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Commentary

THE figures, so far as they are available, show that 1957 has been another year of considerable prosperity for this country's manufacturing industries, and the electronic industry has joined in this prosperity to the full. This is shown in the production and, particularly, export figures, which in most spheres of the industry are at record levels. For the first ten months of the year the total exports of all radio and electronic equipment were valued at £36·2M against £32·8M for the same period of the previous year, so that for the complete year this should give a total of some £43M against the record of £40M achieved in 1956. The growth of interest in British electronic equipment has also been evidenced by the large attendances at the various exhibitions held throughout the year, and here one may remark on the success of the first Instruments, Electronics and Automation Exhibition which provided the industry with an excellent opportunity to show some of the more serious sides of its activities.

From the scientific aspect it has, perhaps, been a year of consolidation and solid progress rather than of spectacular discovery or invention; though it must always be borne in mind that the full import of a discovery or invention is seldom realized until some considerable time after its inception.

In the manufacturing industries the use of electronic devices has continued to grow rapidly and in many industries the use of various forms of electronic equipment is completely accepted as a means for measuring and controlling both the quality and quantity of production. Progress has also been made in the electronic control of machine tools and during the year several new forms of control element have made their appearance. In the past the electronic control has usually been added to existing designs of machine tool but milling machines, etc. designed specifically for electronic control are now in production.

Although considerable research on new switching and memory elements is still proceeding the basic design of the digital and analogue computer is fairly well established and the main emphasis is now towards the useful and profitable application of these machines and their integration into complete systems. In the commercial world, payroll accounting is still the most profitable application for the digital machine and a number of the larger firms are now making use of them. As for example, the Ford Motor Company who are using LEO 1 for computing the payroll for their 19 000 employees. The G.P.O. has also ordered two National Elliot '405' computers to handle the payroll accounting for some 112 000 employees while similar installations are being made by several town and county councils. The scientific use of both analogue and digital machines has also continued to expand and installations have been made for such diverse applications as weather forecasting, biophysical research and wind tunnel data processing. The weakest part of a computer installation is, in general, the input and output which is usually slow compared with the speed of the machine itself. To overcome this and thus increase the range of application of the computer considerable work has been done; notable examples being the Solartron 'Electronic Reading Automaton' which will 'read' printed or typewritten characters at a speed of 120 characters/second, and the Powers-Samas 'Samstronic' tabulator which is capable of printing characters at a speed of 700 characters/second.

The entertainment side of the industry has also had a most successful year. Compared with 1956 the total sales of radio receivers has risen by about 20 per cent radio gramophone and audio equipment by 26 per cent and television receivers by 2 per cent. The sales of radio receivers has, no doubt, been stimulated by the introduction of f.m. broadcasting, while the small increase in the sales of television receivers is probably due to the fact that this is rapidly becoming a replacement only market. The ever-increasing popularity of recorded music is, however, rather more difficult to account for but is probably due, in part, to the very high standard of recording and reproduction that is now offered. One of the most notable events during the year has of course been the 21st birthday of the BBC television service; a service which now caters for 98 per cent of the country's population. With the virtual completion of their Band I programme the BBC is now actively pursuing the possibilities of colour with test transmissions in Band I and are also carrying out a series of tests in Band 5. On the manufacturing side the year has seen considerable progress in the introduction of transistors and printed circuits into domestic equipment; the latter may make for cheaper and more foolproof production but it may also present a somewhat formidable problem to the service engineer.

In the purely scientific sphere radio astronomy has undoubtedly taken the year's honours with the opening of the large steerable radio telescope at Jodrell Bank and the Mullard Observatory at Cambridge. Both of these events have been closely linked to the launching of the Russian earth satellites which has, without doubt, been the most outstanding scientific achievement of the year. Unfortunately little is yet known of the electronic equipment carried by these satellites but, from the evidence available, it would appear that the Russians are not so advanced as ourselves in this branch of science. But then, until the launchings took place, nobody, and most of all the Americans, would have conceded that the Russians had such a clear lead over the rest of the world in rocket propulsion.

From the foregoing brief résumé of the past year it may fairly be concluded that it has been a successful and prosperous one for our industry but that however great ones progress may be it does not pay to be complacent about ones achievements.

Special Timing Techniques Employed on Guided Missile Ranges

By R. J. Garvey*, B.Sc.(Eng.)

A central equipment consisting essentially of a crystal oscillator and a number of frequency dividing circuits provides timing and reference signals to the various measuring stations dispersed over a guided missile test range. The signals serve as time markers for the various recordings made and are also concerned with the control and firing of the missile.

A general description of the timer is given including the method of transmitting the timing pulses to the remote stations. One of the timer signals is coded so that elapsed time, with respect to an arbitrary zero, can be continuously marked on long runs of recording film.

Two equipments associated with the timer are concerned with the automatic control of the ground instrumentation during the launching and flight of the missile and with the control of a number of cine-theodolites. These will be described in later articles.

IN flight testing guided missiles, performance data is recorded at a number of stations dispersed over the range while other stations are concerned with the control of the missile and associated equipment both before and during the flight. In general terms the instrumentation consists of:

Air to ground telemetry.
Radar trackers.
R.F. doppler system.
Cine-theodolites.
Cine-cameras.
Guidance and control system.
Firing circuits.

The measurements and control must be made with reference to an accurate and common time scale; a central timer accordingly supplies timing pulses to each station as shown by Fig. 1. This timer has two functions; it provides time makers for the various recording instruments and switching pulses which operate both recording and control instruments at specific times during a shoot.

Time markers are fed direct to the various recording stations, but switching pulses are supplied via a 'programme switch' as shown on Fig. 1. This unit employs uniselector switches which are stepped by a 1p/s signal from the timer; these switch the various range instruments on and off automatically during a shoot, according to a pre-arranged programme set up for the particular missile.

A clock in the range control room is driven by the timer; this counts off the seconds during a shoot, the times being announced at suitable intervals over the range communication loudspeakers by the control officer.

Cine-theodolites are used on the range to determine the trajectory and velocity of the missile. The cine mechanisms are triggered by timing pulses which are fed via a cine-theodolite controller. This controller phases the pulses from the timer so that the shutters of all the cine-theodolites operate in synchronism and at a regular rate.

Each flight test involves the subsequent analysis of many long runs of recording film. Normal time markers on the long records are not practical since it would be necessary to count 100 or 1 000p/s marks along the edge of the film over lengths of several hundred feet, in some cases, in order to relate the recorded events to the zero mark at the beginning of the film. Timing signals are therefore provided which are either marked off in decades to give a 'ruler' time scale or coded to give an indication of elapsed time with respect to the arbitrary zero.

Experience in the design of the timing equipment has shown that although it is necessary to have a central equipment it is not practical to meet every requirement of the

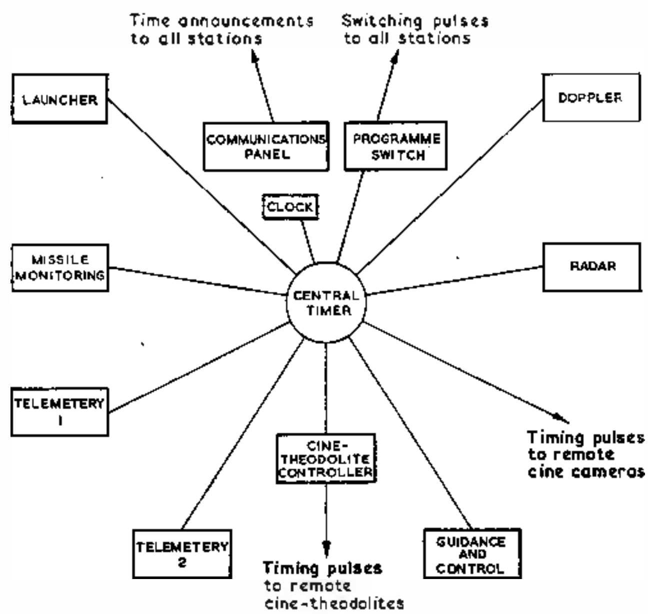

Fig. 1. The central timer and range instrumentation

range instrumentation with one comprehensive unit. To do so would result in a very complex unit, difficult to service and maintain; furthermore much of the equipment would become obsolete as range requirements change. The best arrangement is to design the timer to provide the more basic signals and to meet varied requirements as to the waveform, phasing, voltage and impedance level of the pulses with auxiliary equipment such as the programme switch and cine-theodolite controller. Such units can be modified or scrapped as range requirements change. Auxiliary units can be sited with the timer, or where convenient at the remote stations local to the particular instruments which they serve. A feature of the design which has proved useful is the provision of two central timing units,

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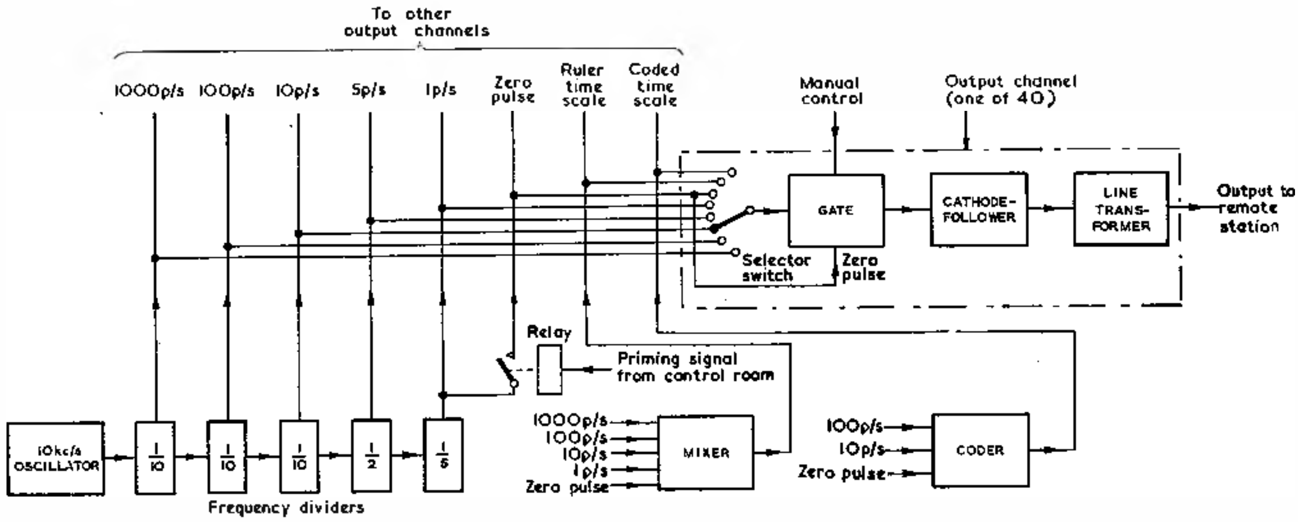


Fig. 2. Schematic diagram of the central timer

one serving as a stand-by in the event of a fault on the other. Extensive monitoring facilities are provided so that a fault is readily detected and the equipment switched out of service. These features are most important since a fault in the timer affects the whole range instrumentation.

Accuracy requirements for the timer are not very stringent. It is sufficient that the pulse repetition frequencies be correct to within $\pm 0{\cdot}01$ per cent since performance measurements made on the missile, such as velocity evaluation, cannot be made in any case to better accuracy than $\pm 0{\cdot}1$ per cent. What is most important is that the various events during a shoot can be co-related in time to within ± 1msec. These requirements are met by deriving all timing pulses by division of a common crystal oscillator frequency.

This article is limited to a description of the basic timing equipment; the auxiliary equipment such as the programme switch and cine-theodolite control system will be described in later articles. The description refers to a specific installation; other equipments very similar to that discussed here are, however, installed at guided missile ranges both in Australia and the U.K. Much of the circuit design is conventional and is not described in detail; the time coding circuits are, however, original and are fully detailed.

General Description

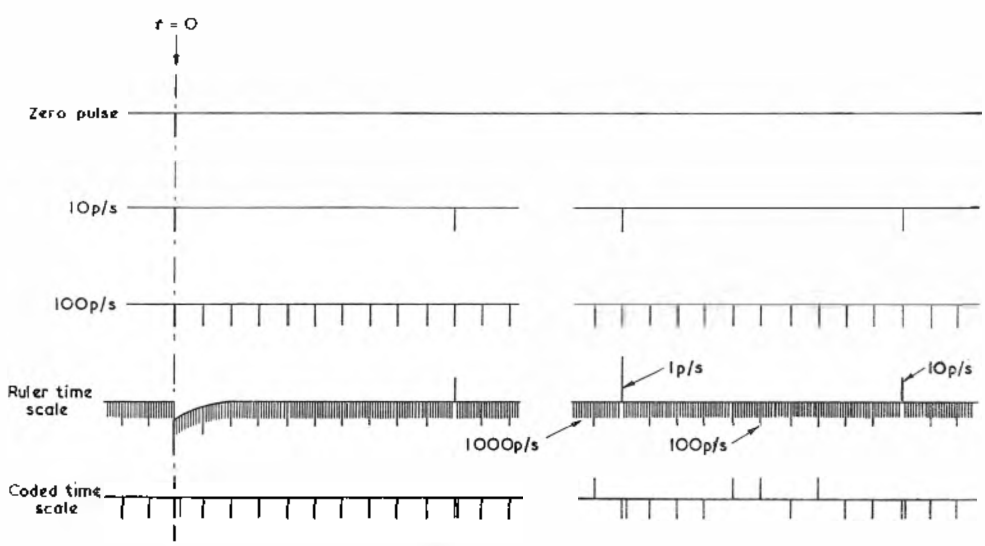
The complete system is shown by Fig. 2. It consists essentially of a 10kc/s crystal oscillator driving a series of dividing circuits which feed pulses at various frequencies to a number of output channels. There are 40 output channels identical with that shown, each one supplying a particular recorder or control equipment. The pulses are transmitted to the remote stations via underground telephone cables.

A gate in each output channel is opened just before the missile is launched so that the start of the transmissions represents zero time. Transmission before zero time can be effected by manual operation of the gates and the zero pulse which triggers them can also be selected and transmitted as a time signal. A relay operated from the control room selects one of the 1p/s pulses as a zero; this ensures that the zero is co-incident with a cardinal marker on all time scales so that it is unnecessary to add an end reading when annotating or analysing film recordings. It is important to mark the zero on all recordings since it is a common reference which permits the various data to be co-related in time.

An additional output is supplied by a mixer. This provides the ruler time scale shown on Fig. 3. It is a 1 000p/s signal with markers superimposed at 1, 10 and 100p/s. Zero time appears as a step in the composite waveform.

The function of the coder, shown on the schematic diagram, is to supply pulses that will serve as time markers and also indicate the elapsed time at any instant with respect to the zero pulse. A typical coded signal is shown by Fig. 12. The coded signal and ruler time scale are used

Fig. 3. Typical timing pulses



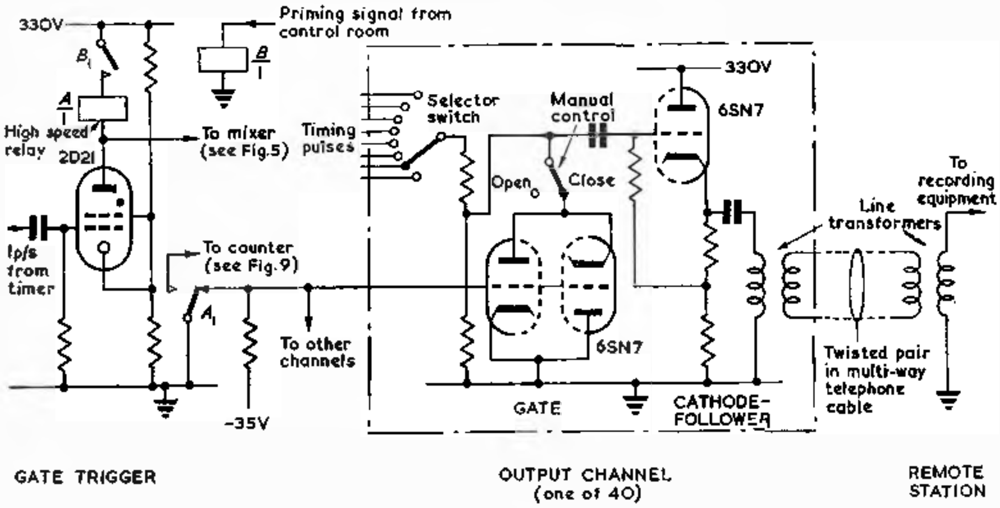


Fig. 4. Output channel

The gate is a double triode with the grids strapped and connected via a relay contact to earth. In this condition signal pulses fed to the grid of the cathode-follower are shorted to earth; the left-hand triode removing positive and the right-hand triode negative pulses. When the relay contact is opened the grids are biased beyond cut-off; this opens the anode circuits and permits signal pulses to reach the cathode-follower. The gate can be opened manually by operating the switch shown in the anode circuit of the triodes. A necessary feature in the design of the gate circuit is its freedom from switching transients. A transient voltage, or change in d.c. level when operating the gate would be fed via the line transformers and would distort the output waveform.

The gates in each of the output channels are operated by a common trigger circuit. This is a thyratron which is fired by one of the 1p/s pulses from the timer. When the thyratron is fired it operates a high speed relay which then removes the earth connexion from all the gate circuits. The thyratron is inoperative until the h.t. circuit is completed by another relay. The relay which primes the thyratron is normally switched on from the control room just before the missile is fired off the launcher. This priming relay is switched off when the flight is over so that the thyratron is reset and ready for another shoot.

where long runs of recording film are involved. The coded time scale is also useful where recording equipments, such as target cameras, are switched on after zero time. It is impossible to record a zero mark under these conditions, but zero time is implied in the code so that the recordings can still be related in time to that of other instruments.

The frequency dividers each consist of four binary circuits, the division being reduced from 16 to 10 by feedback connexions. An additional output is taken from the last decade to give the required 5p/s signal. The circuit design of these dividers, and of the crystal oscillator, is conventional and will not be described here. Design details of the other circuits, in particular those of the coder, are given below.

Output Channel

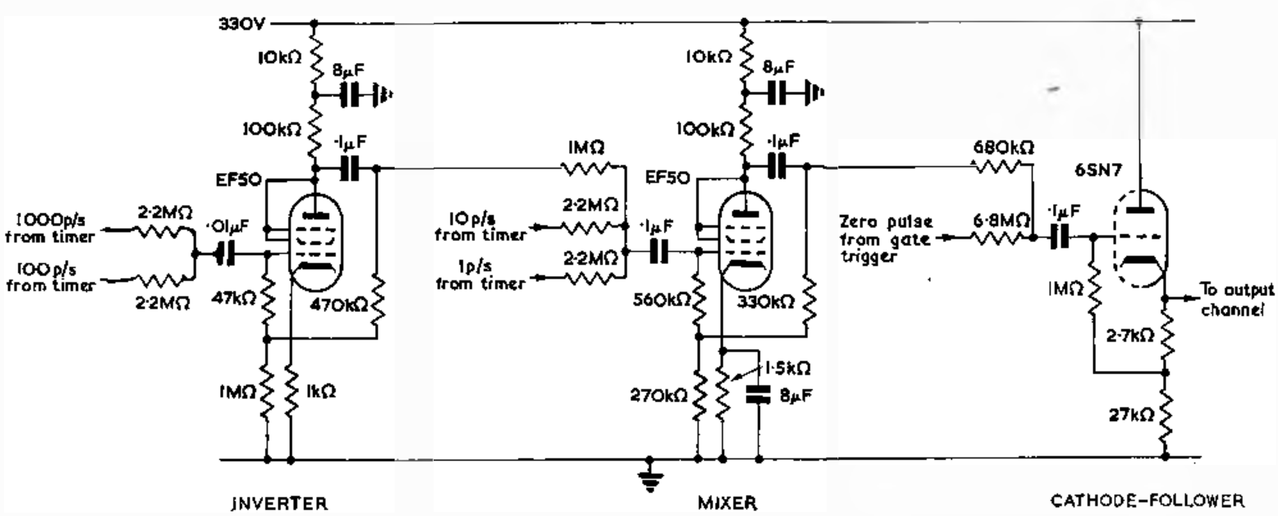
This is detailed by Fig. 4. There are 40 output channels, each one consisting of a selector switch, gate, cathode-follower and a Post Office type line transformer. The cathode-follower and line transformer effect the necessary impedance transformation for transmitting the signals along cables to the remote stations. The cathode-follower has an output impedance of about 400Ω which is reduced to 53Ω by the transformer.

Line Transmission

Timing pulses are transmitted to the remote stations via underground telephone cables. These are 20lb, multi-core, lead sheathed cables; each pair of cores having the following characteristics:

Attenuation: 1 to 2·5dB/mile at frequencies of 1kc/s to 10kc/s.

Fig. 5. Mixer circuit which provides the ruler time scale



Characteristic Impedance: 450 to 170Ω at frequencies of 1kc/s to 10kc/s.

Propagation Delay: 10μsec/mile.

The longest cable run at present on the range is about 4 miles and it is possible to transmit the timing pulses over this distance without excessive attenuation, distortion or delay. Lines longer than 6 miles could involve the use of repeaters depending on the requirements of the remote equipment.

In addition to the timing pulses the multi-core cables carry speech signals and other pulse transmissions. All signals are carried over twisted pairs of lines which are terminated in transformers so that the signals and lines are balanced about earth potential. This eliminates crosstalk between the various transmissions; the attenuation between any two signals being greater than 50dB/mile.

The general transmission of pulses by radio has been considered, but this would involve the use of a multi-channel radio transmitter and a receiver at each remote station. The radio installation would, therefore, involve complex terminal equipment compared to that used with lines. It would be more difficult to service in the field particularly with radio receivers installed at remote and cross-country observation posts. With lines the terminal equipment is limited to line transformers and in some cases a.f. amplifiers at the receiving end. Radio transmission is justified when a line installation is impracticable or too expensive. It would be useful for seaward or very distant observation posts. With the range installation being described a single channel radio link is used only for transmitting pulses to a mobile observation post, where lines would be impractical.

Mixer

This circuit provides the 'ruler' time scale and is illustrated by Fig. 5. Pulses at frequencies of 100 and 1 000p/s are fed from the decade dividers to the grid of the first valve. These pulses being inverted are added to the 1 and 10p/s pulses at the grid of the second valve. The composite signal is then fed via a cathode-follower to the 40 output channels. A zero pulse is added at the grid of the cathode-followers; this appears as a step in the composite waveform as shown by Fig. 3. Negative feedback is employed on the two amplifying valves to reduce the unwanted gain.

Coder

General Description

The function of the coder is to supply pulses that will serve as time markers and also indicate the elapsed time at any instant with respect to the zero pulse. A typical display of the coded time scale is given by Fig. 12. The pulses are spaced at 0·01sec intervals with a double pulse every 0·1sec. The groups of pulses between the double marks are read as binary numbers; a positive pulse being 1 and a negative pulse 0. The binary number indicates the number of tenth seconds that have elapsed up to the preceding 0·1sec marker. An alternative display is shown by Fig. 13. Here positive and negative pulses are replaced by dots and dashes; a dash being read as 1 and a dot as 0.

The general arrangement of the coder is illustrated by Fig. 6. Pulses at a frequency of 10p/s are fed via a gate to a nine stage binary counter. An electronic switch driven at a 100 steps per second scans the counter once after each operation. This scanning switch delivers an output pulse each time it sweeps across a counter that has been switched on.

The signal from the scanner is fed to a mixer where it is mixed with 10 and 100p/s signals to provide the required coded output; the mixing being illustrated by the waveforms of Fig. 11. The positive pulses from the scanner are twice the amplitude of the 100p/s pulses so that the two signals add to produce a positive pulse in the mixer output. The 10p/s signal is fed to the mixer via a delay circuit; it is then out of phase with the 100p/s signal and so produces double pulses in the output. Normally all pulses in the coder circuits are in phase since they are derived from the common 10kc/s oscillator and dividing circuits of the timing system.

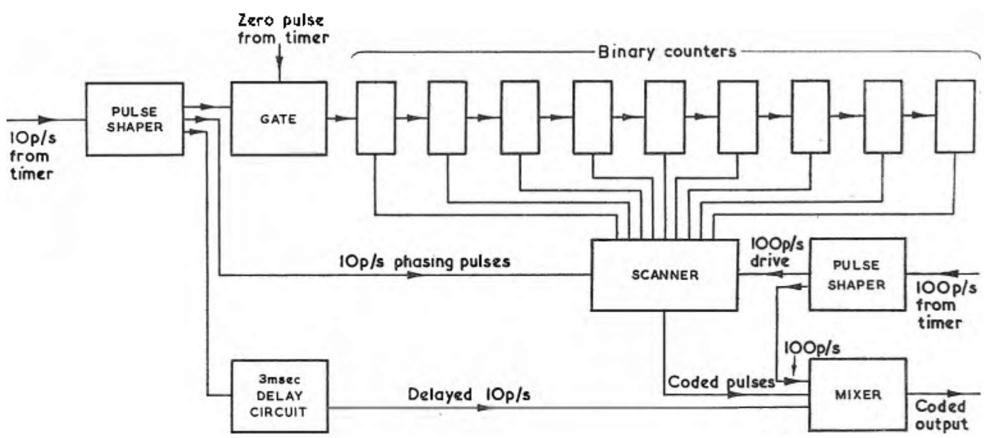


Fig. 6. Schematic diagram of the coder

The interpretation of the coded output signal follows from the fact that positive pulses indicate that certain binary stages are switched on at a particular instant; the position of the pulses with respect to the 0·1sec marker indicating which stages are concerned. Thus the groups of pulses between the double marks can be read as binary numbers, which indicate the number of tenth seconds that have elapsed up to the preceding 0·1sec marker. The elapsed time is measured from the instant at which the gate preceding the binary counter is opened. This gate is opened by a pulse coincident with zero time. Closing the gate at the end of the run resets the counter to the zero position; all the binary stages being switched to the off position.

The negative pulses shown in the output waveform are not strictly necessary for interpreting the code since the absence of a pulse could be read as 0. However, the absence of these pulses would result in long gaps in the time trace, the record would not be so convenient to read and interpolation would be necessary to measure to 0·01sec.

The scanner can be regarded as a ten-way rotary switch driven at 10 revolutions or 100 steps per second so that it rotates in synchronism with the 10p/s signal to the counter. Nine outlets of the switch are connected to the counter, the tenth being unused and serving as an idle position. A 10p/s signal phases the operation so that the switch reaches the idle position each time the counter is operated. This phasing ensures that a complete and successive scan of the binary stages is made after each counter operation.

With the arrangement described, the coder counts up to 51·1sec and re-cycles. This meets most requirements at the range for which this equipment was designed since the flight time of missiles fired there rarely exceeds 30sec. The coder can, however, be used to count seconds rather than tenth seconds when the counter will have a capacity of 511sec.

The output signal from the coder is normally displayed is therefore made by a small unit local to the particular recording equipment.

Pulse Shapers

These circuits which are detailed by Figs. 7 and 8 shape the 10p/s and 100p/s pulses supplied from the external timing system so that they are suitable for operating the counter and scanner, both of which require fast, well defined waveforms. They also feed suitably shaped pulses into the mixer circuit.

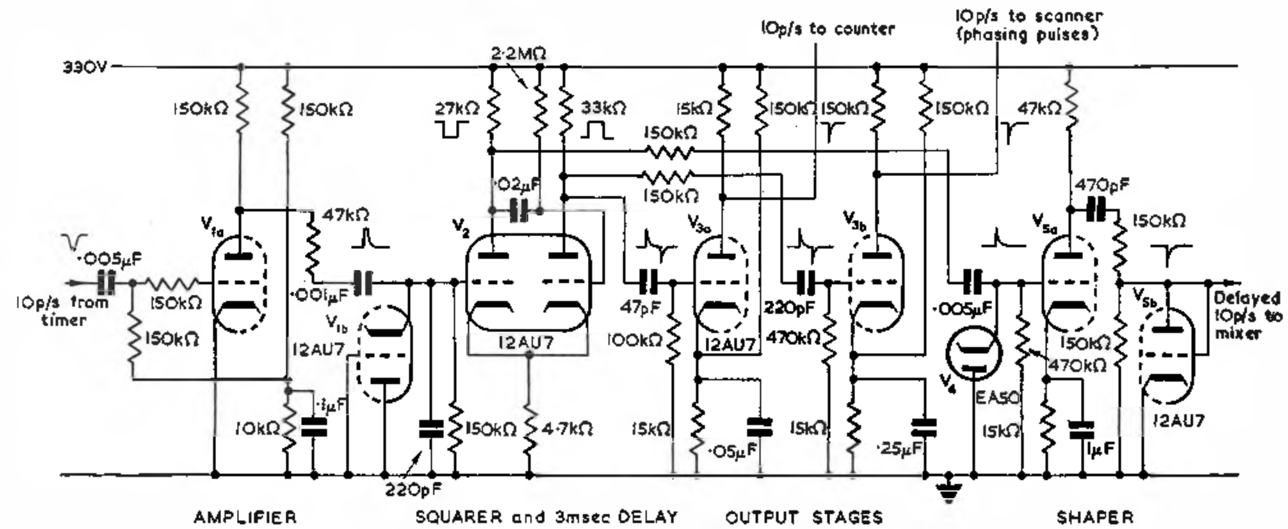


Fig. 7. 10p/s pulse shaper

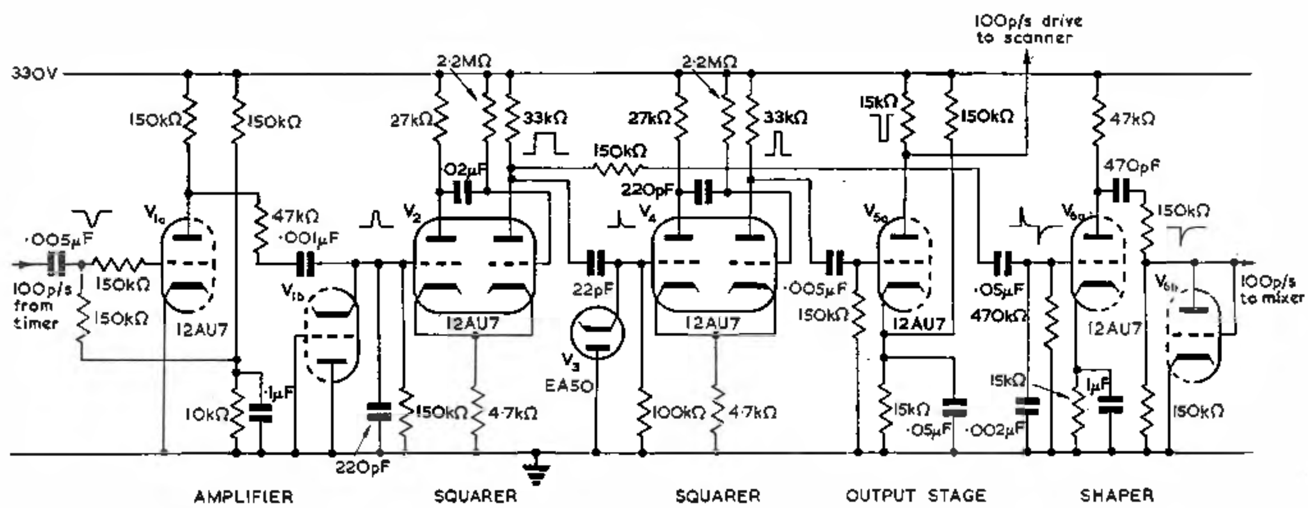


Fig. 8. 100p/s pulse shaper

as a time trace on c.r.o. recorders. With some recording equipment it is more convenient to record the time trace with a neon or argon filled flash lamp. The positive and negative pulses cannot be recorded as such on a single lamp so they are converted to square pulses which operate the lamp to produce a dash-dot time trace as shown by Fig. 13. Fig. 14 is a typical application where a dash-dot trace is recorded along the edge of a cine camera film. It is not practical to transmit the square pulses along the underground cables without excessive distortion; the conversion

The negative 10p/s pulses are fed to the grid of V_{1a}. This is positively biased so that it clips any small amplitude, unwanted signal from the input. The amplified pulses from V_{1a} trigger a flip-flop; fast, square waves being available at the anode of V_{2b}. The diode connected to the grid of the flip-flop removes any negative component of the signal while the 220pF capacitor renders the flip-flop insensitive to any induced h.f. transients. The square waves at V_{2b} are differentiated and fed to the amplifiers V_{3a} and V_{3b} which supply pulses to the counter and scanner. These amplifiers are biased to clip off the negative pulses corresponding to the trailing edge of the flip-flop square waves. The 150kΩ resistor at the anode of V_{2b} prevents the capacitive load of V_{3b} spoiling the fast waveform required by the counter.

Negative square waves available at the anode of V_{2a} are fed to the shaper V_5. Here the square waves are differentiated and the negative pips removed by the diode V_4. The 10p/s pulses at the anode of V_{5a} coincide with the trailing edge of the square waves from V_{2a} and they are therefore 3msec out of phase with the other 10 and 100p/s pulses. This circuit thus corresponds to the 3msec delay unit shown on the block diagram of Fig. 2. The network and diode tiated and the positive pulses removed by V_{6b}. The remaining negative pulses are fed to the mixer.

Counter

This circuit which is shown on Fig. 9 is a conventional nine stage binary counter.

In the reset condition the left-hand resistance chain of each binary is connected via a common rail and a 4·7kΩ resistor to earth. This raises the grid potential of the right-hand triodes so that they remain in the conducting condition with the left-hand triodes cut-off; the 10p/s input signal is then ineffective. The counter starts to operate at

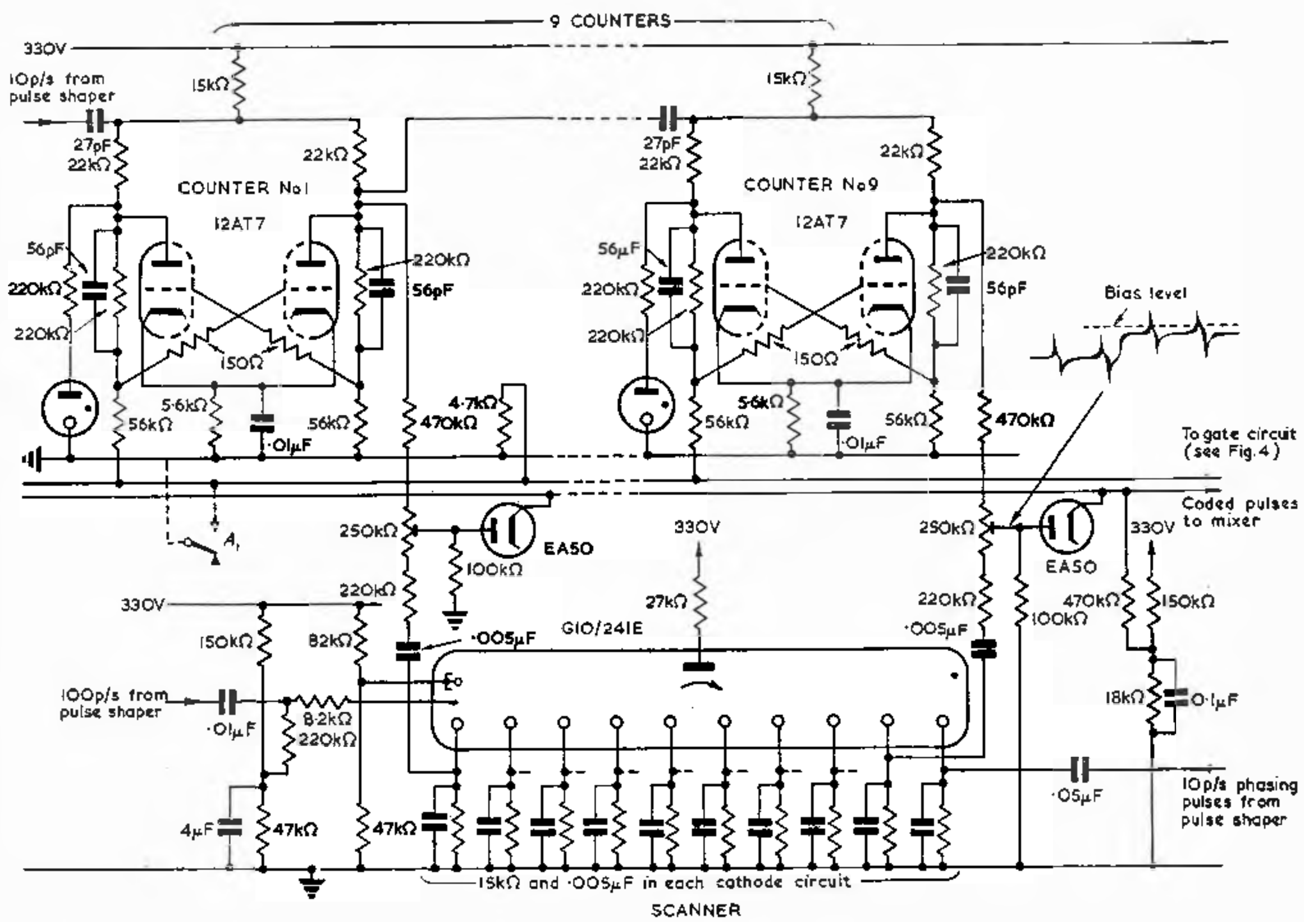


Fig. 9. Counter and scanner

connected to the anode of V_{5a} differentiate and clip the output pulses again so that a suitably shaped signal is available for feeding to the mixer.

The 100p/s pulses are amplified by V_{1a} and trigger the flip-flop V_2 as for the 10p/s signal. The square waves at the anode of V_{2b} are differentiated and the negative pulses shorted by V_3. The remaining positive pulses trigger a second flip-flop V_4 which provides the 40μsec pulses for driving the scanner. These pulses are available at the anode of V_{4b} where they have a peak amplitude of about 200V; they are fed to the grid of V_{5a} which is biased to clip the top and bottom off the input square pulses. The amplifier also inverts the flip-flop output; negative pulses being required to drive the scanner.

The output at the anode of V_{2b} is also fed to V_{6a} which functions as an invertor so that negative-going square waves are available at the anode. The square waves are differen-

zero time when the gate trigger relay, shown on Fig. 4, switches the reset rail to earth.

The neon lamps connected to each binary counter are mounted on the front panel of the equipment and give a useful indication of the counter operation; they are switched on and off by each binary stage.

Scanner

This is also shown on Fig. 9. It is a cold-cathode glow discharge tube in which the anode to cathode discharge is switched to successive cathodes by suitable trigger pulses fed to the transfer electrodes[1]. The transfer electrodes are supplied in this case with a 100p/s signal so that the discharge rotates at a 100 steps or 10rev/sec.

Negative pulses at a frequency of 10p/s and coincident with those operating the counter are fed to the tenth cathode, overriding the normal operation and pulling the

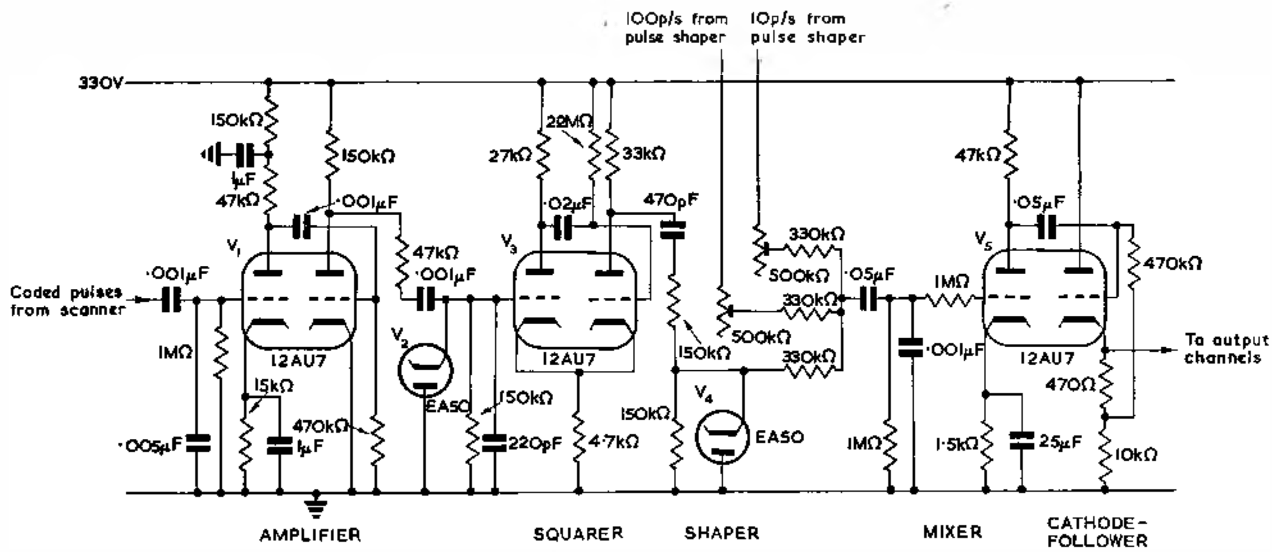


Fig. 10. Mixer and output stage

discharge to this cathode if it does not happen to be there when the counter operates. This phases the scanner so that a complete and successive scan is made after each counter operation. The phasing comes into effect when power supplies to the circuits are first switched on, it is then superfluous since the scanner rotates in synchronism with the counter until the equipment is switched off.

The current flowing successively through the 15kΩ resistors produces 10p/s pulses at each scanner cathode. The nine cathodes are connected respectively to the nine counter stages via a capacitor and resistance chain. A typical waveform for one of these connexions is shown. The square pulses at the cathode are differentiated and added to a platform voltage from the counter stage. The resulting waveform is fed via a diode to a common output rail. This diode is biased, by a d.c. voltage on the rail, so that pulses from the scanner cathode are only fed out when the particular counter stage is switched on. The resulting output is a train of pulses; the phase of each pulse showing which counters are switched on during each sweep of the scanner. A typical output for two successive scans is shown by Fig. 11.

The 250kΩ preset resistors in each resistance chain are used to adjust the signal amplitudes with respect to the bias voltage and thus balance out inequalities of valve characteristics and resistance values.

Mixer

This circuit is detailed by Fig. 10; its function is to shape the pulses delivered by the scanner and to mix them with the 10 and 100p/s signals.

The pulses from the scanner are amplified by V_{1a} and V_{1b} and then trigger the flip-flop V_3. The square pulses at the anode of V_{3b} are differentiated, the negative pip removed with a diode and the positive one fed to the mixer V_{5a}. The 10 and 100p/s signals are added here as shown in Fig. 11. It will be seen that the coded pulses are twice the amplitude and of opposite polarity to the 100p/s pulses so

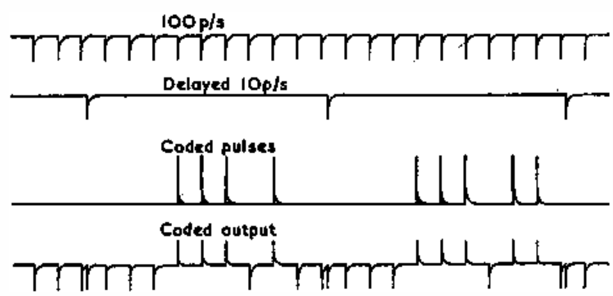


Fig. 11. Waveforms in mixer circuit

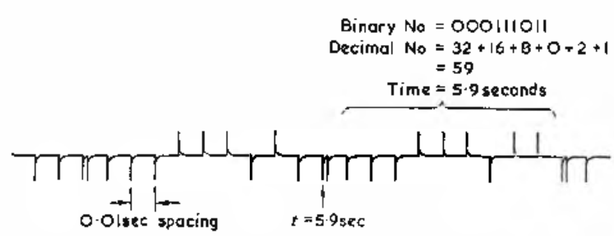


Fig. 12. Coded time scale

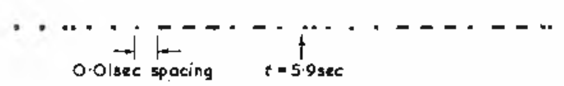
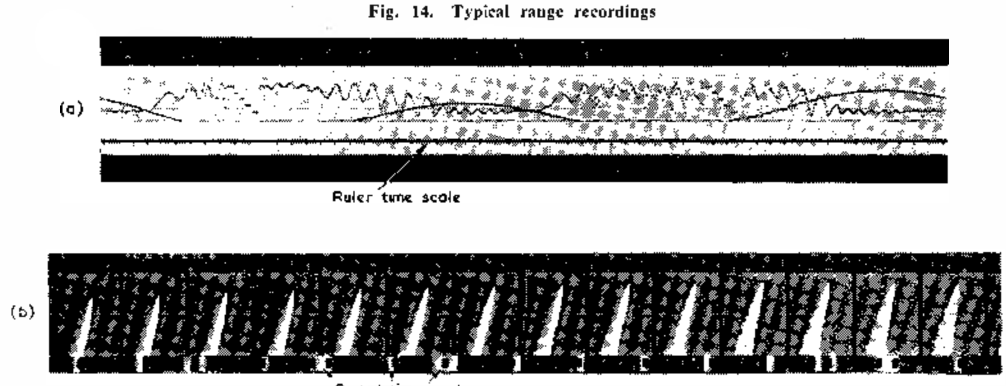


Fig. 13. Alternative coded time scale

Fig. 14. Typical range recordings



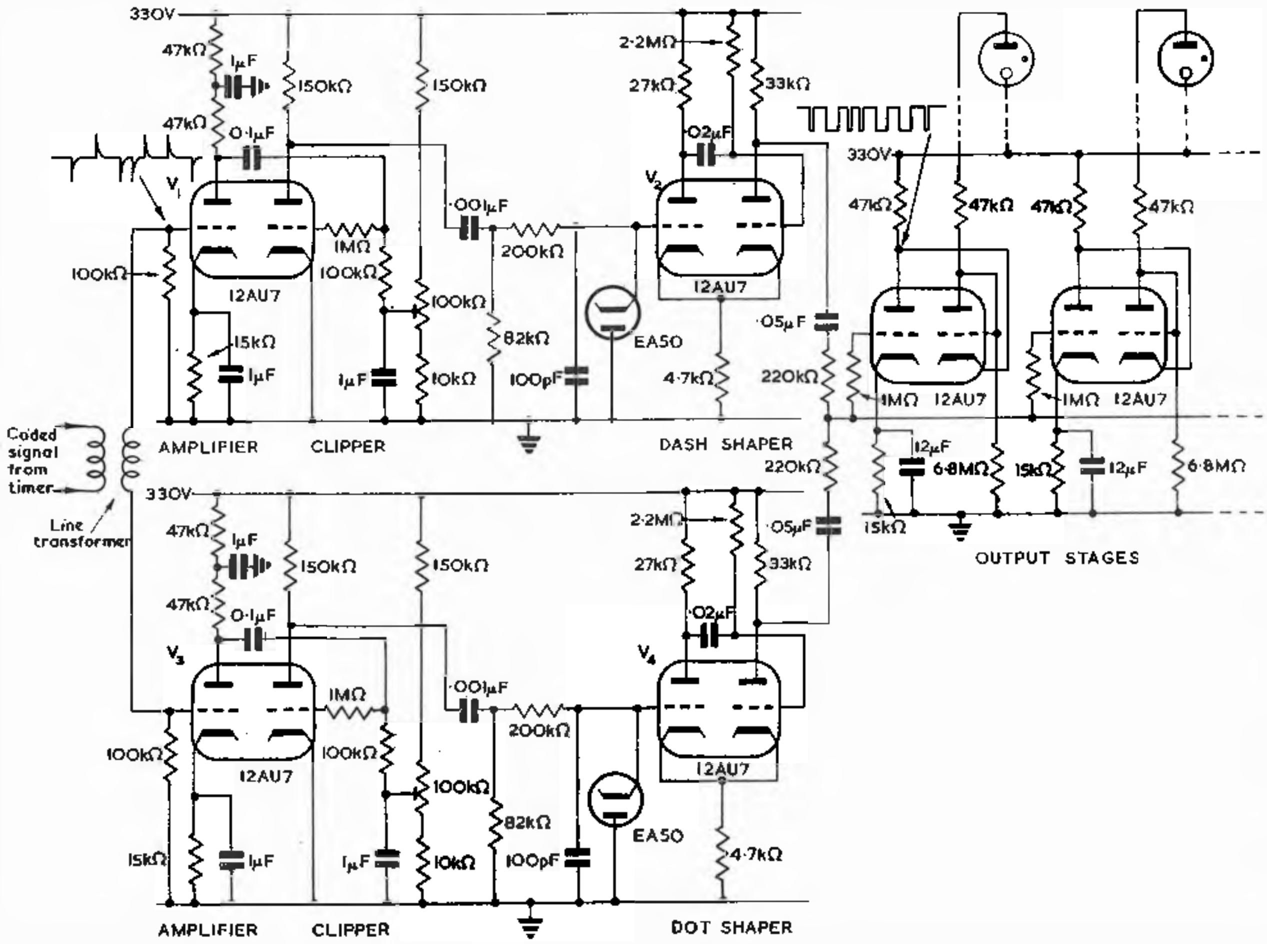


Fig. 15. Local amplifier

that the two signals add to produce positive pulses.

The coded pulses are adjusted to the required amplitude by the 500kΩ preset resistor shown. The 10p/s signal is out of phase with the other pulses and produces the double markers which bracket the pulses into groups of nine.

The shaping network connected to the anode of V_{3b} and an identical circuit in the 100p/s pulse shaper ensures that the two sets of pulses have the necessary similar waveforms. Slight discrepancies in the waveform and phasing of the pulses are corrected by the capacitors connected to the grids of V_{5a} (Fig. 10) and V_{6a} (Fig. 8).

The capacitors shunted across the grids of V_{1a} and V_{3a} in the mixer circuit eliminate any h.f. transients coupled into the circuits; these would cause irregular operation of the flip-flop V_3.

The final coded signal at the anode of V_5 is fed via a cathode-follower to the output channels of the timer.

Local Time Coder

As mentioned above it is convenient with some range equipment to record the time trace on a single neon or argon filled flash lamp. The positive and negative pulses of the 'coded' signal are therefore converted to square pulses which operate the lamp to produce a dot-dash time trace. It is not practical to transmit the square pulses along the underground cables to the remote recorder without excessive distortion; the conversion is therefore made by a small unit local to the recorder.

The circuit of this unit is detailed by Fig. 15. The equipment side of the line transformer is connected as a phase splitter so that a given signal pulse appears with opposite polarities at the grids of V_{1a} and V_{3a}. The signal is amplified by V_{1a} and fed to V_{1b} which is biased to reject positive, and amplify, negative pulses. The positive pulses from the anode of V_{1b} trigger the flip-flop V_2; this provides square pulses, each of 4msec duration, corresponding to the positive pulses fed to the grid of V_{1a}. The circuit, consisting of valves V_3 and V_4 operates in a similar manner to that just described. It operates, however, on pulses of opposite polarity, as seen on the line, and the flip-flop V_4 produces square waves of only 1msec duration.

The square waves from V_2 and V_4 are added and fed to the output valves. These amplify the signal so that it operates the neon recording lamps. The 6·8MΩ resistors keep the lamps ionized which improves the speed and reliability of their operation. The diodes isolate the driving and ionizing circuits. The unit is designed to operate timing lamps for three recording instruments if required.

Typical input and output waveforms are shown on the diagram; where the wide square waves correspond to negative and the narrow ones to positive input pulses. The time trace produced by the neon lamps is shown on the records of Fig. 13 and 14; the latter being a high speed cine camera film.

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Continuous Recording of Waveforms on Photographic Film

By V. B. Hulme*, M.A.

Limits to the information capacity of photographic film as a medium for recording electrical waveforms are examined. The limitations of the system using conventional cathode-ray tubes are discussed. The advantages of variable-area recording are pointed out, with particular reference to a new type of tube, in which the beam cross-section has a ribbon shape, producing a fluorescent line instead of a spot.

Using the new tube, various possible arrangements of the display are outlined. These include class-A push-pull, class-B push-pull, and the use of mirror images. It is concluded that whereas with conventional tubes the performance of the system is limited by the tube, with the new tube the limits set by the medium should be approached.

IT is common practice to record electrical analogues of physical phenomena by photographing the fluorescent spot of a cathode-ray tube. The spot is deflected in sympathy with the electrical waveform, and the information is recorded in the form of the distance between the spot and a datum point on the line of deflexion. To record phenomena which are continuously variable, the film is moved continuously in a direction normal to the recorded deflexion, so that a graph is drawn of the variable on a time scale.

This method of recording arose naturally from the obvious convenience in merely putting together the cathode-ray oscilloscope and the cine-camera, two well-known pieces of apparatus, with the minimum of modification. The system has both advantages and disadvantages, when compared with other possibilities and different media; to discuss these is not the object of the present article, but rather to draw attention to the major limitations and, more particularly, to some interesting suggestions for improvement, employing a ribbon-beam cathode-ray tube[1].

Limitations of the Deflected Spot System

The final limitations to the amount of detail or information which may be obtained from the recording ideally is given by the product of the resolution, or information capacity, of the film stock used in the camera, and the footage run. In practice, however, the film is not the limiting factor.

The number of elements into which the amplitude of the spot deflexion can be resolved, after recording, may depend on:

(a) The resolution of the film and the width of the track available for excursions of the spot image.

(b) The ratio of the tube spot diameter to the deflexion amplitude.

A typical fast negative film will resolve 30 lines/mm, a slow positive stock 100 lines. The track width available between sprocket holes on 35mm film is 25mm, so that between 1 500 and 5 000 amplitude elements can be resolved giving a maximum accuracy of 0·067 to 0·02 of 1 per cent of full deflexion. However, the best electrostatic-deflexion tubes have a spot diameter to maximum deflexion ratio of not less than 1/250, so that the amplitude resolution of the system is limited entirely by the tube, a figure which at best, assuming optimum focus and brightness, is a whole order coarser than the ideal limit imposed by the film.

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Even assuming, however, that a much improved tube could be made available, the ideal limit can never be approached with the deflected spot system, since the trace must have a minimum thickness in order to be detectable. A trace thickness of 0·016mm, corresponding to 30 lines/mm, would be practically invisible, and recovery of the information by electronic scanning systems would also be severely handicapped.

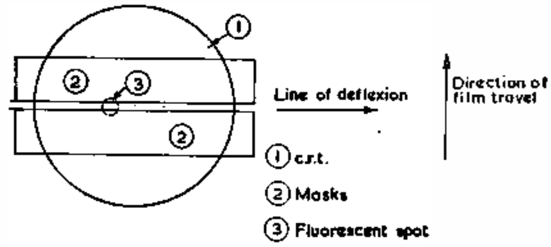

Fig. 1. Masked c.r.t. display

The number of independent elements which can be resolved per second of recording time, may depend on:

(1) The resolution of the film and the speed at which the film is propelled in the camera.

(2) The decay time of the phosphor.

(3) The effective width of the image of the spot in the direction of film movement.

The second factor may be dismissed with the observation that phosphors are readily available, as used in flying-spot scanning processes, with decay times less than 1μsec and allowing up to 10^7 elements/sec to be recorded without difficulty. Turning to the other two limiting factors, ideally the spot image should be equal to (or less than) the smallest element which the film can resolve, in order that the system performance should be limited only by the film and the speed of propulsion. This desideratum is rarely achieved in practice, but it is possible to mask the spot edges with plates having edges parallel to the line of deflexion, as in Fig. 1, and so to reduce the effective spot width as desired.

The velocity of propulsion is limited in continuous recording by practical considerations of bulk and cost only. Thus a typical camera employed for this purpose has a maximum speed of about 4m/sec, which gives $0{\cdot}4 \times 10^6$ elements/sec at a resolution of 50 lines/mm. Theoretically, however, a limit will be dictated by the necessity for adequate exposure of the film, since exposure duration is inversely proportional to velocity. To evaluate this limit with a deflected spot system, allowance must be made for a large component of velocity due to deflexion. This increase in velocity due to deflexion may be determined by considering the maximum excursion of the spot during two elementary intervals of time, giving the maximum average velocity at which the position of the spot on the amplitude scale must be detectable at half-time t_2 (Fig. 2). Since it is assumed that system performance is limited by the film only, the elements have the same dimensions on the amplitude and time scales, and the increase in velocity is therefore given by $N/2$, where N is the number of amplitude elements.

The duration of exposure T is then given by

$T = 2d_i/NV$, where

d_i = width of image

V = velocity of film

and the product of the number of elements per second and the number of elements per track width, which is a

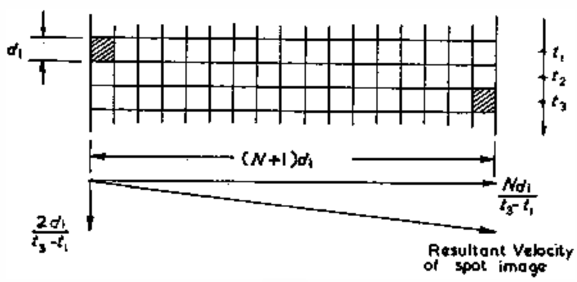

Fig. 2. Image velocity diagram

measure of the information capacity of the system, is given by

$$NV/d_i = 2/T$$

Now $T = E/I = 4EF^2/\pi B$

where I = image illumination

F = f/No. of lens

B = brightness of spot on tube face

E = minimum exposure

giving

$$NV/d_i = \pi B/2EF^2$$

neglecting loss of light in the lens and reduction of effective lens aperture due to the optical magnification, since these will lead only to small errors in practical cases[3].

Owing to the failure of the reciprocity law at short exposure times it is difficult to obtain a figure for the minimum exposure E, which must be considerably larger than in normal photographic practice. It is customary to evaluate emulsion speed in km/sec under specified conditions of tube operation and film development, assuming a lens aperture of $f/1$, for a trace density of 0·1 above fog level, i.e. for a trace which is barely perceptible. While a typical figure is 2km/sec, special methods applicable only to 'one shot' records of high speed transients have enabled writing speeds of several thousands of km/sec to be attained. For day-to-day recording of continuous phenomena a conservative figure must be taken, and it is found that with the maximum film velocity quoted for a typical camera (4m/sec), there is little scope for bandwidth extension by masking the spot, since this reduces the exposure below the minimum level. This would not be so, however, were it not for the reduction in effective exposure due to deflexion velocity.

Another considerable embarrassment is that the deflexion velocity varies from zero to a maximum several orders higher than that of the film, giving rise to exposure modulation in the ratio of the resultant image velocity to the film velocity. The intensity of the spot plotted against distance from the centre approximates the Gaussian probability law, so that the width of the recorded trace varies by an order or more with exposure, and it is consequently impossible to obtain optimum amplitude resolution over a large exposure range.

If the use of a ribbon-beam tube[1] in place of the conventional tube were to eliminate exposure modulation and all its disadvantages, the information capacity of the system would be extended by several orders.

The Ribbon-Beam Tube

The ribbon-beam tube is described in detail elsewhere, but briefly the cathode area has been extended along a

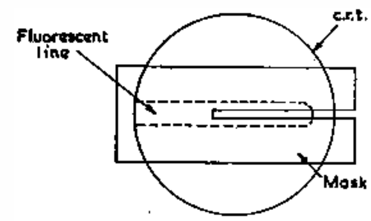

Fig. 3. Masked ribbon beam tube display

Fig. 4. Ribbon beam recording (film negative)

radial line, so that, without deflexion, a fluorescent line appears on the screen[1,2]. The intensity of this line at any point along its length is of the same order as that of a conventional tube spot. The beam is deflected in a direction parallel to the length of the line so that, in recording, the image of an elementary part of the line is replaced on deflexion, with that of another part, for the purpose of exposing an elementary film area, so that no exposure modulation can result, except in association with the end of the line. The information recorded is basically the position of the ends of the line with respect to a datum. Since the end of the line marks a transition from high to low intensity, it is convenient to limit the track width, by masking one side, so that the excursions of only one end of the line are accommodated, thus producing a variable area recording. The masked edge then forms a convenient datum line, and it is preferable that the mask should be on the cathode-ray tube so as to avoid errors due to camera vibration and lateral film movement. The edges of the line may also be masked as suggested for the deflected spot system, illustrated in Fig. 3. The form of the recording is shown in Fig. 4.

Variable-Area Recording

The advantages of the variable-area form of recording are:

(1) The electrical waveform may be reproduced auto-

matically using simple apparatus (light source, optics and photocell).

(2) The information is recorded in the position of a density transition. The sharpness or slope of this transition is not limited as it is with a deflected spot system by the vanishing thickness of a line.

(3) Exposure modulation is reduced, if not eliminated, allowing optimum adjustment of exposure and extended information capacity.

Variable-area recordings have been made by deflecting the spot of a normal cathode-ray tube at a rate high compared to the velocity of the film and the maximum rate of deflexion due to the recorded signal. The line so produced can then be modulated either in length or in position[4,5,6]. Some of the advantages of variable-area recording are obtained with such methods, but it is plain that the effective exposure of the film is reduced (per unit area exposed) by the dissipation of the energy available from the tube spot over a larger area. However, it is possible to operate the tube under optimum conditions, since modulation of the exposure by the rate of change of input is reduced, and there is no risk of burning a hole in the phosphor when the latter is zero. This risk is a certainty when high intensity beams are used in a deflected spot system, unless arrangements are made to remove the spot from the screen.

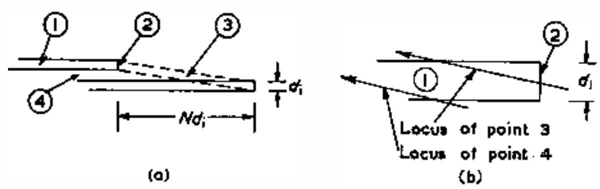

Fig. 5. Image velocity diagram, ideal ribbon beam

In a variable-area recording intended for electrical reproduction (e.g. of sound sequences) the photographic process is closely controlled so as to achieve the best overall system performance, and many subtleties enter into the determination of optimum values for the many variable parameters. In general the object is to produce a density transition across the track on the final print, from a value giving negligible attenuation to one giving negligible transmission, so that the emulsion characteristic is used only as an on-off device, and the photocell output is proportional to the area of unexposed track and not on variable parameters of the photographic process.

To assess the performance limits of such a system it is necessary to consider the brightness transition at the end of the fluorescent line on the tube, and the corresponding density transition produced on the film. There is no reason why the brightness transition should differ greatly from that for a conventional tube, following a radial line from the centre of the spot outwards. In the absence of modulation therefore it will be possible to locate the position of either the spot or the line with equal precision, if marked by the occurrence of the same arbitrary density transition in both cases. When the spot is deflected, however, the reference density may not be obtained at all, owing to velocity modulation, and certainly will not appear at the same distance from the trace centre. This applies even if velocity modulation does not carry the exposure out of the linear part of the film characteristic.

Consider now the image of a line (1) (Fig. 5) having a perfectly defined end (2), and of width d_i equal to an amplitude element. Suppose a deflexion Nd_i occurs in time $2d_i/y$. The exposure at the halfway point (3), assuming a constant deflexion velocity $Nv/2$, will be half that at the point (4), where no velocity occurs. The error in determining the position of the end of the line at the halfway point will be small, because the density transition on the film is produced by a wide exposure range, of the order of $1:2^7$. It can be shown by superposition methods that if the end of the line is imperfectly defined, the same reasoning applies, and the error due to deflexion velocity is no greater (Appendix 1).

Hence the variable-area system, using a ribbon beam tube, should have a great advantage over the deflected spot system, in that negligible velocity modulation results from deflexion. Consequently the bandwith of the system is not likely to be limited by a minimum film exposure, and the information capacity will be limited only by the film resolution.

Multi-Channel Variable-Area System

In a multi-channel variable-area system, the mask at the unmodulated end of the fluorescent line can be replaced by a mirror surface in a plane normal to the line of

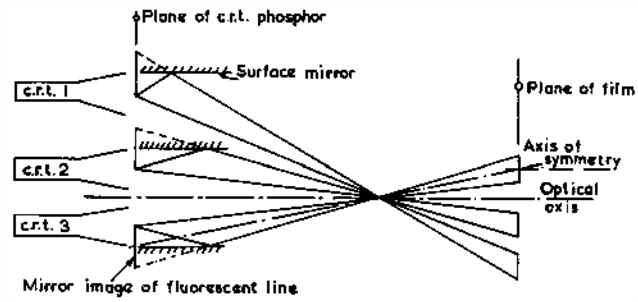

Fig. 6. Multi-channel variable-area recording system

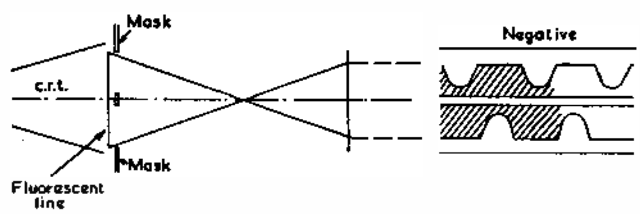

Fig. 7. Push-pull recording system

deflexion, and parallel to the optical axis[7] (Fig. 6). The modulated end of the line must be placed between the mirror and the optical axis, so that it is not masked. The length of the mirror is made adequate to produce a reflected image of the line, but no larger, otherwise excessive safety gaps will be required between channels. This arrangement results in bilateral area modulation, and has the advantage that the exposure of the film is doubled by the reflected light. This advantage is nearly always valuable, e.g. by doubling the optical reduction and the width of the fluorescent line, a film stock can be used having less sensitivity and greater resolution.

Push-Pull Variable-Area Systems

When a high quality bi-directional (centre zero) channel is required, both ends of the fluorescent line can be used, producing push-pull variable-area tracks (Fig. 7). The fluorescent line is of constant length, and occupies a substantially central position in the absence of deflexion.

Deflexion modulates the area exposed on each half of the track which is presented to a pair of photocells, each being affected by one half of the track, their outputs being combined in antiphase. It is advisable to mask a small portion of the line in the centre, as shown, so that a safety gap is produced between the tracks, and small lateral movements of the film do not produce a spurious output. The advantage of the push-pull arrangement is that spurious components common to both channels are reduced or eliminated, including even harmonic distortion which might result from non-linear deflexion characteristics of the tube. The system described would be analogous to a class-A push-pull amplifier or modulator.

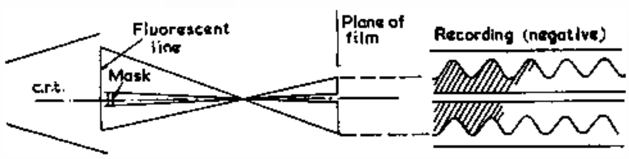

Fig. 8. Quiescent class-B push-pull system

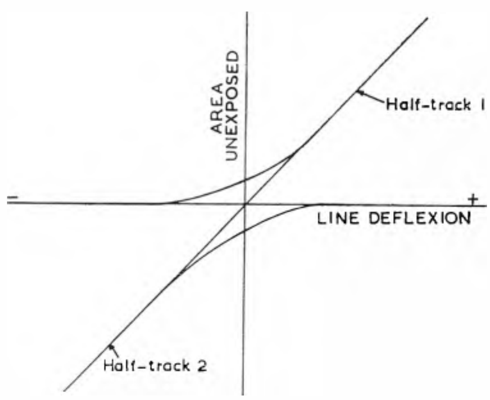

Fig. 9. Recording characteristic for system of Fig. 8

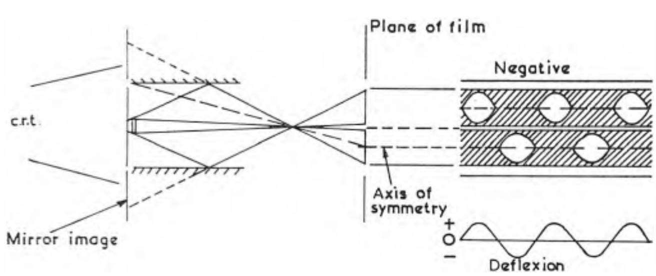

Fig. 10. Bi-lateral push-pull recording with mirrors

By masking the ends of the fluorescent line so that the space between the masks is just filled by the undeflected line, a quiescent class-B type system results. If the negative is used, as in Fig. 8, noise due to the exciter lamp, photocells and foreign matter adhering to the film, which forms a large part of the background noise in a film record replay, is proportional to the modulation (not constant) so that high signal-to-noise ratios are maintained at all modulation levels. Although the length of the line is critical in relation to the distance between the end masking plates, neither of these are likely to drift, once set up. Drift in the zero position of the line results in a d.c. output error only.

When the ends of the line are not clearly defined conditions are similar to those obtaining with a valve having a characteristic curved near cut-off. With the arrangement analogous to class-A operation, no distortion results if excursions are limited to the straight part of the characteristic. In the class-B push-pull arrangement, the curved part of the characteristic cannot be avoided, but distortion can be avoided if the masking is set up so that the curvature due to one end is compensated by that due to the other (see Fig. 9 and Appendix 2).

The advantages of the mirror arrangement can also be realized with class-B operation, by substituting mirrors for the masking plates at both ends of the line (Fig. 10). The film record takes the form of bilateral push-pull tracks (quiescent). Owing to the thickness of the glass of the tube face, the mirrors cannot reach the plane of the phosphor, and a small gap must be present in the centre of each bilateral track. It can be shown that distortion from this cause is avoided if the mirrors are placed so that the distance between them is the same as the length of the line (Appendix 3).

Conclusions

It has been shown that there are serious fundamental limitations to the performance of the conventional cathode-ray trace recorder. These limitations can be more or less completely avoided by employing the ribbon-beam tube and variable-area techniques, performance then being limited only by the information capacity of the photographic medium. There appears to be no valid reason why the ribbon-beam tube should be appreciably more expensive than conventional tubes[1]. Future development will, however, depend on the relative popularity of the photographic medium and its competitors, a subject which has not been discussed. The trend is towards digital techniques, and if these can be applied to film recording so as to extend the information capacity of the medium, it is possible that superior performance and the advantage of a visible and more permanent record will ensure a future for this method of recording.

Appendix 1

In Fig. 5(a) the deflexion of the line image (1) from one extreme position to the other was illustrated for a line having a sharply defined end (2). The deflexion occupied a time interval ($t_3 - t_1$), during which the distance travelled by the film was twice the width of the line image ($2d_i$). The exposure at the point (3), which is defined as the position of the centre of the line end on the trace at time t_2, assuming constant deflexion velocity, was seen to be half that at point (4), since $t_2 - t_1 = t_3 - t_2$. In Fig. 5(b) the moving line image was, for convenience, fixed in the diagram, and the locus of points (3) and (4) drawn as lines crossing the image.

Consider now a line having a graded brightness transition at the end. Such a line may be regarded as the resultant of a large number of lines having sharply defined ends, the ends being located in the neighbourhood of the gradient and distributed symmetrically in pairs about a point of half maximum brightness b, Fig. 11. Two of these overlapping lines are shown, having ends at a and c, and the locus of the point (3) (Fig. 5) is drawn across the ends, the position at times t_1, t_2, t_3 being indicated.

Since ends a and c are symmetrically spaced about b, the lines contribute a mean exposure at point (3) equal to that from a line ending at b. There will be in fact an infinite number of such pairs distributed over the gradient, contributing a combined exposure at point (3) half that at point (4), Fig. 5. If the position of the line end (2) Fig.

5 is taken as that of the half maximum brightness point on the gradient, then the exposure at point (3) is the same as at point (2) in the absence of deflexion. Therefore the velocity of deflexion will not modulate the exposure, and no error will result from this cause in measuring the position of the 'end' of the line, when the end has a brightness gradient of the type postulated.

APPENDIX 2

The conditions for linear class-AB push-pull operation

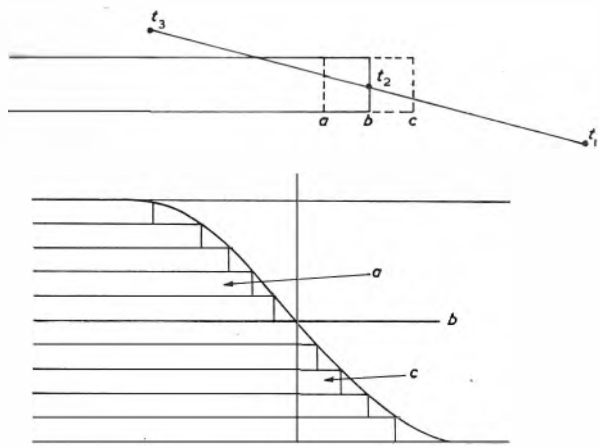

Fig. 11. Graded brightness transition

are easily found. Let the output from track (1) be an arbitrary function of deflexion x.

$$y_1 = F(x)$$

and that from track (2)

$$y_2 = F(-x)$$

For a linear balanced system the resultant output is

$$y_1 - y_2 = Kx$$

Expanding $F(x)$ in a power series

$$F(x) = a + bx + cx^2 + dx^3 + ex^4 + dx^5 \ldots.$$

But $F(x) - F(-x) = 2\,[bx + dx^3 + fx^5 \ldots.]$
$= kx$

$\therefore\ k = 2b$

and d, f etc. $= 0$

i.e. apart from a constant term and the term in x, odd powers of x are excluded from $F(x)$. Similarly, even powers of x are excluded from the slope of $F(x)$, $F'(x)$, which exhibits skew symmetry about $x = 0$.

APPENDIX 3

In a class-B quiescent push-pull system, both ends of the line must be masked. The length of the fluorescent line is critical, and must be equal to the distance between the end masks, so as to avoid a discontinuity in the characteristic at zero deflexion. When mirrors are used instead of end masks, the mirror surface cannot extend to the ends of the line since the thickness of the tube glass at the face intervenes.

Referring to Fig. 12, the mirror suface EF fails to meet the line OB by the distance FB. The light ray BL meets the plane of the film, after passing through the lens L, at the centre of track (1). Owing to the gap BF, however, the division of the track into two parts a, b, exposed to the mirror image and the direct image respectively, is marked by a point (or line) somewhat off centre, where the direct ray GFL and the reflected ray HFL meets the film. Since EF is normal to OB, $OB = BC$ and $HB = BG$.

Fig. 13 is a graph relating the unexposed area of the negative to the deflexion of the line from its central position (with the end at B). Curve a indicates the contribution of the reflected image (on part a of track (1)), and meets the zero area axis at u, which point corresponds to deflexion from B to H. Curve b gives the contribution of the direct image, meeting the zero area axis at r corresponding to deflexion from B to G.

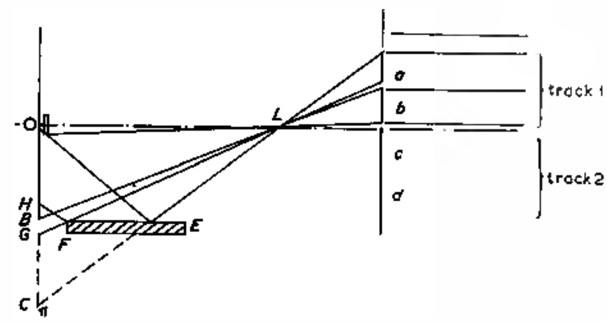

Fig. 12. Geometry of mirror optics

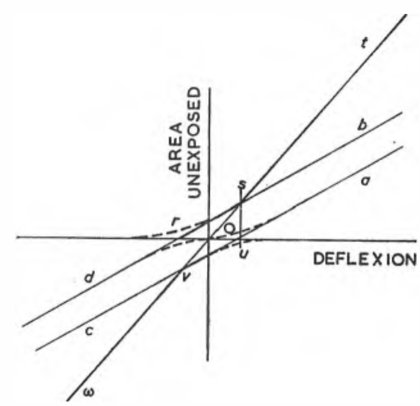

Fig. 13. Recording characteristic of system of Fig. 10

Similarly curves c and d indicate direct and reflected contributions to track (2) from the other half of the fluorescent line. If the geometry of the arrangement for track (2) has mirror symmetry with that for track (1), about the optical axis OL, then curves c and d meet curves a and b at u and r respectively and have the same slope, so that the overall push-pull characteristic is given by the straight line $w\ v\ o\ s\ t$.

In practice the ends of the fluorescent line have a finite brightness gradient. If the gradient conforms to that postulated in Appendix 1, and the distance between half brightness points at the two ends is made equal to the distance between the two mirrors, then a linear overall characteristic is obtained, although the curves representing individual contributions from parts a, b, c, d deviate from linearity, as shown by the broken lines, near the zero area axis.

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High Frequency Parameters of Transistors and Valves

By J. Zawels*

High frequency equivalent circuits for thermionic valves in the common-cathode, common-grid and common-anode (cathode-follower) configurations are developed, analogous to the common-emitter, common-base and common-collector configurations of junction transistors. From these, the exact open-circuit (z), short-circuit (y) and hybrid (h) parameters for each of the above circuits are tabulated. The approximations permitted by the relative magnitudes of various elements are also indicated.

THE behaviour of a device may be greatly clarified by using an equivalent circuit consisting of 'ideal' elements to represent it. In such a circuit approximations to simplify subsequent calculations may be made by inspection. If frequency dependent passive or active elements are introduced, however, the advantages of an equivalent circuit may be lost, since the behaviour of the circuit becomes obscured by the mathematical expressions describing the frequency dependence of the elements, and the use of the complex mathematical expressions for the open-circuit (z) of short-circuit (y) or hybrid (h) parameters[1], may in fact be simpler.

It is well known that at low frequencies the phase relationships in junction transistors and in vacuum triodes, (or in pentodes), are similar. The analogy exists between the base, collector and emitter electrodes on the one hand and the control grid, anode and cathode on the other. It would naturally be an advantage towards the understanding of the behaviour of both devices if their equivalent circuits could be made to correspond topologically. Indeed it was for this reason that a π equivalent circuit was first proposed to represent the low frequency behaviour of transistors[2]. Because the phase relationships at higher frequencies differ, however, the equivalent circuits of the two devices are unlikely to correspond without the introduction of frequency dependent elements.

The primary object in deriving the equivalent circuits used in this article was to use only frequency independent elements[3,4]. It was also found that there are significant advantages in developing special circuits for the three modes of operation, viz. common-emitter, common-base and common-collector. Fortuitously, accurate valve circuits having topological resemblance to these circuits are possible. The use of special equivalent circuits with valves is also advantageous.

In this article the equivalent circuits of both junction transistors and valves are presented, and the exact z, y and h parameters of both devices are tabulated. Approximations which may be made as a result of differing impedance levels are also indicated.

Triode at Low Frequencies

Fig. 1(a) shows a common-grid triode and Fig. 1(b) shows its conventional low frequency equivalent circuit which utilizes a fictitious generator. In practice this circuit is most useful in the analysis of common-cathode circuits.

For common-grid operation it will be shown that it is often advantageous to consider the generator and grid as a special three terminal device, as shown in Fig. 1(c). It is evident that this device has equal input and output currents and, for all terminations, has a voltage output which is a constant multiple ($\mu + 1$) times the input voltage. The usefulness of such a device, to be termed a K-amplifier[3,4], is that it has impedance transforming properties, and in this respect it can be treated as a conventional ideal two-winding transformer, with an impedance transformation ratio of $(\mu + 1) \equiv K$.

For the analysis of common-anode (cathode-follower) circuits, use is made of the impedance transforming properties of the K-amplifier by first transferring the anode resistance to the input side as shown in Fig. 1(d), its value now becoming $r_a/(\mu + 1)$. The K-amplifier is then replaced

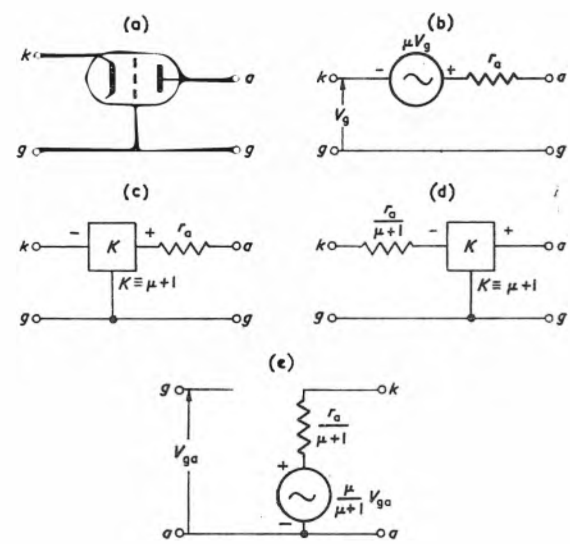

Fig. 1. Low frequency triode configurations

by a fictitious generator whose voltage V_a refers to the new input terminals, and finally the resultant circuit is redrawn for convenience to obtain Fig. 1(e). It is seen that this circuit closely resembles the conventional circuit (Fig. 1(b)), but for cathode-follower analyses is considerably simpler.

Triode at Medium and Very High Frequencies

For the high frequency case, consider complex impedances z_a, z_{ag}, and z_{gk} to be present between the electrodes. The equivalent circuits for the three configurations shown in Figs. 2(b), 2(a) and 2(c), follow by extension from Figs. 1(b), 1(d) and 1(e) respectively. The elements of these impedances may be very accurately obtained, by comparing Fig. 2(b) with Fig. 3, which is the 'intrinsic' equivalent circuit of a triode, and applies from zero frequency up to that frequency where transit time is appreciable compared with the period of the impressed signal[5]. This circuit is referred to as the intrinsic circuit since it is only valid for the active part of the valve, i.e. the portion of the valve elements that intercepts electrons[6]. For the complete circuit inter-electrode capacitances C_{ag}'', C_{gk}'' and C_{ak}'' must be added in parallel with the electrodes, the lead inductances

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in series with the electrodes. From the above, at very high frequencies, (neglecting lead inductance)

$$z_a = \frac{r_a}{1 + sr_a(C_{ak}' + C_{ak}'')} \quad \text{(1a)}$$

$$z_{ag} = \frac{1 + sC_{ag}'r_{ag}'}{s(C_{ag}' + C_{ag}'') + s^2 r_{ag}'C_{ag}'C_{ag}''} \quad \text{(1b)}$$

$$z_{gk} = \frac{1 + sC_{gk}'r_{gk}'}{s(C_{gk}' + C_{gk}'') + s^2 r_{gk}'C_{gk}'C_{gk}''} \quad \text{(1c)}$$

and

$$\mu' = \frac{\mu}{1 + sr_a(C_{ak}' + C_{ak}'')} \quad \text{(1d)}$$

where $s \equiv j\omega$. Note that the complex quantity μ' is used for convenience in tabulation, and arises out of the combination of r_a and $(C_{ak}' + C_{ak}'')$ as a single impedance, z_a. In

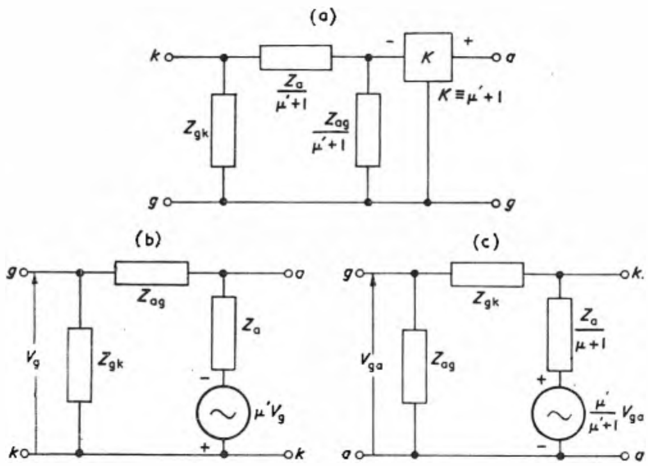

Fig. 2. High frequency triode configurations

the medium and low frequency ranges $r_{ag}' \ll 1/sC_{ag}'$ and $r_{gk}' \ll 1/sC_{gk}'$, and thus the above quantities simplify to

$$z_a = \frac{r_a}{1 + sr_aC_{ak}} \quad z_{ag} = (1/sC_{ag})$$

$$z_{gk} = (1/sC_{gk}) \text{ and } \mu' = \frac{\mu'}{1 + sr_aC_{ak}} \quad \text{(2)}$$

Transistors

The equivalent circuit of a junction transistor, termed the natural equivalent circuit[3], is shown in Fig. 4(a) and except for the presence of the element r_{bb}' its topological resemblance to Fig. 2(a) is evident. As in Fig. 3, the inter-electrode capacitances and lead inductances have been neglected. However, if any of these are to be included, it should be noted that the collector-to-base inter-electrode capacitance is the most important, by virtue of an analogous Miller effect. Also, collector and emitter lead resistances have been neglected (of the two only the latter is of possible significance). Now, similarly to Figs. 1 and 2, Fig. 4(a) is most suitable for common-base, Fig. 4(b) for common-emitter, and Fig. 4(c) for common-collector analyses. Fig. 4(d) shows an alternative common-emitter circuit which is premised on the fact that generally for transistors in the useful frequency operating range $y_{11} \gg y_{12}$ and $y_{21} \gg y_{22}$[3,4]. (See also assumption (4a) below.) It has the advantage over Fig. 4(b) in that its fictitious generator is proportional to a terminal quantity, viz., the collector current I_c, whereas that of Fig. 4(b) refers to an internal quantity, viz. voltage $V_{b'e}$. On the other hand, however, it introduces two more elements, the current generator and the one-to-one phase reversing transformer.

The natural equivalent circuit elements are r_{bb}', C_e, g_e, C_c', g_c', r_d, L_d and K. The element L_d is responsible for the common emitter short-circuit current amplification factor,

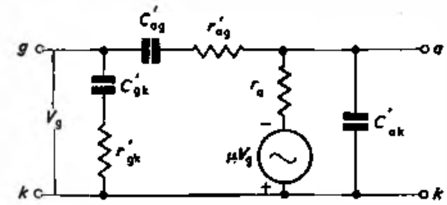

Fig. 3. Very high frequency intrinsic equivalent circuit for triode

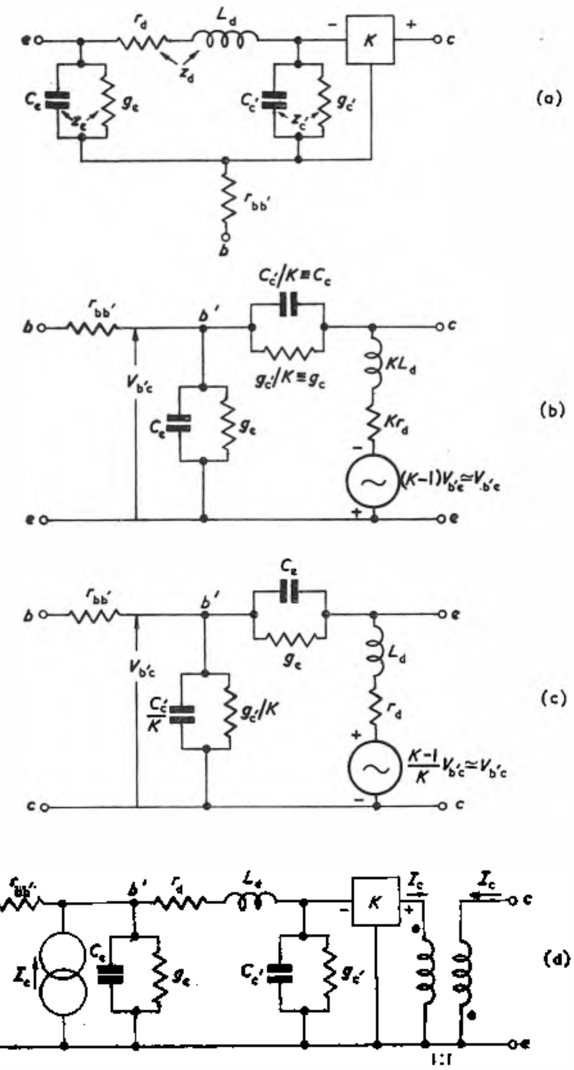

Fig. 4. Junction transistor configurations

α_{cb}, being capable of a phase shift greater than 90°, and is of importance only at the highest frequencies of the useful frequency range. It is seen that in the configurations of Figs. 4(b) and 4(c) the quantities C_c'/K, g_c'/K etc. appear. They arise from the impedance transforming properties of the K-amplifier and for convenience may be referred to as C_c and g_c, respectively.

Due to the topological resemblance of the equivalent circuits of the thermionic and transistor triodes, any expression in terms of the parameters of the one, are easily converted to the other. Table 1 gives the exact expressions for the z, y and h parameters for transistors for the common-emitter, common-base and common-collector modes of operation. The expressions are in terms of $r_{bb'}$, z_e, $z_{c'}$, z_d, and K where (s = jω),

$$z_e = \frac{1}{g_e + sC_e} \qquad z_{c'} = \frac{1}{g_{c'} + sC_{c'}} \quad \text{and } z_d = r_d + sL_d \qquad \ldots\ldots\ldots\ldots (3)$$

The usefulness of the concept of the K-amplifier in the common-base configuration is here illustrated by the fact that the z-parameters, for example, may be written down by inspection.

Table 3 applies to thermionic triodes and is easily derived from Table I, simply by letting $r_{bb'} = 0$, and replacing z_e by z_{gk}, z_d by $z_a/(\mu' + 1)$ and $z_{c'}$ by $z_{ag}/(\mu' + 1)$.

By utilizing the information on the relative magnitudes of the elements of a particular equivalent circuit, considerable simplification is possible. For example, in pentodes usually, $C_{ag} \ll C_{gk}$, and hence $z_{ag} \gg z_{gk}$. This may simplify some of the expressions in Table. 3. In transistors the following assumption may very often be made: $1/r_d g_e \gg 1$, $C_{c'} > C_e \gg C_{c'}/K$ $(= C_c)$, $1/g_e > r_{bb'}$, also g_e and g_e are of the same order of magnitude, while K is a large number (varying from say 500 to 10 000). Thus z_d, z_e, $z_{c'}$ and $r_{bb'}$ are usually at most no more than one order of magnitude removed from each other. Thus the following (not necessarily mutually exclusive) expressions will hold:

$$K \gg |1 + (z_d/z_{c'})| \quad \ldots\ldots (4a)$$

$$K \gg |z_e/z_{c'}| \quad \ldots\ldots (4b)$$

$$K \gg \left|\frac{r_{bb'}(z_e + z_d + z_{c'})}{z_{c'}(z_e + z_{d'})}\right| \quad \ldots\ldots\ldots\ldots (4c)$$

$$K \gg \left|\frac{r_{bb'}(z_e + z_d + z_{c'})}{z_{c'}z_e}\right| \quad \ldots\ldots\ldots\ldots (4d)$$

$$K \gg \left|\frac{z_e}{z_e + z_d}\right| \quad \ldots\ldots\ldots\ldots (4e)$$

$$K \gg \left|\frac{r_{bb'}z_e}{z_{c'}(r_{bb'} + z_e)}\right| \quad \ldots\ldots\ldots\ldots (4f)$$

Accordingly Table 2 gives the approximate expressions for the z, y and h parameters for transistors; the bracketed numbers indicate which of the above assumptions were made in the derivation of the particular parameter.

TABLE 1. 'Exact' Z, Y and H parameters for transistors

$[\Sigma \equiv Z_e + Z_d + Z_{c'};\ A \equiv KZ_{c'}(r_{bb'} + Z_e) + r_{bb'};\ B \equiv KZ_{c'}(Z_e + Z_d) + r_{bb'}\Sigma]$

PARAMETER		COMMON-EMITTER		COMMON-BASE		COMMON-COLLECTOR	
z_{11}	z_{12}	$r_{bb'}+\frac{Z_e(Z_d+Z_{c'})}{\Sigma}$	$\frac{Z_dZ_e}{\Sigma}$	$r_{bb'}+\frac{Z_e(Z_d+Z_{c'})}{\Sigma}$	$r_{bb'}+\frac{Z_eZ_{c'}}{\Sigma}$	$r_{bb'}+\frac{KZ_{c'}(Z_e+Z_d)}{\Sigma}$	$\frac{KZ_{c'}Z_d}{\Sigma}$
z_{21}	z_{22}	$-\frac{Z_e[(K-1)Z_{c'}-Z_d]}{\Sigma}$	$\frac{Z_d(KZ_{c'}+Z_e)}{\Sigma}$	$r_{bb'}+\frac{KZ_eZ_{c'}}{\Sigma}$	$r_{bb'}+\frac{KZ_{c'}(Z_e+Z_d)}{\Sigma}$	$\frac{KZ_{c'}(Z_e+Z_d)-Z_{c'}Z_e}{\Sigma}$	$\frac{Z_d(KZ_{c'}+Z_e)}{\Sigma}$
y_{11}	y_{12}	$\frac{KZ_{c'}+Z_e}{A}$	$-\frac{Z_e}{A}$	$\frac{B}{Z_dA}$	$-\frac{r_{bb'}\Sigma+Z_eZ_{c'}}{Z_dA}$	$\frac{KZ_{c'}+Z_e}{A}$	$-\frac{KZ_{c'}}{A}$
y_{21}	y_{22}	$\frac{Z_e[(K-1)Z_{c'}-Z_d]}{Z_dA}$	$\frac{r_{bb'}\Sigma+Z_e(Z_d+Z_{c'})}{Z_dA}$	$-\frac{r_{bb'}\Sigma+KZ_eZ_{c'}}{Z_dA}$	$\frac{r_{bb'}\Sigma+Z_e(Z_d+Z_{c'})}{Z_dA}$	$-\frac{KZ_{c'}(Z_e+Z_d)-Z_{c'}Z_e}{Z_dA}$	$\frac{B}{Z_dA}$
h_{11}	h_{12}	$r_{bb'}+\frac{KZ_eZ_{c'}}{(KZ_{c'}+Z_e)}$	$\frac{Z_e}{(KZ_{c'}+Z_e)}$	$\frac{Z_dA}{B}$	$-\frac{r_{bb'}\Sigma+KZ_eZ_{c'}}{B}$	$r_{bb'}+\frac{KZ_eZ_{c'}}{KZ_{c'}+Z_e}$	$\frac{KZ_{c'}}{KZ_{c'}+Z_e}$
h_{21}	h_{22}	$\frac{Z_e[(K-1)Z_{c'}-Z_d]}{Z_d[KZ_{c'}+Z_e]}$	$\frac{\Sigma}{Z_d(KZ_{c'}+Z_e)}$	$\frac{r_{bb'}\Sigma+Z_eZ_{c'}}{B}$	$\frac{\Sigma}{B}$	$\frac{KZ_{c'}(Z_e+Z_d)-Z_{c'}Z_e}{Z_d(KZ_{c'}+Z_e)}$	$\frac{\Sigma}{Z_d(KZ_{c'}+Z_e)}$

TABLE 2. Approximate Z, Y and H parameters for transistors

(For assumptions see equation (4))

PARAMETER		COMMON-EMITTER		COMMON-BASE		COMMON-COLLECTOR	
z_{11}	z_{12}	$r_{bb'}+\frac{Z_e(Z_d+Z_{c'})}{\Sigma}$	$\frac{Z_dZ_e}{\Sigma}$	$r_{bb'}+\frac{Z_e(Z_d+Z_{c'})}{\Sigma}$	$r_{bb'}+\frac{Z_eZ_{c'}}{\Sigma}$	$\frac{KZ_{c'}(Z_e+Z_d)}{\Sigma}$ (c)	$\frac{KZ_{c'}Z_d}{\Sigma}$
z_{21}	z_{22}	$-\frac{KZ_eZ_{c'}}{\Sigma}$ (a)	$\frac{KZ_{c'}Z_d}{\Sigma}$ (b)	$\frac{KZ_eZ_{c'}}{\Sigma}$ (d)	$\frac{KZ_{c'}(Z_e+Z_d)}{\Sigma}$ (c)	$\frac{KZ_{c'}(Z_e+Z_d)}{\Sigma}$ (e)	$\frac{KZ_{c'}Z_d}{\Sigma}$ (b)
y_{11}	y_{12}	$\frac{1}{r_{bb'}+Z_e}$ (b)	$-\frac{Z_e}{KZ_{c'}(r_{bb'}+Z_e)}$ (f)	$\frac{Z_e+Z_d}{Z_d(r_{bb'}+Z_e)}$ (c)	$-\frac{r_{bb'}\Sigma+Z_eZ_{c'}}{KZ_{c'}Z_d(r_{bb'}+Z_e)}$ (f)	$\frac{1}{r_{bb'}+Z_e}$ (b)	$-\frac{1}{r_{bb'}+Z_e}$ (f)
y_{21}	y_{22}	$\frac{Z_e}{Z_d(r_{bb'}+Z_e)}$ (a)(f)	$\frac{r_{bb'}\Sigma+Z_e(Z_d+Z_{c'})}{KZ_{c'}Z_d(r_{bb'}+Z_e)}$ (f)	$-\frac{Z_e}{Z_d(r_{bb'}+Z_e)}$ (d)	$\frac{r_{bb'}\Sigma+Z_e(Z_d+Z_{c'})}{KZ_{c'}Z_d(r_{bb'}+Z_e)}$ (f)	$-\frac{Z_e+Z_d}{Z_d(r_{bb'}+Z_e)}$ (f)	$\frac{Z_e+Z_d}{Z_d(r_{bb'}+Z_e)}$ (c)
h_{11}	h_{12}	$r_{bb'}+Z_e$ (b)	$\frac{Z_e}{KZ_{c'}}$ (b)	$\frac{Z_d(r_{bb'}+Z_e)}{Z_e+Z_d}$ (c)	$-\frac{Z_e}{Z_e+Z_d}$ (c)	$r_{bb'}+Z_e$ (b)	1 (b)
h_{21}	h_{22}	$\frac{Z_e}{Z_d}$ (a)(b)	$\frac{\Sigma}{KZ_{c'}Z_d}$ (b)	$\frac{r_{bb'}\Sigma+Z_eZ_{c'}}{KZ_{c'}(Z_e+Z_d)}$ (c)	$\frac{\Sigma}{KZ_{c'}(Z_e+Z_d)}$ (c)	$1+\frac{Z_e}{Z_d}$ (b)(e)	$\frac{\Sigma}{KZ_{c'}Z_d}$ (b)

TABLE 3. Exact Z, Y and H parameters for vacuum tube triode

$[\Sigma \equiv (\mu' + 1)Z_{gk} + Z_a + Z_{ag};\ \mu' = \mu Z_a/r_a]$

PARAMETER		COMMON-CATHODE		COMMON-GRID		COMMON-ANODE	
z_{11}	z_{12}	$\frac{Z_{gk}(Z_a+Z_{ag})}{\Sigma}$	$\frac{Z_aZ_{gk}}{\Sigma}$	$\frac{Z_{gk}(Z_a+Z_{ag})}{\Sigma}$	$\frac{Z_{gk}Z_{ag}}{\Sigma}$	$\frac{Z_{ag}[Z_{gk}(\mu'+1)+Z_a]}{\Sigma}$	$\frac{Z_{ag}Z_a}{\Sigma}$
z_{21}	z_{22}	$-\frac{Z_{gk}(\mu'Z_{ag}-Z_a)}{\Sigma}$	$\frac{Z_a(Z_{ag}+Z_{gk})}{\Sigma}$	$\frac{Z_{gk}Z_{ag}(\mu'+1)}{\Sigma}$	$\frac{Z_{ag}[Z_{gk}(\mu'+1)+Z_a]}{\Sigma}$	$\frac{Z_{ag}(\mu'Z_{gk}+Z_a)}{\Sigma}$	$\frac{Z_a(Z_{ag}+Z_{gk})}{\Sigma}$
y_{11}	y_{12}	$\frac{1}{Z_{ag}}+\frac{1}{Z_{gk}}$	$-\frac{1}{Z_{ag}}$	$\frac{(\mu'+1)}{Z_a}+\frac{1}{Z_{gk}}$	$-\frac{1}{Z_a}$	$\frac{1}{Z_{ag}}+\frac{1}{Z_{gk}}$	$-\frac{1}{Z_{gk}}$
y_{21}	y_{22}	$\frac{\mu'Z_{ag}-Z_a}{Z_aZ_{ag}}$	$\frac{1}{Z_a}+\frac{1}{Z_{ag}}$	$-\frac{\mu'+1}{Z_a}$	$\frac{1}{Z_a}+\frac{1}{Z_{ag}}$	$-\left(\frac{\mu'}{Z_a}+\frac{1}{Z_{gk}}\right)$	$\frac{(\mu'+1)}{Z_a}+\frac{1}{Z_{gk}}$
h_{11}	h_{12}	$\frac{Z_{ag}Z_{gk}}{Z_{ag}+Z_{gk}}$	$\frac{Z_{gk}}{Z_{ag}+Z_{gk}}$	$\frac{Z_aZ_{gk}}{(\mu'+1)Z_{gk}+Z_a}$	$\frac{Z_{gk}}{Z_{gk}(\mu'+1)+Z_a}$	$\frac{Z_{ag}Z_{gk}}{Z_{ag}+Z_{gk}}$	$\frac{Z_{ag}}{Z_{ag}+Z_{gk}}$
h_{21}	h_{22}	$\frac{(\mu'Z_{ag}-Z_a)Z_{gk}}{(Z_{ag}+Z_{gk})Z_a}$	$\frac{\Sigma}{Z_a(Z_{ag}+Z_{gk})}$	$-\frac{(\mu'+1)Z_{gk}}{(\mu'+1)Z_{gk}+Z_a}$	$\frac{\Sigma}{Z_{ag}[Z_{gk}(\mu'+1)+Z_a]}$	$-\frac{(\mu'Z_{gk}+Z_a)Z_{ag}}{(Z_{ag}+Z_{gk})Z_a}$	$\frac{\Sigma}{Z_a(Z_{ag}+Z_{gk})}$

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A Very High Speed Precision Tachometer

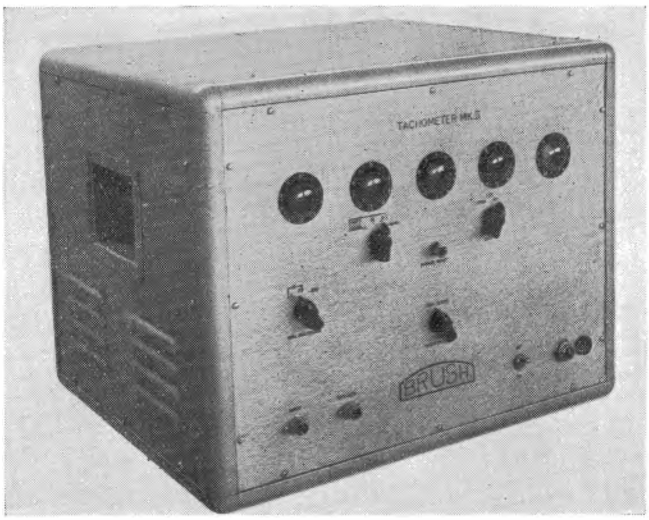
By J. K. Goodwin*, A.Inst.E.

The development of new high speed counting tubes prompted the design of this instrument. Although it is used mainly as a tachometer it may also be used as a general purpose frequency measuring instrument of high accuracy.

IN 1954 Bland and Cooper[1] described an integrating type of tachometer capable of reading up to 80 000rev/min in two decade steps.

After gaining considerable experience with this tachometer it was decided to develop another similar instrument which would be portable and as small as possible consistent with ease of servicing. The first tachometer was housed in a console as part of an installation.

Since the first tachometer was built new counter tubes have been developed having a maximum counting rate of 20 000p/s. The instrument described in this article will count up to 40 000rev/min with the extra decade step giving 400 000rev/min. The tachometer has been designed bearing in mind that it is possible to use the instrument for numerous applications. The instrument may also be used as a counter, counting up to 20 000p/s to an accuracy of ±1p/s or 200 000 ±10p/s and for counting random pulses over any time period up to a maximum number of 999 999. An external timing circuit can be used with the instrument to provide counting for periods other than the 1 and 2sec provided by the internal timing source. The circuit operation is still exactly as stated for this latter use. The accuracy limitation is inherent in this type of counting equipment because of the necessity for sampling the input signal by means of some form of gate.



The complete tachometer

Basic Operation

The overall circuit operation is very similar to the previous tachometer. Briefly the operation is as follows: input pulses are applied to a gating pentode, controlled by an accurately known timing source. The input pulses from the gate are passed to a series of Dekatron counting tubes. After the pulses have been displayed on the counting tubes for a certain time, the Dekatron 'glows' are returned to zero ready for the action to begin again. The circuit will be described from the block schematic diagram shown in Fig. 1.

INPUT AND LIMITING STAGE (Fig. 2)

The requirements of the input circuit are as follows: the input voltage shall be reasonably low in order that the instrument will be versatile and not normally require external pre-amplifiers. The input impedance must be high over the entire frequency range covered by the instrument.

A double triode (V_8) is used for the input amplifiers, one half as a class-A amplifier and the other half as a cathode-follower. The overall response of V_8 is 3dB down at 1·6c/s and 180kc/s. A minimum input of 1·5V peak-to-peak is required to operate the counter. The maximum input voltage should be limited to 10V. R_{43} is included in the circuit in order that the signal source is not loaded should V_{8a} be driven into grid current by excessive input voltage.

V_{8a} is *RC* coupled to V_{8b} and operated under near cut-off conditions as a cathode-follower. The output from the cathode-follower is positive going and feeds a Schmidt trigger which produces a rectangular wave with a very short rise-time.

In order to produce a rectangular waveform of very short rise-time it is necessary that the trigger valve is driven hard into its unstable state, that is grid current is drawn by V_{9a} and by V_{9b} during the period when the circuit is changing from its unstable to its stable state. The action of the circuit is as follows: in the stable state it is arranged so that V_{9a} is cut off by the voltage developed across R_5 and V_{9b} is conducting.

Resistors R_{50} and R_{51} form a potential divider so that when V_{9a} is cut off then the grid of V_{9b} is at cathode potential and taking a small amount of grid current.

As the cathode potential of V_{8b} rises it takes the grid of V_{9a} with it and at some point during this rise the grid potential of V_{9a} becomes sufficiently positive to allow V_{9a} to conduct; as V_{9a} begins to conduct the voltage across R_{48} increases and the grid voltage of V_{9b} falls. The action is cumulative and continues until V_{9b} can no longer conduct and it becomes cut-off and V_{9a} is conducting.

It will be clear from Fig. 3 and the circuit description that the width of the output wave from the trigger will vary with the input pulse width and amplitude. This is of no importance however, since the output from the trigger is differentiated by $C_{26}R_{53}$ and the negative going pulse is removed by the shunt diode across R_{53}. Then the output is fed directly to the gating valve; or the output is fed via C_{25} and R_{56} when the trigger output is connected to the scale of ten divider. In the latter case, however, the negative pulses are used to trigger an Eccles-Jordan divider, the positive pulses being removed by the diodes in the binary dividers.

The rise-time of the rectangular wave from the trigger is approximately 1µsec. By switch selection the output from the trigger can be fed to either the scale of ten divider or direct to the gating valve.

SCALE OF TEN DIVIDER

The divider consists of four Eccles-Jordan binary

* *The Brush Electrical Engineering Company Limited.*

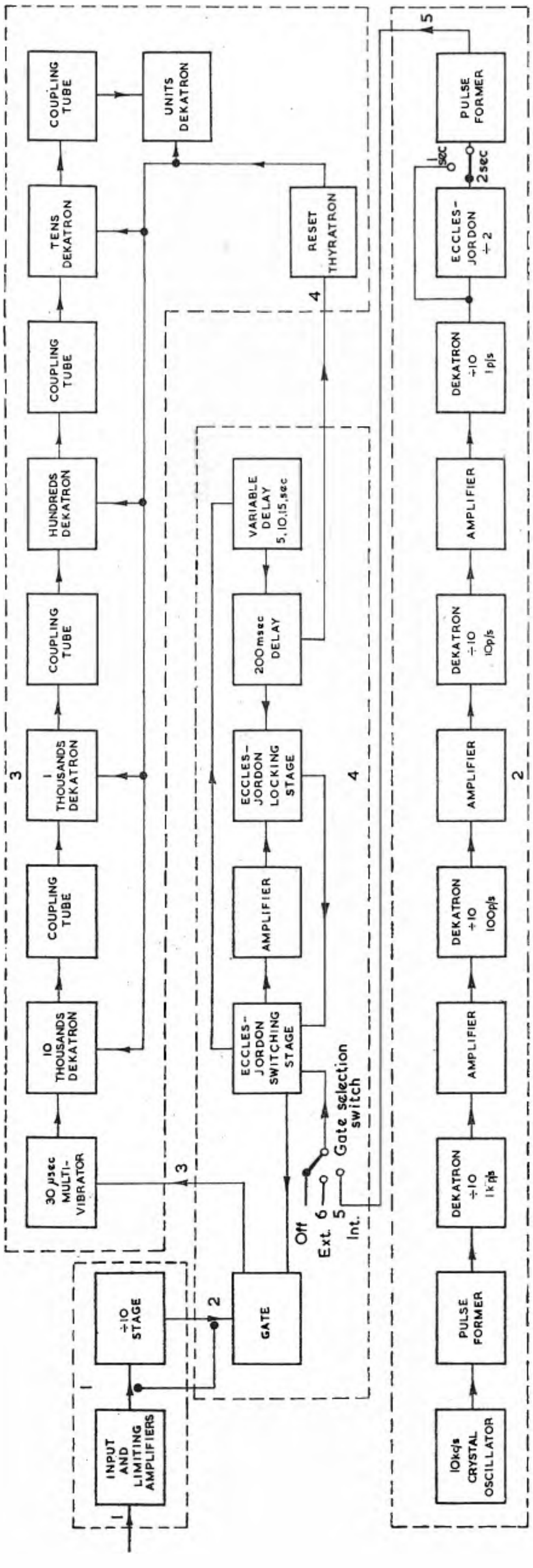


Fig. 1. General arrangement of the tachometer

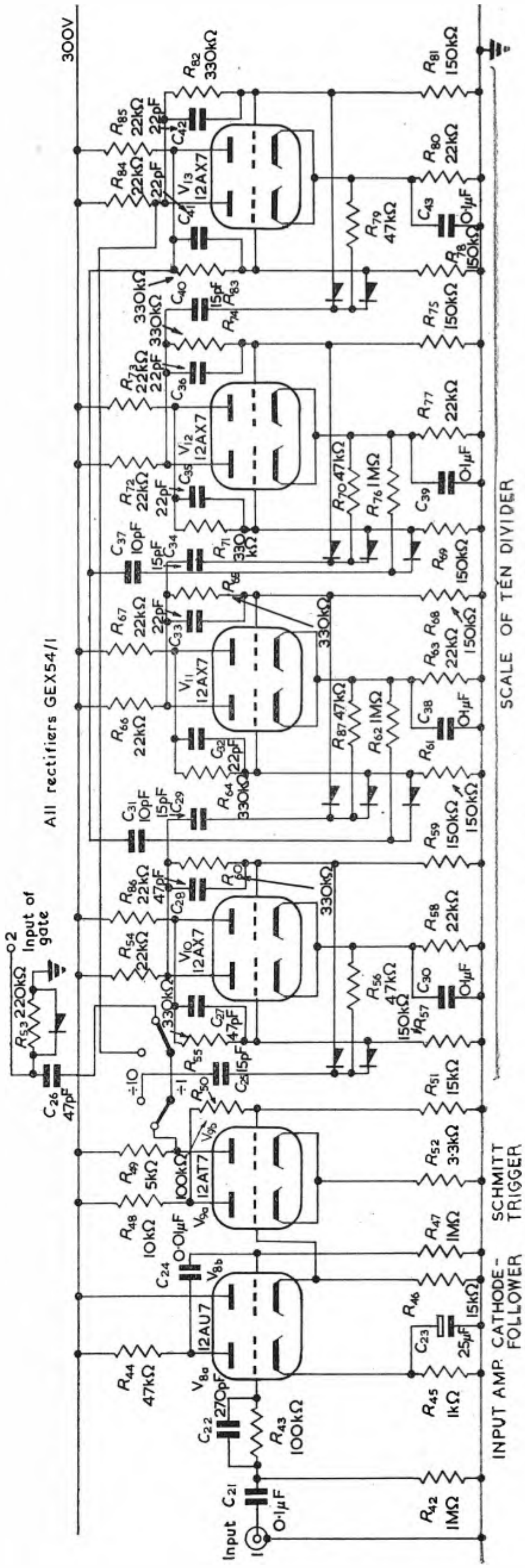


Fig. 2. Input and limiting stage—scale of ten divider

dividers connected in cascade, this arrangement gives a division ratio of 16. By suitable feedback loops the division ratio is reduced to 10. The circuit will operate up to a maximum frequency of 500kc/s. When the divider is operating correctly the operating code is as follows:

(0)	0000	(5)	0101
(1)	0001	(6)	0110
(2)	0010	(7)	0111
(3)	0011	(8)	1110
(4)	0100	(9)	1111

Where the left-hand digits represent the number of input

Fig. 3. Grid and anode waveforms of V_9

pulses and the right-hand row represent the valve or valves which are conducting for a given number of input pulses.

From the input and limiting stage and the scale of ten divider there is now a series of pulses corresponding to either the input signal or one-tenth of the input which can be fed into further stages of the tachometer

Timing Panel (Fig. 4)

The purpose of this unit is to provide accurate timing pulses for operating the gate.

The accuracy of the whole instrument depends on this unit and it is of paramount importance that it is unfailing in its accuracy and reliability.

The whole unit has been completely re-designed, a 10kc/s crystal is used as the frequency standard instead of a 100kc/s crystal which was previously used. This simplifies the divider considerably and reduces the number of valves used in the unit.

A variable-mu pentode (V_{14}) is used as the drive source for the crystal. The output from the valve is amplitude controlled by means of a voltage developed across a crystal diode. The stage is operated at a very low voltage (50V) as well as having the amplitude control to prevent the crystal being overdriven and damaged. C_{44} is used to 'pull' the crystal to an accurate 10kc/s; this is necessary since the crystal is cut to an accuracy of 0·005 per cent and this tolerance would cause inaccuracies particularly at high frequencies. The crystal frequency being set up with the aid of a valve maintained tuning fork.

The output from the crystal oscillator stage is fed into the cathode-follower to prevent loading of the oscillator. This cathode-follower drives an amplifier which is normally conducting hard. Positive going pulses from the anode of V_{15b} cause V_{16a} grid to rise and V_{16a} conducts. Negative going pulses from this valve drive the first Dekatron divider, a high speed type Ericsson GC10D dividing from 10kp/s to 1kp/s. Three sub-miniature Dekatron tubes reduce the 1kp/s pulses to 100p/s, 10p/s and 1p/s. In the future it is hoped to replace the hard valve coupling between the GC10D's by trigger tubes to reduce h.t. consumption and heater current.

Pulses of 1sec duration from the last divider (V_{23}) are fed either direct to a one-shot multivibrator (V_{25}) pulse forming stage, or through an Eccles-Jordan divider (V_{24}) and then to the multivibrator. Thus pulses of either 1 or 2sec repetition frequency at a constant pulse width are available. This unit will provide a very accurate frequency standard which may be used for other purposes in the laboratory.

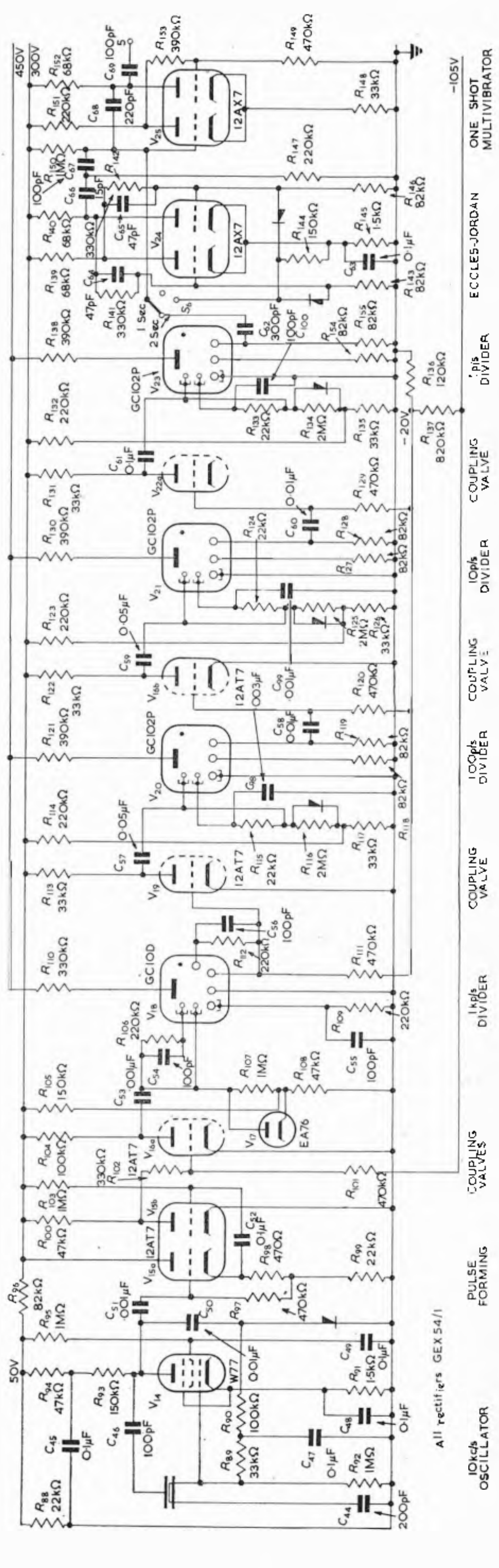


Fig. 4. Timing panel

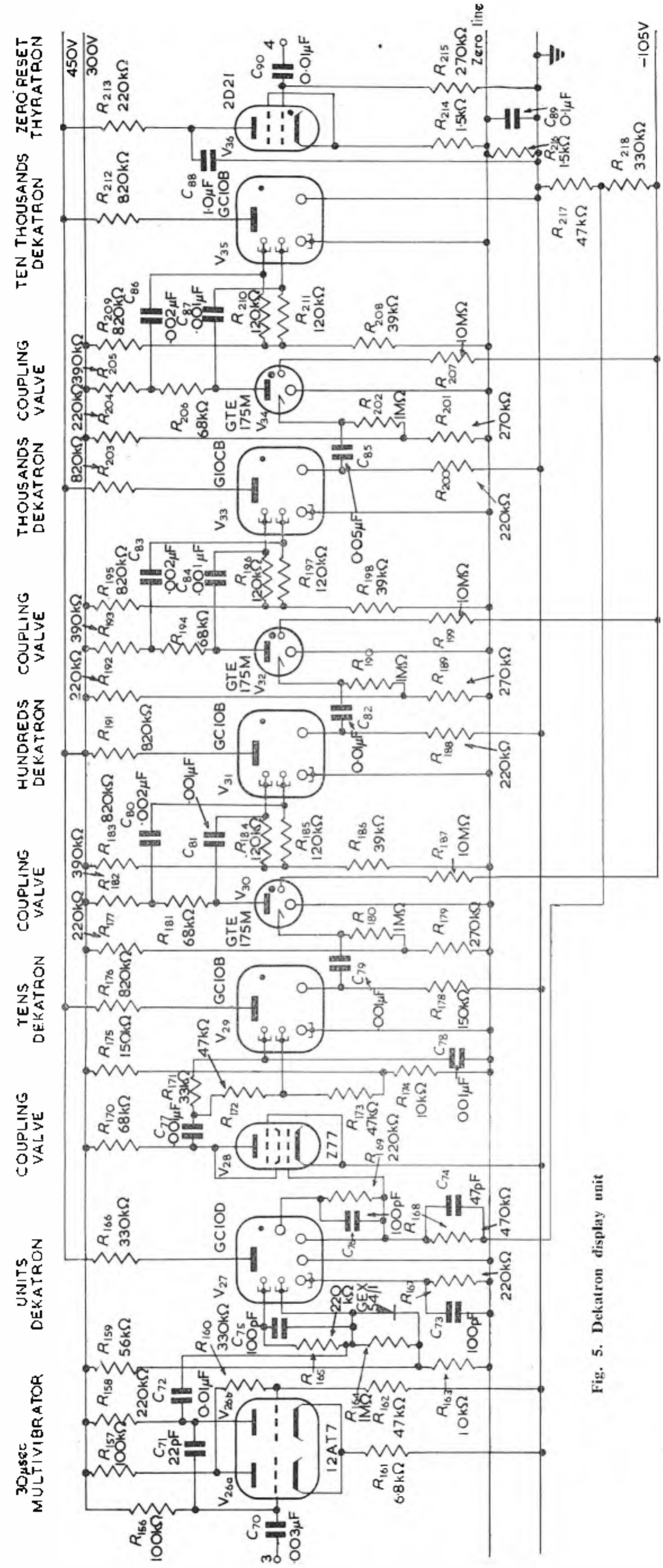


Fig. 5. Dekatron display unit

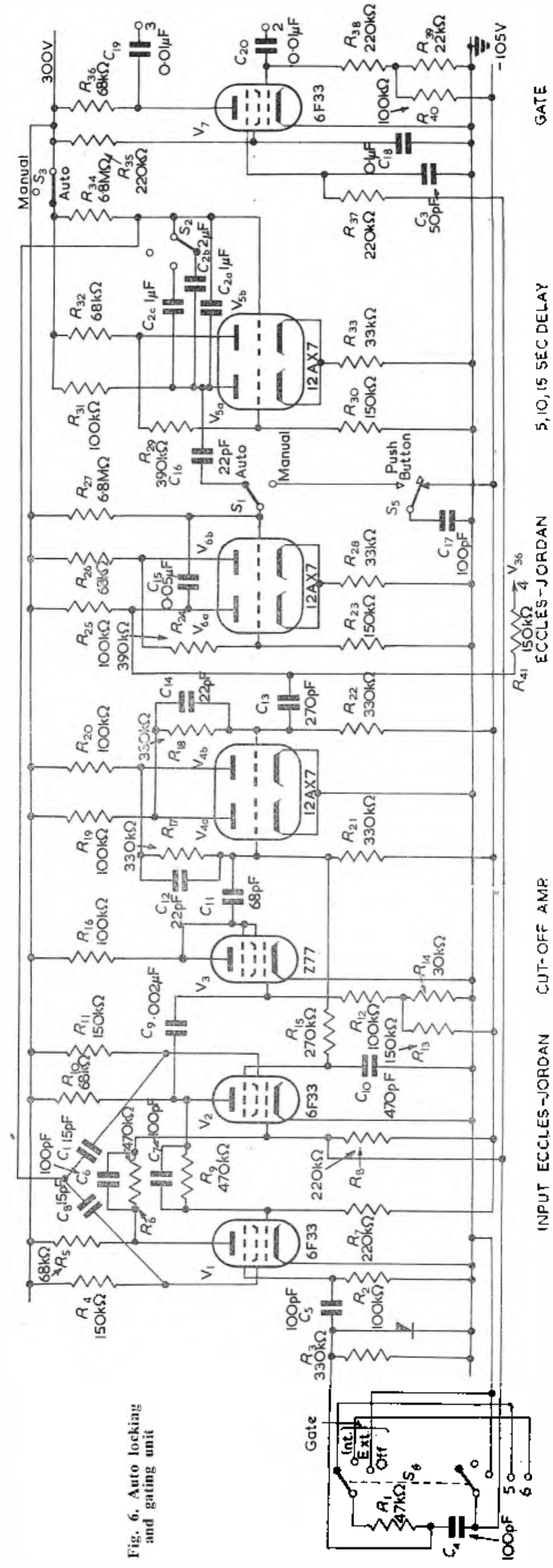


Fig. 6. Auto locking and gating unit

DEKATRON DISPLAY UNIT (Fig. 5)

The display unit has five Dekatron tubes arranged in cascade. When the timing period is 2sec then it is possible to count to 40 000. The limit being set by the maximum speed of the first Dekatron. Negative going pulses from the anode of the gating valve are fed to a 30μsec one-shot multivibrator. The anode of V_{26} will go negative for the 30μsec period, this pulse is fed into a GC10D high speed Dekatron. The maximum operating frequency for this tube is stated to be 2 000p/s which represents a periodic time of 50μsec, therefore the pulse forming multivibrator (V_{26}) must produce a pulse width of less than 50μsec in order to secure reliable running at a maximum speed.

The final cathode of the GC10D (V_{27}) is fed to a triode connected Z77 (V_{28}) biased to cut-off. The output from V_{28} which consists of 140V negative-going pulses, with a maximum p.r.f .of 2 000, is fed into a GC10B Dekatron (V_{29}). This Dekatron is coupled to a third, (V_{31}) by a trigger tube V_{30} arranged to operate as a self quenching tube, so that it is extinguished after each input triggering pulse. The last two Dekatrons V_{33} and V_{35} are coupled by GTE175M tubes, V_{32} and V_{34}.

In order to return the Dekatrons to the zero position at the end of each counting and display period, it is necessary to make the 'zero line' positive with respect to earth for a short period. The guide electrodes of the Dekatrons are all returned to a common line, the zero line. By making this line positive all these electrodes are made positive with respect to the final cathode and the 'glows' are returned to these cathodes. A tetrode thyratron V_{36} (2D21) is normally biased to cut-off and with the application of a positive pulse to its control grid derived from the auto locking stage (see below), the thyratron conducts, discharging a large capacitor, C_{39}, in its anode circuit. A large pulse of current flows through R_{207} raising the zero line some 100V positive for 20msec, thus returning all the Dekatron 'glows' to the zero position.

Provision has been made to use another display unit remote from the main instrument. It consists of a unit identical to the display unit described above, with the exception of the re-setting thyratron. The thyratron (V_{39}) situated in the main counter is used to reset both units.

AUTO LOCKING AND GATING UNIT (Fig. 6)

In order that the instrument can count a series of pulses for a given time, some form of gating circuit must be used. After the counting period, in this case either 1 or 2sec, the gate must be closed and remain closed for a sufficient time to enable the display to be read. When the display period is complete the Dekatrons must be returned to the zero position, so that the action can be repeated.

The display time is made a variable period of either 5, 10 or 15sec (S_2) to suit different operators and uses to which the instrument may be put. A push-button is also fitted so that instantaneous counts can be made. So that the push-button can be used the delay circuit controlling the display time must be switched out of circuit (S_3) thus, after the count has been taken, the display will remain on the display tubes until the push-button is again pressed when the tubes will be returned to zero and automatically

TABLE 1

VALVE	INITIAL STATES	AFTER 1ST GATING PULSE	AFTER 2ND GATING PULSE	AFTER DELAY PERIOD OF 5, 10 OR 15 SEC	AFTER 200MSEC DELAY TIME FOR ZEROING DISPLAY
V_1	ON	OFF Positive pulse through C_8 does not operate V_5 but ensures it is in its correct state	ON	ON Stages locked by bias on V_2 suppressor from V_{4a}	ON
V_2	OFF Gate shut	ON Gate open	OFF Screen falls, negative pulse through C_1 changes V_5 to its unstable state	OFF	OFF
V_3	OFF	OFF	ON Pulsed	OFF	OFF
V_{4a}	ON	ON	OFF V_3 suppressor biased to cut-off by bias on V_{4a} grid	OFF	ON Removes locking bias from suppressor of V_2
V_{4b}	OFF	OFF	ON	ON	OFF
V_{5a}	OFF	OFF	ON Delay period	OFF Display period completed	OFF
V_{5b}	ON	ON	OFF Ditto	ON Ditto	ON
V_{6a}	OFF	OFF	OFF	ON 200msec delay V_{6a} anode instantly rises firing the re-setting thyratron	OFF
V_{6b}	ON	ON	ON	OFF Stage initiated by a negative pulse from V_{5a} anode	ON
V_7	OFF Controlled by V_2	ON	OFF Gate shut	OFF	OFF

a new count is taken. The push-button S_3 is normally in the lower position as shown (Fig. 6) so that capacitor C_{17} is charged to -105V. When the push-button is pressed V_{6D} grids are driven negative starting the action of the 200msec delay and the circuits associated with it.

The unit consists of two pentodes (V_1 and V_2) connected as an Eccles-Jordan trigger, the gating pulses from the timing unit initiate the circuit action on the suppressor of V_1 and the control grid of V_2. V_1 and V_2 are short suppressor base pentodes type 6F33.

S_4 selects the gating pulses, these can be either internal (INT) or external (EXT) or, in the 'OFF' position, the gate is locked open so that random pulses can be counted over an indefinite period of time.

With the switch S_4 in the INT position positive pulses are fed through R_1 to limit the peak current through the germanium diode shunting R_2, this diode removes the negative going pulses produced by the differentiating action of C_{60} and R_3.

V_3 is biased to cut-off coupling V_2 to V_4, an Eccles-Jordan trigger is used to lock V_2 in the off or non-conducting state.

V_5 and V_6 are two one-shot multivibrators connected in series. V_5 is a variable delay valve providing the delay of 5, 10 or 15sec for the reading time of the display. The actual delay time for the reading depends on the interval between the timing pulses. The multivibrator V_5 is actually set for 3, 8 and 13sec so that in the 2sec timing period the actual delay is 5, 10 and 15sec; while in the case of the 1sec timing period it is 4, 9 and 14sec.

V_6 provides a 200msec delay during which time the Dekatrons in the display unit are returning to zero. After this 200msec period V_4 is switched over and the cut-off bias on the suppressor of V_2 is removed and the circuit is returned to a state ready for the next series of pulses to be counted. The complete action of the unit is shown in Table 1.

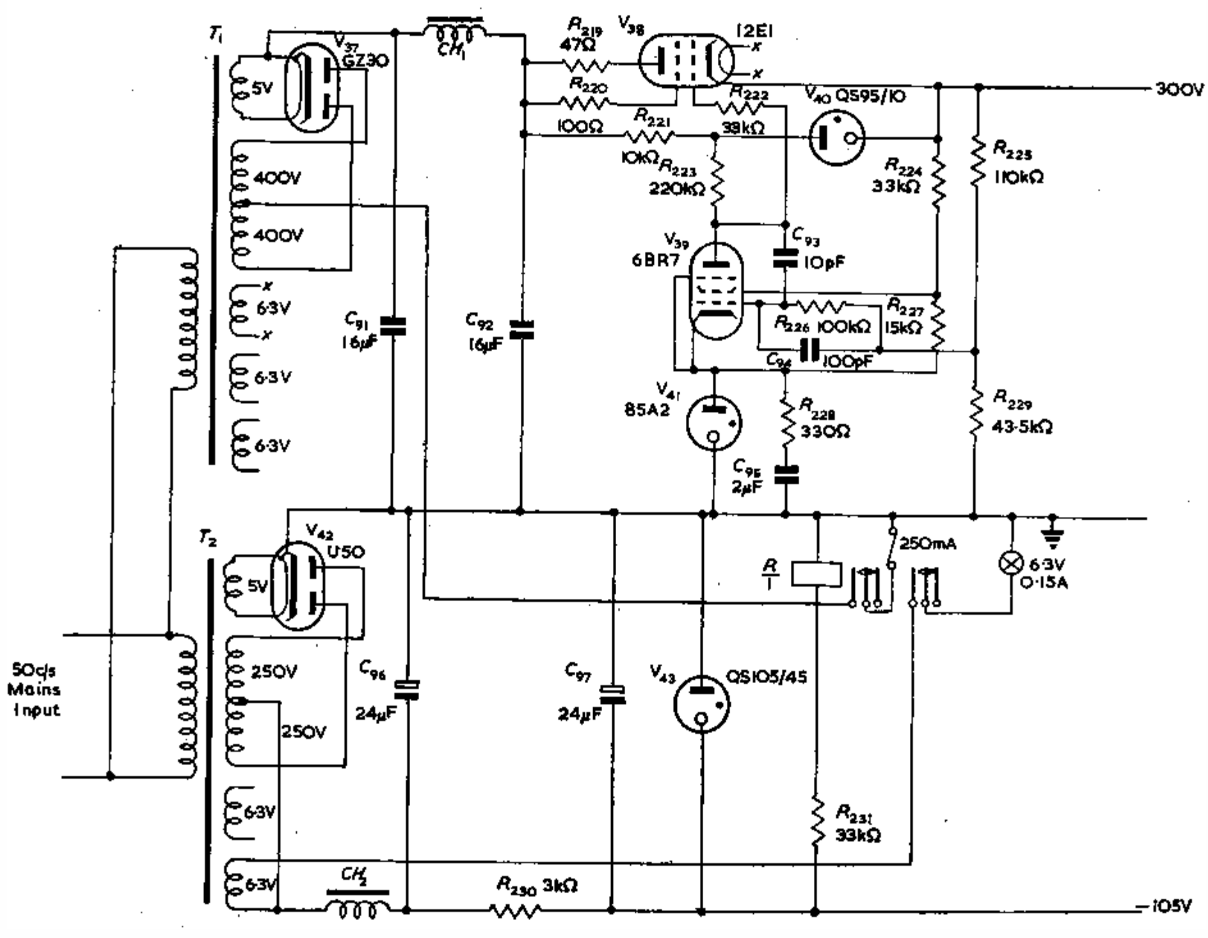
Pentode Gate

A short suppressor base pentode is used for the gating valve. The operation of the gate is controlled by the suppressor grid which is normally biased to cut off; the bias being produced by the control grid of V_2. When V_1 and V_2 are changed over by the application of the first timing pulse V_2 control grid rises to earth potential and the bias is removed from the suppressor of V_7. The valve can now operate as a normal amplifier.

Input pulses from either the input limiting amplifiers, or from the scale of ten divider are fed through the blocking capacitor C_{20} to the control grid of V_7. The control grid is biased beyond cut-off for two reasons, firstly to prevent a pedestal being formed in the anode circuit as the suppressor bias is removed, and secondly, to suppress spurious pulses from the scale of ten divider. The spurious pulses are produced as a result of suppressing the division ratio from 16 to 10 by the feedback loops. These pulses are accentuated by the differentiating action of C_{26},R_{53}. They are, however, of smaller amplitude than the main signal and can be effectively removed by biasing the control grid of V_7 beyond cut-off.

When the suppressor bias is removed the grid signals are passed to the anode circuit and hence to the display unit. After the timing period V_1 and V_2 change over and V_2 control grid is biased to cut-off and therefore so is V_7 suppressor and the control grid signals are blocked from the display unit.

Fig. 7. Power supply



Power Supply (Fig. 7)

The power supply is in two sections.

(1a) providing 300V at 100mA. (1b) 420 to 450V.

(2) -105V bias supply.

The 300V supply is stabilized against variations in mains supply voltage and has an output impedance of approximately 1Ω, so that no interstage decoupling is required. The 450V supply is unstabilized and is used to supply the anodes of the Dekatrons.

The 105V negative bias supply is stabilized by means of V_{42} (QS105/45). The bias rectifier is a directly heated type V_{43} (U50) so that the bias supply is available before the main h.t. supply to prevent damage to the valves in the main unit. A relay is also incorporated in the supply to guard against either a short-circuit line in the counter or a failure of the bias supply. In either case the relay will not operate and the main h.t. power supply cannot be switched on. An indicator lamp is connected across a pair of relay contacts so that when the bias is on the indicator lamp is lighted.

Performance and Operation

The reason for using a two second timing period may seem obscure without reference to Bland and Cooper[1].

In order to measure the speed of rotating machinery a slotted rotor having 30 teeth evenly spaced around the periphery rotates within a stator also having 30 slots cut into its inner edge. The rotor is coupled to the machine shaft: when the machine is rotating the capacitance change from the transducer frequency modulates an oscillator and the resulting frequency change is converted by means of a discriminator, to voltage pulses corresponding to 30 per revolution. Hence by counting over a two second period a direct reading in revolutions per minute is obtained. Although the chance of requiring to measure above 40 000rev/min is remote the extra decade multiplier has been found very useful for general laboratory work.

The instrument was constructed as a number of plug-in units. Each unit is as shown in the theoretical diagrams and fits into a framework inside the enclosing cabinet. The overall size of the cabinet being 10in high, 14in long and 8in deep. Ventilation being by convection only.

Acknowledgments

The author wishes to acknowledge the work of Mr. B. J. Cooper on the Timing Panel, and his preliminary work on the Auto Locking Unit. Also Mr. J. Blanchard who made the final assembly of the instrument, introduced several modifications and carried out extensive tests on the tachometer.

The author also wishes to express his thanks to the Brush Electrical Engineering Company Limited for permission to publish this article and to his colleagues for help and constructive criticism during the design and development of the instrument and the preparation of this article.

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An Internal Defibrillator with Current Measuring Facilities

By B. J. Perry*, B.Sc., Ph.D., A.R.C.S., and
R. E. Trotman†, B.Sc., Grad.Inst.P.

The principle of giving a substantial shock to the exposed heart to stop ventricular fibrillation has been known for some time[1,2]. Originally, short pulses of current of up to 5A r.m.s. were delivered through the heart by direct connexion to the 115V a.c. power supply. Various improvements and additions to the basic method have since been made and are included in this design for the Westminster Hospital Group.

Basic Circuit

The diagram of the basic circuit is shown in Fig. 1. Current pulses are derived from the secondary of the isolating transformer T and are controlled in length by a foot, hand or timer operated contact S. Current control is by the variable resistor R_1, there being a small value fixed resistor R_2 in the secondary circuit at all times. The secondary voltage of the transformer is 120V r.m.s. and the transformer is rated to give an output of one pulse of 0·2sec duration every second to a maximum r.m.s. value of 5A.

As an alternative to providing a variable resistor R_1 to vary the average current during a pulse, the primary of the isolating transformer can be fed from a variac or similar device or the secondary of the transformer can be provided with a series of voltage tappings and the pulse current is varied by tap switching[3].

Current Measurement

The average r.m.s. current passed during a pulse can be measured on a conventional moving-iron or moving-coil ammeter if the length of pulse is sufficient[1,2]. The usual pulse length is, however, of the order of 0·2sec and an ammeter is unsatisfactory. A simple valve memory circuit has, therefore, been devised which measures the peak value of current

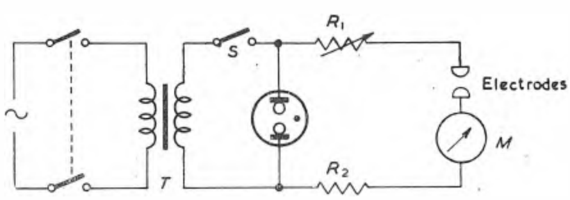

Fig. 1. The basic circuit

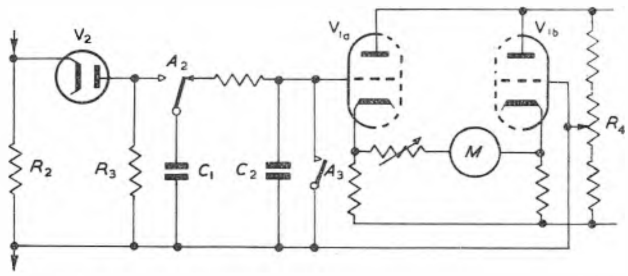

Fig. 2. The principle of the current measuring memory circuit

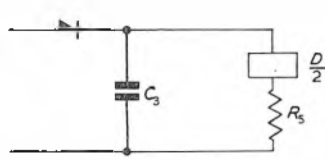

Fig. 3. The timing device

during a pulse and is calibrated in terms of the r.m.s. value of current passed during the pulse.

During a pulse the capacitor C_1 (Fig. 2) is charged through the diode V_2 and the contacts A_2 from the voltage developed across R_2. Contact A_3 shorts C_2 ensuring that no charge remains on it. At the end of a pulse contacts A_2 and

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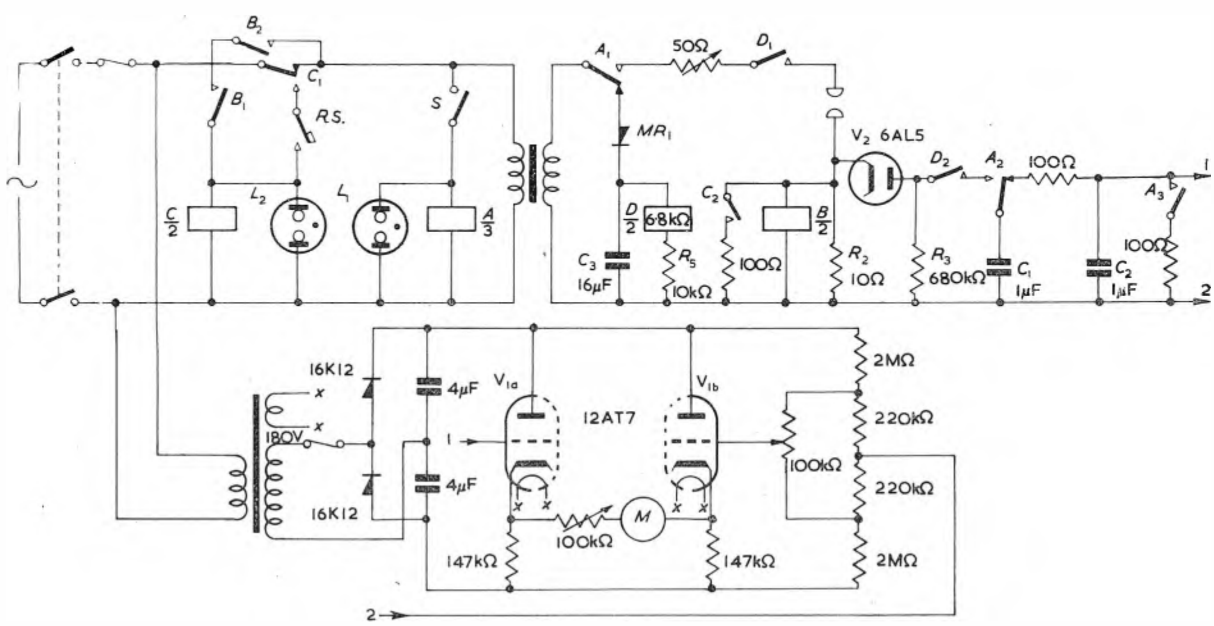

Fig. 4. The complete circuit diagram of the instrument

A_3 are released thus allowing the charge on C_1 to be shared with C_2. The resultant voltage across C_1 and C_2 is measured by the valve-voltmeter formed by V_{1a} and V_{1b} and meter M. The meter M (1mA f.s.d.) thus gives a measure of the peak current of the pulse, the meter reading being adjusted to give a full scale deflexion for a current of 5A r.m.s. through R_2. R_3 provides a diode load and also allows some of the residual charge on C_1 to leak away during a pulse if the peak current of the pulse is much lower than the peak current of the previous one.

The meter gives an indication of pulse current imme-

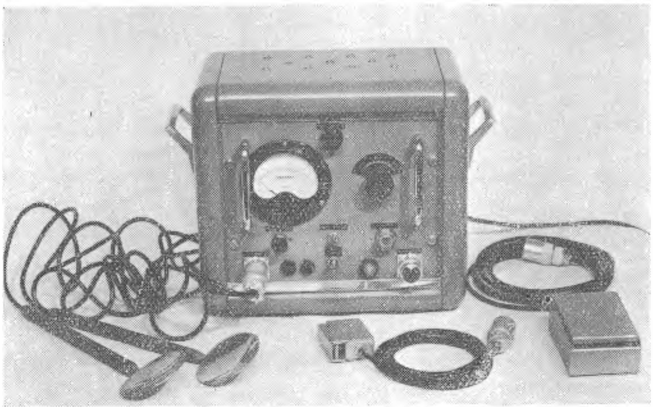

Fig. 5. The internal defibrillator and accessories

diately after the pulse has occurred and will hold this reading for some time due to the large capacitance of C_1 and C_2 (now in parallel) and the low value of grid current of V_{1a}. About 4 per cent of the full scale reading is lost in 10min. The zero stability is good due chiefly to its balanced arrangement and the high level of output voltage corresponding to full scale deflexion. If zero adjustment is necessary it is achieved by altering the setting of potentiometer R_1.

Timing of Pulse

The optimum time of pulse required is 0·2sec and it is almost impossible to judge that order of time accurately when using a hand or foot switch. It is therefore desirable to have some sort of timing device. The time of application may be controlled as here by an RC circuit (Fig. 3), or by a slugged relay[4] or a clock controlled switch, depending on the length of time required.

The Complete Circuit

The complete circuit of the defibrillator is shown in (Fig. 4). The foot or hand switch S operates relay A. Contact A_1 completes the secondary circuit and the current passes to the patient. Contacts A_2 and A_3 operate as before. The length of pulse is defined by relay D. This is normally on due to the connexion via the back contact of changeover contacts A_1. On pressing switch S the charge on capacitor C_3 begins to decay and the time-constant C_3 $(R_5 + D)$ is arranged so that relay D remains on for 200msec after relay A operates. When relay D releases, contact D_1 opens the transformer secondary circuit and contact D_2 prevents C_1 discharging through R_3 during the time that relay D is released, but A_2 not yet opened.

Relay B is an overload relay adjusted to operate at a pulse current of 5A. The operation of this relay brings on the locking relay C. The contacts of relays B and C are arranged so that relay C is fully held on before B is released and the primary circuit of the transformer opened. This arrangement ensures that there is no contact race and is a precaution against relay chatter and premature and uncertain operation. Reset switch RS prepares the circuit for immediate use.

The neon indicator L_1 lights whenever a pulse is passed and L_2 indicates an overload.

The whole apparatus is shown in Fig. 5 including the large curved electrodes and both a hand and foot switch. These accessories are attached to the instrument by push lock plugs and sockets.

Acknowledgments

Acknowledgment must be made to Dr. C. W. Wilson in whose department this instrument was developed and to Mr. David McNish for his constructional help. The apparatus is covered by British Patent No. 16505.

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VIBRATO CIRCUITS

for Electrical Musical Instruments

By Alan Douglas, M.I.R.E., M.Amer.I.E.E.

The judicious use of vibrato enhances the value of many musical sounds. In this article it is shown how electrical or mechanical devices attempt to provide some, at least, of the characteristics of this embellishment.

A GOOD deal of misconception exists as to what is meant by vibrato or tremolo. There is further confusion because the stop causing this effect is called a tremulant on a pipe organ. From a strictly musical point of view, vibrato means a form of frequency modulation while tremolo is not a modulation at all, but a rapid reiteration of a note without any change in pitch and without specific reference to time. This is often accomplished on the violin, for example, by rapid up and down bowing.

The tremulant on an organ is a corruption of the word 'tremblant', introduced by the French who were the first to apply a mechanism to cause a regular cyclic change in wind pressure. In present day practice, the apparatus produces both frequency and intensity changes; it is a true vibrato.

Therefore, although some writers refer to tremolo as an intensity modulating system for an electronic sound source, it does not produce a true musical tremolo which may be—and often is—a series of quite discontinuous notes.

It is evident that tone sources, the frequency of which is not readily varied, such as vibrating reeds and constant-speed rotating generators, may have to put up with a form of tremolo if a waver in the sound is required. This is a thoroughly bad thing; it never occurs in natural sounds and while at first it may seem to soften the effect of a sustained note, the invariable rate of repetition and the constant depth of switching soon takes control of the sound which becomes domineering and aggressive because of the monotony.

We can disregard this type of level modulation and concentrate on vibrato. This has its origin in emotional or biological tremors and is a musical asset of the greatest value when properly applied. Indeed no orchestral or vocal music can exist without some contribution from vibrato. Too much space would be occupied if the whole process of the pschycological evolution of vibrato was analysed[1], but the following should be noted as essential attributes—

(1) For preference, both frequency and intensity should be altered.

(2) The rate of vibrato is roughly proportional to the frequency of the noted played.

(3) The intensity should be related to the loudness of the note.

(4) Powerful sound sources do not lend themselves to the application of vibrato.

The foregoing are equally applicable to the singing voice, although this does not concern us. The full beauty of vibrato is exemplified by a solo violincello, the greatest abuse by the cinema organ played 'all out' with tremulants

Related to vibrato but quite separate is the choir effect due to many sounds of constantly changing phase heard simultaneously; and the voix celestes and unda maris, a strictly pipe organ effect employing two complete ranks of pipes having each pair (sounding the same note) tuned to give a beat of one or two per second with each other. These elusive and romantic effects cannot be synthesized electrically, or only at very great expense, although they are to some extent present in electronic organs having quite independent oscillators for each note; such instruments have therefore a greater richness of tone quality than organs using frequency division or mechanically linked generators.

We have, then, to make the best of frequency or phase modulation in one form or another. Great though the value of vibrato is to orchestral or vocal music, it is of even more value with electronic tone generators. These have (except for certain very advanced monophonic instruments) a steady waveform and a steady harmonic content for any one note at any instant. Vibrato smooths the harmonics and modulates the waveform, thus greatly modifying the tonal presentation. Against this must be offset the fact that an electrical vibrato is not progressively and continually adjustable, except in certain instances[2,3]; therefore caution is required in applying this effect, but this would be expected from a musical standpoint.

It is conceded that a modulation rate between 5 and 8c/s is most agreeable, and that vibrato should not be applied to notes much below 130c/s. This is because of the dissonance associated with low-frequency beats[4].

There is no general agreement over the depth of vibrato; since with a tremolo there is no deviation from pitch, this may be very intense; but with the frequency excursions due to vibrato, too great a swing is unpleasant. Occasionally this may be noticed in vocal music. It would appear that about 10 cents or one tenth of a semitone is a reasonable pitch deviation, and if this is not proportional to the accompanying changes in level, the latter can be quite considerable; but a true frequency modulation can be effective if there is no change in level at all. It all depends on the application.

Taking the simplest case first, most forms of multivibrator are susceptible to frequency change if the h.t. line voltage is varied. Fig. 1(a) shows such a circuit, suitable for a monophonic instrument. If the anode supply to one half of V_2 is fed through the potential divider formed by R_1 and R_2 in the anode circuit of a low-frequency oscillator, the supply to R_3 undergoes a cyclic variation and vibrato is produced at the output C_1. The depth of modulation is adjusted by varying the ratio of R_1 to R_2, but in such a circuit R_2 forms part of the frequency determining circuit of V_2 and hence cannot be altered to change the degree of vibrato. If this condition is required, the vibrato oscillator must be isolated from the tone generator by a valve as is Fig. 1(b). In this instance the cathode current is modulated but the vibrato generator may be switched to several rates or degrees of modulation without affecting the tuning of the tone generator.

Oscillators used to generate the low-frequency vibrato voltage should have a symmetrical and smooth waveform. Phase shift circuits are very suitable. It will be appreciated that square and other steeply rising waveforms may themselves modify the waveform from the tone generator. A sine wave is desirable. One such circuit[5] which is particularly good is shown in Fig. 1(c). This has the advantage

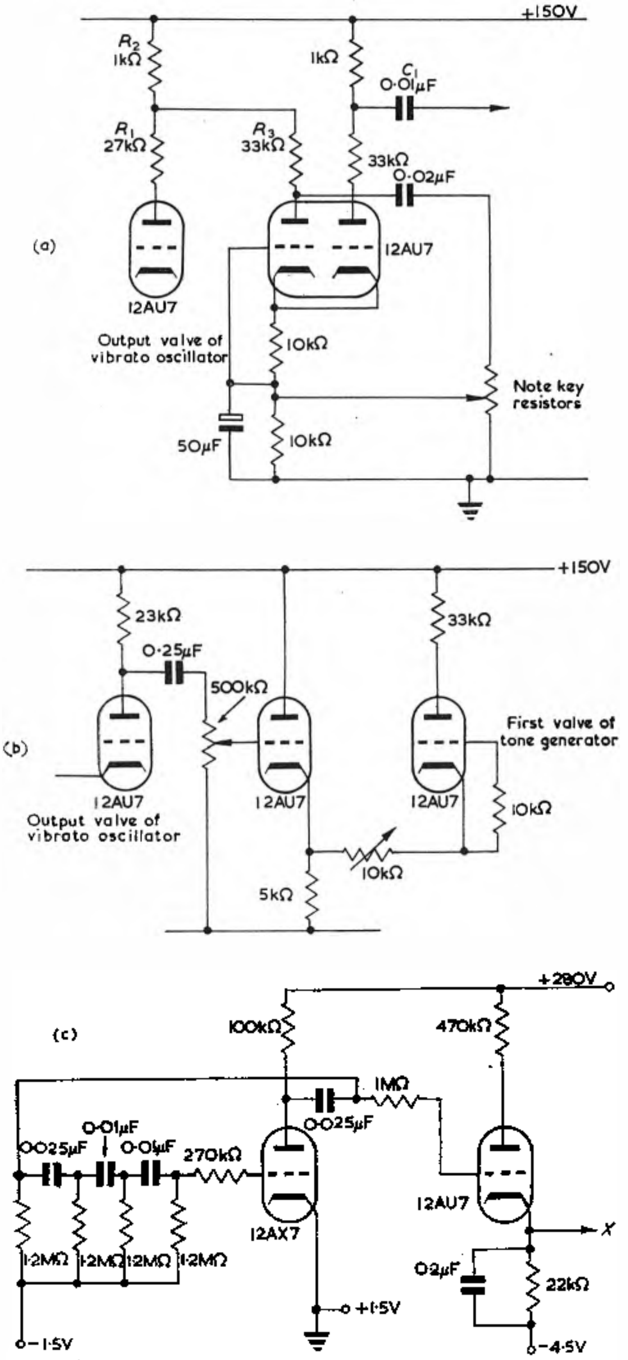

Fig. 1. (a) Tone multivibrator with vibrato
(b) Vibrato circuit with buffer valve
(c) Phase shift vibrato oscillator

that d.c. modulation can be carried out in the cathode circuit of V_2, so that grid bias control is available for other kinds of oscillator; for example that of Fig. 2.

This type of oscillator may be adjusted to be little influenced by normal changes in h.t. line voltage, so that grid modulation is the most effective method of applying vibrato. If the grid resistors are connected to the switch at the point marked X in Fig. 1(c), vibrato may readily be started or stopped without any change in the steady frequency.

Such a system can be used in a musical instrument employing frequency dividers, since the dividers are locked too tightly to be directly influenced by vibrato circuits. It does, however, lead to the condition common to most electrical musical instruments that the one vibrato frequency is applied over the whole compass. This is not compatible with the effect as employed by orchestral players, where the rate of vibrato varies with the frequency. As is well known, the multiple phase changes brought about by superimposing the many vibratos leads to an extraordinary richness of tone. It is evident that one cannot economically have an infinite number of such phase variations, although it is quite feasible to provide vibrato generators for each

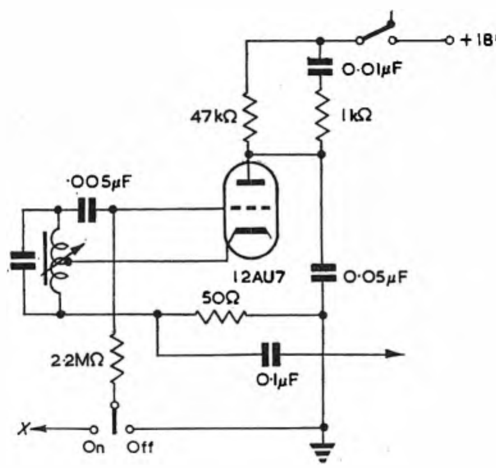

Fig. 2. LC tone oscillator with vibrato switch

octave from tenor C upwards; each oscillating at a slightly different frequency.

A considerable step forward was taken by Hanert in designing the line vibrato circuit[6] of Fig. 3. While the delay from the static line does vary linearly with frequency, it is further modulated by a scanning element in the form of a variable capacitor having 16 fixed sections connected to selected parts of the line, with a single set of rotating vanes operating at 412 rev/min. The signal is retarded in phase by each line section and each group of sections is modulated by the rotary scanner so that the whole line is scanned in about 1/14sec. This rate of change causes about 1·5 per cent decrease in frequency and this is constantly low to that extent so long as the capacitor is rotating at a uniform speed. Switching arrangements permit of the whole, half and one third of the line being scanned at will, thus altering the depth of vibrato.

While the effect from this current is much more realistic than that from fixed frequency vibratos, the most interesting result arises from mixing some of the signal through the line with some of the signal before it has been applied to the delay line. An extremely complex system of small amplitude beats is thus set up, and some approximation to a true chorus is obtained; since no one can tell in advance the phase relationships of the generators at the instant of keying the signal. In the present state of the art, this arrangement is less tiring and much less cyclically evident than any other electrical vibrato system for polyphonic instruments.

As stated above, some tone systems cannot be frequency modulated owing to their inherently high stability and in

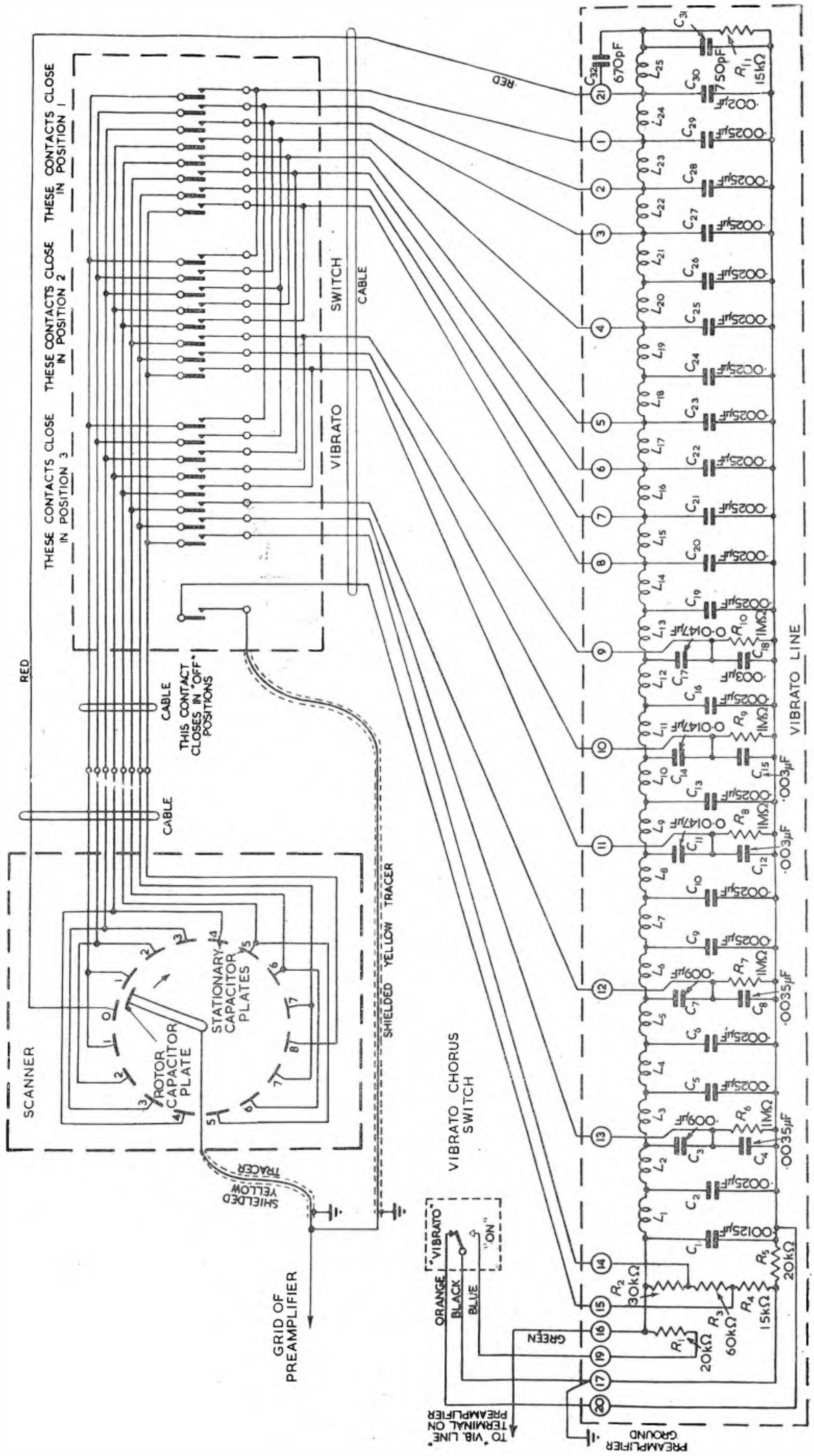

Fig. 3. Delay line vibrato circuit

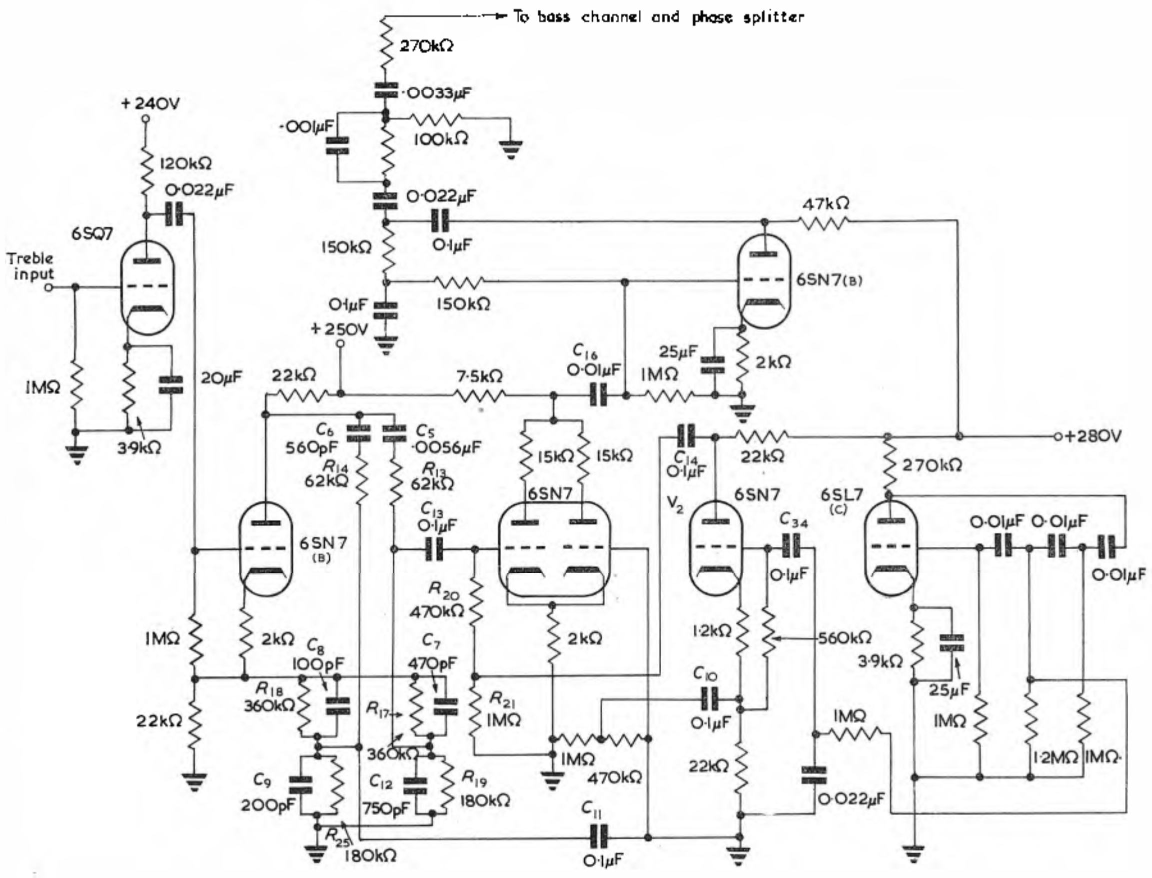

Fig. 4. Wurlitzer vibrato circuit

such cases a circuit which can be applied after the generator may be very useful. There are now many circuits for producing a phase shift proportional to frequency, but the arrangement used in the later Wurlitzer instruments is interesting. Free reeds do not appreciably change their pitch with a change in wind pressure, this being taken advantage of in the 'expression' stop in harmoniums and the Mustel reed organ. This stop disconnects the wind reservoirs and puts the reed chests in direct communication with the bellows. Thus by 'expressive' pedalling, the volume can be regulated from a whisper to full, without change in pitch; often to be heard from the Mustel reed organ in the Palm Court Orchestra broadcasts.

In Fig. 4, which shows the Wurlitzer circuit, the signal is first divided so that notes below 130c/s are not modulated. With vibrato, the treble signal passes through the input valve to half of a 6SN7 connected as a phase invertor. Again, the out of phase signals from the anode and cathode divide between the networks $C_5 R_{13} C_6 R_{14}$; and $C_7 R_{17} C_8 R_{18}$. These four circuits are connected by $C_9 R_{25} C_{12} R_{19}$ to form two output circuits. The phase-angle between each of these networks and the input voltage increases almost linearly with frequency between 500c/s and 15kc/s. The outputs of these networks come through $C_{13} C_{11}$ to the grid of the 6SN7 mixer valve, 90° apart.

The phase shift oscillator using half a 6SL7 (c) supplies V_2 through C_{34}. This is another phase invertor, the outputs of which are connected through $C_{10} R_{21} C_{14} R_{20}$ to the grids of the 6SN7 mixer. The low-frequency signal thus controls the conductivity of the two sections of this 6SN7. When section (1) is conducting, section (2) is nearly cut off. The output signals are thus alternately delayed by about 90°. This shifting to and fro between the two signals produces an apparent frequency change to the ear. The output from the mixer is then fed to one half of the 6SN7 (B) feedback filter stage which cuts off sharply below 130c/s. This is to prevent the phase-shift oscillator frequency from appearing as part of the musical signal. The modulated signal then joins the bass amplifier and so proceeds to the output stage.

It has often been suggested that a mechanical modulator be used at the actual loudspeakers. Such arrangements can be very effective if the mechanism can be so arranged that it does not interfere with the free sound field when not in use. Any such scheme must be based on the Doppler effect, the frequency f at the point of observation being

$$= \frac{V}{V - V_s} \cdot f_s$$

where V = velocity of sound in the medium

V_s = velocity of the source

f_s = frequency of the source.

The sound therefore rises in pitch if a rotating paddle is placed in front of the loudspeaker cone, and as this turns in the direction of the observer; equally the sound falls in pitch as the paddle turns towards the cone. Incidentally this is an instance where the frequency and the pitch are not numerically equal.

The Allen organ company used a loudspeaker rotating on an elliptical axis at the top of a long wooden tube, like a stopped organ pipe, for a long time. In their more recent installations, medium and high frequency units are rotated on a plane baffle[7]. It is clear that very fine control of the effect is possible from an infinitely variable speed motor. The method is very successful, if cumbersome.

A scheme dating back thirty years is shown in Fig. 5. The signal is applied to the 'armature' and removed from the field windings. It is rather difficult to ensure that the degree of magnetic coupling is exactly the same each time the device stops, for unless this is so, the level will change.

A quite new approach to the problem of adding some chorus effect to a frequency divider organ is due to the

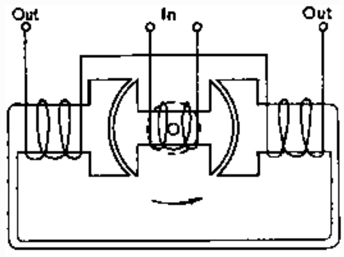

Fig. 5. Early rotary vibrato

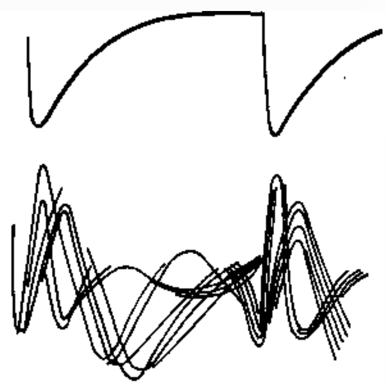

Fig. 6. Waveforms of Baldwin system

Baldwin company. The signal is fed into a circuit which produces a three-phase signal from the input. This passes through a single sideband suppressed carrier modulator unit. The modulator contains an oscillator, which in effect determines the amount and the direction in which the original signal should be altered in frequency. Thus not only is frequency modulation applied but also, some degree of amplitude variation.

As an example, a 375c/s note would be reduced by 1c/s, a 750c/s note increased by 2c/s; a 1 500c/s note reduced by 4c/s; and a 3 000c/s note increased by 8c/s. This continual phase and amplitude shifting over the compass (at least, from 200 to 4 000c/s) results in a complex changing wave pattern as in Fig. 6, which shows a string tone waveform as generated and when passed through the modulator unit. Such circuits add considerable richness to the tone quality.

Monophonic or melodic instruments may be fitted with more exact forms of vibrato control. In the Trautonium, for example, a liquid resistor in the form of a long tube

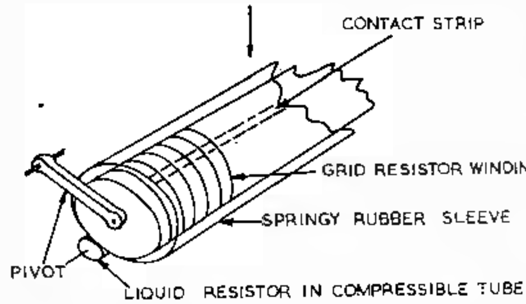

Fig. 7. Trautonium variable touch device

can be depressed by the 'manuals', to form infinite gradations of volume and so, a vibrato. This idea holds great possibilities (see Fig. 7). Playing keys have been devised which can move longitudinally as well as up and down, such sliding action being used to control feedback capacitors or inductors to regulate a vibrato voltage. The monophonic instrument would appear to offer many opportunities for ingenuity in design here; but so far as polyphonic keyboard instruments or those which must conform to strictly conventional design are concerned, the application of a true vibrato in the orchestral or temperamental sense remains unsolved.

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High Altitude Rockets

A Skylark high altitude research rocket was fired in darkness at the Woomera rocket range recently. It reached a height of 83 miles. It was the fourth of six proving rockets scheduled to be fired before the beginning of a joint programme of upper atmosphere research by the Royal Society and the Royal Aircraft Establishment.

The first three firings having shown the performance of the rocket vehicle and its Raven solid propellant motor to be very satisfactory, attention has been turned to trials of rocket-borne equipment and instrumentation, as well as special ground equipment. Skylark No. 4 carried, in addition to routine instrumentation, the equipment for four separate high altitude measurements, each under test but complete and capable of making the required measurements. The four equipments were:—

Grenades. Eighteen grenades each carrying about 1lb of explosive composition, to be ejected from the rocket at three second intervals above 100 000ft altitude. On the ground there were microphones to record the time of arrival of the sound waves, ballistic cameras to record the positions in space of the flashes and a detector to record the instant of each flash. The experiment which is being made by University College, London, aims at measuring atmospheric temperatures and high altitude wind velocities. In this flight three different types of grenade were being tested, and the microphones, cameras and flash detectors were in use together for the first time.

Window Cartridges. Fourteen cartridges of window (well known in its wartime role as a confuser of radar) to be released explosively in two clouds at different heights above 100 000ft; these clouds were then to be followed by tracking radars. The experiment, devised by Imperial College, London, is designed to measure winds at altitudes up to 50 miles.

Dielectric Experiment. A section of the rocket nose was insulated from the remainder of the body to form a capacitor of fixed value. This capacitor was made part of a 5Mc/s oscillatory circuit which can detect the presence of electrons in the ionosphere the results being telemetered to the ground. This very simple but ingenious method of measuring ionospheric electron densities, which has been devised by the University of Birmingham, will be more effective in later rockets which will penetrate further into the ionosphere.

Pressure Gauges. Three Pirani pressure gauges were carried in the rocket cone, one measuring ram pressure in the very tip of the nose, the other two measuring cone wall pressures. These measurements, being made by the Royal Aircraft Establishment, enable the undisturbed ambient pressure at various altitudes to be determined.

Besides these main items of equipment, Skylark No. 4 carried timing gear and switches, telemetry sender, power supplies, Doppler transponder and microwave beacon, the last two items being used for tracking purposes to give velocity and position. The flight was made at night for the convenience of the grenade experiment. The rocket nose, which carried a payload of 130lb, was exhibited publicly at the Symposium on High Altitude Rockets and Satellites at Cranfield in June of last year.

A Feedback Electrometer Amplifier

By D. Allenden*, A.M.I.E.E.

The drift problem in direct-coupled electrometers is discussed briefly, and a design for a direct-coupled feedback electrometer amplifier which uses its inherent feedback loop gain to stabilize the electrometer valve filament supply is presented and described in detail. It is concluded that direct-coupled amplifiers with zero stabilities of the order of a few millivolts per hour can be realized using economical circuits and simple power supplies.

ONE of the major objections to the use of direct-coupled amplifiers is the existence of zero-level drift due to changes of effective heater-cathode potential. This problem is particularly acute in direct-coupled electrometers since the electrometer valve usually gives little, if any, gain, and hence drift in the second stage of the amplifier is as significant as in the first. The magnitude of the drift is normally of the order of 10mV per 1 per cent change in heater voltage[1], resulting in a probable total zero instability of 20 per cent of f.s.d. in the typical case of a current measuring amplifier in which 1V is developed across the input resistor. With a low gain input stage, variations of h.t. supply voltage are more significant than usual, but there

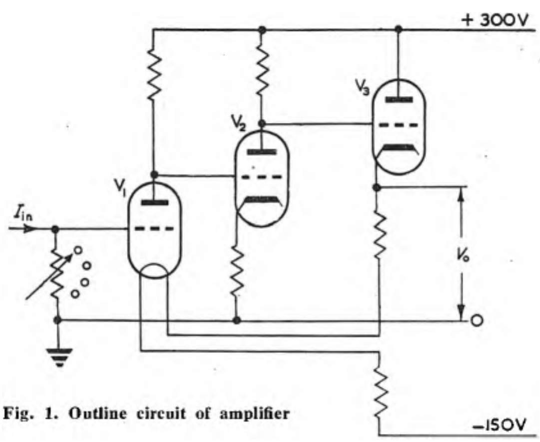

Fig. 1. Outline circuit of amplifier

is no real problem here, since h.t. voltage stabilizers of adequate performance present no design problems. The drift figure quoted is commonly reduced by a factor of 5 or so by the use of balanced stages: better results can be obtained by effective stabilization of the heater supply, a practice which has been little resorted to in view of the inconvenient voltage and current level at which the average valve heater operates. In the electrometer field, the only alternative amplifier is the vibrating-capacitor electrometer[2], which may have zero drifts less than 100μV per day: the cost of this instrument, however, tends to preclude its use except in cases where this order of stability is essential.

There exists therefore a demand, particularly in the field of vacuum physics, for an electrometer with zero stability of the order of a few millivolts with greater linearity and calibration stability than can be realized with simple non-degenerative balanced circuits. The recent development of pentode electrometer valves such as the CV2348 (ME1403), which feature both significant gain and very low filament consumption (8·2mA at 1·4V), have made such electrometers not only practicable but realizable with considerable economy of valves and components. This article gives details of a design of which some half-dozen examples have been constructed and which have given trouble-free service for over a year.

* *Associated Electrical Industries Ltd.*

General Circuit Arrangement

Outline of Operation

An outline circuit is given in Fig. 1. The current to be measured is passed through a suitable high-value resistor and the resultant voltage drop amplified in a 'ring-of-three' feedback amplifier. The electrometer valve, V_1, operates with an anode voltage of 15V and is directly coupled to the second stage which is in turn directly coupled to the cathode-follower, V_3. V_3 anode current is chosen to be 8·2mA, so that feedback and filament supply to V_1 can be effected by including V_1 filament in V_3 cathode resistor chain. A negative supply line is necessary to permit correct

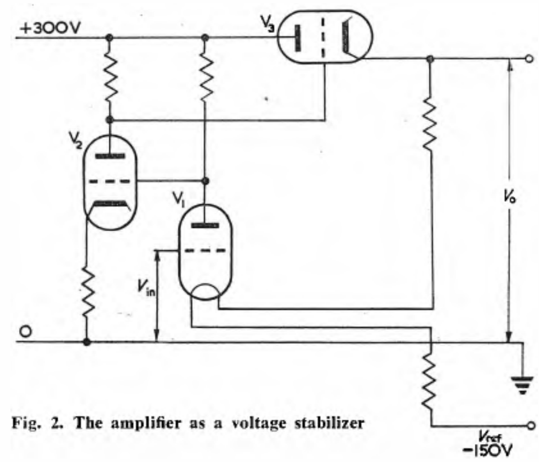

Fig. 2. The amplifier as a voltage stabilizer

d.c. voltage levels to be established without sacrifice of loop gain.

Zero Stability Considerations

The factors which influence the zero stability of the circuit are perhaps better understood by reference to the re-drawn circuit of Fig. 2. It is here seen that in the absence of input voltage the system is a degenerative voltage stabilizer in which the negative supply line is the reference potential relative to which the voltage level of V_3 cathode is stabilized, and that V_1 filament is supplied from this stable source. A signal applied to V_1 grid changes the effective reference voltage of the system, so changing the output voltage. The fractional variation of V_1 filament current in the absence of input is, at most, about double the fractional variation of the reference line. The fact that this is a larger variation than would result from feeding the filament from the reference line is not of great consequence, since the stability easily realizable is such that the residual drift is due to causes other than filament voltage changes. Furthermore, a direct filament feed would complicate the feedback arrangements and inevitably reduce the loop gain, although a successful application of this method has been described[3]. The filament current obviously changes when a signal is applied, but changes of this nature are systematic, and their effects are automatically included in the calibration; it is hence necessary only to ensure

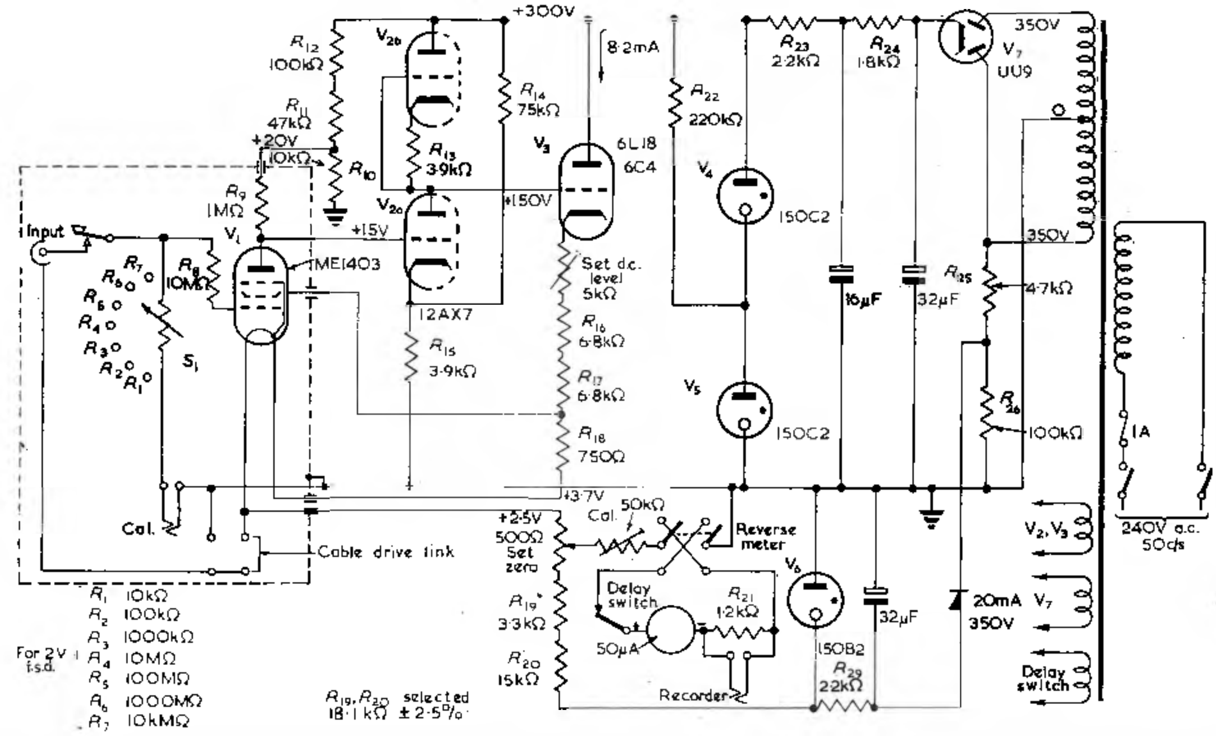

Fig. 3. Electrometer amplifier

that the allowable variation (±1mA) in nominal filament current is not exceeded. Adverse effects on valve life are thereby avoided.

Detailed Circuit Description

Fig. 3 gives a complete circuit of an amplifier with a full-scale voltage sensitivity of 2V, for measurement of ion or electron currents down to 2×10^{-12}A (f.s.d. 2×10^{-10}A), though this does not represent the maximum sensitivity realizable with this design.

Electrometer Stage

V_1 operates with an anode voltage of 15V and a 1MΩ load resistor, and under these conditions a gain of at least 10 can be realized. Grid bias for V_1 is obtained by arranging that its filament sits some 2·5V positive of earth, the system being self-regulating. The screen-grid of V_1, which for true pentode operation at zero frequency should have its potential fixed relative to that of the cathode, is fed from a tapping 6·25V above the filament. There is a small though insignificant amount of positive feedback inherent in this method of supply. V_1 anode supply is taken from a potential divider to limit V_1 anode excursion should the valve be cut off.

Intermediate Stage

With a gain of 10 in V_1, heater-cathode drifts in V_2 can be just significant if no stabilization or compensation is applied to this stage. V_2 therefore employs series-valve compensation[4], a method which is one of the more successful of the available techniques. A 12AX7 twin triode is employed, one-half of the valve acting as anode load for the other. The two triodes have large degenerative cathode resistors of equal values, so that the stage consists essentially of an equivalent generator of output impedance $(r_a + (1 + \mu)R_k)$ feeding a load of the same value; a gain of $\mu/2$ ($\simeq$50) is thus realized in spite of the presence of R_{15}, which is in any case necessary to maintain the correct d.c. level at V_{2a} cathode. If heater-cathode variations are the same in both triodes, there is complete cancellation. In practice, this is unlikely to be the case, though complete cancellation can be realized by variation of V_{2b} cathode resistor; with such adjustment, drifts as low as 150μV/h have been realized. Even without this refinement, the series compensation technique is inherently better than most others; this is probably because the two triodes must operate at the same anode current and voltage, and their dynamic impedances are largely determined by the relatively stable $(1 + \mu)R_k$ term.

Output

V_3, the output cathode-follower, is designed to operate at an anode current equal to the filament current required in V_1, and its cathode load resistor is returned to a 150V negative line. The d.c. operating level of V_3 cathode is determined by the effective β-ratio of the equivalent stabilized supply, and it is desirable to provide pre-set adjustment of this level. Any such adjustment is better located in the upper end of the cathode chain, since in this position it does not alter V_1 filament current; this is determined by the resistance of the lower part of the chain, and the components here (R_{19}, R_{20}) should have suitably small tolerances.

The output meter is connected between the 'earthy' point of V_3 cathode chain and earth. Since the β of the feedback loop is affected by the meter, and also since meter current is drawn through V_1 filament, a high-sensitivity (50μA or 100μA) movement is desirable. In the circuit shown, provisions for meter reversal and insertion of a high-sensitivity recorder or external meter are included. The zero stability with the meter connected at this point is exceptionally good, since, for all variations other than those due to an input signal, the meter feed point is only a volt or so removed from the virtual earth of the stabilizer.

Input Circuit

Input arrangements are conventional. A selector switch S_1, which must naturally be of a high insulation type, selects the desired measuring resistor. The earthy ends of the resistors are returned to the common line via a closed-circuit jack which enables calibration or backing off-signals to be introduced on any range. The whole input stage is enclosed in a light-tight screening box. The input cable capacitance can be driven from V_1 cathode, giving a significant improvement in bandwidth. The cable could be overdriven by taking the drive from a tapping on V_3 cathode chain somewhat above the filament, and this would allow the effect of the residual input capacitance across the resistor (not normally subject to feedback) to be reduced. Instability can result if excessive drive is applied, however, and its adjustment is likely to be critical. The residual capacitance can also be reduced to some extent by a guard-ring driven from V_1 cathode, and more elegant multiple guard-ring systems of greater efficacy have been described[5,6].

POWER SUPPLY

The stability of the negative supply line is obviously of importance, since V_1 filament current and, to a large extent,

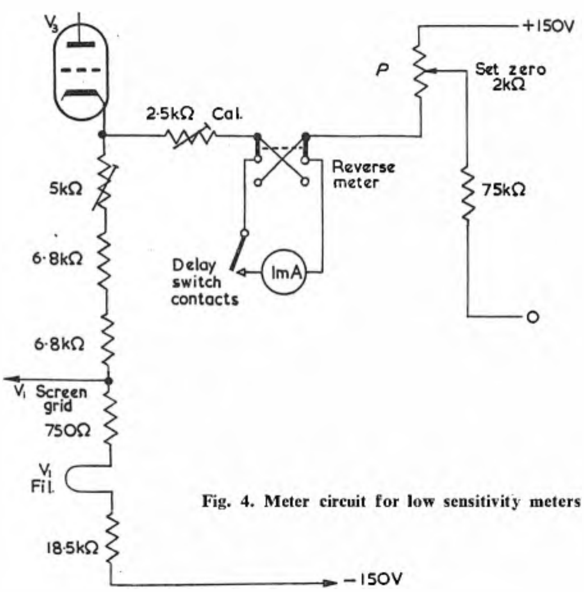

Fig. 4. Meter circuit for low sensitivity meters

the zero stability of the whole system, depends on this. Since the current drain on this line is constant, the requirements are met by a single high-stability reference tube.

The single-ended circuit of the amplifier necessitates some consideration of the effects of positive-line fluctuations if elaborate stabilizer circuits are to be avoided; thus the effect of a change of 1·5V on the +300V line is just about equivalent to a 1 per cent change in V_1 filament current if its effect on V_1 anode potential is uncompensated.

By arrangement of power feeds to V_1 and V_2 such that small variations are largely self-cancelling, the power supplies may be simplified to the extent that ordinary gas-filled reference tubes give adequate performance.

Warming-up effects also need consideration. Although V_1 filament can never be over-run, grid current can flow, particularly if the negative line voltage is established before the +300V supply. Previous work[7] with the CV2348 has established that such grid current is not harmful if limited to a few microamperes or so. The series grid resistor R_g fulfils this requirement. The meter movement is disconnected during warm-up by a thermal delay switch.

Performance

The performance of the circuit is very good, judged by normal direct-coupled amplifier standards. Grid current in the CV2348 is usually well below 10^{-14}A, giving less than 1 per cent zero shift even with input resistors as high as $10^{12}\Omega$. Where sensitivities requiring resistors $>10^{12}\Omega$ are contemplated, the 'rate-of-rise' technique, using an input capacitor[8] is preferred. Departures from linearity are within the permitted values for a first-grade meter movement alone. Zero stability, at a full-scale voltage sensitivity of 2V is better than ± 1 part in 1 000 over several hours, after an initial drift of $\simeq 1$ to 2 per cent in the first ten minutes. Fig. 5 shows a typical zero-stability record.

Output Circuit for Low Sensitivity Meter

In some cases, it may be desirable to use a meter of lower sensitivity than 100μA; in such circumstances, the reduction in loop gain and variations of V_1 filament current caused by taking the meter feed from the virtual earth point would be intolerable. On the other hand, an additional cathode-follower feeding the meter is also undesirable, since it introduces a source of heater-cathode variation that is outside the feedback loop. A modified output circuit shown in Fig. 4 has been devised for use in such cases. The output is taken directly from V_3 cathode and returned to a potential divider connected to the +150V line via a zero-setting arrangement which renders the calibration sensitivity independent of zero-adjustment. Here zero stability may be somewhat

The Complete Amplifier

inferior, since it is necessary for both −150V and +150V lines to be stable, and a high-stability reference tube is thus employed for the latter source, and its loading should be arranged to be the same as that of the −150V tube. A zero stability of 1 part in 300 has been realized with this circuit.

Conclusions

The designs presented have permitted the realization of direct-coupled electrometer amplifiers having a performance intermediate between that of earlier electrometer valve circuits and capacitive-modulation electrometers. This performance has been achieved with simple and economical circuits and without recourse to compensation techniques requiring frequent adjustment.

Acknowledgments

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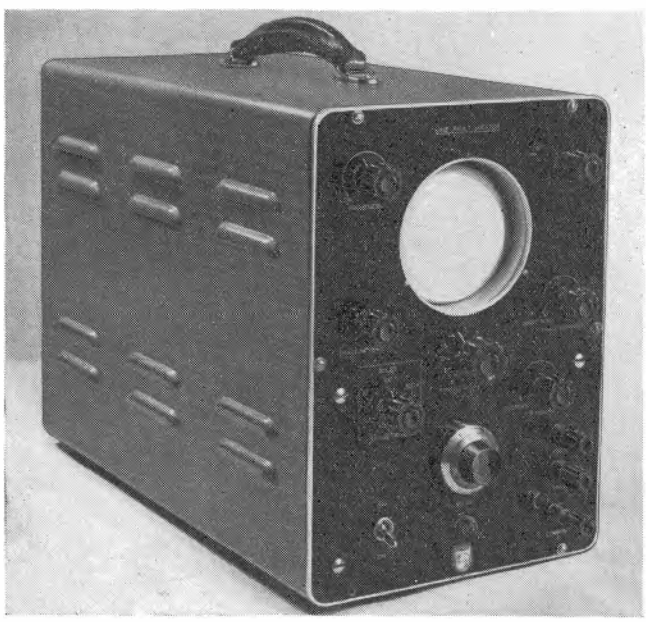
An Electronic Fault Locator for Overhead Lines

By G. Hitchcox*, M.I.E.E., M.Brit.I.R.E., and W. A. Neighbours*

The general character of the problem is outlined and the comparative inadequacy of the various loop tests associated with the testing of underground cables is stressed. The body of the article describes an instrument which displays faults along the time-base trace of a cathode-ray oscilloscope together with a novel monitoring system accurately to determine their distance from the point of test. It is particularly suitable for the detection of multiple and intermittent faults. Finally, the alternative frequency modulation system based on the radio altimeter is briefly described.

EVERY authority which operates overhead lines, either for power distribution or in communications systems, is well aware of the pressing need for rapid and accurate location of faults. The ever-present possibility of sudden faults and the importance of speed in effecting repairs may require the maintenance of expensive service teams, on duty day and night. Anything which can reduce the time spent in searching for trouble and which can indicate its character is of the highest value. In some overseas countries natural conditions make the problem exceptionally difficult; New Zealand is a case in point. Much of the territory is mountainous, densely covered in bush, and remote from roads or even tracks. Storms and gales are frequent, the wind velocity in some parts of the country greater than 40 mile/h for one-tenth of the year. Intervals between control points are often long, perhaps fifty or one hundred miles; a sparse population necessitates a simple and inexpensive mechanical construction. At the same time, adequate distribution of electric power and reliable communications systems are vital to the continued existence of scattered communities, and the need for efficient maintenance is correspondingly high.



The Philips-EDAC electronic line fault locator

Of the various methods commonly used for testing underground lines—the Blavier test, and the Murray, Varley, and Fisher loops—none is primarily suitable for aerial lines. They are inconvenient if the line includes short buried sections of different resistance, as often happens in practice; they are even more difficult to apply where branch lines are connected to the main run or where there are intervening or terminating transformers. The three loop tests require a second line for comparison, often not available. None of them easily detects intermittent contacts, probable results of growing foliage or overhanging branches of trees. All are confused by multiple faults, which appear as a single 'ghost' in a false position. None of them can indicate series faults at all. Complete rupture of the line, as distinct from high resistance joints, can be located by measurement of earth capacitance, but seldom accurately, since such capacitance is often variable, and again confusion is caused by either branches or intervening transformers.

In recent years there have appeared several instruments based upon radar techniques, specifically intended for the testing of aerial lines. The pulse injection method is the most common. If a series of short, widely separated, high amplitude pulses is injected into the line, the presence of either series or parallel faults induces reflections which, after suitable amplification, can be made to modulate (deflect vertically) the time-base trace of an oscilloscope, each stroke initiated by the generation of each transmitted pulse. Since the time interval between generation and reception of the reflection is a measure of the distance between the fault and the point of test, the oscilloscope can be directly calibrated in terms of distance. Each fault appears as a sharp vertical 'pip', its distance from the origin proportional to the distance of the fault along the line. The ability to view the whole length of the line at one time facilitates the detection of both multiple and intermittent faults. Either series or parallel faults can be observed, and since they appear in opposite sense, are readily distinguishable from each other.

Pin-pointing the Fault

Indication of the presence of a fault is of little value unless its position can be accurately determined; its presence may well be obvious from its direct effects. With oscilloscopic display, accurate indication requires two facilities; expansion of the trace, or of that part of it where the fault 'pip' occurs, and a calibration or monitoring system, to reduce or eliminate errors due to variable amplifier gain, non-linearity in the oscilloscope tube, and to parallax in observing it. Expansion is necessary if a tube of practical size, say 12 to 15cm in diameter, is to cover distances of 100 miles or more; without it, one mile of line would occupy only about 1·2mm of the trace, too small for accurate observation.

In one calibration system, the time-base trace is

* *E.D.A.C. Ltd., Wellington, New Zealand.*

monitored by the addition of marker spikes, derived from a crystal controlled multivibrator oscillator. With a constant frequency, successive intervals between spikes indicate equal transmission distances; there is, however, some risk of confusion when testing long lines. If errors due to non-linearity and to inaccurate interpolation are to be reduced to insignificance, intervals between markers must be kept short, of the order of one to five miles. This postulates a large number of them and some uncertainty in deciding which is which, especially when looking for intermittent faults. It helps if every fifth or tenth marker is stressed, either by increasing its size or by inverting it, this demanding a considerably more complicated oscillator system.

In the instrument described below, a different system is adopted. The time-base is generated by a phantastron circuit, chosen because of its high inherent stability. The phantastron is triggered by the same oscillator as that which drives the injection pulse generator and is locked in such a way that the origin of each time-base stroke exactly coincides with the leading edge of the transmitted pulse. The duration, as distinct from the repetition frequency, of the time-base can be varied very sensitively and accurately by adjusting a multi-turn potential divider forming part of the phantastron circuit. In operation, the time-base is adjusted to a time greater than the transmission plus reflection interval of the line under test; if any fault pip is observed, the time-base is then 'peeled back'—that is, its duration reduced—until its end exactly coincides with the peak of the fault pip, as shown in Fig. 1. In view of the sharpness of the apex of the pip, this adjustment can be effected most sensitively, by shortening the time-base until the tip disappears. Fault distance is read by observing the setting of the multi-turn potential divider which, suitably

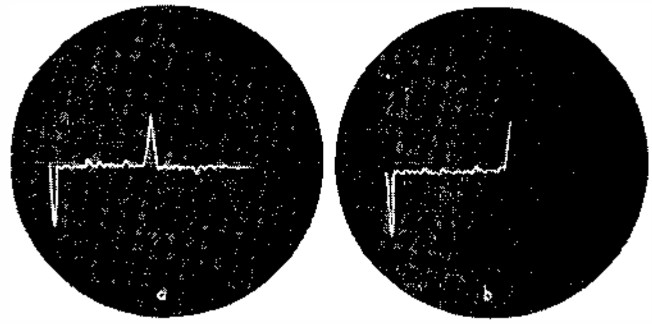


Fig. 1. Double oscilloscope trace

(a) Time-base fully extended on 0 to 100 miles range with major parallel fault at 50 miles (b) Time-base retracted to peak of fault echo, its controls then indicating the distance to the fault

arranged, can be directly calibrated in terms of distance. This combination of calibration monitoring system and time-base is believed to be unique; its advantage is the ease and sensitivity with which monitoring can be carried out.

A Practical Instrument

The circuit of a recently introduced commercial instrument using the variable duration time-base principle is shown in Fig. 2. All power supplies have been omitted for simplicity, together with the gun circuits of the cathode-ray oscilloscope; they follow conventional practice.

The left-hand section of double-triode V_1 forms part of a blocking oscillator the pulsed output from which both triggers the phantastron time-base generator and drives the pulse injection amplifier valve V_6. Pulse duration is approximately 2·5μsec and repetition rate 200p/s, implying a

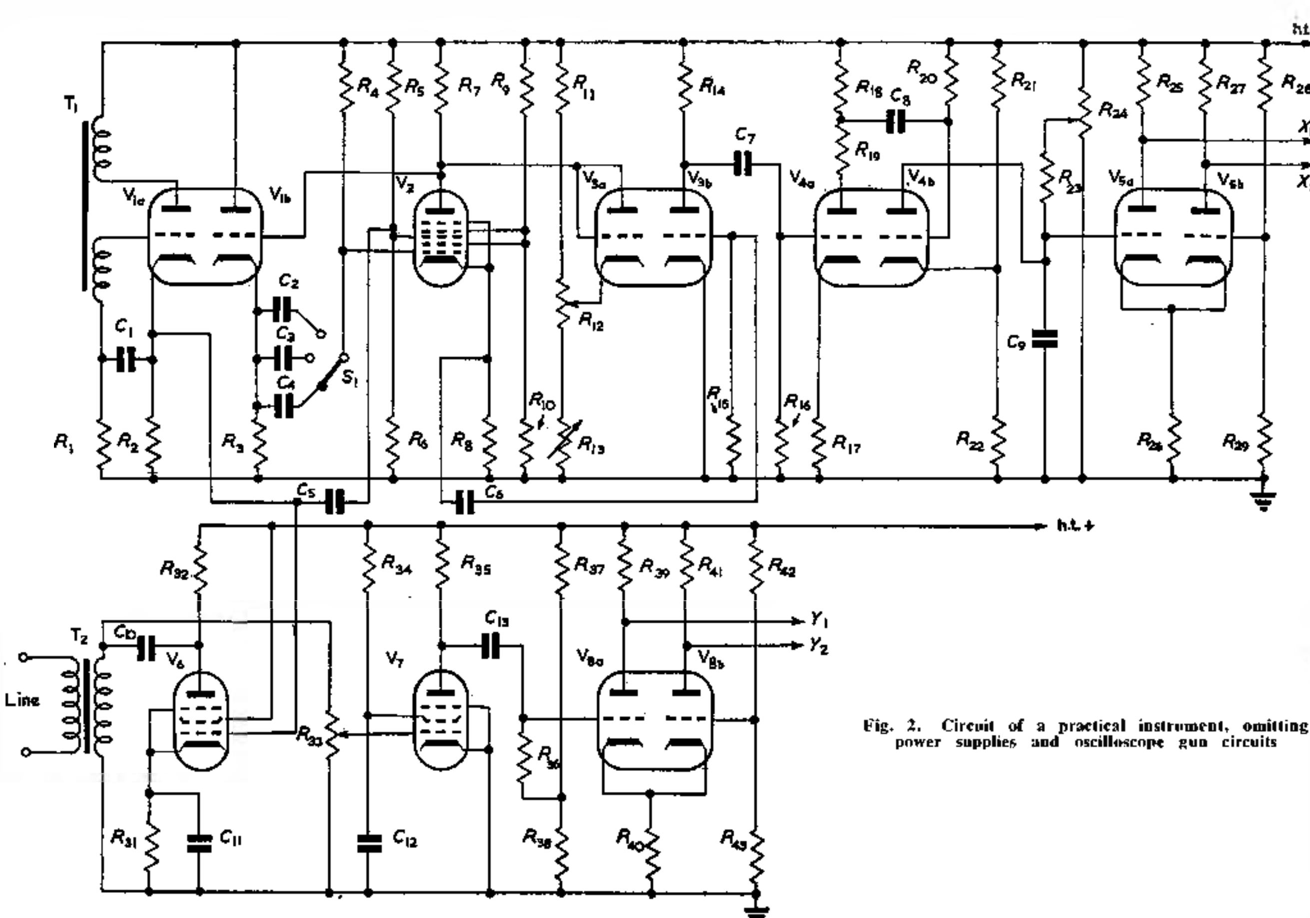


Fig. 2. Circuit of a practical instrument, omitting power supplies and oscilloscope gun circuits

mark-space ratio as far as injection into the line is concerned of 2·5 to 5 000 = 1 to 2 000. The upper limit to the pulse repetition rate is set by the longest length of line which the instrument is intended to test, in this case 300 miles, corresponding to a transmission plus reflection time of about 3·2msec. The separation interval between successive pulses must, of course, exceed this time to avoid overlap. The output from the pulse generator stage is taken through the isolating capacitor C_5 to the third grid of the heptode valve V_2 in the phantastron circuit, and to the control grid of high power pentode valve V_6, the output from which is fed into the line through matching transformer T_2. Normally V_6 is run at a very low anode current; its output pulse is negative-going in order not to paralyse the reflection receiver amplifier, described later.

The heptode V_2 together with cathode-follower V_{1b} and locking diode V_{3a} (a strapped triode) forms the phantastron time-base generator. The merits of a phantastron are: its ability to generate very widely variable 'mark' intervals, high stability of generated interval against power supply changes, and a high degree of linearity. The stability of the 'mark' interval is such that no special precautions need be taken to control either heater or h.t. supplies. Linearity—that is, a uniform relationship between generated time interval and setting of the adjustment control—is necessary if this control is to be accurately calibrated in terms of distance. Three time ranges are provided, according to the value of the timing capacitance selected by the range switch S_1, the corresponding distance ranges being 0·3 to 10 miles, 3 to 100 miles, and 10 to 300 miles. The continuously variable control R_{12} is a 10-turn spiral potentiometer directly calibrated in miles; its setting can be adjusted and read to 1 part in 1 000 of its maximum value. A degree of flexibility is provided by the pre-set adjustment R_{13} to allow the instrument to be directly calibrated for the actual line under test; it is set up by adjusting it accurately to indicate the distance of some known point, such as the end of the line.

It may be stressed that adjustment of the phantastron controls affects only the mark-to-space ratio of its output; repetition rate is fixed by the pulse generator which triggers it, and remains constant at 200p/s. The duration of the mark interval is variable over the three ranges from about 3·2msec to 3·2 msec. The nearly rectilinear output is heavily amplified by the triode amplifier valve V_{3b} and is then further 'squared' by the following limiter stage based upon valve V_{4a}, the gain of which is reduced by tapping down the anode load. The signal, now nearly perfectly rectilinear, is fed to triode V_{4b}, its anode-cathode impedance directly in parallel with the sweep generation capacitor C_9 which, during the 'space' interval, is thereby practically short-circuited. In the 'mark' period, the heavy negative signal applied to the grid of V_{4b} transforms it to an open-circuit and C_9 is charged from the positive h.t. supply through R_{22} at a rate determined by the adjustment of sweep expansion control R_{24}. The rate of rise is substantially constant since the level at which C_9 is short-circuited, at the end of the 'mark' period, is normally small relative to the positive voltage derived from R_{24}. The effect of R_{24} is to vary the amplitude of the series of triangular signals developed across C_9, their duration being determined by the adjustment of the phantastron. These signals are applied to the phase-splitter stage based upon double-triode V_5, the balanced outputs from which are directly connected to the horizontal deflexion plates of the oscilloscope tube.

Reflection signals induced by faults in the line return through the matching transformer T_2 and are controlled by the continuously variable attenuator R_{33}. After amplification by high gain pentode V_7, they are applied to phase-splitter V_8 and then fed to the vertical deflexion plates of the oscilloscope. It has been mentioned that the output from injection amplifier pentode V_6 is such that the polarity of the high amplitude injection pulse is negative relative to earth, to avoid paralysis of the receiver. During the 'space' interval, the beam current of the cathode-ray tube is cut off, the connexion from the pulse squarer stage to the grid of the tube not being shown in this simplified circuit.

Operation

The instrument is connected to the line under test, the operating conditions of the cathode-ray tube adjusted in the ordinary way, and the phantastron controls set so that the time-base trace covers a distance rather more than the total length of the line. The presence of discontinuities in the line such as faults induces sharp 'pips' on the horizontal trace, in the same sense as the transmitted pulse for series faults, and opposite to it for parallel faults. The distance separating any fault from the point of test can be determined by shortening the time-base until it ends exactly on the sharp tip of the corresponding 'pip', the setting of phantastron controls S_1 and R_{12} then directly indicating distance. Sensitivity of adjustment is such that any fault can be located to within 1 500ft, much finer discrimination being possible on the shorter ranges, especially where the operator is familiar with the characteristics of the line under test.

Where distance to the fault is small relative to the maximum of the range in use, the horizontal trace can be expanded with advantage, using R_{24}, which determines time-base amplitude. To adjust the instrument as a whole to the characteristics of any particular line, the phantastron controls should be set to indicate the separation distance of some accurately known point; the master trimmer R_{13} is then adjusted to make the end of the time-base trace exactly coincide with the tip of the corresponding reflection 'pip'. The instrument will then correctly indicate this or any other distance within its working range.

Accuracy and Limitations

One merit of this system is that it avoids the need for either lineal measurement of the tube trace or interpolation, its accuracy also being unaffected by variation of gain in either deflexion amplifier or by non-linearity in the oscilloscope tube itself. Operational accuracy is primarily determined by the stability and linearity of the phantastron time-base generator and while such stability cannot reach the perfection of a crystal-controlled monitoring oscillator, accuracy of operation can be rapidly and simply checked against a single known point on the line actually under test. In this case, a certain degree of flexibility can be of positive advantage, since transmission velocity is slightly dependent on the physical construction of the line.

An instrument of this kind cannot reveal potential as distinct from actual faults; for example, it cannot detect the presence of obstructions close to the line but not touching it, which might be found from a high voltage flash-over test. Nor can it display high resistance parallel faults exceeding 10kΩ, perhaps due to damaged but not broken insulators, since the reflection signal then becomes too small for observation. The use of a pulsed test signal of very low average power content has at the same time some advantages; it cannot cause personal shock or damage to ordinary apparatus inadvertently connected to the line, and it is unlikely to induce serious interference in other lines which may be running parallel to that under test.

Finally, while this instrument can detect either series or major parallel faults, or combinations of them, it obviously cannot reveal faults on the far side of a total rupture of the line.

Frequency Modulation Systems

In conclusion, a brief reference to the very interesting alternative frequency modulation system may be of interest. The basic principle is the same, indication of fault distance by measurement of transmission time, but the mechanism is very different. A continuous constant amplitude radio frequency signal is injected into the line. Its frequency is modulated on a sawtooth pattern, increasing at a constant rate $df/dt = M$ for a period greater than the transmission plus reflection time of the line under test, then suddenly reverting to its original value to start the modulation cycle again. As before, the presence of a fault induces a reflected signal; when this reaches the instrument its frequency f_r will be lower than the transmission frequency f_t, the difference given by:

$$f_t - f_r = 2tM$$

where t is the transmission time from the instrument to the fault. This difference frequency can be extracted by heterodyning and measured, the result indicating transmission time and hence fault distance. This method, borrowed from the radio altimeter, is not in very wide use since, without an oscilloscope display to indicate spatial relationships, it cannot conveniently detect either multiple or intermittent faults.

Acknowledgment

This instrument was originated by the second author and is manufactured by E.D.A.C. Limited (a Philips company) in Wellington, New Zealand, and thanks are given to that company for permission to describe it.

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Measurement of the Characteristic Impedance of a Coaxial Cable

By L. B. D'Alton*, M.E., A.M.I.E.E.

A method is described for measuring the characteristic impedance of a coaxial cable and the velocity of propagation, the only equipment required being a radio receiver, a signal generator and a calibrated capacitor.

THE input impedance Z_s of a dissipationless two-wire transmission line terminated in a load Z_R is given by the well-known expression

$$Z_s = R_c \frac{Z_R + jR_c \tan \beta d}{R_c + jZ_R \tan \beta d} \quad \text{.......... (1)}$$

where R_c is the characteristic resistance, β the phase shift per unit length and d the length. If the line is open-circuited ($Z_R = \infty$) this reduces to

$$Z_s = -jR_c \cot \beta d \quad \text{.............. (2)}$$

and if the line is short-circuited ($Z_R = 0$) it reduces to

$$Z_s = jR_c \tan \beta d \quad \text{................ (3)}$$

Thus, a method commonly prescribed for the measurement of R_c is that of measuring Z_s first with the line open-circuited and secondly with the line short-circuited. Then

$$R_c^2 = Z_{s(o.c.)} \cdot Z_{s(s.c.)} \quad \text{.............. (4)}$$

However, this system (or any other system based on equation (1) barring a v.s.w.r. technique necessarily excluded in the case of a coaxial cable) calls for an r.f. impedance measurement; the measurement of Z_s at audio frequencies would lead to intolerable errors. To meet the case where an r.f. bridge, slotted line, or other standard method is not available, or as an alternative to its use, the author has devised the following method.

Method

It is obvious from equation (2) that providing $\cot \beta d$ is positive, i.e., providing $\beta d < 90°$, the cable looks like a pure capacitance of value C given by

$$(1/\omega C) = R_c \cot \beta d \quad \text{.............. (5)}$$

The method of determining C experimentally is indicated in Fig. 1. A modulated signal is fed through R_1, and the switch S either to the cable (open-circuited) or to the calibrated capacitor K and the latter is adjusted until the meter M connected to the audio output of the receiver gives the same reading for either position of S. The effective capacitance of the cable is then that indicated by K. It remains to determine βd, but before passing on there are some points worth noting.

R_1 forms a potential divider with the cable and/or K and it can easily be shown that in order that a given disparity between the apparent capacitance of the cable and the capacitance of K should produce the largest possible percentage change in M the value of R_1 should be at least

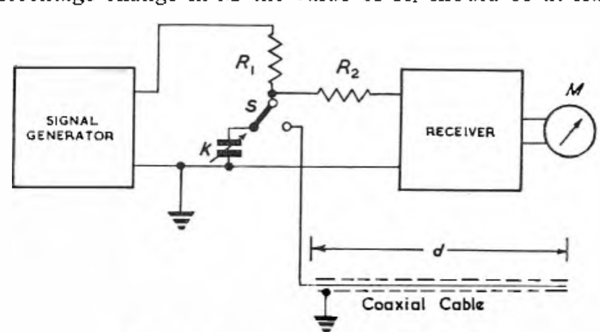

Fig. 1. Arrangement of the equipment

five times the reactance of K. The function of R_2 is simply to prevent the input capacitance of the receiver from swamping K and/or to prevent K from tuning the receiver, which effect if permitted results in an ambiguity in the setting of K. It should therefore also be large compared with the reactance of K.

It is further worth noting that all the advantages of a substitution method accrue viz., the impedance seen by the signal generator is unaffected by the setting of S as is also the impedance seen by the receiver; errors arising from a variation of these quantities are thus avoided. Non-linearities in the receiver introduce no inaccuracy as they

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would if differing voltages were being measured; it is not even necessary that the receiver should be stable.

In regard to the determination of βd let it be assumed first that the velocity of propagation (i.e., the phase velocity) along the cable is known. (In point of fact the manufacturers often specify this quantity to a fair degree of accuracy though the characteristic impedance is specified to no better than 10 per cent). Then, if the velocity is p per cent of the velocity of light c and if λ is the wavelength

$$\lambda = (pc/f)$$

and since

$$\beta = (360/\lambda)^\circ$$

$$\beta d = (360fd/pc)^\circ \quad \ldots\ldots\ldots\ldots\ldots \quad (6)$$

Thus cot βd is known and since C and ω are known R_c may be determined from equation (5).

The quantities f and d must, of course, be chosen so that $\beta d < 90^\circ$ and it is desirable that βd should lie between 10° and 80°. It is also desirable that the cable should be so long that inaccuracies in the measurement of its length introduce no appreciable error. A reasonably long cable also meets another desideratum viz., that f should be kept as low as possible; for high values of f may introduce losses in the variable capacitor K unless it is of extra high quality and are also liable to introduce errors as the circuit configuration is changed by the switch. Connexions to the latter should be kept short.

On the other hand if the cable is very long the error introduced by the attenuation will become serious (see Appendix).

The author chose a value of 4m for d for the Hackethal (German) Type AI 0·6/7·2 100Ω cable and measured R_c at seven different frequencies between 1Mc/s and 10Mc/s. The manufacturers specified p as 66 per cent. The average result was 100·5Ω and the maximum deviation from this value was as little as 0·6Ω, i.e., 0·6 per cent.

Measurement of Velocity

The only remaining problem is the determination of βd or the propagation velocity when the latter is unknown.

Suppose that at a frequency f_1 the cable has an effective capacitance C_1 and that the value of β is β_1. Then equation (5) becomes

$$\frac{1}{2\pi f_1 C_1} = R_c \cot \beta_1 d \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

Similarly at a frequency f_2

$$\frac{1}{2\pi f_2 C_2} = R_c \cot \beta_2 d \quad \ldots\ldots\ldots\ldots\ldots \quad (8)$$

$$\therefore \quad \frac{f_2 C_2}{f_1 C_1} = \tan \beta_2 d \cot \beta_1 d \quad \ldots\ldots\ldots\ldots \quad (9)$$

This equation can be solved only by a graphical method and graphical methods are tedious. But if f_2 is made equal to $2f_1$ then equation (6) shows that $\beta_2 = 2\beta_1$ and equation (9) becomes

$$2C_2/C_1 = \tan 2\beta_1 d \cot \beta_1 d$$

$$= \frac{2 \tan \beta_1 d}{1 - \tan^2 \beta_1 d} \cot \beta_1 d$$

$$= \frac{2}{1 - \tan^2 \beta_1 d}$$

$$\therefore \tan \beta_1 d = \sqrt{(1 - (C_1/C_2))} \quad \ldots\ldots\ldots \quad (10)$$

from which the propagation velocity may be computed if required.

Condition For Minimum Error

It is worth noting again that if the frequency f_1 is low, $\tan \beta_1 d$ is small and from equation (10) C_1 is very nearly equal to C_2. A very small error in C_1 or C_2 will then give rise to a very large error in $\beta_1 d$. Thus, if C_1 and C_2 differ by only 1 per cent then an error of -1 per cent in C_1 gives rise to an error of $+40$ per cent in $\tan \beta_1 d$. If, however, C_2 is of the order of twice C_1 then an error of -1 per cent in C_1 gives rise to an error of less than $+1$ per cent in $\tan \beta_1 d$. Therefore for high accuracy $\tan \beta_1 d$ should be not less than $1/\sqrt{2}$

$$\therefore \ \beta_1 d \geqslant 35{\cdot}3^\circ$$

$$\therefore \ f_1 d \geqslant (pc/10)$$

If p is assigned a representative value of 75 per cent then in general $f_1 d$ should be arranged to be of the order of 23×10^6m/sec.

Using the method outlined the author derived a value of 66·4 per cent for p as compared with the manufacturer's figure of 66 per cent.

Measurements of R_c and p made on a cellular polythene 75Ω television downlead cable yielded similar results.

Summary

The procedure may be summarized as follows:

Arrange the circuit as in Fig. 1 and having chosen a suitable length of cable d (preferably not less than about 2m) set the signal generator and receiver each to a frequency f_1 such that

$$f_1 d \simeq 23 \times 10^6 \text{m/sec}$$

and make R_1 and R_2 of the order of 1·5kΩ each. Adjust K until the reading of M is independent of the switch position. If the propagation velocity is known R_c may now be calculated from equation (5). If not, reset the signal generator and receiver to a frequency $f_2 = 2f_1$ and adjust K as before. The value of R_c may now be computed from equations (10) and (5) and the propagation velocity, if required, from equation (10).

It is, of course, necessary that the r.f. losses in K should be negligible and that its calibration should be accurate. (If such a capacitor is not available an ordinary 500pF tuning capacitor may be used and measured at any setting against a 1kc/s bridge or any suitable standard). It is also necessary that the frequency calibration of the signal generator should be accurate.

It is shown in the Appendix that the error involved in the assumption that the cable is dissipationless is negligible.

Appendix

The input impedance of a cable which is not dissipationless is given by

$$Z_s = Z_c \coth (\alpha + j\beta)d \ \ . \ . \ . \ \text{(open-circuited)}$$

where Z_c is the characteristic impedance and α the attenuation in nepers per unit length. This may be written

$$Z_s = Z_c \ \frac{1 + j \tanh \alpha d \tan \beta d}{\tanh \alpha d + j \tan \beta d}$$

Even at 10Mc/s the highest attenuation for the smallest diameter cellular polythene television cable is of the order of 0·0057N/m. If p and d are assigned representative values of 80 per cent and 4m respectively then $\alpha d = 0{\cdot}0228$N and $\beta d = 60^\circ$; and since Z_c is very nearly the same as R_c for a low-loss cable, substitution of these values gives

$$Z_s = R_c(0{\cdot}031 - j\,0{\cdot}577)$$

The operator will automatically adjust K to make its reactance equal to the modulus of Z_s, i.e.,

$$(1/\omega C) = 0{\cdot}5781 R_c$$

The true value of R_c is thus $1/0{\cdot}5781\omega C$ and the operator proceeding on the assumption of a dissipationless cable measures it as $1/\omega C \cot 60^\circ = 1/0{\cdot}5774\omega C$.

Thus the apparent value of R_c is

$$\frac{0{\cdot}5781}{0{\cdot}5774} R_c = 1{\cdot}001 R_c$$

so that the error is 0·1 per cent. This error increases as the cable length is increased, but is still only 0·2 per cent for $d = 13$m.

A Versatile Pulse Pattern Generator

By Peter H. Cutler*, B.Sc., and L. R. Peters*

Two independent pulse pattern generators are used to produce any sequence of pulses of either polarity; either sequence contains up to ten pulses, and the pulse duration may be varied from 45μsec to 1sec.

The two separate patterns may be combined to give a third pattern whose maximum length is ninety pulses.

DURING an experimental investigation into the use of multi-cathode gas tubes in computer type circuits, the pulse pattern generator described below, was designed and constructed.

Two multi-cathode selector gas tubes are used each driven by a common variable-frequency oscillator.

Each produces a pattern of pulses whose frame length may separately be varied, the patterns being determined by means of selector switches on the front panel. These outputs are available with either polarity and if desired the pulse length may be reduced so as to produce a gap between adjacent pulses.

Suitable combining circuits of the type commonly used for carrying out logical operations in electronic computers enable pattern cycles up to ninety pulses in length to be obtained. The generator has been in operation in the authors' laboratory and has been found to be very useful and reliable.

* *S.R.D.E.*

Brief Circuit Description (see Figs. 1 and 2)

(a) The Counters

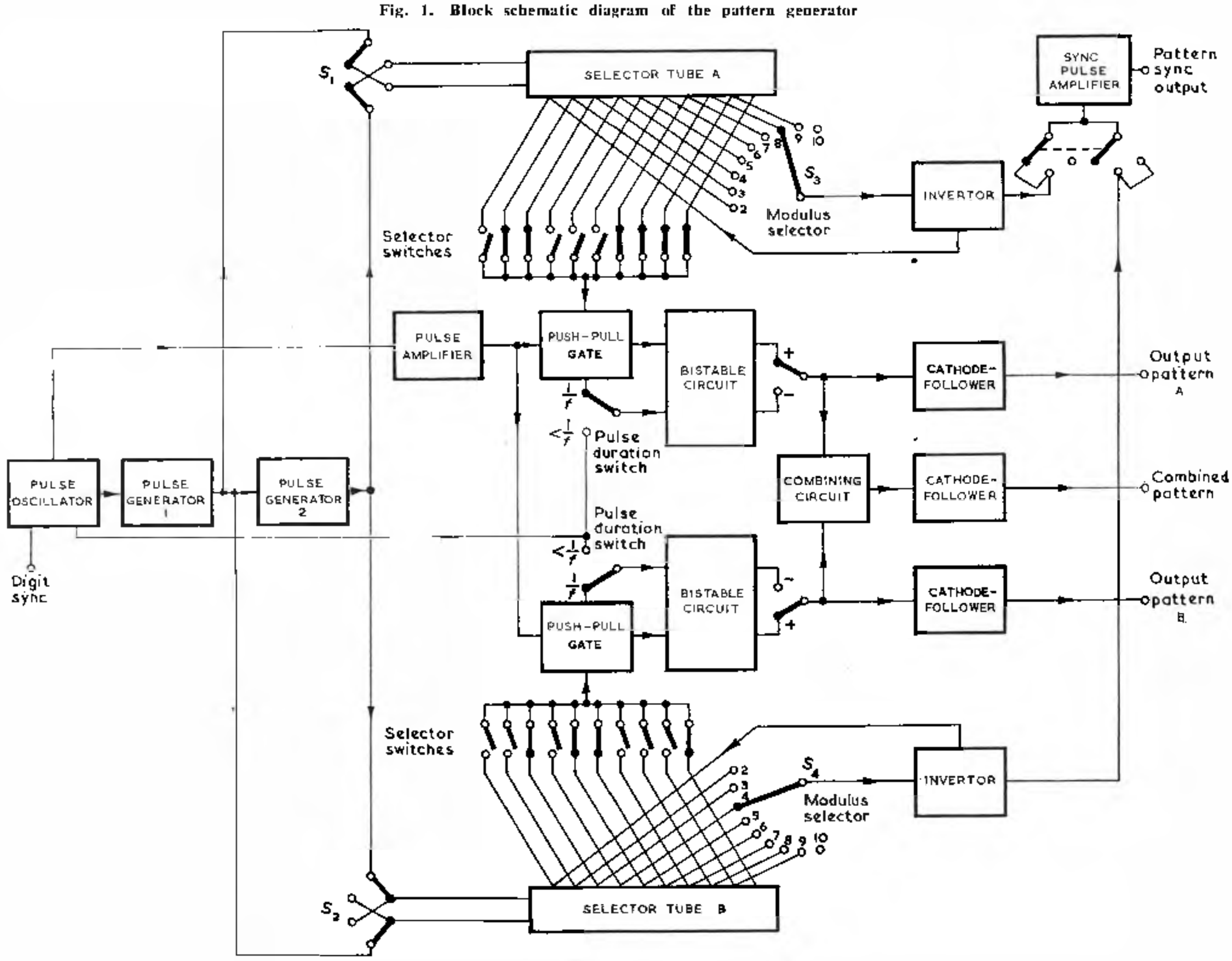
The basic unit of the pattern generator consists of a counter type circuit which is driven by periodic pulses. The Ericsson GS10D multi-cathode selector tube has been chosen for the counter and acts as a pulse pattern generator. This valve and suitable circuits for driving it have been described in the literature[1].

A pair of pulses to drive the GS10D tubes is derived from a multivibrator (V_1 and V_2), whose frequency may be varied from 1c/s to 22kc/s. These pulses cause a single "glow" to rotate about the anode disk, the direction of rotation being determined by the sequence of the drive pulses.

Switches S_1 and S_2 have been inserted into the drive leads of each valve so that each may be driven in either direction, as desired.

When a 'glow' invests one of the ten output cathodes, a positive voltage pulse occurs across the cathode load

Fig. 1. Block schematic diagram of the pattern generator



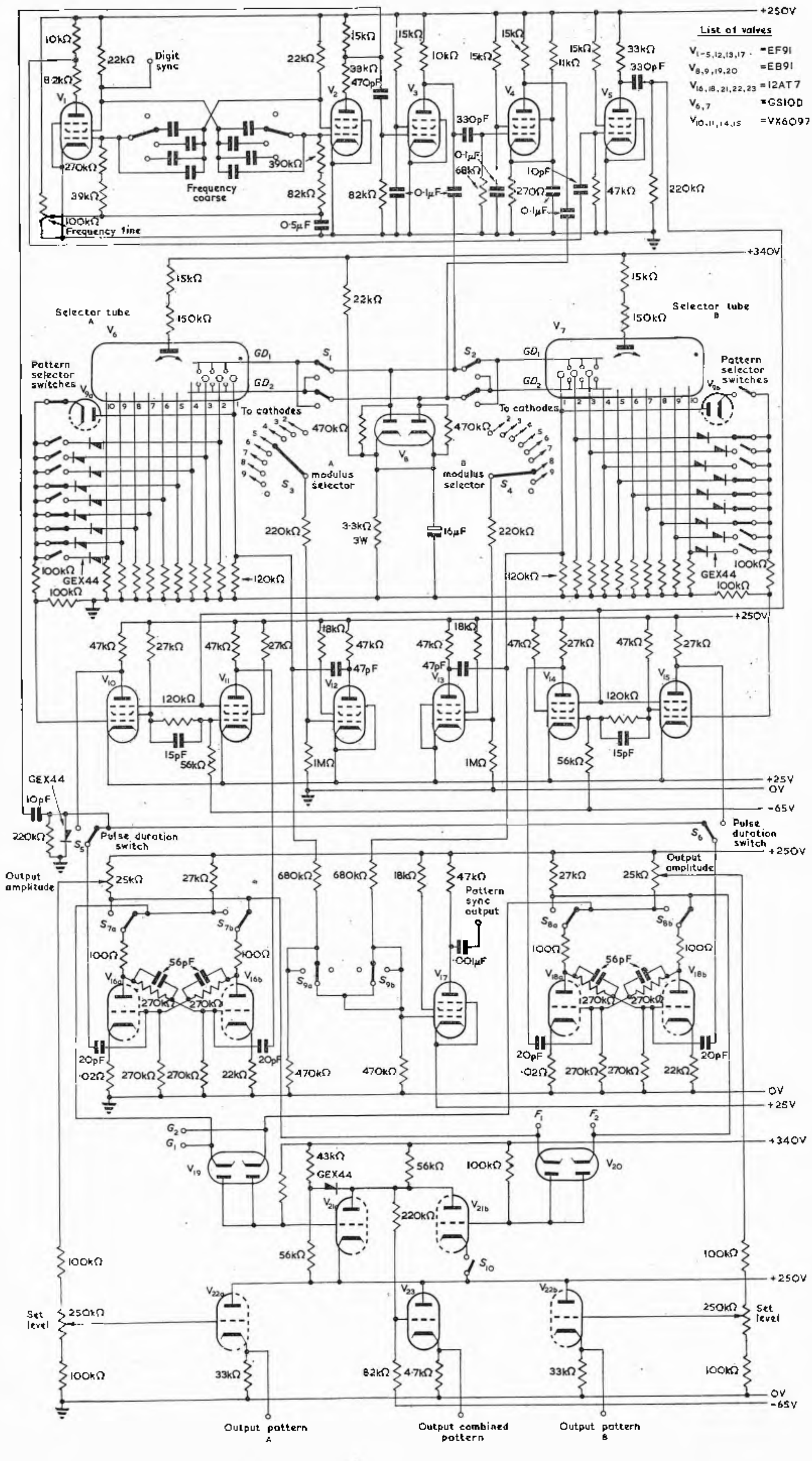
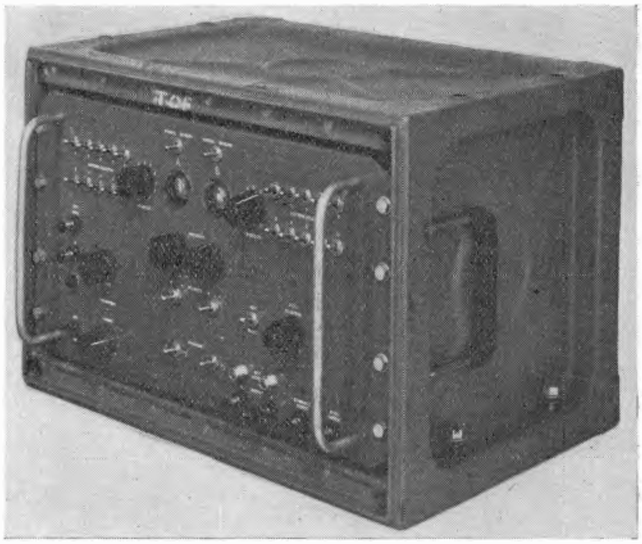


Fig. 2. Circuit diagram of the pattern generator

resistor. Any cathode may be connected via a rectifier and switch to a common line, the rectifiers preventing coupling between cathodes. Thus any pattern of pulses up to a maximum of ten may be obtained by setting the appropriate switches associated with a counter. Since two counters are incorporated, two separate patterns may be obtained.

The modulus of a counter may be varied from two to ten by selecting an output from any cathode (switches S_3 or S_4) and using it, after suitable amplification and inversion, to reset the 'glow' in that counter to its first position.

The outputs of the counters are separately used to gate clock pulses derived from the multivibrator, which, in turn, control bistable circuits. Variable amplitude output pulses of both polarity are obtained from these circuits and fed to cathode-followers (V_{22a} and V_{22b}) to present a low output impedance at the terminals. If desired, pulses which occur between the normal clock pulses may be fed (via switches S_5 and S_6) into one input of the bistable circuits, so causing a return to zero at the end of each output pulse.



The complete pattern generator

(b) The Combining Circuits

The usefulness of the pattern generator has been very greatly increased by the inclusion of simple diode combining circuits (V_{19} and V_{20}). The three basic functions which the diodes perform are (a) an AND gate, (b) an OR gate, and (c) a NOT EQUIVALENT gate. To avoid confusion in terminology, truth tables of these operations are shown in Table 1.

As an example, imagine that the counters have moduli of six and seven, and are producing outputs as shown diagrammatically in Fig. 3(a) and 3(b) (an X represents a mark, either positive or negative). The voltage at the anodes of V_{19} (point C) (Fig. 2) will only rise positively when both G_1 and G_2 are positive at the same time, so that if the waveforms are considered to be a series of positive marks, the voltage at C (shown diagrammatically in Fig. 3(d) will represent the effect of performing an AND operation. Similarly if the waveform is considered to consist of negative marks the voltage at C will represent the effect of an OR operation. Either polarity of mark may be obtained by selecting the appropriate anodes of the bistable circuits by means of switches S_7 and S_8. Fig. 3(d) and (e) show the possible outputs which may be obtained using the inputs shown in Fig. 3(a) and (b).

As mentioned above, either counter may be reversed in direction. Thus, taking the examples given in Fig. 3, the output shown in (b) may be reversed giving the output 3(c). Using, then, the outputs 3(a) and (c), the combining circuit will give 3(f) or (g). This reversing facility therefore provides a simple and quick method of obtaining different patterns containing the same number of marks and spaces. The appendix gives formulæ enabling the numbers of marks and spaces to be quickly calculated.

The triode V_{21a} acts as a slicer and its output is passed to a cathode-follower V_{23}.

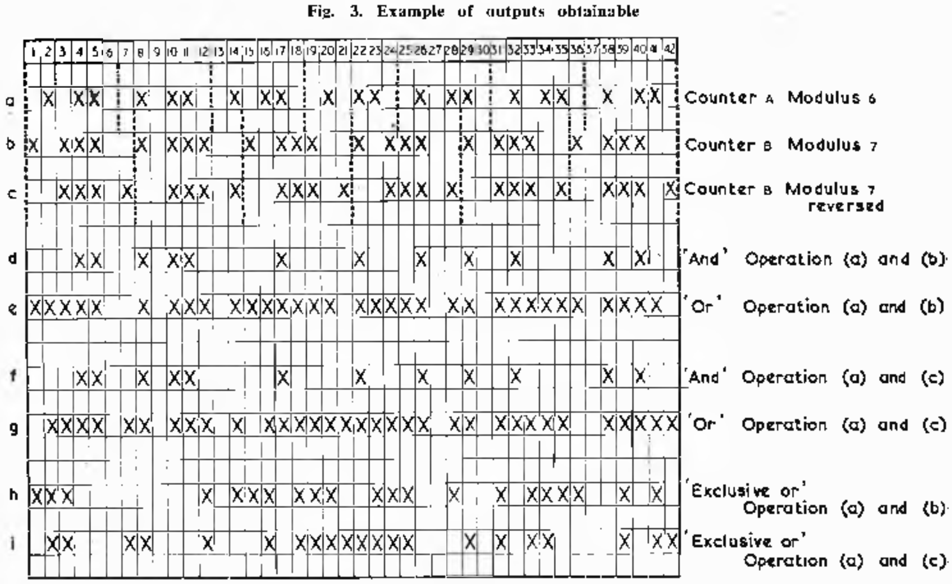
By incorporating another diode gate and feeding to this second gate (see Fig. 2, F_1 and F_2) the inverted patterns of the waveforms applied to the first, the voltage at the anodes of V_{20} (point D) will only rise positively when both inputs to the first gate are negative at the same time. Thus the waveforms at C and D taken together will indicate the effect of a NOT EQUIVALENT operation on the input waveforms. By closing the switch in the cathode lead of V_{21b} the inverted output of the addition of points C and D is obtained at the common anode of V_{21}. Fig. 3(h) and (i) show the outputs which may be obtained by using the waveforms shown in Fig. 3(a) and (b) or (c) as inputs.

As an additional refinement, direct current couplings are used from the bistable circuits onwards so that the low frequency content of the waveform is maintained, the actual level of the positive portion of the output waveforms being

TABLE 1

INPUT		OUTPUT		
A	B	AND	OR	NOT EQUIVALENT
0	0	0	0	0
0	1	0	1	1
1	0	0	1	1
1	1	1	1	0

Fig. 3. Example of outputs obtainable



adjustable by means of the preset controls in the grid circuits of V_{22}.

The photograph shows the external appearance of the pattern generator. The GC10D's are mounted horizontally so that the glow on the electrodes of the valves may be observed from the front, thus affording a quick method of checking the length of the two basic pattern cycles.

Appendix

It will be observed in Fig. 3 that each position of the modulus six counter coincides once and once only with each position of the modulus seven counter. This is obviously true in general.

Let the counter moduli be m_1 and m_2

Let the number of marks be n_1 and n_2
and the number of spaces be s_1 and s_2

Then $n_1 + s_1 = m_1$ and $n_2 + s_2 = m_2$

The total length of the cycle $= m_1 \,.\, m_2$

'AND' OPERATION

The number of marks present $= n_1 \,.\, n_2$

The number of spaces present $= m_1m_2 - n_1n_2$

'OR' OPERATION

The number of marks present $= m_1m_2 - s_1s_2$

The number of spaces present $= s_1s_2$

'NOT EQUIVALENT' OPERATION

The number of marks present $= m_1m_2 - s_1s_2 - n_1n_2$

The number of spaces present $= s_1s_2 + n_1n_2$

N.B.—If the two moduli m_1 and m_2 contain a common factor the total length of the cycle will be proportionately reduced, e.g. moduli 6 and 8 give a cycle of $\frac{6 \times 8}{2} = 24.$

Table 2 shows the number of marks and spaces for the examples shown in Fig. 3.

TABLE 2

$m_1 = 6,\ m_2 = 7 \quad \therefore$ total length $= 42$

$n_1 = 3,\ n_2 = 4 \quad \therefore\ s_1 = 3,\ s_2 = 3$

OPERATION	FORMULA FOR NO. OF MARKS	EXAMPLE NO. OF MARKS
AND[1]	$n_1 \,.\, n_2$	12
OR[2]	$m_1m_2 - s_1s_2$	33
NOT EQUIVALENT[3]	$m_1m_2 - s_1s_2 - n_1n_2$	21

1. (See Fig. 3(d) and (f))
2. (See Fig. 3(e) and (g))
3. (See Fig. 3(h) and (i))

REFERENCE

1. Cold Cathode Tubes (Ericssons Telephones Ltd).

A Chart for Return Loss Determination

By L. Kitajewski*, B.Sc.

Formulæ are derived and a set of curves plotted which enable the return loss or reflection coefficient to be quickly determined from bridge measurements of impedances. Merits of the chart are discussed.

DESIGN of telecommunication equipment involves, in most cases, matching to a specified impedance.

The degree of matching is usually called 'return loss' and is expressed in decibels.

Return loss can be measured quickly and accurately with the help of a return loss tester arranged in a special circuit. This method, however, is more suitable for production testing purposes than for development work where information as to the actual modulus and angle of impedance are often required. In development work it is more customary to measure the impedance by a bridge and then find the return loss value from suitable charts.

The purpose of this article is to discuss a chart for use with parallel bridge measurements of impedance.

Reflection coefficient, ρ, is a measure of the degree of matching, but a more convenient way of expressing that degree of matching is to take $20\log 1/\rho$, which is then called return loss.

As,

$$\rho = \frac{Z_o - Z_R}{Z_o + Z_R}$$ where Z_o = characteristic impedance, and Z_R = terminating impedance.

* *The General Electric Co. Ltd.*

Hence, return loss $= 20\log \dfrac{|Z_o + Z_R|}{|Z_o - Z_R|}$

Let $\dfrac{|(Z_o + Z_R)|}{|(Z_o - Z_R)|} = N$ and $Z_o = R_o + jX_o$, $Z_R = R + jX$

$$\therefore N = \frac{|(R_o+R)+j\,(X_o+X)|}{|(R_o-R)+j\,(X_o-X)|} \quad \therefore\ N^2 = \frac{(R_o+R)^2+(X_o+X)^2}{(R_o-R)^2+(X_o-X)^2} \qquad (1)$$

From equation (1), on multiplying out,

$$N^2R_o^2 - 2RR_oN^2 + N^2R^2 + N^2X_o^2 - 2XX_oN^2 + N^2X^2 = R_o^2 + 2RR_o + R^2 + X_o^2 + 2XX_o + X^2$$

$$\therefore R^2(N^2-1) + X^2(N^2-1) - 2R(N^2+1)\ R_o - 2X\,(N^2+1)\,X_o + R_o^2\,(N^2-1) + X_o^2\,(N^2-1) = 0$$

$$\therefore R^2 + X^2 - 2R\,\frac{N^2+1}{N^2-1}\,R_o - 2X\,\frac{N^2+1}{N^2-1}\,X_o + R_o^2 + X_o^2 = 0 \qquad (2)$$

which is an equation of a circle with R and X variables and N, R_o, X_o constants.

To find the co-ordinates of the centre to this circle and its radius equation (2) can be transformed to:

$$\left(R - \frac{N^2 + 1}{N^2 - 1} R_o\right)^2 + \left(X - \frac{N^2 + 1}{N^2 - 1} X_o\right)^2$$
$$= \left(\frac{N^2 + 1}{N^2 - 1} R_o\right)^2 + \left(\frac{N^2 + 1}{N^2 - 1} X_o\right)^2 - (R_o^2 + X_o^2) \quad .. \; (3)$$

From equation (3) co-ordinates of the centre are obtained:

$$\left(\frac{N^2 + 1}{N^2 - 1} R_o, \; \frac{N^2 + 1}{N^2 - 1} X_o\right)$$

and the radius $= \sqrt{\left[\left(\frac{N^2 - 1}{N^2 + 1}\right)(R_o^2 + X_o^2) - (R_o^2 + X_o^2)\right]}$

if Z_o is purely resistive, then $X_o = 0$

hence, co-ordinates of the centre are: $\left(\frac{N^2 + 1}{N^2 - 1} R_o, \; 0\right)$

and radius $= \sqrt{\left[\left(\frac{N^2 + 1}{N^2 - 1} R_o\right)^2 - R_o^2\right]}$

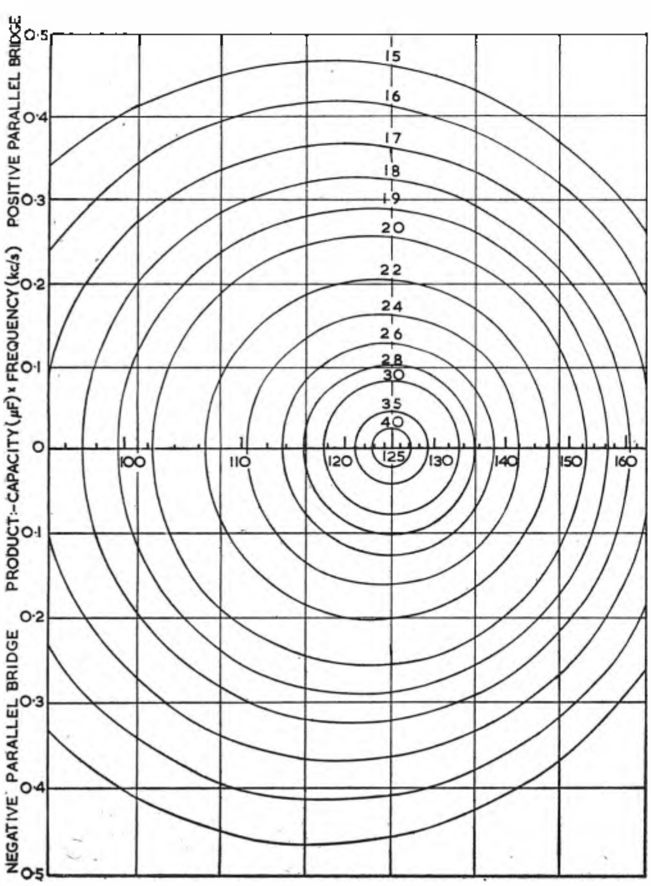
$$= \sqrt{\left[\frac{(N^2 + 1)^2 R_o^2 - R_o^2(N^2 - 1)^2}{N^2 - 1}\right]}$$

$$= \sqrt{\left[\frac{2 \times 2N^2 R_o^2}{(N^2 - 1)^2}\right]} = \frac{2N}{N^2 - 1} R_o$$

If a parallel bridge is used, it will measure resistance in parallel with reactance, then

$$\text{Return loss} = 20 \log \frac{|Y + Y_o|}{|Y - Y_o|} \quad \ldots\ldots\ldots\ldots \; (4)$$

Fig. 1. Return loss curves with respect to 125 $\angle 0°\Omega$. Parallel bridge



Where Y and Y_o are admittances $\begin{array}{l} Y_o = G_o + jB_o \\ Y = G + jB \end{array}$

Where G = conductance. B = susceptance.

Denoting $\frac{|Y + Y_o|}{|Y - Y_o|} = N$ and proceeding as previously one obtains an equation of a circle for a constant value of N. The co-ordinates of the centre to the circle are now given by

$\left(\frac{N^2 + 1}{N^2 - 1} G_o, \; 0\right)$ and radius by $\frac{2N}{N^2 - 1} G_o$.

For impedance of 125 $\angle 0°$, $Y_o = (1/125)$.

Also, if one measures reactance in terms of capacitance only, then

$B = \omega C = 2\pi f C$ or if f is in kc/s and C in μF

$B = 2\pi \times$ kc/s $\times \mu$F $\times 10^{-3}$ and $\frac{B}{2\pi \times 10^{-3}} =$ kc/s $\times \mu$F

For a return loss of 15dB $N = 5{\cdot}62$ from equation (4).

Hence,

$$\text{Centre} = \left(\frac{5{\cdot}62^2 + 1}{5{\cdot}62^2 - 1} \times (1/125) \times \frac{1}{2\pi \times 10^{-3}}, \; 0\right)$$

or (1·356, 0)

and radius $= \frac{2N}{N^2 - 1} \times (1/125) \times \frac{1}{2\pi \times 10^{-3}} = 0{\cdot}468$

Hence the scale is chosen to represent 0·1 of the product (kc/s) $\times$ (μF) by 1 inch.

Product of (kc/s) $\times$ (μF) is put on the vertical scale.

Centre for the infinite return loss, i.e. $(1/125) \times \frac{1}{2\pi \times 10^{-3}}$ or (1·272, 0) is taken as the reference origin. Other centres (for other return losses) are set off from it.

The horizontal scale is calibrated in $G \times \frac{1}{2\pi \times 10^{-3}}$ to comply with vertical scale units.

G = 125 is on the scale at $(1/125) \times \frac{1}{2\pi \times 10^{-3}}$ or 1·272 which is the reference.

Other values of G are set off from 1·272 in the reverse direction. For example:

$R = 120\Omega \quad \therefore G = (1/120) \quad \therefore G \times \frac{1}{2\pi \times 10^{-3}} = 1{\cdot}328$

$\therefore 1{\cdot}328 - 1{\cdot}272 = 0{\cdot}055.$

Hence 120 is marked off to the left of the reference origin by 0·055 of the 1in = 0·1 scale.

This chart can be used only in the particular case when return loss is required between an unknown impedance and $Z_o = 125 \angle 0°\Omega$. Thus, in comparison with Smith charts its use is limited. But in telecommunication engineering characteristic impedances are sufficiently standardized to warrant the use of such a specialized chart.

The advantage of this chart over Smith charts lies in its simplicity and quickness of use.

Acknowledgments

The author wishes to express his thanks to Mr. N. Head for suggesting this method and to the G.E.C. Telephone Works for permission to publish this article.

BBC High Power Experimental U.H.F. Television Transmissions

ON 11 November the BBC began a series of high power test transmissions on 654·25Mc/s to collect further data relating to the u.h.f. bands (Bands IV and V) which were allocated to television at the International Radio Conference at Atlantic City 1947, but have not so far been used for television broadcasting in this country.

The v.h.f. and u.h.f. bands allocated to sound and television broadcasting are as follows:—

Band I	41 to 68Mc/s	v.h.f.	used for BBC television
Band II	87·5 to 100Mc/s	v.h.f.	part used for BBC sound programmes
Band III	174 to 216Mc/s	v.h.f.	part used for I.T.A. television
Band IV	470 to 585Mc/s	u.h.f.	
Band V	610 to 960Mc/s	u.h.f.	

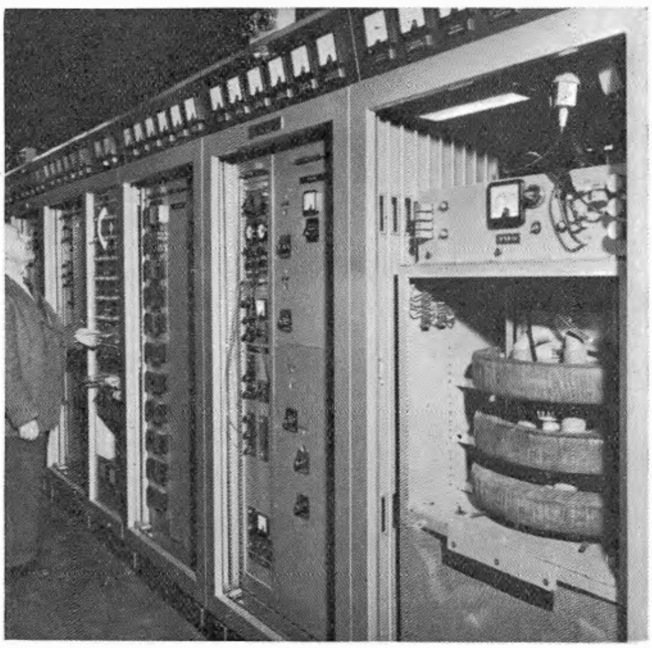


Fig. 1. The 10kW Band 2 vision transmitter showing the klystron output stage in the foreground

Following earlier laboratory work, the BBC embarked on its first series of u.h.f. propagation measurements in 1955. These were directed towards the investigation of propagation on frequencies between 479 and 960Mc/s over distances representing a normal service area, and over longer distances to obtain data necessary for the evaluation of co-channel interference. The initial tests were for the latter purpose and employed transmitters, modulated by square waves, manufactured for the BBC by Mullard Limited. The transmitters were installed at various television transmitting stations, the aerials being erected high up on the television masts. Regular field-strength measurements were made over long periods at various locations, some as far away as the Shetland Islands. These transmissions are still in progress

The long-distance tests were followed in 1956 by a series to determine propagation conditions within a typical service area using a transmitter at the Crystal Palace on 495 Mc/s, working into a Yagi aerial and radiating a peak power of 1kW over a fairly narrow beam when modulated with square waves; pulse modulation was also used for some of the tests. The bearing of the aerial, which was at a height of 440ft, was changed from time to time so that field-strengths could be measured over the whole circle from Crystal Palace. These measurements have been completed and the resulting information concerning field-strength contours and an assessment of shadow and echo effects are now being studied.

The information collected in this way was however insufficient to determine fully the best way of using Bands

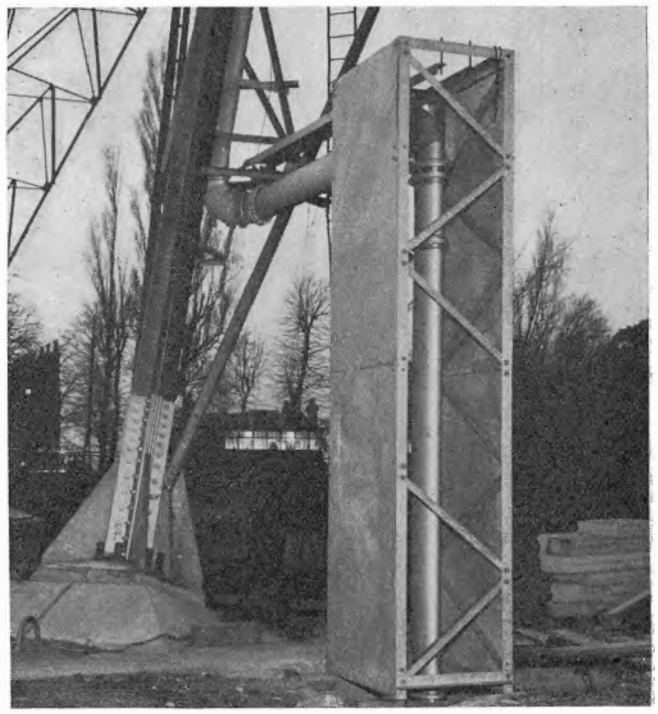


Fig. 2. The elliptical waveguide at the front of the tower showing the concertina expansion couplings and the construction of the angle bends. The partly constructed cover is designed to protect the waveguide from mechanical damage

IV and V for television broadcasting, and at the request of the Television Advisory Committee the BBC decided earlier this year to embark on a more ambitious series of experiments using a high power transmitter and radiating full television signals, initially on 405 lines and later on 625 lines (C.C.I.R. standards). These tests have been planned by the BBC in co-operation with the Committee and the Radio Industry.

For this purpose the BBC has installed at the Crystal Palace a 10kW peak-white u.h.f. vision transmitter and a $2\frac{1}{2}$kW carrier power sound transmitter manufactured by E.M.I. Electronics Ltd, the vision frequency being 654·25 Mc/s. The equipment which is shown in Fig. 1 is low-power modulated on both sound and vision channels and employs Eimac 3K50000LF klystrons in both audio and video final stages. These klystrons use three external cavity resonators and operate as linear amplifiers with a power gain of approximately 100. They are driven by a modulated amplifier stage operating with a cathode modulated circuit. The output of the transmitters is combined in a circuit of the filter bridge type constructed in rectangular section waveguide. The combined output is then conveyed to the aerial by a 12in by 6in elliptical waveguide. Fig. 2 shows a view of the waveguide at the foot of the tower. The elliptical waveguide is made of 99·5 per cent aluminium in 12ft lengths. At the top of the television mast the waveguide is transformed into a 5in concentric feeder to take power to the four driving points of the helical aerial, the pole supporting the aerial being arranged to form the outer of the concentric feeder.

The helical aerial is made of ½in diameter copper rod and comprises four bays mounted one above the other on the same vertical axis, each having a linear height of five wavelengths. Each bay (Fig. 3) is fed at the centre, the helix being wound from the centre point of the bay in opposing directions to cancel the vertical component of radiation. In the four bays there is a total of 48 turns, each turn being approximately two wavelengths long. The aerial is mounted at the summit of the Crystal Palace tower, the top of the 6½in diameter pole supporting the aerial being 707ft above the ground, while the centre line of the aerial is 691ft above the ground. The aerial has a power gain of 20 and after allowing for losses in the feeder and waveguide system, the effective radiated power of the vision signal is of the order of 125kW peak-white in the horizontal plane. The complete helical aerial prior to erection is seen in Fig. 4. Provision is made for de-icing the aerial by electrical heating.

The Band V transmitter is in operation for several hours a day radiating pictures on the 405-line standard. The pictures are the same as those radiated by the Band I transmitter installed in the same building. Later on, pictures on 625 lines will be produced at Lime Grove from flying spot telecine equipment supplied by Cinema-Television Ltd, and sent over a specially equipped coaxial cable to the Crystal Palace.

The tests on 405 lines will continue until about March 1958, when there will be an interval for the transmitter to be adjusted to radiate on 625 lines. A second series of tests will then begin and will continue for a period not yet decided. For these latter tests, the programme will usually be different from that being radiated by the Band I transmitter, but at certain times duplicate copies of films scanned on the two systems will be radiated by the two transmitters simultaneously.

The BBC, the Radio Industry, the Post Office, the D.S.I.R. and the I.T.A. are organizing comprehensive studies of the received pictures. A number of experimental receivers of various types will be used, some being representative of the designs that might be expected to become available to the public in the future. The BBC hopes that the information which will be gained from these tests will throw light on the problems to be considered by the Television Advisory Committee when preparing its recommendations for the future development of television in the United Kingdom, in particular on the problems which would be encountered were it decided to provide television service in the u.h.f. bands, and on the effects of a possible change of United Kingdom television standards for these bands to conform with the standards used on the continent. There is of course no intention of making any change in the 405-line standards used in Bands I and III.

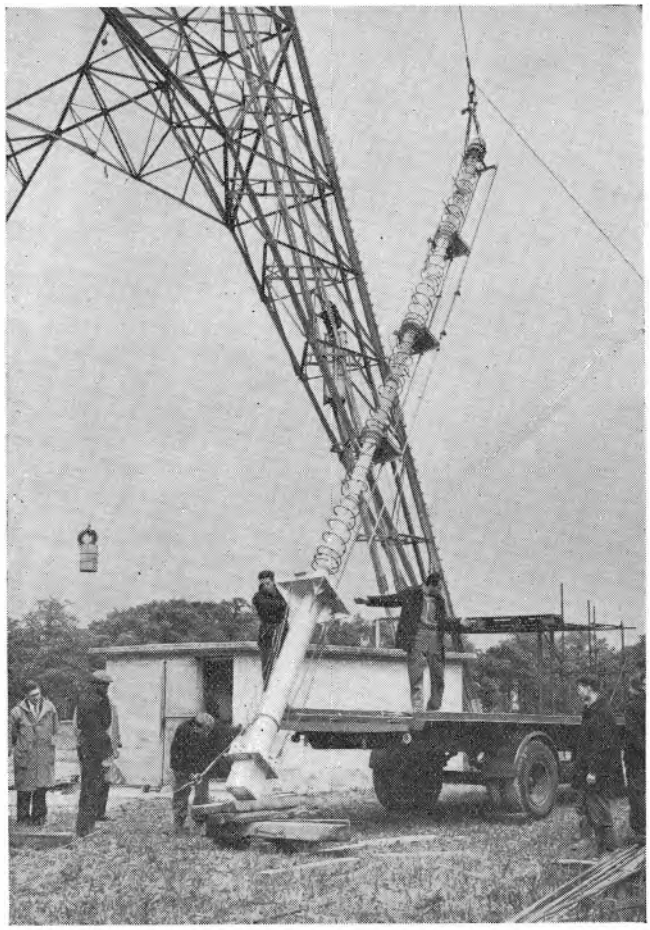


Fig. 4. The four bay helical aerial about to be hoisted to the top of the existing BBC tower at Crystal Palace

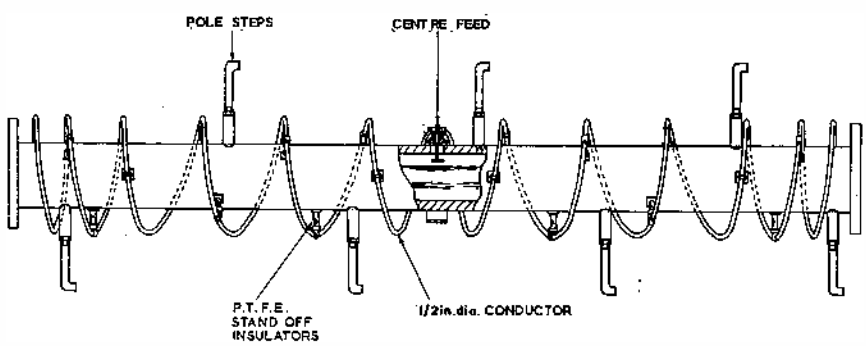


Fig. 3. A single bay of the helical aerial

LETTERS TO THE EDITOR

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

The Schmitt Trigger Circuit

DEAR SIR,—A useful modification to the 'Schmitt Trigger Circuit' enables the realization of a square wave generator using emitter coupled junction transistors. In the circuit of Fig. 1 the regenerative switching action is made to occur by suitably biasing X_1. If X_1 is initially conducting, X_2 is cut off. Connecting the capacitor to the base of X_1 turns X_1 off and X_2 on. The capacitor then charges up via R_1 and R_2 and turns X_1 on again and X_2 off. Since base current now flows, the capacitor discharges (effectively via R_1 and $\alpha CB\ R_E$) and the collector voltage of X_1 rises, turning X_2 on and X_1 off once more.

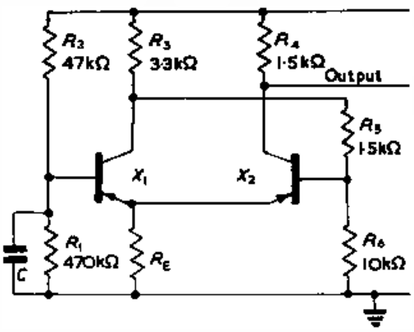


Fig. 1. The modified Schmitt trigger circuit

A straightforward linear analysis enables suitable component values to be chosen—the values given below are typical.

Compared with normal multivibrator circuits, this device has the advantage of a relatively free terminal at the collector of X_2. That is, the performance is relatively independent of the value of R_4. The output at the collector is a square wave with transition times limited only by the high frequency effects in the transistors themselves.

The repetition frequency depends upon the value of C and operation is satisfactory up to about 10kc/s. Because of the high values of R_1 and R_3 the circuit is inherently unsuitable for high temperature operation.

Yours faithfully,

A. G. KLEIN

Department of Electrical Engineering,
University of Melbourne.

Calibration of Vibration Pick-ups

DEAR SIR,—Mr. Foster's article which appeared in the October 1957 issue does not make reference to the upper frequency limit or the accuracy attainable by the methods described for calibration. It is thought that the frequency limit for these methods is 600c/s.

There is a further method which is often regarded as the most accurate for the calibration of pick-ups especially at high frequencies. This method involves the use of a Michelson Interferometer. If one mirror is fixed and the other mirror attached to the vibrating system[1], it can be shown that the fringe pattern vanishes at certain well defined amplitudes of vibration[2].

This method has now been in use in our laboratories for two years and has shown itself to be consistent and reliable, with an accuracy of better than 5 per cent for frequencies up to 10kc/s.

Yours faithfully,

J. G. HAYNES.

Environments Division,
E.M.I. Electronics Ltd.

REFERENCES

1. DAVIES, S. M. Calibration of Vibration Pick-ups at High Frequencies using a Michelson Interferometer. *Nature, 178*, 161, (1956).
2. OSTBERG. *J. Opt. Soc. America* (1932).

The author replies:

DEAR SIR,—With reference to Mr. J. S. Haynes' letter and his first query concerning the upper frequency limits of the methods described, I would like to point out that any frequency limitation is created not so much by the method, as by the vibration source and the pick-up under test. The limitation created by the method is one of amplitude, and is, of course, the minimum amplitude which can be measured with reliability. We can, for instance, measure an amplitude of ± 0·00025in in comfort by means of Fopple's triangle and even with an early type of moving-coil vibration generator such an amplitude is well within its range at 1kc/s. The theoretical maximum frequency for either the methods with unaided present day vibration equipment would be about 1·4kc/s.

It is possible, however, to extend this range very considerably by fitting resonant reeds to the vibration generator and an upper limit of 10kc/s can be reached by this means with ease and accuracy. It is of course necessary to use a very light moving-coil vibration generator with a light load such as a barium titanate pick-up.

Using synchronized stroboflash methods the normal commercial synchroflash is rated at about 500c/s maximum frequency, but even these have been used with special triggering circuits up to 1kc/s. The reflected light beam and strain gauge methods can also be used at frequencies well into the kc/s region.

The interference fringe method can be extended, by using an image convertor tube in place of the standard synchroflash lamp, and by this means its theoretical limit of frequency is set only by the decay time of the fluorescent screen. Unfortunately, at the moment I cannot quote any figures in this respect, but it seems likely that it is a matter of many megacycles per second.

Regarding the second point in Mr. Haynes' letter, the article was not intended to be comprehensive and many alternative methods have been omitted. The Michelson Interferometer is of course an excellent method of making static measurements, but for vibration measurements its accuracy, in practice, is somewhat open to doubt. This is due to the fact that the interferometer uses four reflecting or refracting mediums and if one or more of these mediums is subjected to vibration transmitted through the mounts serious, though not obvious, errors occur. Otherwise the method is very good.

Yours faithfully,

K. C. FOSTER,

Billericay, Essex.

PUBLICATIONS RECEIVED

THE BRITISH COUNCIL ANNUAL REPORT 1956-57 states that widespread attention has been secured overseas by the opening of the first full-scale nuclear power station at Calder Hall. Such publicity needs to be supported by a sustained flow of information, it adds, and an increasing proportion of the Council's work consists in helping to disseminate knowledge of British scientific achievement and serving as a link between British experts and professional bodies and their opposite numbers overseas. In the last five years the Council has assisted some 7 000 visitors in the sciences, many of whom wished to survey instruments and equipment designed and built in Britain. The British Council, 65 Davies Street, London, W.1.

'WIRELESS WORLD' DIARY 1958 published by T. J. & J. Smith Ltd., in conjunction with 'Wireless World.' There are eighty pages of reference material plus the usual diary pages of a week to an opening. The diary is now in its 40th year of publication and the reference section includes base connexions for 700 current valves, design data, addresses of radio organizations in this country and abroad, licence regulations, etc. Price 6s 3d (Leather), 4s 6d (Rexine). Dorset House, Stamford Street, London, S.E.1.

COMPUTERS is the title of a new monthly by The Bureau of Technical Information, 38 bibliographical bulletin on the subject produced Farringdon Street, London, E.C.4. The Bulletin, published on the first of each month, will provide a complete record of everything published in this country and overseas on computers, fully annotated with all bibliographical details. The annual subscription is £6 6s.

RECEIVING AERIAL SYSTEMS by I. A. Davidson describes some of the technical problems associated with aerials particularly as they apply to domestic installations. It is divided into four parts. The first gives an outline of the subject in general terms; the second is concerned with particular types of aerials; the method of choosing the correct aerial for a given site is also explained; and finally there is a short description of the method of making aerial measurements as an illustration of the work that goes on behind the scenes when manufacturers give technical data on their products. Heywood & Co. Ltd, Ingersoll House, 9 Kingsway, London, W.C.2. Price 21s.

RADIO VALVE DATA. The latest edition of this reference book is over 25 per cent larger than the previous edition. It contains full operating data on over 2 500 types of British and American radio valves, 37 transistors, and 300 cathode-ray tubes. An index enables any valve to be found immediately in the tables, while the full list of equivalents, a new feature introduced in the fourth edition, has been retained and enlarged. Iliffe and Sons Ltd, Dorset House, Stamford Street, London, S.E.1. Price 4s. 6d.

MAN AND AUTOMATION by L. Landon Goodman is divided into two parts; technical aspects, and social economic, and other aspects. Under the first heading will be found sections on the applications of automation not only in engineering but also in industry generally. The human implications of automation are also discussed. Penguin Books Ltd, Harmondsworth, Middlesex. Price 3s. 6d.

Short News Items

Marconi Instruments recently held a private exhibition at St. Albans of specialized test equipment for multi-channel links. The main exhibits comprised u.h.f. Test Set OA 1248, which provides comprehensive testing facilities for radio link equipment in the 1 700 to 2 300Mc/s band; White Noise Test Set OA 1249, used for the measurement of noise and intermodulation distortion in multi-channel telephony equipment; and Derivative Test Equipment OA 1259. The latter consists essentially of a sweep generator and a cathode-ray tube display unit, and has been designed primarily to enable the linearity controls of Marconi HM40 and HM60 Modern Racks to be adjusted for best performance. The applicability of these test instruments to the exact needs of link equipment is the result of close collaboration between the design engineers of Marconi Instruments and of the Communications Division of Marconi's Wireless Telegraph Co.

BBC's Crystal Palace Television Tower. An unusual test was made on the BBC's new 709ft tower on 20 November. As part of an investigation into the aerodynamic behaviour of the tower and to check the theoretical design calculations a battery of ten rockets were mounted on the tower at a height of 625ft above ground level and fired in sequence at predetermined intervals by remote control. The rockets were not detached from the tower when fired, but held fixed, and each exerted a thrust of half a ton. The movement of the tower produced by the reaction was recorded on electronic equipment in a mobile laboratory on the ground. It was the first time that this form of test has been made on a high tower in this country, and it is hoped that the results will add considerably to present knowledge of the behaviour of structures of this type in steady winds of relatively light force. The BBC's main contractors for the Crystal Palace tower, British Insulated Callender's Construction Co Ltd, placed the test in charge of Dr. A. R. Flint who conducted it in collaboration with the National Physical Laboratory and the Royal Aircraft Establishment.

The Instruments, Electronics and Automation Exhibition, to be held at Olympia in April, will be almost one third larger in floor area than the 1957 exhibition. With more than 60 new exhibitors, the British Section is completely sold out, and the organizers have a waiting list for space. At a special Exhibitors' Meeting, held at the close of the 1957 exhibition, it was decided that the 1958 I.E.A. Exhibition should become international, and already in the Overseas Section of the exhibition stand space has been booked for exhibitors from Belgium, Denmark, France, Germany, Holland, Sweden, Switzerland and the U.S.A. In addition to the trade associations who sponsored the 1957 I.E.A. Exhibition—The British Electrical and Allied Manufacturers' Association, The British Industrial Measuring and Control Apparatus Manufacturers' Association, The British Lampblown Scientific Glassware Manufacturers' Association, The Drawing Office Material Manufacturers' and Dealers' Association, The Scientific Instrument Manufacturers' Association of Great Britain—the Radio Communication and Electronic Engineering Association has joined this impressive list in promoting the 1958 exhibition. I.E.A. 1958 will again be accompanied by a conference, taking place at Olympia at the same time as the exhibition itself. This year's conference will be international in character, and scientists, engineers and industrialists from many continental countries, Canada and the U.S.A. have already asked for details. The exhibition is again organized on behalf of the promoting trade associations by Industrial Exhibitions Ltd, 9 Argyll Street, London, W.1.

The South East London Technical College, Department of Electrical Engineering and Applied Physics is organizing a course of six lectures on Modern Electric Network Theory and Design, commencing on 22 January. The course is intended to give a survey and summary of the analysis of communication networks and of modern design methods. The fee is 10s. and further information may be obtained from the Head of the Department concerned at Lewisham Way, London, S.E.4.

PLUTO, the new research reactor which recently commenced operation at Harwell, is equipped with Ekco nuclear instrumentation. Ekco Electronics Ltd designed, installed and commissioned the complete nuclear instrumentation and control circuits for this latest-type heavy water moderated reactor. Ekco Electronics have made substantial contributions to the design principles of the safety system of PLUTO, with the aim of achieving the highest possible standard of safety and reliability. Current Ekco projects include the complete nuclear instrumentation of the Australian HIFAR reactor at Lucas Heights and the DMTR reactor now being commissioned at Dounreay, Scotland.

Easco Electrical (Holdings) Ltd have recently been awarded a contract by the London County Council for the manufacture and installation of twelve combined broadcast equipments and staff communication systems for installation in twelve new Secondary Schools. This is believed to be one of the largest contracts placed for this type of equipment.

The University of Birmingham announces that a Graduate Course in Information Engineering will be repeated in the 1958-59 session. Applicants wishing to be considered for a D.S.I.R. grant for the session 1958-59, covering the fee of £81 and a maintenance allowance, should apply to the Electrical Engineering Department before 3 February. Copies of the syllabus of the course, or other information, may be obtained from The Supervisor of Graduate Courses, The Electrical Engineering Department, The University, Birmingham, 15.

The Institute of Physics has arranged a three-day conference on Nuclear Fuel Cycles to be held at The Institution of Civil Engineers, London, form 15-17 January. Full details may be obtained from The Institute of Physics, 47 Belgrave Square, London, W.1.

The Second International Analogue Computation Meeting will take place in Strasbourg in September 1958 under the patronage of the Association Internationale pour le Calcul Analogique. The sessions will take place from 1-9 September. On Sunday, 31 August, the partakers will be welcomed. An exhibition of analogue computation machines and equipment will be open from 30 August to 10 September. The programme of this meeting will cover the various Analogue Computation methods (differential analyzers, rheoelectrical analogies, network analysers, simulators, special calculators, etc.) and their applications to science and industry. Owing to the increasing tendency of associating analogue to digital methods, the Organizing Committee has decided to create a new group devoted to electronic digital computation, especially as regards its connexions with analogue computation. Those interested in further details of this meeting, or those intending to attend, or to participate by submitting a paper, are requested to communicate with Monsieur F. H. Raymond, Secretaire Delegue General de l'Association Internationale pour le Calcul Analogique, 138 Boulevard de Verdun, Courbevoie (Seine) France. The representative of the Association in this country is Professor S. C. Redshaw of Birmingham University.

The Nuffield Talking Book Library for the Blind is appealing for volunteers to assist in the repair of the specially designed portable electric or battery gramophones used for the work. There are over 4 000 blind readers in Britain at present and only about 300 helpers, of which 100 are in Scotland; a large number of these are radio and television engineers. Help is urgently needed in London as well as numerous towns and areas in England. Further information may be obtained from Mr. D. Finlay-Maxwell, (Hon. Organizer of Voluntary Helpers), J. Gladstone & Co Ltd, Galashiels.

BOOK REVIEWS

The Services Textbook of Radio and the Services Textbook of Electrical Engineering. Volume 1: Electrical Fundamentals

G. R. Noakes. 645 pp. 70 figs. Demy 8vo. Her Majesty's Stationery Office. 1957. Price 30s.

TWO series of textbooks are being written for the use of Service technical personnel. The first series consists of seven volumes dealing with radio topics while the second is made up of five volumes on electrical engineering (Volume 3 in the Radio Series was reviewed in *Electronic Engineering* in August 1956). The book under review, which deals with the fundamentals of electrical science; is common to both series. The standard and scope of the majority of this volume is appropriate to the first year of a university course in electrical engineering. However, certain more advanced topics, such as harmonic-analysis, coupled-circuits and semi-conductors also receive space. In line with current practice in university courses and in Higher National teaching, the m.k.s. system of units is used throughout. One of the difficulties in working with m.k.s. units is that although the main electrical quantities such as volts, amperes, ohms, farads and henries are the same in m.k.s. units as in the everyday work, this does not apply to the units of mass and length. Thus conversion is usually needed to get the result of a calculation into British Units. This conversion often causes a little difficulty and it would have been helpful to have included a short appendix dealing with the conversion of quantities between the m.k.s. and the British system.

The book is written on two levels, the main text is intended for a reasonably able reader while those parts of it that may give difficulty to a beginner are "starred." The more mathematical proofs are given in the appendices, however, no particular high standard of mathematical ability is required to follow these proofs as only simple calculus is required.

The first chapter describes the mechanical units of work energy, etc. followed by a chapter describing the behaviour of charged bodies and of permanent magnets. Chapter III deals with the electric current and includes Ohms law, Kirchhoff's Law and the Wheatstone bridge. The remaining circuit theorems could well have been introduced at this stage instead of being left to much later in the book. Chapters on electrostatics, electromagnetism and electromagnetic induction are followed by a uesful chapter on conventional magnetic materials (excluding ferrites). The rudiments of electro-chemistry and the action of primary and secondary cells are then explained. Chapter IX is on the ionization of gases and here the good standard of the earlier chapters is not maintained as the treatment is quite inadequate. No attempt is made to classify the various sorts of gaseous discharge and further, although naturally-occurring lightning is dealt with the glow-discharge is not mentioned.

The material so far discussed represents the first half of the book, and covers the basic electrical knowledge needed to understand the second half, which deals with electric motors and generators and a number of special topics. The understanding of the earlier work would be made easier if exercises had been included at the end of each chapter, and the space taken up by the special topics would perhaps have been better employed for this purpose. The text is illustrated by numerous simple sketches and, as is appropriate to a book on fundamentals, no attempt has been made to give illustrations of actual hardware.

For students who are taking a regular course of lectures the book provides a useful and readable introduction to the fundamentals of electrical engineering. However, the absence of numerical examples and the "two-level" approach make it somewhat less suitable for individual study particularly for examination purposes.

V. H. ATTREE

Telecommunications

By W. Fraser. 772 pp. 65 figs. Medium 8vo. Macdonald & Co. (Publishers) Ltd. 1957. Price 65s.

THE standard aimed at is described by the author as that of an introductory text book for students reading for an Engineering Degree or who intend to take the examinations in Telecommunications of the Higher National Certificate, the City and Guilds of London Institute or the Institution of Electrical Engineers; the book is also claimed to be of value to "practising electrical engineers who wish to work in the field of electrical communications."

About half of the book is concerned with the basic theory of telecommunications, e.g. the theory of networks, circuit elements including transmission lines, valves, amplifiers, oscillators, modulators and demodulators, etc., and the theoretical aspects of the transmission of information; the remainder of the book is devoted to the elements of telecommunication systems proper, including telephony (manual and automatic); telegrahpy, including multi-channel voice-frequency telegraphy; facsimile; line-carrier, coaxial-cable and microwave radio links for multi-channel telephony and television; long-distance shortwave radio links; sound and television broadcasting.

This is an extremely wide field to attempt within the scope of a single volume and the author is to be congratulated on his courage in attempting it and the degree of success he has attained.

From the students point of view the book has many attractive features, the style is concise, clear and unencumbered by excessive detail, there is a good selection of worked examples and of questions from recent examination papers in Telecommunications set by London University and the Institution of Electrical Engineers; the references at the end of each chapter are well chosen, with a predominance of British publications which are likely to be more accessible to students than the American literature. Furthermore, the student is given in one volume the substance of several specialized text books and a good deal of material not otherwise available in text book form; apart from the question of practical convenience this is an economic advantage of special appeal to students, who are notoriously short of money.

From the view point of the practising engineer this book is perhaps less satisfactory since the wide range makes it impossible to more than touch on the practical engineerings aspects of telecommunication systems and there are some notable omissions. For example, microwave communication systems are now widely used in many countries for the transmission of television and multi-channel telephony over long distances; in the U.S.A. more than threequarters of the total television link mileage between cities and a quarter of the total trunk telephone circuit mileage are provided by this means. However, in *Telecommunications* some ten lines only (page 719, Chapter XXII) are devoted to such systems and it is said that "ultra-high frequencies are now used in long-distance transmission of television signals, and multi-channel telephony may be a future development." M.F. and v.h.f. broadcasting systems and v.h.f. mobile radio-communications systems also receive only a brief mention.

Summing up, it may be said that while this book provides a useful survey of the principles and theory of telecommunications well adapted to the needs of students, a wholly satisfactory text book on the practice of telecommunications has yet to be written; such a work would no doubt require the collaboration of several specialists in view of the wide scope of the subject, and a number of volumes.

A minor protest must be registered at the thickness of the present volume, it is more than three inches thick which is really too great for convenient handling and calls for the use of thinner paper or a sub-division into two volumes.

W. J. BRAY

Technical Aspects of Sound: Volume II

Edited by E. G. Richardson. 412 pp. 206 figs. Demy 8vo. D. van Nostrand Co. Inc., New York, Cleaver-Hume Press, London. 1957. Price 70s.

THIS companion volume to the first part of the work deals with ultrasonics and underwater acoustics. The co-ordination of subject matter and uniformity of presentation are considerably better than in Part I. An errata slip corrects the few unavoidable errors.

Whilst there is much published literature on ultrasonics, this book devotes no fewer than 80 pages to the characteristics and design of transducers, in addition to 374 references to such applications. This section alone establishes the value of the work. Undoubtedly the most interesting sections are those contributed by Dean E. Meyer on air and gas bubbles in water, and the techniques and apparatus for the measurement of underwater sounds. These are paralleled by two further sections on absorbent systems for water-borne sounds, representing extensions of naval research in Germany during the war. Dr. Richardson adds the results of investigations into propellor noises and the various modes of cavitation in water.

Three chapters on aircraft propellor and jet noises by a team of eminent American experts complete the contents of this volume, thus bringing together a veritable mine of detailed and up to date information on this branch of acoustics. The 570 references cited ensure that the specialist in any branch of ultrasonics will have the means to amplify the contents of the 406 pages of this extremely well produced and readable compendium.

ALAN DOUGLAS

Industrial Electronic Circuits

By R. Kretzmann. 200 pp. 206 figs. Demy 8vo. Philips Technical Library, Holland. Cleaver Hume Press Ltd, London. 1957. Price 35s.

THIS book deals with the circuits of industrial electronic apparatus. The functions of the various circuit elements are described and information is also supplied on the component values. Examples are given of photo-electric control devices, counting circuits for various purposes, stabilizing circuits, switching and control circuits, amplifiers and oscillators, rectifying circuits and motor controls.

Der Transistor. Ein neues Verstärkerelement (The Transistor. A New Amplifier Element)

By Joachim Dosse. 207 pp. 101 figs. Medium 8vo. (2nd revised and enlarged edition). R. Oldenbourg, München. 1957. DM.19·80.

THIS is another instance where the rapid development of the subject necessitated the preparation of a second edition within not much more than one year after the publication of the first edition. This was reviewed in ELECTRONIC ENGINEERING in September 1956. The new editions shows not only an increase in the number of pages but also in size and the number of coloured plates has been increased from four to eight.

While the general explanation of the physical principles remains on the whole unchanged there is a more detailed discussion on the behaviour of the junction transistor in base connexion with sudden increase or decrease of emitter current as dependent on operation in the amplifier and saturation ranges of the volt/ampere characteristic. This is of particular importance for impulse amplifiers and relaxation connexions. A brief explanation of the diffusion of charge carriers is given.

The methods adopted for increasing the cut-off frequency are explained in greater detail and the diffusion transistor is described for which a cut-off frequency of more than 1 000Mc/s is predicted. The section on modified transistors is largely extended and particular attention is given to the avalanche transistor.

Perhaps the largest changes in comparison with the first edition will be found in the chapter dealing with the technical properties of the transistor. The treatment is somewhat more mathematical than before, particularly in the sections dealing with characteristic values and equivalent diagrams for amplifying weak signals and for noise. It is based partly on the use of equivalent diagrams which afford a good insight into the physical phenomena and partly on the more formal use of matrices in quadripole theory. A large number of numerical examples is given. Also the chapter on connexions and circuits is materially enlarged and brought up to date.

While in the first edition the main application consisted of hearing aids the transistors have meanwhile found a much larger field of application and are used e.g. in amplifiers for subscribers' apparatus, call and duplex telephony, channel amplifiers, impedance convertors, generators for carrier frequency and telegraphic purposes, studio amplifiers, series amplifiers for relays and measuring instruments and — last but not least — for computers.

The bibliography has been extended to cover the literature till January 1957 and contains now more than 630 references. The enlargements incorporated in this second edition greatly enhance its value, in particular for the designing engineer.

R. NEUMANN

Industrial Television

By H. A. McGhee. 120 pp. Demy 8vo. George Newnes Ltd. 1957. Price 15s.

THE first section of this book deals with the practical aspects of governing the choice of cameras as well as the principles of pick-up devices. The author then covers the design of the associated control equipment, including synchronizing and blanking pulse generation, time-bases, supply circuits and remote control facilities. In the second part of the book, specific applications of such television systems are described and illustrated. Typical applications of television as a remote viewing aid in industry, including atomics and aviation, commerce, research and education, together with information on the capital and maintenance costs of closed circuit systems are described in the final chapter.

ELECTRONIC EQUIPMENT

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

ELECTRONIC STETHOSCOPE

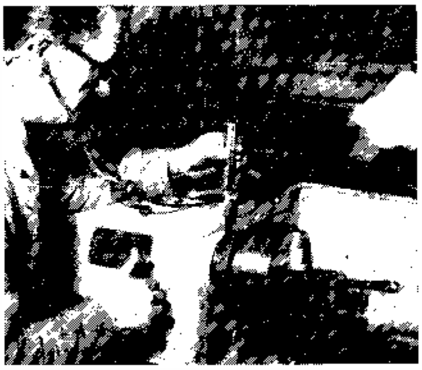
(Illustrated below)

Airsonic Ltd, 14 Old Queen Street, London, S.W.1

THE 'Soniscope' is an electronic stethoscope for carrying out investigations on machinery, pipe-lines, etc. The amplifier has a gain of 90dB with low distortion and can be tuned to any frequency between 25c/s and 20kc/s.

Operated from two miniature batteries within the robust metal instrument case, the overall dimensions are 4¾in by 3¼in by 1⅜in. Two sets of stethoscopes can be connected simultaneously to one instrument, or one outlet connected to a meter, with or without loudspeaker. Several people can therefore study and interpret the signal at the same time. The Soniscope weighs 24oz.

Apart from its specific use as a stethoscope, the instrument will actuate most types of measuring instruments, graphers and tape recorders.



SELECTOR UNIT

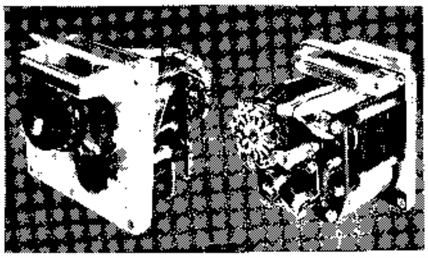
(Illustrated above right)

The Plessey Co. Ltd, Vicarage Lane, Ilford, Essex

A NEW light-weight selector unit, known as the type 9008, has been developed by the Plessey Company Limited in collaboration with the Royal Aircraft Establishment. This remotely controlled device has been designed for rotating a shaft to any one of twelve predetermined angular positions contained within one 320° turn, and locking it in the selected position.

Primary movement of the unit, which has been granted R.A.E. approval, is provided by an electric motor operated from a 48V d.c. supply which may vary between ±6V. A relay is used to control the operation of the motor and to reverse its rotation. Selection of the required preset position is provided by means of a single-pole 12-position switch at the remote operating position.

Each of the angular positions can be set individually without affecting the remaining eleven settings, and facilities are provided at the rear of the unit for limiting the arc in which the twelve angular positions are contained. Accurate settings can be obtained by fine adjustments being made to the output or tuning shaft with a geared-down control knob.



The unit is capable of driving an external frictional load of 2¼lb. in and resets within nine minutes of arc on any single position. The average reset error obtained from all twelve positions is better than six minutes of arc. The output shaft, at the rear of the unit, extends through the front panel to carry a dial co-operating with a pointer. Another dial, which is visible through a circular window in the front panel, indicates which of the twelve positions is in use.

Overall dimensions of the unit, which has been designed for mounting in equipment protected against the free entry of dust, are approximately 4in by 4in by 4½in depth. Weight is approximately 2¼lb.

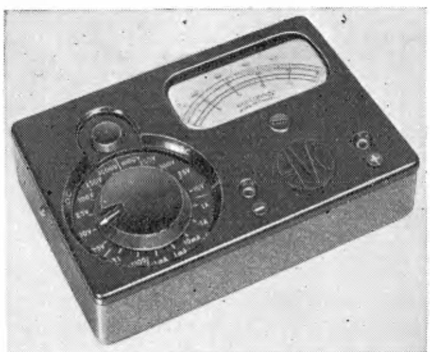
MULTI-RANGE METER

(Illustrated below)

Avo Ltd, 92-96 Vauxhall Bridge Road, London, S.W.1

TWO versions of this meter are available, the Multiminor model 1 for use in temperate climates, and the model 2 for use under adverse climatic conditions.

All components which could be moulded, have been made on precision moulding presses, the panclimatic version, using a recently developed moulding powder, which endows the finished product with a very high factor of electrical insulation, even when working under conditions of extreme humidity. Moreover, this new material resists the growth of fungus, an enemy of instruments used under humid conditions, and a very common cause of failure. The indicating movement has been constructed in a dust-proof casing, and all wiring is termite proof.



Range selection is by means of a rotary switch, there being only two connexion sockets for all measurements. The instrument has 19 ranges, a single scale being provided for current and voltage measurements and another for resistance measurements.

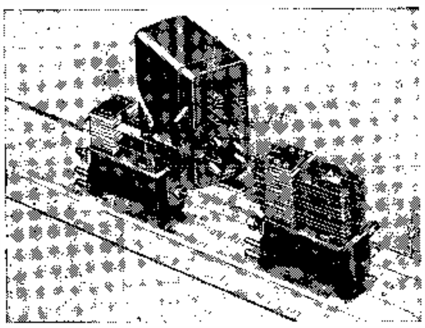
The voltage ranges cover from 0 to 1000V d.c. and a.c. and the current range 0 to 1A d.c. The lowest ranges are 0 to 100mV d.c., 0 to 10V a.c. and 0 to 100μA d.c. The d.c. sensitivity is 10kΩ/V and the a.c. sensitivity 1kΩ/V.

SUB-MINIATURE RELAYS

(Illustrated below)

D. Robinson & Co., 58 Oaks Avenue, Worcester Park, Surrey

A MARK 2 version of the Hopwood-Par type P3 sub-miniature relay is just announced. It can now have up to six changeover contacts for use in control



circuits, or two special changeover contacts with very low capacitances for such uses as 'send-receive' and crystal switching.

In addition it can now be used in temperatures up to 100°C: has a new shock-proof knive edged armature mounting and as it has all tags and fixings at one end it can be mounted through a chassis (a piercing tool is available) or on a printed circuit.

The design target for the P3 was to provide as many of the features of the versatile "3000-type" as miniaturization allows and in consequence the P3 has buffered blades with twin contacts; excellent sensitivity and if required a residual adjusting screw by which the fall off point can be set to between 20 per cent and 60 per cent of the operate point.

Coils can be single or double wound; slugged and up to 9kΩ in value. Consumption can be as low as 100mW and typical contact ratings are 2A at 25V d.c. and 1A at 250V a.c.

The low capacitance version has low loss insulation and a capacitance between open contacts of under 1pF. The tag layout has been arranged to keep capacitances to ground low and constant.

LIGHTWEIGHT SOLDERING IRON

(Illustrated above)

Wolf Electric Tools Ltd, Pioneer Works, Hanger Lane, London, W.5

OF entirely new design, the Wolf No. 92 lightweight soldering iron has been developed to meet the requirements of continuous small assembly and instrument work in the radio, television, electronic and allied industries.

The element is wound on a flat mica strip and folded with an interposing mica insulating piece. It is then encased in mica and enclosed within a rectangular section steel casing.

The dull, nickel plated copper bit is designed for easy replacement, being secured to the barrel by means of a special locking nut. A split moulded handle is used, with all terminals easily accessible. The cable and sleeve are securely clamped into position and an earth symbol permanently marked, conforms to various International regulations. Slots are provided in the chromium plated barrel to ensure that the temperature of the handle is maintained within comfortable limits.

The consumption of this iron is 25W and it is available for a wide range of voltage between 25 and 250V. The net weight of the iron is 3oz.

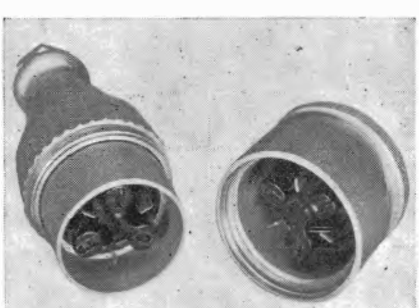
PLUGS AND SOCKETS

(Illustrated below)

Modern Acoustics Ltd, Manor Way, Boreham Wood, Hertfordshire

THIS new range, which meets the stringent electrical and mechanical requirements of the latest electrical equipment, will be manufactured in the United Kingdom under licence from Tuchel Kontakt, Germany. All world-marketing rights outside Europe for the patented Tuchel design have been assigned by the Plessey Company Limited to Modern Acoustics Limited.

Basically, the contact consists of a series of tubes of decreasing diameters, opened longitudinally with two or more transverse slits, assembled one within the other. In operation, the bevelled contact blade is inserted into the first split cylindrical spring bank and as the blade is thicker than the gap, the spring assembly is forced apart and each of the three springs grip the blade pincer fashion. As the blade is introduced, the front edges of the spring assembly rub against it cutting through existing dirt or oxide film. This process is repeated through the second and third banks respectively, until the contact blade is fully inserted. An interesting feature is that the spring banks are not rigidly secured, but float in the connector assembly. This floating action ensures that the pressure exerted by the individual springs upon each side of the contact blade is always balanced. Other important features of this range of plugs and sockets are: a multiple number of independent contact points for each pole, low contact resistance with large contact area, unaffected by vibration, effective in a wide range of temperature, plus simple engagement. The variety of plugs and sockets available ranges from small ½in diameter units, 36 poles per unit, each carrying 60A at 500V.



MARKING MACHINES

(Illustrated below)

Rejafix Ltd, 81-83 Fulham High Street, London, S.W.6

THE diversity of components, their shapes, sizes and the various materials used in their manufacture, present a considerable marking or stamping problem in the electronic and radio component industry. The Rejafix combined marking and printing machine is so designed as to be able to mark direct on the surface of flat, round or irregular shaped components as well as articles having recessed surfaces.

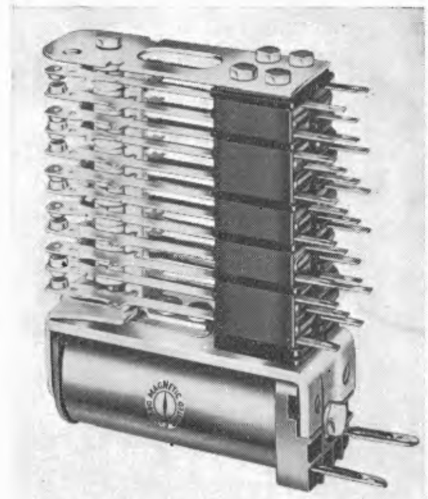
The Rejafix method is simple and it employs the use of positive conventional printing blocks. The printing block or type is automatically inked, and then brought into controlled light contact with the surface of a flat resilient transfer pad; thereby transferring a negative inked impression to the pad's surface.

The article to be marked is then rolled or presses on to this pad, and the inked impression is transferred to the surface of the article. The result is most efficient, and long continuous runs of marking can be carried out, giving the same quality of marking throughout the run.

On the hand-operated models up to 800 pieces per hour can be marked, while on the power-driven models it is up to 3 600 pieces according to the article.

For use in conjunction with their machines Rejafix Ltd supply a range of high quality marking inks in different grades to suit various materials such as glass, paper, cardboard, metal, rubber, wood, p.v.c., Bakelite, Perspex and many other kinds of plastics, as well as far marking on painted surfaces. A ceramic ink is also available for indelibly marking on glass and porcelain.

Illustrated is a 12 position turret model being used to stamp capacitor plates.



D.C. RELAYS

(Illustrated above)

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

THE series 285 relay is extremely compact and has been designed specifically to meet the requirements of the industrial, Electronics and automation fields.

It has a lifter comb and buffer block between the contact stacks and the high insulation properties of the relay are ensured by the encapsulating of the contact units, at their clamped ends.

The original design of lifter comb and buffer block allows for up to 10 pole changeover light duty contacts. Negligible contact bounce is a further feature of the series 285 relay.

With a 12750Ω coil and 2 pole changeover contact assembly the sensitivity is approximately 3mA. These relays are available with a wide range of coil resistances and contact assemblies.

U.H.F. ATTENUATOR PADS

Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex

THE type A83 attenuators are coaxially pads with rod and disk type resistors in T networks. These pads are the same units as used in the turrent attenuator type A63. Five units are available in steps of 10dB and each unit is fitted with Plessey minor coaxial fittings so that they may be joined together in the same line.

These attenuators are capable of handling up to 0·5W of sine wave power. In sharply pulsed conditions their rating is limited by the peak voltage of the signal. The limit of the voltage is of the order of 100V for pulses of a few microseconds duration.

Meetings this Month

THE BRITISH COMPUTER SOCIETY

Date: 20 January. Time: 6.15 p.m.
Held at: York Hall, Caxton Hall, Westminster, London, S.W.1.
Lecture: The Physical Principles of Xerography.
By: N. Levin.

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 29 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: Ultra-High-Speed Oscillography.
By: I. Maddock.

West Midlands Section

Date: 8 January. Time: 7.15 p.m.
Held at: The Wolverhampton and Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: Instrumentation of Space Vehicles.
By: N. R. Nicoll.

North Eastern Section

Date: 8 January. Time: 6 p.m.
Held at: The Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Lecture: The Earth Satellite Project.
By: P. H. Tanner.

South Midlands Section

Date: 31 January. Time: 7 p.m.
Held at: The Winter Gardens, Malvern.
Annual General Meeting, followed by Lecture: Digital Computers.
By: R. Deighton.

BRITISH SOUND RECORDING ASSOCIATION

Date: 17 January. Time: 7.15 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.
Lecture: The Electrical Production of Music.
By: Alan Douglas.

THE INSTITUTE OF NAVIGATION

Date: 17 January. Time: 5.15 p.m.
Held at: The Royal Geographical Society, Kensington Gore, London, S.W.7.
Lecture: The Influence of Atmospheric Conditions on Radar Performance.
By: J. R. Saxton.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Ordinary Meeting

Date: 9 January.
Lectures: The Development of Variable-Speed High-Power Drives for large Wind Tunnels and A Variable-Frequency Power Installation for Large Wind-Tunnel Drives.
By: P. McKearney, L. S. Drake and E. G. Mallalieu.
Speed Control of Large Wind Tunnels with particular reference to the R.A.E. 8ft by 8ft High-Speed Wind Tunnel.
By: L. S. Drake, J. H. Fox and G. H. A. Gunnell.
Data-Setting and Electrical Control Gear for the Flexible Nozzle and Diffuser of a High-Speed Wind Tunnel.
By: C. R. Dunham.

Measurement and Control Section

Date: 21 January.
Lecture: The Effect of Modern Industrial Requirements on Electrical Instrument Design.
By: G. F. Tagg.

Radio and Telecommunication Section

Date: 22 January.
Lecture: Special Problems of Broadcasting in Sweden.
By: E. Esping.
Date: 27 January.
Informal Evening: An Inquiry into the Specification of Transistors.
By: F. F. Roberts.
Date: 28 January. Time: 2.30 p.m.
Symposium: Long-Distance Propagation Above 30Mc/s.
(a) Ionospheric Forward Scatter Propagation.
Time: 5.30 p.m.
(b) Tropospheric Propagation Beyond the Horizon.

East Midland Centre

Date: 7 January. Time: 6.30 p.m.
Held at: The College of Arts and Crafts, Waverley Street, Nottingham.
Lecture: Earth Electrode Systems for Large Electric Stations.
By: J. H. Humphries.

North-Eastern Centre

Date: 13 January. Time: 6.15 p.m.
Held at: Neville Hall, Newcastle-on-Tyne.
Lecture: Earthing of Low- and Medium-Voltage Distribution Systems and Equipment.
By: F. Mather.
Date: 14 January. Time: 7 p.m.
Held at: Workington College of Further Education.
Repeat of above Lecture.

North-Western Centre

Date: 7 January. Time: 6.15 p.m.
Held at: The Engineers' Club, 17 Albert Square, Manchester.
Lectures: The Development of Variable-Speed High-Power Drives for Large Wind Tunnels and: A Variable-Frequency Power Installation for Large Wind-Tunnel Drives.
By: P. McKearney, L. S. Drake and E. G. Mallalieu.

North-Western Measurement and Control Group

Date: 21 January. Time: 6.15 p.m.
Held at: The Engineers' Club, 17 Albert Square, Manchester.
Lecture: Power System Protection, with particular reference to the Application of Junction Transistors to Distance Relays and: A Dual-Comparator Mho-Type Distance Relay Utilising Transistors.
By: C. Adamson and L. M. Wedepohl.

North-Western Supply Group

Date: 28 January. Time: 6.15 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: The Digital Computer Applied to the Design of Large Power Transformers.
By: W. A. Sharpley and J. V. Oldfield.

North Lancashire Sub-centre

Date: 6 January. Time: 7.15 p.m.
Held at: The North Western Electricity Board Demonstration Theatre, 19 Friargate, Preston.
Lecture: The BBC Sound Broadcasting Service on Very-High Frequencies.
By: E. W. Hayes and H. Page.

Northern Ireland Centre

Date: 13 January. Time: 7.30 p.m.
Held at: Sir William Whitla Hall, Queen's University, Belfast.
Lecture: Engineering Education in the Soviet Union.
By: P. E. Sleight and E. Giffen.

North Scotland Sub-centre

Date: 9 January. Time: 7 p.m.
Held at: Electrical Engineering Department, Queen's College, Dundee.
Lecture: The Remote and Automatic Control of Semi-Attended Broadcasting Transmitters.
By: R. T. B. Wynn and F. A. Peachy.
Date: 10 January. Time: 7.30 p.m.
Held at: Robert Gordon's Technical College, Aberdeen.
Repeat of above Lecture.

South-East Scotland Sub-centre

Date: 8 January. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Chairman's Address: Too Much and Too Little.
By: M. G. Say.
Date: 20 January. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: Some Aspects of Half-Wave Magnetic Amplifiers.
By: G. M. Ettinger.
Date: 21 January. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Informal lecture: The Importance of Research in Hearing and Seeing to the Future of Telecommunication Engineering.
By: E. C. Cherry.

South Midland Radio and Measurement Group

Date: 27 January. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Transistor Circuits and Applications.
By: A. G. Milnes. (To be read by G. W. E. Stark.)

Rugby Sub-centre

Date: 29 January. Time: 6.30 p.m.
Held at: Rugby College of Technology and Arts.
Lecture: Recent Developments in X-Ray and Electron Microscopy with Some Applications to Radio and Electronics.
By: C. W. Oatley and V. E. Cosslett.

Southern Centre

Date: 1 January. Time: 6.30 p.m.
Held at: C.E.A., Portsmouth.
Lecture: Cathodic Protection.
By: L. B. Hobgen, K. A. Spencer and P. W. Heselgrave.
Date: 8 January. Time: 6.30 p.m.
Held at: R.A.E. Technical College, Farnborough.
Lecture: Colour Television.
By: C. J. Stubbington.
Date: 15 January. Time: 6.30 p.m.
Held at: Brighton Technical College.
Lecture: The BBC Sound Broadcast Service on Very-High Frequencies.
Date: 24 January. Time: 6.30 p.m.
Held at: South Dorset Technical College, Weymouth.
Lecture: Gyroscopic Stabilizers.
By: H. Neal.

Western Centre

Date: 13 January. Time: 6 p.m.
Held at: Bristol University Engineering Laboratories.
Lecture: The BBC Sound Broadcasting Service on Very-High Frequencies.
By: E. W. Hayes and H. Page.

Western Utilization Group

Date: 27 January. Time: 6 p.m.
Held at: South Western Electricity Board, Bristol.
Informal Discussion on The 13th Edition of the I.E.E. Wiring Regulations.
Opened by: Forbes Jackson.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 6 January. Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Lecture: The Introduction of Subscriber Trunk Dialling.
By: A. J. Thompson and N. Walker.
Date: 22 January. Time: 5 p.m.
Held at: The Conference Room, 4th Floor, Waterloo Bridge House, London, S.E.1.
Lecture: Subs. Line Concentrators—Their Development and Application.
By: A. J. Barker and G. M. Blair.

THE TELEVISION SOCIETY

Date: 9 January. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: A French Portable Television Camera.
By: J. Polonsky.
Date: 23 January. Time: 7 p.m.
Held at: The Royal Institution, 21 Albemarle Street, London, W.1.
The Fleming Memorial Lecture: Crystal Valves.
By: T. R. Scott. (Admission by ticket only.)

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