter l'analyse de fréquence rapide et précise de signaux complexes tels que ceux qui surviennent dans les mesures de bruits et de vibrations. Il peut, cependant, être employé également comme voltmètre à lampe pour la lecture de l'intensité efficace ainsi que pour l'analyse harmonique de signaux électriques.

C'est donc, fondamentalement, un analyseur de fréquence qui divise la bande de fréquences de 35Hz à 35KHz en gammes d'un tiers d'octave, au moyen de 30 filtres passe-bande d'une très grande sélectivité (élimination de 40dB au double de la fréquence nominale passe bande) et d'une atténuation de bande passante zéro.

Les contacts d'entrée des filtres sont parallèles et le choix de la bande fréquences s'effectue au moyen d'un interrupteur dans le circuit de sortie des filtres. L'interrupteur peut être actionné à la main ou automatiquement au moyen d'un moteur d'entraînement extérieur. Lorsqu'on emploie, à cet effet, la commande par moteur de l'Enregistreur Graphique de Niveau type 2304, l'opération de commutation peut être synchronisée avec la commande du papier d'enregistrement de l'Enregistreur de Niveau de façon à enregistrer automatiquement les spectrogrammes (diagrammes fréquence-amplitude) sur du papier d'enregistrement étalonné et pré-imprimé. En raison du pré-chargement des filtres, la durée de suroscillation est réduite au minimum et un spectrogramme complet peut être enregistré en moins de 20 secondes.

Le Spectromètre présente l'avantage important de pouvoir mesurer non seulement la valeur moyenne arithmétique du signal d'entrée, mais également la valeur efficace réelle (de signaux dont le facteur de crête peut atteindre 5) et la valeur de pointe.

L'instrument peut aussi faire fonction d'amplificateur linéaire et de voltmètre à lampe dans la bande de fréquences 2Hz à 35KHz. Deux amortisseurs différents sont fournis afin d'assurer, même aux basses fréquences, la précision des mesures et une lecture facile de l'indicateur.

Enfin, l'instrument comprend les quatre réseaux à atténuation prédéterminée pour les mesures de niveaux sonores, mondialement normalisés, qui peuvent être branchés à n'importe quel moment.

EE 5758 *pour plus amples renseignements*

TUBE À RAYONS CATHODIQUES MICRO-SPOT

(Illustration à la page 52)

Ferranti Ltd, Hillinwood, Manchester, Lancashire.

UN nouveau tube micro-spot à rayons cathodiques pouvant résoudre 5 000 lignes—le seul du genre au monde—vient d'être mis sur le marché par le Service d'Electronique de la Société Ferranti.

D'un diamètre de 127mm, le Tube CM5/71, ainsi qu'on le dénomme, a été conçu en vue d'applications aériennes, telles que la cartographie aérienne. Le diamètre du spot lui-même est sensiblement inférieur à 0,025mm.

Le pouvoir élevé de résolution du tube a été rendu possible par l'emploi d'un écran très fin, ainsi que d'un nouveau genre de canon à électrons comportant deux éléments de focalisation, dont l'un est électromagnétique et à l'extérieur du tube, tandis que l'autre est électrostatique et de longueur focale fixe.

Le tube a une face à surface optique plate en verre non assombrissant et une courte partie cylindrique revêtue—sauf sur la surface de l'écran—d'une épaisse couche de résine plastique, permettant de l'employer dans des conditions atmosphériques défavorables, à savoir 30kV d'intensité à 24 000m d'altitude, sans risque de panne de très haute tension. De même, le tube n'a pas besoin de support, les conducteurs à son extrémité arrière étant enfermés dans des capsules de manière à ce qu'il puisse être monté facilement sur son ensemble de focalisation et de balayage.

Quoique le nouveau tube soit surtout destiné à présenter ou à photographier des renseignements se répétant, il peut aussi photographier des images transitoires isolées à des vitesses d'enregistrement qui ne sont limitées que par les conditions requises par la bobine d'exploration.

Le phosphore courant utilisé dans le tube est à fluorescence verte, déclinant jusqu'au niveau de 1/e en l'espace de 10 microsecondes lorsque l'excitation est supprimée. Il se prête aussi à l'usage avec de nombreux types de matériel photographique.

EE 5759 *pour plus amples renseignements*

APPAREIL DE MESURE À PLUSIEURS GAMMES

(Illustration à la page 52)

Taylor Electrical Instruments Ltd. Montrose Avenue, Slough, Buckinghamshire.

LA caractéristique principale de cet appareil de mesure à plusieur gammes (Modèle 100A) consiste dans sa sensibilité de 100kΩ/V. Il comporte un mouvement électrodynamique à pôle central de 8μA, un disjoncteur à maxima fonctionnant sur toutes les gammes, ainsi qu'un mécanisme de polarité inverse.

La gamme de tensions continues s'étend de 0,5V à 2,5V et la gamme de tensions alternatives va de 10V à 2,5kV. Les gammes de courant vont de 10μA à 10A c.c., tandis que les gammes de résistance vont de 2kΩ à 200MΩ d.g.n. (déflection grandeur nature). Un adaptateur est fourni pour augmenter la gamme de tensions jusqu'à 25kV c.c.

EE 5760 *pour plus amples renseignements*

RADIOCOMPAS SUBMINIATURE

(Illustration à la page 52)

Elliott Brothers (London) Ltd. Elstree Way, Borehamwood, Hertfordshire.

LA Société Elliott Brothers fabrique sous licence de la Société Française Air-Equipement d'Asnières, un nouveau radiocompas léger ' subminiature ' pour avion. Cet appareil est destiné principalement aux applications où le poids, l'encombrement et le prix sont des facteurs primordiaux. Tout en étant donc fort léger, il est, cependant, hautement précis et sensible, et convenant particulièrement à l'équipement de tous les modèles d'avions civils et militaires, ainsi que des avions de commerce de grandeur moyenne et des hélicoptères.

L'ensemble de l'appareillage comprend trois éléments: le récepteur, le cadre et l'indicateur de relèvement, assurant au pilote une indication automatique continue du gisement relatif et la réception d'informations. Ces éléments sont utilisés pour les ondes entretenues modulées et les émissions radiotéléphoniques dans la gamme de fréquences 200 à 800KHz.

Le cadre qui peut être installé soit au dessus soit au dessous du fuselage, est enroulé sur un noyau de ferrite et fixé sur une monture d'isolement ne débordant que de 38mm du fuselage.

La commande du cadre et le système indicateur de transmissions sont logés dans cet élément qui mesure 113 × 89 × 127mm, l'assemblage entier étant hermétiquement fermé.

Le récepteur, logé dans la boîte de contrôle et mesurant 113×140×165mm, comporte les sous-éléments suivants:

(a) Elément haute fréquence, employant des lampes subminiature.

(b) Elément moyenne fréquence avec détecteur, oscillateur local et circuits anti-fading (l'amplificateur moyenne fréquence et les circuits anti-fading emploient des lampes subminiature, les circuits de l'oscillateur local et du détecteur étant transistorisés).

(c) Elément servo-amplificateur, entièrement transistorisé.

(d) Elément basse fréquence, entièrement transistorisé.

EE 5761 *pour plus amples renseignements*

Résumés des Principaux Articles

Un étalon d'hyperfréquences. par B. H. L. James et M. T. Stockford

Résumé de l'article aux pages 2 à 7

L'article qui suit décrit un étalon produisant des signaux positivement identifiés pour l'étalonnage d'hyperfréquences dans la gamme de 7 kMHz à 20 kMHz. Ces deux limites peuvent être étendues par l'usage d'un appareil auxiliaire,

Les plus petits intervalles qui s'obtiennent ne dépassent pas 0,03% de la fréquence d'étalonnage, les signaux étant "synthétisés" à partir d'un oscillateur piézoélectrique basse fréquence, à température contrôlée, ayant une stabilité à court terme de plus d'une partie sur 10⁹, qui peut être comparée directement avec une fréquence de la B.B.C. connue avec précision. Les divisions de fréquences jusqu'à 50 Hz donnent plusieurs basses fréquences précises et permettent l'emploi d'une montre afin de vérifier la stabilité à long terme de l'oscillateur.

Un générateur à transistors de formes d'ondes de télévision. par F. Rozner.

Résumé de l'article aux pages 8 à 12

Cet article décrit un générateur de formes d'ondes de télévision concu pour être employé avec le matériel portatif de télévision. Son encombrement, son poids et sa consommation électriques des plus réduits sont dûs à l'emploi de transistors et à l'élimination totale de tout tube électronique. Ce générateur de formes d'ondes fait usage de procédés numéraux et, quoique cela exige un grand nombre d'éléments actifs, le fonctionnement qui en résulte est, dans une grande mesure, indépendant des caractéristiques des transistors individuels, des variations de haute tension, etc. tandis que la stabilité et la précision du générateur sont parfaitement suffisantes pour les émissions de télévision.

Un phasemètre de basse-fréquence. par N. Hambley.

Résumé de l'article aux pages 13 à 15

L'appareil décrit dans cet article mesure avec rapidité et précision la différence de phase entre deux formes d'ondes. Les formes d'ondes actionnent des vannes qui permettent à des compteurs numéraux électroniques d'indiquer la différence de phase en degrés. Cet appareil de mesure exige un niveau d'entrée de 1 à 10 Volts r.m.c. (racine de la moyenne des carrés). Sa gamme de fréquences est de 1 à 100 Hz mais la limite maximum peut être portée, par une simple modification, à 3 kHz.

Un filtre passe-bas à variation continue et tension régulatrice. par A. J. Seyler et A. Korpel.

Résumé de l'article aux pages 16 à 22

Cet article décrit un filtre passe-bas à variation continue, conçu et construit selon la méthode des séries temporelles et la technique des lignes à retard. La caractéristique de coupure de ce filtre correspond à peu près à une pente de 6 décibels/octave et sa fréquence de perte de 3 décibels peut être variée en appliquant une tension de contrôle. Le modèle expérimental permet une variation de fréquence de coupure de 1:1000 (4,5 kHz à 5 MHz) pour une tension de contrôle variant de—85 Volts à 0 Volt.

Un indicateur visuel de réflectomètre à hyperfréquences de 7 500 à 11 000Mc/s. par J. C. Dix et M. Sherry.

Résumé de l'article aux pages 24 à 29

Cet article décrit un indicateur visuel de reflectometre pour bande de 5 200 à 11 000 MHz de réalisation récente. L'appareil en question donne une image continue de l'adaptation d'un accessoire de guide d'ondes pour bande de 5 200 à 11 000 MHz à une bande de fréquences de 7 500 à 11 000 MHz.

Le principe de base est que lorsqu'on change de fréquence, la variation du coefficient de réflexion d'une charge d'essai est indiquée par le changement de niveau de l'onde réfléchie, si l'amplitude de l'onde incidente demeure constante et indépendante de la fréquence.

Les auteurs examinent aussi les causes d'erreurs qui pourraient éventuellement se produire dans le système et ils donnent des détails de certaines expériences faites pour s'assurer de la précision du fonctionnement de l'instrument.

L'article ne traite que d'un réflectométre pour bande de 5 200 à 11 000 MH/z, mais il attire l'attention, en conclusion, sur le fait que puisqu'il existe un choix d'oscillateurs à onde de retour pouvant s'appliquer à toute la gamme d'hyperfréquences, soit de 1 600 MH/z à 11 500 MH/z, le système pourrait facilement être étendu à n'importe quelle autre bande de fréquences comprise dans ces limites.

Un analyseur à tête vibrante pour contrôler et répertorier les enregistrements magnétiques. par J. S. Gill

Résumé de l'article aux pages 30 à 31

Un dispositif ordinaire de réenregistrement ne peut être utilisé pour la lecture d'une bande magnétique fixe ou tournant très lentement car l'intensité acoustique est proportionnelle à la vitesse et elle tombe à zéro lorsque la bande est arrêtée. L'article ci-après décrit un dispositif de réenregistrement à tête vibrante permettant la lecture de bandes magnétiques fixes. Le principe a été appliqué avec succès pour séparer l'excitation du larynx de l'enregistrement sonore par l'interprétation visuelle directe de la forme d'onde.

Un calculateur de rapport électronique. par A. D. Boronkay.

Résumé de l'article aux pages 32 à 34

On emploie rarement des circuits de rapport entièrement électroniques dans les machines à calculer électroniques en raison de la difficulté de trouver l'analogie exacte d'un quotient. Le système que décrit cet article n'est pas basé sur une analogie physique rigoureuse mais plutôt sur une approximation valable dans certaines limites, dont le principe est qu'il existe un rapport quasi linéaire entre l'angle de phase de la somme des deux signaux et le rapport d'amplitude. Le dispositif de mesure du rapport se compose d'un circuit totalisateur qui somme le numérateur et le dénominateur, d'un circuit sensible aux différences de phase, fournissant une quantité électrique proportionnelle au vecteur résultant, et d'un redresseur commandé.

Un circuit de relais à phototransistor et température stabilisée. par J. C. Anderson et T. Winer.

Résumé de l'article aux pages 36 à 37

On peut employer un phototransistor de jonction pour commander directement un relais mais le fonctionnement qui en résulte varie sensiblement en fonction de la température ambiante. Cet article décrit un circuit employant un phototransistor et un transistor de jonction de type courant sous forme de dispositif conjugué "à longue queue". Grâce à ce système, toute variation du courant d'obscurité en fonction de la température ambiante, se trouve considérablement réduite, tout en assurant une haute sensibilité. La réponse transitoire est également bonne.

Contrôle de fréquence différentielle d'oscillateurs résistance-capacité. par D. L. A. Smith.

Résumé de l'article aux pages 38 à 39

Les oscillateurs résistance-capacité n'ont pas été pourvus jusqu'ici de stabilisateurs de fréquence différentielle à variation continue, étalonnés avec précision. Cet article montre qu'en ajoutant un circuit de déphasage supplémentaire au circuit de déphasage ordinaire, on peut faire varier différentiellement, sur une petite échelle, la fréquence produite. L'article démontre aussi que lorsqu'on veut un accroissement relativement important, cela entraîne une erreur dans l'accroissement qui est, en premier lieu, proportionnelle à la constante de temps du circuit de déphasage principal. Un moyen pratique de corriger systématiquement cette erreur est également indiqué.

La structure superficielle des filaments de diodes saturées. par F. A. Benson et M. S. Seaman

Résumé de l'article aux pages 40 à 41

La durée d'une diode à filament de tungstène fonctionnant sur alimentation filament c.c. n'est pas nécessairement égale à celle d'une diode alimentée en courant alternatif. Les filaments fonctionnant sur courant continu acquièrent une structure superficielle en gradins tandis que les filaments fonctionnant sur courant alternatif deviennent lisses, sauf près des supports. Un nombre de filaments de diode ont été examinés et l'article ci-dessous analyse brièvement les résultats obtenus.

Un compteur de décades inversible. par D. L. A. Barber.

Résumé de l'article aux pages 42 à 43

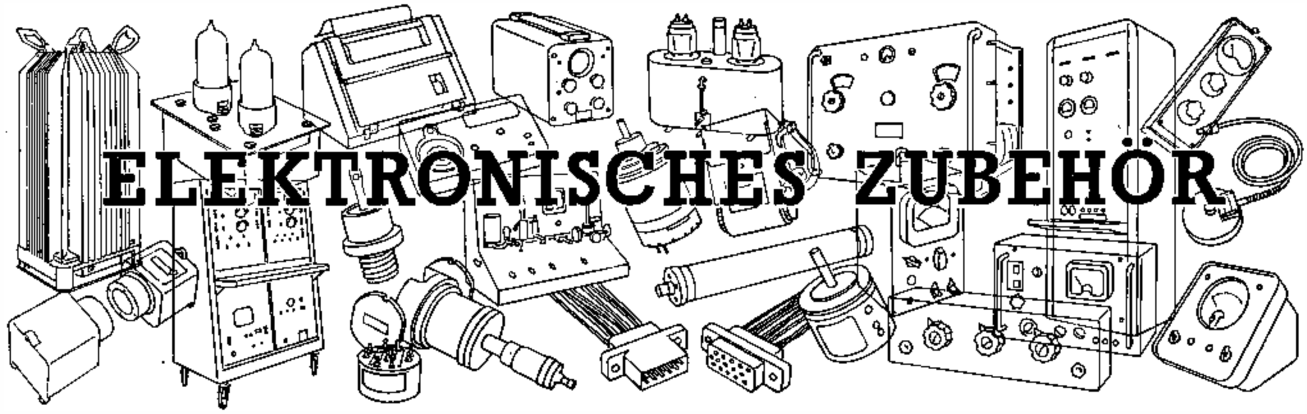
Cet article décrit un compteur de décades pouvant aussi bien additionner que soustraire. Ce compteur comprend un circuit à transistors joint à des lampes et à un "Dekatron". N'importe quel nombre de circuits de décades peut être relié en cascade.

Un cardiomètre à transistors. par L. D. Kitchen.

Résumé de l'article aux pages 44 à 45

Un simple appareil transistorisé pour la lecture directe continue du rythme du coeur durant les opérations chirurgicales.

D'un encombrement réduit, il a une sensibilité d'entrée de 250µV et il est relativement insensible aux interférences de 50 Hz. Il comprend un générateur d'impulsions d'étalonnage.



ELEKTRONISCHES ZUBEHÖR

Beschreibungen, zusammengestellt auf Grund von Mitteilungen aus Herstellerkreisen von neuen Bestandteilen, Zubehöreinrichtungen und Prüfgeräten. *(Übersetzung der Seiten 50 bis 52)*

MINIATUR-UNISELEKTOR

(Abbildungen auf Seite 50)

Siemens Edison Swann Ltd, 155 Charing Cross Road, London, W.C.2

DIESER Uniselektor-Kleinsteckdosenschalter wurde zur Verwendung in automatischen Systemen auf den Markt gebracht: Er hat 36 Anzapfstellen, misst 95 × 55 × 30mm und weight 341g. Der Uniselektor ist zur Verwendung in 22 oder 50V Schaltungen entworfen und ist mit einer Eigenfunklöschvorrichtung versehen. Dieses Aggregat nimmt für sich in Anspruch, zwei Millionen Arbeitsvorgänge zu ermöglichen, ohne dass eine Justierung vorgenommen werden muss, und eine Standzeit von mindestens acht Millionen Umdrehungen zu haben, bevor irgendein Bestandteil ausgewechselt werden muss.

Die drei Hauptelemente sind: Ein Antriebsmechanismus und eine Kontaktbürstenmontage; eine Kontaktträgerplatte, über der die Kontaktbürsten mit einer Geschwindigkeit von 60 Stufen/Sek. absuchen, und eine vorgeschaltete Trennklinke, die an der Fassung angebracht ist und das Ein- und Ausstecken des Uniselektormechanismus ermöglicht, ohne irgendeine Schaltung zu behindern.

Die Trägerplatte umfasst 36 Kontaktsegmente, die in 3 Stufen angeordnet sind. Die Trägerplatte ist so konstruiert, dass einer der Arme der Kontaktbürstenmontage immer ein besonders angebrachtes massives Kontaktsegment berührt, und durch diese Anordnung werden Zubringerfedern überflüssig gestaltet.

EE 5751 *für weitere Einzelheiten*

KRAFTZUFUHRAGGREGAT

E.I.R. Instruments Ltd, 329 Kilburn Lane, London, W.9

DIESES Gerät ist eine zweckdienliche kontinuierlich regelbare Kraftquelle für Wechsel- oder Gleichspannung und Wechsel- oder Gleichstrom.

Das Aggregat hat Netzantrieb und umfasst zwei getrennte Kraftquellen, die in eine gemeinsame Schaltbrett- und Chassismontage eingebaut sind; diese Kraftquellen lassen sich sowohl auf 'Spannungs'—als auch 'Strom'—Ausgangsleistung umschalten.

Wenn das Gerät auf 'Gleichstrom'—Ausgangsleistung umgeschaltet wird, setzen sich Ausgangsmetallgleichrichter und Siebkreise in Betrieb.

Dieselbe kontinuierlich regelbare Steuerung kommt sowohl bei den 'Spannungs'—als auch bei den 'Strom'—Ausgangsstellen zur Verwendung, und wird wie erforderlich von einer zur anderen umgeschaltet.

Für diesen Zweck werden keine regelbaren Transformatoren verwendet, sondern die Steuerung wird mit Hilfe von 'Grob'—und 'Fein'—Hochleistungs-Potentiometern vorgenommen, die über die Eingangsstelle mit dem Aggregat verbunden sind. Auf diese Weise lässt sich sowohl die Ausgangsspannung als auch der Ausgangsstrom praktisch von Null bis zum Höchstwert regeln.

Die Ausgangsleistungen sind:—Wechsel- oder Gleichstrom von 0—10A. Wechsel- oder Gleichspannung von 0—1kV bei 60mA.

EE 5752 *für weitere Einzelheiten*

RICHTKOPPLER

(Abbildungen auf Seite 50)

Sir W. G. Armstrong Whitworth Aircraft Ltd, Baginton Aerodrome, Nr. Coventry, Warwickshire

DIESER Koppler ist ein 'Schleifen'—Typ und ist für Messungen des Radiofrequenzstroms und stehenden Wellenverhältnis in Koaxialkabeln und Antennensystemen für kleine und mittlere Kraftwerte geeignet.

Der Richtkoppler kan in Serienanordnung an jeder beliebigen Stelle der Leitung eingeführt werden, in der der Kraftfluss zu messen ist. Bei Messungen von Kraftflüssen über 3W empfiehlt es sich jedoch, den Einführungspunkt soweit als möglich vom Generator entfernt zu wählen.

Obwohl dies eine Breitbandvorrichtung ist, richten sich die Meterablesungen nach der Übergangsfrequenz, da die mit jedem Gerät gelieferte Eichkurve nur bei der auf der Eichtabelle angegebenen Frequenz als genau angesehen werden darf. Die Richteigenschaften beiben jedoch grösstenteils unbeeinflusst von Frequenzänderungen, so dass der Koppler als ein Hilfsgerät zum Erreichen von Höchstendstufen eines 52Ω Koaxialsystems für Frequenzen bis zu 600MHz verwendet werden kann.

EE 5753 *für weitere Einzelheiten*

KRISTALLUHR

(Abbildungen auf Seite 50)

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey

DIESES Gerät (Typ TSA) wurde als eine kompakte und tragbare Frequenzquelle hoher Genauigkeit konstruiert. In diesem Gerät kommen durchwegs Transistoren zur Verwendung, und eine synchronisierte 50Hz-Uhr mit Sekundenzeiger ist eingebaut.

Auf Wunsch kann an Stelle der Innenuhr ein 12V 50Hz-Synchronmotor geliefert werden.

Schaltbare Standardfrequenzen von 10kHz, 100Hz und 50Hz stehen über eine am Vorderschaltbrett angebrachte Koaxialsteckdose zur Verfügung.

Die Genauigkeit hält sich innerhalb 1 Sekunde pro Woche bei Betrieb in Raumtemperatur.

EE 5754 *für weitere Einzelheiten*

UNSCHÄDLICHES ISOLIERUNGSPRÜFGERÄT

(Abbildungen auf Seite 51)

A/S Danbridge, Frederikssundsvej 60C, Kopenhagen, Dänemark

DER Isolierungsprüfer Typ JP1 umfasst einen HF-Generator für die Beschaffung der Prüfspannung, Dieselbe lässt sich mittels eines Potentiometers kontinuierlich bis zu einem Höchstwert von 12kV Gleichspannung regeln.

Die Ausgangsspannung wird an einem Meter mit 2 Bereichen angezeigt, um genaue Messungen bei niedereren Spannungen zu ermöglichen.

Die Prüfspannung wird durch ein abgeschirmtes Koaxialkabel an eine isolierte Prüfsonde eingespeist. Ein Sicherheitsschalter ist in den Griff der Prüfsonde eingebaut. Beim Schliessen dieses Schalters wird die Hochspannung durch ein Relais eingeschaltet, wodurch ein Höchstmass an Betriebssicherheit für den Bedienungsmann gegeben ist.

Am Gerät befindet sich ausserdem eine Steckdose, so dass, falls erforderlich, das Relais ferngesteuert werden kann. Die 'nieder' Klemme der Prüfspannung ist direkt geerdet, so dass geerdete Gegenstände leicht geprüft werden können, und ausserdem Brummstörungen grösstenteils ausgeschaltet sind.

Falls im Prüfgegenstand Ionisation auftritt, entwickelt der sich ergebende Strom kleine Spannungsimpulse in der Primärschaltung eines in Serienanordnung mit der Prüfspannung verbundenen Transformators.

Die Sekundärspannung wird in einen Dreistufenverstärker geleitet, der den Betrieb eines elektrostatischen Lautsprechers bewirkt. Auf diese Weise ergeben selbst winzigste Entspannungen ein hörbares Rauschen im Lautsprecher.

Der Verstärkungsgrad kann mittels eines Potentiometers dem örtlichen Störpegel angepasst werden. Etwaige Kriechströme im Prüfgegenstand können mittels eines Mikroammeters am Vorderschaltbrett kontrolliert werden. Falls erforderlich kann der Widerstand aus den Spannungsmesser- und Mikrostrommessungen errechnet werden.

EE 5755 *für weitere Einzelheiten*

ZWEIPHASEN-NIEDERFREQUENZ-DEKADENSOSZILLATOR

(Abbildungen auf Seite 51)

Muirhead & Co. Ltd, Beckenham, Kent

DER Frequenzbereich des D-880-A Zweiphasen-Dekadenoszillators liegt zwischen 0,01Hz und 11,2kHz. Über diesen gesamten Bereich ist die Ausgangsleistung innerhalb ±1dB konstant.

Die Schaltung dieses Oszillators, mit dem ein neuer Weg zur Lösung des Problems der Versorgung von Selektivschaltungen mit sehr niederen Frequenzen beschritten wird, gründet sich auf eine Schaltung, die unter das 'National Research Development Corporation' (Nationalen Körperschaft für Forschung und Entwicklung) verliehene britische Patent Nr. 781374 fällt; das Gerät wird im Rahmen eines Konzessionsabkommens hergestellt und schliesst mehrere Verbesserungen in sich ein.

Das Grundprinzip der Arbeitsweise besteht in der Lösung einer Differentialgleichung zweiter Ordnung mittels Dämpfung unter Verwendung zweier Integratorenstufen und eines Vorzeichenwandlers. Einfacher ausgedrückt, die für das Auftreten der Oszillation erforderliche Gesamtphasenverschiebung von 360° wird durch zwei elektronische integrierende Schaltungen—mit einer Phasenverschiebung von je 90°—und einer Vorzeichenumwandlungsschaltung mit einer Phasenverschiebung von 180° erreicht. Zwei Ausgangsleistungen—90° von einander entfernt—werden den Integratorenschaltungen entnommen und stehen bei einem Spannungspegel von 10V zur Verfügung. Beide Ausgangsleistungen sind direkt gekoppelt und der Gleichstromgehalt wird mittels vorgewählter Steuerungsvorrichtungen auf einen sehr niederen Wert eingestellt. Die '0°' Ausgangsleistung ist zur Einspeisung in eine Ladung von 600Ω geeignet, und kann mittels eines Ausgangsdämpfers bis auf—62dB ref 10V eingestellt werden; die 90° Ausgangsleistung ist für eine Ladung von 10kΩ geeignet und mittels eines Potentiometers bis auf Null einstellbar.

Abgesehen von seiner Verwendung als Formteil des 'Muirhead' Transfer-Funktionanalysators findet der Oszillator viele Anwendungen auf dem servotechnischen Gebiet, wo das genaue Untersuchen von Niederfrequenzen von steigender Wichtigkeit ist.

Weitere Merkmale des Instruments sind seine Frequenzgenauigkeit, sein äusserst niedriger Oberwellengehalt, und die Gleichheit, oder Proportionalität, der beiden Ausgangspegel.

EE 5756 *für weitere Einzelheiten*

GROSSBEREICH-OSZILLATOR

(Abbildungen auf Seite 51)

Dawe Instruments Ltd, 99 Uxbridge Road, London, W.5

DER 'Dawe' Grossbereich—Oszillator, Typ 441, ist eine RC-Tuner—Konstruktion, die alle Frequenzen von 5Hz bis 600kHz in fünf Bereichen mit sowohl abgeglichener als auch unabgeglichener Ausgangsleistung deckt. Dasselbe Gerät kann auf diese Weise dazu verwendet werden, eine vollständige Erfassung von Subaudio—bis zu niederen Radiofrequenzen zu erreichen. Der weite Bereich zusammen mit dieser Art von Abstimmung erzielbaren niederen Verzerrung und konstanter Ausgangsleistung bietet beträchtliche Vorteile für Schwingungsuntersuchungen, sowohl für das Prüfen von Breitbandverstärkern, Umsetzervorrichtungen, medizinischen Geräten und Trägernachrichtensystemen, als auch für die Mehrzahl der üblichen Messungen innerhalb des Tonbereichs.

Dieses Aggregat umfasst ausserdem eine abgewandelte Wien-Robinson Brücke, die als Frequenzregelnetz wirkt, und deren Ausgangsleistung einen abgeglichenen Gegentaktverstärker mit Kathodenmitlaufs-Ausgangsleistung für entweder abgeglichenen oder unabgeglichenen Betrieb speist. Die Ausgangsleistung beträgt 1,0W in 600Ω wobei die Leerlaufspannung 15,5V beträgt.

EE 5757 *für weitere Einzelheiten*

TONFREQUENZ-SPEKTROMETER

(Abbildungen auf Seite 51)

Bruel & Kjær, Noerum, Dänemark. Vertrieb durch:—B & K Laboratories Ltd, 4 Tilney Street, London W.1.

DIESES Gerät wurde im Grunde dafür entwickelt, die rasche und genaue Frequenzanalyse von zusammengesetzten Signalen, wie sie z.B. bei Rausch- und Schwingungsmessungen auftreten, zu erleichtern. Es kann jedoch auch als Röhrenspannungsmesser mit getreuen Effektivstromablesungen und für die Oberwellenanalyse von elektrischen Signalen verwendet werden.

Es ist im Grunde genommen ein Frequenzanalysengerät, das das Frequenzband von 35Hz bis 35kHz in 1/3 Oktavenbereiche unterteilt, und zwar mittels 30 Bandpassfiltern mit sehr hoher Selektivität (40dB Unterdrückung bei zweimaliger Passbandmittenfrequenz) und Nullpassbanddämpfung.

Die Eingangsstufen zu allen Filtern sind parallelgesschaltet, und die Frequenzbandwahl wird mittels eines Schalters in der Filterausgangsschaltung vorgenommen. Der Schalter kann entweder handbetätigt oder automatisch mittels eines Aussenmotorantriebs betätigt werden. Wenn der Motorantrieb des Typs 2304 des schreibenden Pegelanzeigers für diesen Zweck verwendet wird, so kann der Schaltvorgang mit dem Papiervorschub des Pegelanzeigers in einer solchen Weise synchronisiert werden, dass Spektrogramme (Frequenzamplitudendiagramme) automatisch auf vorgedrucktem, geeichten Papier registriert werden. Dank der Vorbelastung der Filter werden Verdopplungszeiten auf ein Mindestmass beschränkt, und ein vollständiges Frequenzamplitudendiagramm kann in weniger als 20 Sekunden registriert werden.

Eine wichtige Eigenschaft des Spektrometers ist seine Fähigkeit, nicht nur den arithmetischen Mittelwert des Eingangssignals, sondern ausserdem sowohl den wahren Effektivwert (für Signale mit Spitzenfaktoren bis zu 5) als auch die Scheitelwerte zu messen. Die gewünschte Anzeigenart wird mittels eines Schalters gewählt.

Das Gerät lässt sich ausserdem zum Betrieb als Linearverstärker und Röhrenspannungsmesser im Frequenzbereich von 2Hz bis 35kHz umschalten. Zwei verschiedene Meterdämpfungskennlinien sind vorgesehen, so dass genaue Messungen und leichte Meterablesungen auch bei den Niederfrequenzen erzielt werden.

EE 5758 *für weitere Einzelheiten*

MIKROLEUCHTFLECK-KATHODENSTRAHLRÖHRE

(Abbildungen auf Seite 52)

Ferranti Ltd, Hollinwood, Manchester, Lancashire

EINE neue Mikroleuchtfleck-Kathodenstrahlröhre, die für sich in Anspruch nehmen kann, die erste ihrer Art auf der Welt zu sein, und die ein Auflösungsvermögen von 5 000 Zeilen besitzt, kann jetzt von der elektronischen Abteilung der Firma Ferranti bezogen werden.

Die unter der Nummer 5/57 CM bekannte Röhre, mit einer Querabmessung von 127mm, wurde für Luftfahrtsverwendungszwecke, wie z.B. Landkartenvermessung vom Flugzeug aus, entworfen. Der Leuchtfleckdurchmesser liegt beträchtlich unter 0,025mm.

Das hohe Auflösungsvermögen der Röhre wurde ermöglicht durch Verwendung eines ausserordentlich feinen Schirms und einer neuartigen Konstruktion einer Elektronenkanone unter Verwendung von zwei Fokussierungselementen. von denen das eine elektromagnetisch an der Aussenseite der Röhre wirkt, während das andere elektrostatisch und mit fester Brennweite ausgestattet ist.

Die Röhre hat eine optisch flache Stirnfläche aus nichtdunkelndem Glas und eine kurze zylindrische Birne, und ist—ausser an der Schirmoberfläche—mit einer starken Kunstharzschicht überzogen, was ihren Betrieb unter ungünstigen atmosphärischen Bedingungen ermöglicht, d.h. 30kV bei 24 000m ohne Gefahr von Höchstspannungsausfall. Gleicherweise wird keine Röhrenfassung benötigt, da die Leitungen zum rückwärtigen Ende der Röhre so eingeschlossen sind, dass dieselbe leicht und unbehindert in ihre Fokus- und Abtastmontage eingesetzt werden kann.

EE 5759 *für weitere Einzelheiten*

ALLBEREICH-MESSER

(Abbildungen auf Seite 52)

Taylor Electrical Instruments Ltd, Montrose Avenue, Slough, Buckinghamshire

DAS Hauptmerkmal dieses Allbereich-Messers (Typ 100A) besteht in seiner Empfindlichkeit von 100kΩ/V. Das Gerät besitzt eine 8μA Mittenpol-Drehspulbewegung, und hat eine mechanische Überlastungsschutzvorrichtung, die sich in allen Bereichen betätigt, und ausserdem eine Polumkehrungsvorrichtung.

Der Gleichspannungsbereich liegt zwischen 0,5V und 2,5V, und der Wechselspannungsbereich zwischen 10V und 2,5kV. Die Strombereiche liegen zwischen 10μA und 10A für Gleichstrom, und die Widerstandsbereiche zwischen 2kΩ und 200MΩ Vollausschlag. Ein Vorsatzgerät steht zur Erhöhung des Spannungsbereichs bis zu 25kV Gleichspannung zur Verfügung.

EE 5760 *für weitere Einzelheiten*

SUBMINIATUR-RADIOKOMPASS

(Abbildungen auf Seite 52)

Elliott Brothers (London) Ltd, Elstree Way, Borehamwood, Hertfordshire

EIN neuer Subminiaturleichtgewichts-Flugzeugradiokompass wird von der Firma Elliott unter Konzession seiner Entwerfer, der Firma Air-equipment, in Asnieres, Frankreich, hergestellt, und ist in erster Linie für solche Anwendungszwecke gedacht, wo Gewicht, Grösse und Preis ausschlaggebende Faktoren sind. Nichtsdestoweniger ist er höchst genau, empfindlich und besonders gut zum Einbau in alle leichten Militär- und Verkehrsflugzeugtypen, und Verkehrsflugzeuge und Hubschrauber mittlerer Grösse geeignet.

Die gesamte Anlage besteht aus drei Aggregaten — Empfänger, Rahmenantenne und Peilungsanzeiger—die, wenn in das Flugzeug eingebaut, dem Piloten eine automatische Daueranzeige der Fahrtpeilung und des Nachrichtenempfangs vermitteln. Diese Einrichtungen können für modulierte ungedämpfte Wellen, und Radiotelefonbetrieb innerhalb des Frequenzbereichs von 200 bis 800kHz verwendet werden.

Die Rahmenantenne, die über oder unter dem Rumpf angebracht werden kann, ist auf einen Ferritkern gewunden und in einer isolierten Fassung befestigt, die nur um 38mm aus dem Rumpf hervorragt.

Das Rahmenantriebs- und Anzeigeübertragungssystem sind in diesem Aggregat untergebracht, dessen Abmessungen 113 × 89 × 127mm betragen; die gesamte Montage ist hermetisch abgeschlossen und vollkommen eingekapselt.

Der in dem eigentlichen Steuerschrank untergebrachte Empfänger misst 113 × 140 × 165mm und besteht aus den folgenden Nebenaggregaten

(a) RF-Aggregat, unter Verwendung von Subminiaturröhren.
(b) ZF-Aggregat mit Detektor, Ortoszillator und automatischer Lautstärkeregelschaltungen (ZF-Verstärker und Lautstärkeregler verwenden Subminiaturröhren, die Ortsoszillatoren und Detektorschaltungen sind transistorisiert).
(c) Servo-Verstärkeraggregat—vollständig transistorisiert.
(d) NF-Aggregat — vollständig transistorisiert.

EE 5761 *für weitere Einzelheiten*

Zusammenfassungen der Hauptartikel

Ein Mikrowellen-Frequenznormgerät. von B. H. L. James und M. T. Stockford

Zusammenfassung des Artikels auf Seite 2-7

Dieser Artikel beschreibt ein Gerät, das positiv ausgewertete Signale für Eichungszwecke in einem Frequenzbereich von 7 kMHz bis 20 kMHz erzeugt. Beide Grenzwerte können mittels Verwendung der Zusatzvorrichtung erweitert werden.

Die verfügbaren Kleinstpausen sind nicht grösser als 0.3% der Eichfrequenz, wobei die Signale aus einem temperaturgeregelten NF-Kristalloszillator mit einer kurzzeitigen Stabilität, besser als ein Teil in 10^7, synthetisiert werden, was einen direkten Vergleich mit einer genau bekannten BBC-Frequenz gestattet. Die Frequenzunterteilung bis zu einem Mindestskalenwert von 50 Hz ergibt mehrere genaue Niederfrequenzen und ermöglicht den Betrieb einer Uhr für die Kontrolle der langzeitigen Stabilität des Oszillators.

Ein Fernsehwellenformgenerator unter Verwendung von Transistoren. von F. Rozner

Zusammenfassung des Artikels auf Seite 8-12

Dieser Artikel beschreibt einen Fernsehwellenformgenerator zur Verwendung mit tragbaren Fernsehsendegeräten. Dank der Verwendung von Transistoren und der vollkommenen Ausschaltung von thermionischen Vorrichtungen wurden sehr kleine Abmessungen, geringes Gewicht und niederer Kraftverbrauch erzielt. Der Wellenformgenerator verwendet Digitalarbeitsvorgänge, und, obwohl dies eine grosse Anzahl von aktiven Elementen erforderlich macht, ist die sich ergebende Leistung zu einem grossen Ausmass unabhängig von den Charakteristiken einzelner Transistoren, Hochspannungsschwankungen u.s.w., und die Stabilität und Genauigkeit sind ausreichend für Sendezwecke.

Ein Niederfrequenz-Phasenmesser. von N. Hambley

Zusammenfassung des Artikels auf Seite 13-15

Das in diesem Artikel beschriebene Aggregat ermöglicht eine schnelle und genaue Messung des Phasenunterschieds zwischen zwei Wellenformen. Die Wellenformen betätigen Schaltungen, die es elektronischen Zählwerken gestatten, den Phasenunterschied in Graden anzuzeigen. Das Phasenmessgerät benötigt einen effektiven Eingangsspannungsstand zwischen 1 und 10 V. Der Frequenzbereich liegt zwischen 1 und 100 Hz, aber der obere Grenzwert kann mit Hilfe einer einfachen Modifikation auf 3 kHz erhöht werden.

Ein spannungsgesteuertes kontinuierlich regelbares Tiefpassfilter. von A. J. Seyler und A. Korpel

Zusammenfassung des Artikels auf Seite 16-22

Hier wird ein kontinuierlich regelbares Tiefpassfilter beschrieben, das unter Verwendung der Zeitserienmethode und Verzögerungsleitungsverfahren entworfen und konstruiert wurde. Die Grenzkennlinie dieses Filters kommt einer 6dB/Oktavensteilheit nahe und seine 3 dB Schwundfrequenz wird durch Einsatz einer Steuerspannung geregelt. Das neu entwickelte Modell gestattet eine Grenzfrequenzregelung von 1:1 000 (4,5 kHz bis 5 MHz) für die Steuerspannung, die sich zwischen −85 und 0 V bewegt.

Ein Mikrowellenreflektometer-Bildsystem für 7 500 bis 11 000 MHz. von J. C. Dix, B.Sc. (Eng) und M. Sherry, M.Sc.

Zusammenfassung des Artikels auf Seite 24-29

Ein X-Band Reflektometer-Bildsystem, das kürzlich entwickelt wurde, wird hier beschrieben. Dieses Gerät ergibt ein kontinuierliches Bild der Anpassung einer X-Band Wellenleitervorrichtung über einem Frequenzband von 7 500 bis 11 000 MHz.

Das Grundprinzip besteht darin, dass wenn die Frequenz geändert wird, die Schwankung des Moduls des Reflektionsfaktors einer Prüfladung durch die Änderung des reflektierten Wellenpegels angezeigt wird, falls die Amplitude der Einfallswelle konstant und unabhängig von der Frequenz gehalten wird.

Der Besprechung der in dem System zu erwartenden Fehlerquellen ist ein bestimmter Platz eingeräumt, zusammen mit Einzelheiten von zur Gewährleistung der Betriebssicherheit angestellten Prüfversuchen.

Der Artikel befasst sich lediglich mit einem X-Band Reflektometer, aber es wird jedoch abschliessend darauf hingewiesen, dass, da ja eine Reihe von Rückwärtswellenoszillatoren für den vollen Mikrowellen-Frequenzbereich von 1 600 MHz bis 11 500 MHz zur Verfügung steht, das System mit Leichtigkeit auf jedes andere Frequenzband innerhalb dieser Grenzwerte ausgedehnt werden kann.

Ein Vibrations-Tonkopf zum Untersuchen und Registrieren von magnetischen Aufnahmen. von J. S. Gill

Zusammenfassung des Artikels auf Seite 30-31

Wenn es sich als notwendig erweist, ein feststehendes oder sich sehr langsam bewegendes Magnetband abzulesen, lässt sich eine übliche Wiedergabevorrichtung nicht verwenden, da die Ausgangsleistung von der Geschwindigkeit abhängt, und bei feststehendem Band auf Null abfällt. In diesem Artikel wird ein Vibrations-Tonkopf beschrieben, der zur Wiedergabe von feststehenden Magnetaufnahmen verwendet werden kann. Der Grundgedanke wurde mit Erfolg zum Ermitteln der Kehlkopferregung aus Sprechaufnahmen durch direkte Sichtauswertung der Wellenform angewandt.

Ein elektronischer Verhältnisermittler. von A. D. Boronkay

Zusammenfassung des Artikels auf Seite 32-34

In elektronischen Rechengeräten werden allelektronische Schaltungen selten verwendet, was auf die Schwierigkeit zurückzuführen ist, einen genauen Analog zu einem Quotienten zu finden. Das in diesem Artikel beschriebene System beruht nicht auf einem genauen physikalischen Analog, sondern auf einer innerhalb bestimmter Grenzwerte gültigen Annäherung, wobei das ausschlaggebende Prinzip darin besteht, dass eine annähernde lineare Beziehung zwischen dem Phasenverschiebungswinkel der Summe der beiden Signale und des Amplitudenverhältnis besteht. Das Verhältnismessgerät besteht aus einer Addierschaltung, die die Endsummen des Zählers und Nenners ermittelt, einer phasenempfindlichen Schaltung zur Beschaffung eines elektrischen Grössenwertes proportional zu dem Summenvektor und einem phasenempfindlich geregelten Gleichrichter.

Eine temperaturstabilisierte Phototransistoren-Relaisschaltung. von J. C. Anderson und T. Winer

Zusammenfassung des Artikels auf Seite 36-37

Ein Abzweigtransistor kann für den direkten Betrieb eines Relais verwendet werden, aber die erzielte Leistung schwankt stark mit der Zimmertemperatur. In diesem Artikel wird eine Schaltung beschrieben, in der ein Phototransistor und ein gewöhnlicher Abzweigtransistor in einer 'Langdraht-Paarableitung' zur Verwendung kommen. Auf diese Weise wird die durch die Veränderung in der Zimmertemperatur bedingte Dunkelstromschwankung stark vermindert, während zu gleicher Zeit hohe Empfindlichkeit erzielt wird. Die Sprungcharakteristik erweist sich ebenfalls als gut.

Inkrementale Frequenzsteuerung von RC-Oszillatoren. von D. L. A. Smith

Zusammenfassung des Artikels auf Seite 38-39

Die Beschaffung einer kontinuierlich regelbaren, inkrementalen Frequenzsteuerung mit genauer Eichung war bisber mit einem Oszillator der RC Type nicht möglich. In diesem Artikel wird gezeigt, dass mittels Beiordnung einer Phasenschieberhilfsschaltung zu der üblichen Phasenschieberschaltung die erzeugte Frequenz inkremental über einen kleinen Bereich geregelt werden kann. Es wird weiterhin gezeigt, dass in Fällen, wo ein verhältnismässig grosses Inkrement erforderlich ist, ein Fehler in dem Inkrement besteht, der in erster Ordnung sich proportional zur Zeitkonstante der Hauptphasenschieberschalting verhält. Eine praktische Methode zur systematischen Korrektur dieses Fehlers ist ebenfalls miteingeschlossen.

Oberflächenstruktur von gesättigten Diodenheizfäden. von F. A. Benson und M. S. Seaman

Zusammenfassung des Artikels auf Seite 40-41

Die Lebensdauer einer Wolframheizfadendiode beim Einsatz mit einer Gleichstromheizfadenquelle ist nicht unbedingt dieselbe wie bei Wechselstrombetrieb. Mit Gleichstrom betriebene Heizfäden entwickeln eine stufenförmige Oberflächenstruktur, während wechselstromerhitzte Heizfäden glatt werden, mit Ausnahme in der Nähe der Befestigungspunkte. Eine Anzahl von Diodenheizfäden wurden untersucht, und die Ergebnisse sind in diesem Artikel kurz besprochen.

Ein reversibler "Dekatron" Dekadenzähler. von D. L. A. Barber

Zusammenfassung des Artikels auf Seite 42-43

Ein Dekadenzähler für sowohl Additions—als auch Subtraktionsvorgänge wird hier beschrieben. Der Zähler umfasst eine Transistorenschaltung in Verbindung mit Röhren und einem Dekatron; jede gewünschte Anzahl von Dekadenschaltungen lässt sich in Kaskadenanordnung anschliessen.

Ein Herzpulsmesser mit eingeordneten Transistoren. von L. D. Kitchen

Zusammenfassung des Artikels auf Seite 44-45

Ein einfaches mit Transistoren bestücktes Gerät zum kontinuierlichen direkten Ablesen des Wärmeaustausches während operativer Eingriffe wird hier beschrieben.

Es ist in kleinen Abmessungen gehalten, und hat eine Eingangsempfindlichkeit von 250 μV, ist jedoch verhältnismässig unempfindlich einer Interferenz von 50 Hz gegenüber. Ein Eichimpulsgenerator ist miteingebaut.