Electronic Engineering

Vol. XXV MARCH 1953 No. 301

Commentary

IT may not have escaped the attention of our more observant readers that the February issue of ELECTRONIC ENGINEERING was number 300, and that we have therefore completed the first twenty-five years of our existence.

The first issue of the journal, from which ELECTRONIC ENGINEERING has descended, was published in March, 1928, under the title of *Television* and was the official journal of the Television Society formed in September, 1927, with Lord Haldane as the Society's first President.

Regrettably space does not allow a review of our own history during these twenty-five troubled years nor can we comment on the amazing growth of radio, television and electronics that has followed. Even in the short space of twenty-five years the pace has been tremendous and one has only to glance through the pages of the first issue to realize how far we have come. Radio—it was "wireless" in 1928—was well established then as a means of communication and broadcasting, and exciting new things were happening in television. The prophets of the day were forecasting all kinds of developments many of which, we may add, have since come true. But tucked away in a corner of our first issue was a statement by Lord Brabazon, then Lieutenant-Colonel Moore-Brabazon, M.P., in which he said he "trembled to think of the responsibilities of a future Postmaster-General when visual sights are transmitted as well as sounds."

* * *

No development in electronic science has, in recent years, so captured the imagination, of both the layman and the scientist alike, as the announcement of the transistor by the Bell Telephone Laboratories some four years ago. Since that time a large amount of literature has appeared, largely in America, but also in this country and in Europe, on the use of these components and the specialized circuits that are necessary for their operation. Indeed, one of the latest articles goes a step further and describes the functions of a germanium tetrode! Due to the difficulty of obtaining samples the initial interest had tended to die down somewhat, except in those research departments engaged on their development.

This interest has, however, been very much revived in the last two months by two events in the U.S.A. The first was the announcement, by an American firm, that they can now supply junction type transistors in production quantities. The second was the end-of-the-year statement by Brigadier General David Sarnoff, the Chairman of RCA. In this he stated, "In recent years a vast new field for exploration and development called 'electronics of solids' has opened in the scientific world. So impressive are the developments, and so important the potentialities for the future, that scientists are acknowledging electronics of solids as one of the most dramatic steps in technical progress." He went on to say, "Recognizing the great potentialities of transistors, RCA research men and engineers are developing them for mass production and are studying the multiplicity of new applications they make possible."

Previous to this, RCA had given, in November, 1952, a demonstration of equipment using transistors exclusively. The exhibits included a complete television receiver (with the exception of the C.R.T.), a radio-gramophone and a car radio receiver.

The Americans are certainly to be congratulated on these achievements although at present it is somewhat difficult to make a true assessment of their merits.

One effect of these announcements, however, has been to prompt engineers in this country to ask what are the British manufacturers doing in the way of transistor development, and are we behind the Americans in this field?

It is, of course, common knowledge that a number of firms in this country are actively engaged in transistor development and that at least one of them demonstrated an all transistor radio receiver as long ago as eighteen months.

Broadly speaking the position at the moment is that research is proceeding on sound lines and that several manufacturers can supply certain types of transistors in reasonable quantities for genuine research and development purposes, but they are definitely not yet available for mass production purposes.

Since the transistor was invented at the Bell Telephone Laboratories it is not of course unreasonable to expect them to still hold some lead. It must also be borne in mind that although an American firm claim to be in quantity production there is, as yet, little information available concerning their characteristics, particularly their temperature range and life characteristic, nor is it known what standard of uniformity is being achieved. It is one thing to produce one batch with similar characteristics, it is another to produce a large number of batches with uniform and pre-determined characteristics. This is very apparent when one considers that the germanium must contain certain impurities, but that these must be controlled to a ratio of less than one atom to every 10 000 000 germanium atoms.

At this point it is perhaps interesting to recall a statement made only two months ago by Dr. E. W. Engstrom, Vice-President in Charge of RCA Laboratories Division. He said, "We haven't yet worked out mass production techniques for transistors, although germanium itself is available, it requires careful processing to get it in the form that gives transistors their remarkable characteristics. Thus, the cost of even those few types of transistors that are available in limited quantities is still high." He also stated that RCA does not expect the transistor to supplant the electron tube any more than radio replaced the phonograph.

From all the evidence so far available we feel, therefore, that it is a sound policy to restrict the availability of transistors until such time that they are a thoroughly reliable component and can be mass produced with every confidence in their quality and uniformity.

The Amplification and Recording of Foetal Heart Sounds

By M. C. Wood,* M.A., M.D., D.M.R.

Knowledge of the condition of a baby before birth and during delivery is most important and is mainly assessed by listening to its heart beats. An apparatus was constructed to amplify these sounds, to record them and to demonstrate their visual characteristics on the screen of a cathode-ray tube. A tachyometer indicating beats per minute and a pen recorder were also incorporated. The apparatus is both an aid to clinical obstetrics and is used for research into foetal heart sounds.

In the field of obstetrics, knowledge of the condition of the fœtus (the baby before birth) is important. The chief method of examination is to listen to the heart beat through the mother's abdomen: alteration in the rate and rhythm in the later stages of pregnancy or during delivery may indicate distress of the baby.

The foetal heart rate is normally 120 to 180 beats per minute and as the rate is usually timed with a watch and the sounds are only faintly audible, it was decided to design and build an apparatus to investigate the foetal heart sounds and to record them.

An extensive survey of the medical literature has not revealed any published work on this subject in this country. Some work has been carried out abroad, mainly in America, and in most of it simple amplifiers were employed. No reference to the employment of cathode-ray tubes, pen recorders or tachyometers for this research could be found. Some of the findings claimed in these previous papers do not agree with the results we have obtained from the cathode-ray tube tracings. It must be emphasized that this work concerns sound only and has nothing to do with electrocardiography which records differences of electrical potential occurring in the cardiac muscle.

This apparatus was built as an experimental model for research and was never intended for routine work in midwifery. It has been in use for some time and some of the results obtained were the subject of a paper read at The Royal Society of Medicine, 28th November 1952, in conjunction with Mr. A. L. Gunn.

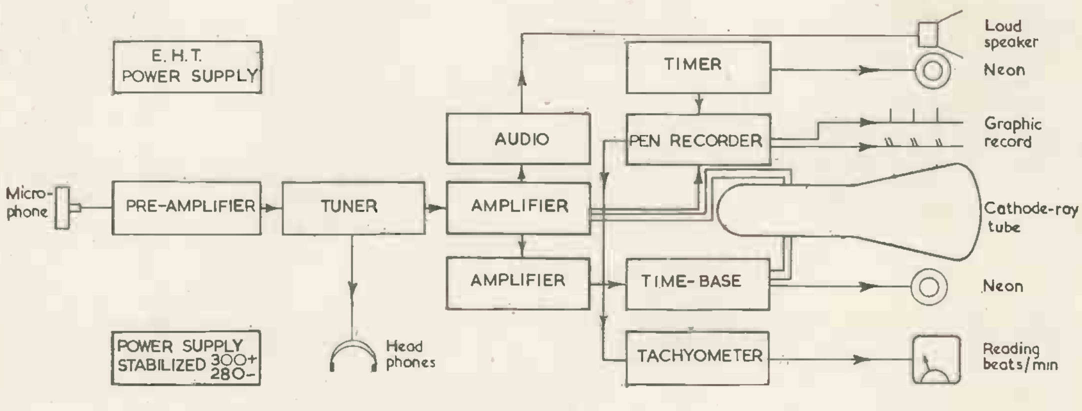

Fig. 1. Block diagram of the circuit

Method of Operation

A crystal microphone converts the audible sounds into corresponding electrical impulses. The signal so received is then amplified, tuned to whatever frequency is wanted and after further amplification may be lead into any of the following channels:

1. It is heard through earphones for monitoring to find the place where the fœtal heart is heard with maximum intensity.
2. It is passed through an audio output stage to drive a loudspeaker.
3. It deflects the Y plates of a built-in cathode-ray tube to demonstrate the visual characteristics of the foetal heart sounds.
4. It works a pen to provide a graphic record of the foetal heart sounds.
5. It passes into a tachyometer to indicate beats per minute on a dial.
6. It causes triggering of the time-base for the cathode-ray tube.
7. It is fed into an external oscilloscope for further inspection and photographing.
8. It illuminates a neon bulb, each flash indicating a beat.

* *Lewisham Hospital.*

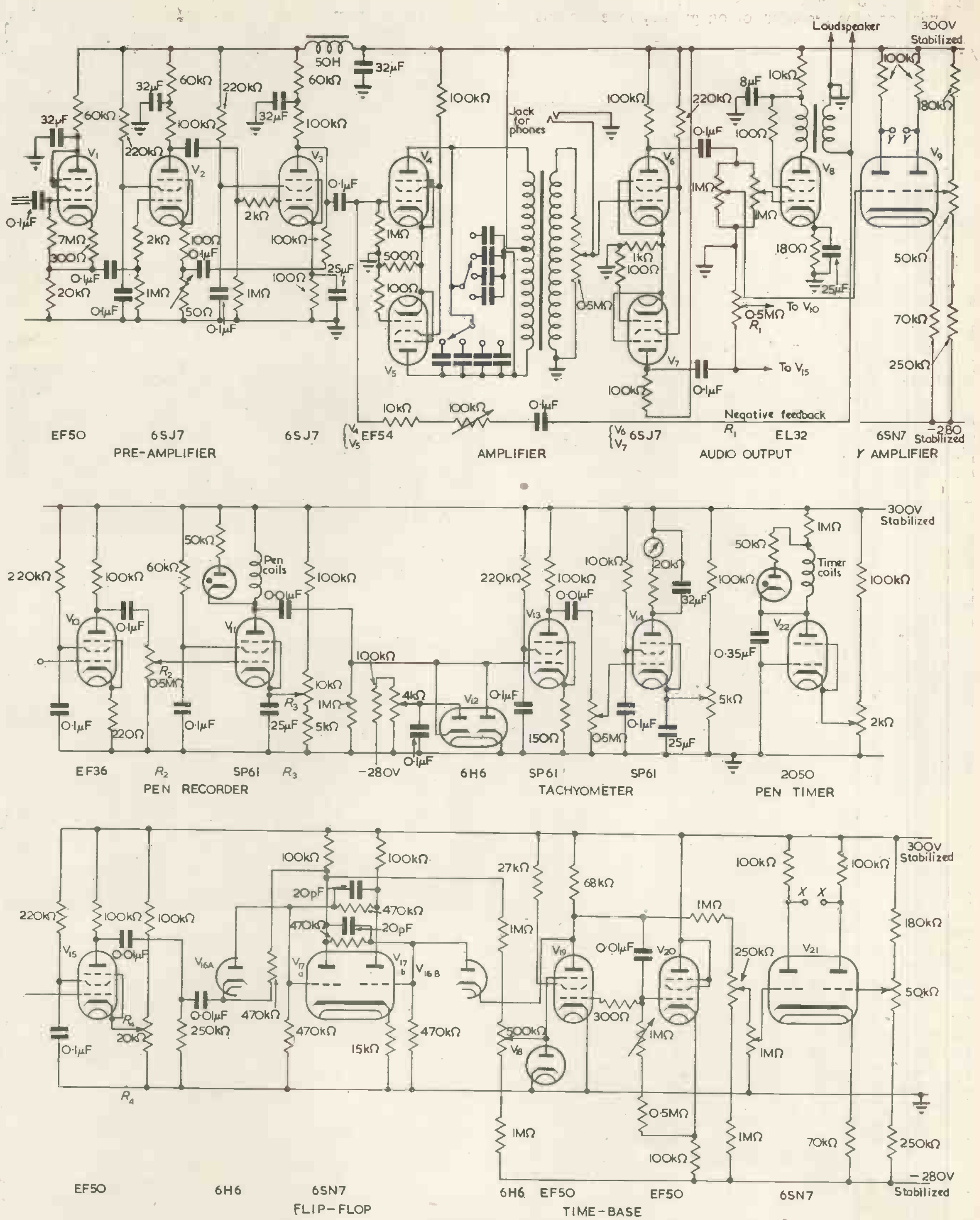

Fig. 2. Circuit diagram

9. The audible foetal heart sounds may be recorded on a wire or tape recorder, or on gramophone records.

Apparatus

This consists of:

1. A microphone and pre-amplifier.
2. Tuner and amplifier.
3. Audio output stage and negative feedback.
4. Pen recorder and pen timer.
5. Trigger circuit and time-base.
6. Cathode-ray tube and its timer.
7. Tachyometer.
8. E.H.T. and stabilized power supply.

A block diagram of the apparatus is shown in Fig. 1 and the circuit of the main units is shown in Fig. 2.

1. MICROPHONE AND PRE-AMPLIFIER[1]

A crystal microphone, enclosed in a special metal case for screening, has been used. This is fitted with a paper diaphragm and there is a small column of air between the skin of the maternal abdomen and this diaphragm. A rubber ring around the metal case maintains an airtight contact with the skin. The high impedance of the microphone is matched by a cathode follower V_1, the signal received is amplified through two pentode stages V_2 and V_3, together forming the pre-amplifier: variable negative feedback is provided over these two stages.

Fig. 3. Photograph of the apparatus seen from the front

2. TUNER AND AMPLIFIER

The main amplifier consists of a pair of cathode coupled pentodes V_4 and V_5 whose anode loads are formed by each half of a centre tapped primary winding of a transformer. Selection of frequencies is effected by tuning the primary circuit. Coarse and fine adjustments are provided. At 80c/s it is possible to obtain a 15 cycle bandwidth. A high quality mu-metal transformer was to hand and used, but its secondary winding was not centre tapped. Output from this secondary winding provided the signal when required for earphones. These are used to find the site on the abdomen for the microphone giving the maximum signal intensity. The signal is further amplified by a second cathode coupled pair of pentodes V_6 and V_7. A long tailed pair, V_9, then provides push-pull deflexion for the Y plates of the cathode-ray tube and also the Y shift.

3. AUDIO OUTPUT STAGE AND NEGATIVE FEEDBACK

This is of conventional type, but negative feedback is obtained from the secondary winding of the output transformer and returned to the input of the main amplifier. This may be so arranged to convert the amplifier to one of constant gain which has been found to be most useful.

4. PEN RECORDER AND PEN TIMER

The signal from the amplifier is again amplified by V_{10} and then passed to V_{11} which is biased well beyond cut-off so as to accept only the peaks of the signal. The coils of the pen recorder form the anode load (P.O. type relay). The recorder was home-built and designed to record an indication for each beat of the foetal heart and not to respond to the individual components of the beat. A small neon with a series resistor is in parallel with the pen coils and gives a flash for each heart beat. The neon, from its brightness also gives a visual indication of the current passing through the coils. This can be adjusted to a reasonably steady level by R_1, R_2 and R_3. A thyratron oscillator is incorporated for use with the pen recorder to give an indication each second and a neon bulb simultaneously flashes. This neon and that working with the pen recorder can be seen in the photograph of the apparatus (Fig. 3) above the cathode-ray tube.

5. TRIGGER CIRCUIT AND TIME-BASE

A Miller time-base has been employed with a cathode-

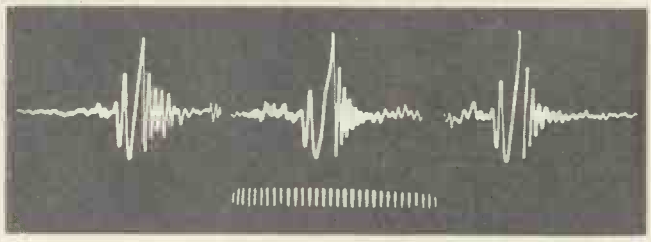

Fig. 4. Photograph of a trace on the cathode-ray tube of three normal foetal heart sounds. Timer indicates 1/50th of a second

Fig. 5. Further photograph of a trace of three foetal heart sounds

follower and a long tailed pair V_{21} to provide push-pull deflexion for the X plates and also for the X shift.

It was found desirable when showing the visual characteristics of the heart sounds to be able to superimpose the trace of each sound upon the previous one. Alternatively 4 or 5 beats were required to be shown in one scan, and with a long persistence tube these also could be superimposed scan upon scan. The start of the first impulse of the sound triggers the time-base and subsequent sounds have no effect until a full scan has been completed.

A time-base with trigger circuit described by Dickenson[2] was after a little modification found convenient. In its quiescent state the Miller valve, V_{19}, is cut off by negative voltage on g_3. The start of the first sound changes the stable state of the flip-flop V_{17}. V_{17b} now conducts, the potential on the anode of V_{17a} rises removing the negative cut-off from V_{13}. V_{19} then starts to conduct, its anode potential falls at the rate controlled by the rate at which the charge on the grid leaks away. After a full sweep the coupling from the anode of the Miller valve to the grid of V_{17b} cuts off the current of this valve returning the flip-flop to its original state, ready to accept the next impulse. R_4 controls the level of the signal at which triggering of the circuit occurs.

6. Cathode-ray Tube and its Timer

A 3FP7 tube was available and has proved satisfactory. To show the time relationship of the foetal heart sounds, a trace on the screen of 50c/s is obtained from the mains supply. On a double beam tube the time trace and the heart sounds could be produced simultaneously and photographed. In this apparatus, however, single beam tubes had to be used, the traces of the heart and 50c/s oscillations were photographed separately and later superimposed as shown in the illustrations.

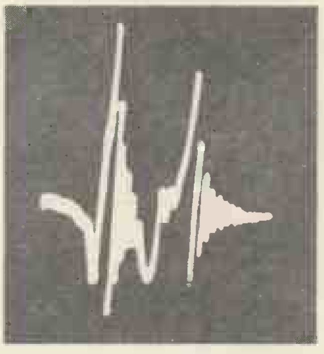

Fig. 6. Photograph of a trace on cathode-ray tube of heart sounds shortly after birth

stabilized negative as well as positive supply. E.H.T. supply has been obtained from a 1 800 volt transformer.

Results

Fig. 4 shows photographs of the trace on the cathode-ray tube of three foetal heart sounds obtained from one case taken at approximately half minute intervals. There are three main components in each sound and the average duration of each sound is $7/50^{th}$ of a second.

Fig. 5 shows the trace of three sounds from another case in which the interval between the first and second components is increased. The pattern differs somewhat from the previous case.

Fig. 6 shows the trace of the heart sound shortly after birth; the adult appearance with two distinct sounds is taken on shortly after birth.

Fig. 7 shows a tracing of the pen recorder over a period of 69 sounds and demonstrates the regularity of the beat. This recording was taken during a uterine contraction.

Fig. 8 shows another trace with greater magnification.

The recording of the foetal heart sounds is not an easy procedure. The signal received from the microphone is small—usually 200 to 500 microvolts, and sounds from

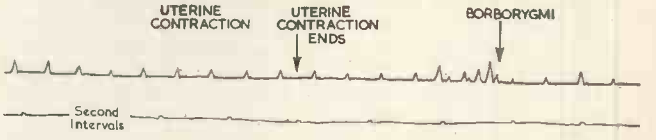

Fig. 8. Photograph of another tracing by pen recorder with greater magnification

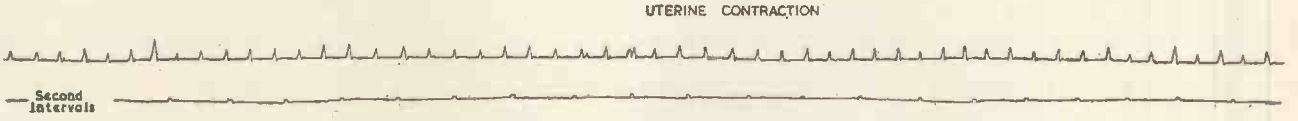

Fig. 7. Photograph of a tracing by the pen recorder indicating the foetal heart sounds

7. Tachyometer

The impulses used for energizing the pen coils are reasonably constant, but in addition V_{12} acts as a limiter with a negative output. This limited output is amplified and used to activate V_{14} which is biased beyond cut-off. A meter in the anode lead of this valve is calibrated to read beats per minute. This has given good results as a cardio-tachyometer.

8. Power Supply

A conventional series valve stabilized power supply described by Dickenson has been employed, providing

the abdomen such as borborygmi and skin noises have a frequency note that is in the same range as the foetal heart sounds.

In the building of this apparatus cost was a limiting factor, so that it had to be constructed almost entirely out of surplus equipment. The valves used were not ideal, but they were the only ones available.

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1. Kelly, S. Piezo Electric Crystal Devices (Part 3). *Electronic Engng. 23*, 173 (1951).
2. Dickenson, C. J. Electrophysiological Technique. (Electronic Engng. Monograph, 1950).

A Synthetic Surveillance Radar Trainer

The Ministry of Civil Aviation Air Traffic Control Experimental Unit has developed a radar trainer that will synthesize the flight paths of aircraft as seen by any modern surveillance radar and, in addition, will permit the flight paths of aircraft to be altered independently at will. The over-all conception of the trainer is based on electro-mechanical calculators which continuously calculate the position in range and bearing of the synthetic aircraft from the radar site. The angular position of shafts representing range and bearing are used to produce a short pulse which appears in the correct position on the normal type Planned Position Indicator. In order to make the trainer independent of actual radar equipment, all the normal data usually obtained from the real equipment is produced artificially—aerial turning information, time-base trigger, range markers and permanent echoes.

The new Trainer will be used at the Ministry's Air Traffic Control School at Hurn. It will effect a considerable financial saving by reducing the period required to train Air Traffic Control Officers as Traffic Directors, from 8 to 6 weeks with more officers on each training course, and aircraft hours for practical work reduced by approximately 25 per cent or 50 hours.

The principles used in this Trainer have been applied to the development of another device, viz., a simulator to assist in experiments and investigations in the development of air traffic control techniques and procedures which have had to be carried out by the use of real aircraft at very great expense. Although such equipment cannot remove completely the need for full-scale tests with real aircraft, it will reduce them to a fraction of what they would otherwise be. It will also enable problems to be simulated that cannot well be reproduced in trials.

Pulse-Operated Time Bases

By H. A. Dell*, Ph.D.

IT is often necessary to examine the fine details of complicated repetitive electrical signals with the aid of a cathode-ray tube. This is difficult if only a normal recurrent time-base is used.

If clear steady images on the tube screen are to be produced, precise synchronization between the time-base sweeps and the signals to be displayed is essential. The fastest time-base that can fulfil this condition is that which displays a single complete cycle of the signal on the screen. The characteristics of a signal as a function of time are usually measured from the geometrical position of detail on the linear time-base trace. In this case an accuracy of only about one two-hundredth part of the repetition period is therefore all that can be achieved, due to the finite spot size, distortions associated with the screen and similar causes.

It is sometimes possible to run the time-base at a higher repetition rate than that of the signal, when only one of several sweeps is truly synchronized. A greater length of time-base is then available, but the general instability makes the measurement of further detail very difficult.

This measurement problem is particularly acute in radar apparatus. The difficulty here is that for accurate range determinations the delay time of reflected pulses must be found, often in the 1μsec to 10μsec region, although the pulse repetition frequency (P.R.F.) of the radar transmitter may be only about 1kc/s. If a normal recurrent time-base were used, synchronized to a transmitter having this repetition frequency, accurate measurement would be impossible.

However, in this case it is only the part of the time-base trace immediately following the transmitter impulse that is of interest. It is not difficult to arrange then, that the time-base circuit, though synchronized with the transmitter pulses, sweeps the cathode-ray spot right across the tube face in the first fraction of the interval between pulses, so that only the first 50μsec (say) of the trace appears on the screen. The measurement of the restricted fraction of the time-base displayed is then possible to the required accuracy.

One way of making the circuit do this is to generate very large beam deflecting signals so that the cathode-ray spot is off the screen for a high proportion of the time. This is wasteful because large valves and high currents are necessary for the period when the spot is never seen. A better method is that of generating a complex time-base signal suitable for sweeping the cathode-ray spot rapidly to one side of the tube and then holding it there until the instant of flyback. Still synchronized with the transmitter impulse, it then returns to the other side and quickly sweeps across a second time. This latter method is economical as large deflecting signals are not necessary, but during the rest period the cathode-ray spot lies on or near the edge of the tube screen. The brilliance of this may well be disturbing, so steps are usually taken to cut off the beam when it would be stationary, by applying suitable potentials to one or more of the intensity modulating electrodes of the electron gun.

From the circuit point of view, it is often more convenient to make the part of the time-base cycle for which the spot is to be stationary lie at the beginning of the trace and not at the end. In other ways the system is then similar in operation.

In taking this step away from the repetitive time-base, a new feature has been introduced in the radar application. The range time-base is now no longer broadly synchronized with the P.R.F. of the transmitter. Each time-base trace is instead precisely started by each transmitter impulse. Thus the time-base circuits are in a stable, or semi-stable state until the receipt of an initiating impulse from the transmitter, when they immediately go through one cycle of operation and settle to stability once more. These circuits, in which the speed of the time-base traverse is independent of the repetition rate, are therefore sometimes called " pulse operated " time-base circuits.

An incidental advantage of this feature is that it becomes possible to synchronize each time-base trace with the transmitter impulse, although the latter either by accident or design may not be perfectly regular.

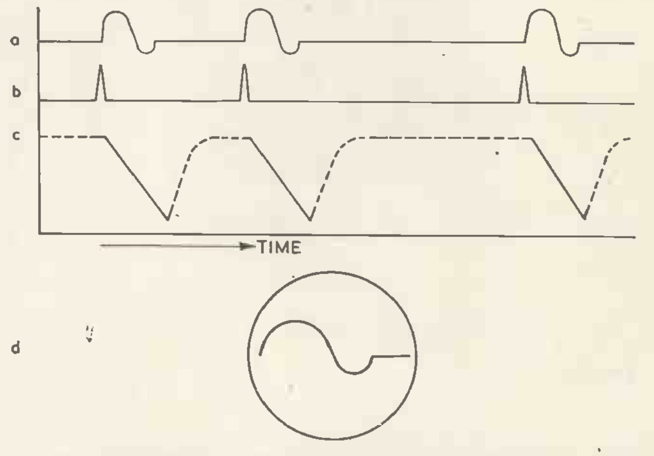


Fig. 1. Waveforms produced with pulse-operated time-bases

(a) Series of randomly spaced signals of similar form.
(b) Pulses produced by signals (a).
(c) Time-base deflexion of cathode-ray spot produced by pulses (b).
(d) Resulting display on the cathode-ray tube screen.

The necessary time-base starting instant must be matched to each transmitter impulse if certainty in range measurement is to be achieved. By designing the time-base to allow for such irregularity in starting, and by initiating it from the transmitter pulse itself, this matching can be ensured.

General Uses of Pulse Operated Time-Bases

Many experimental observations which are strictly of a transient character can be regarded as fundamentally repetitive, but with no fixed repetition period. The waveforms generated in irradiated Geiger-Müller tubes, the waveforms associated with spark breakdown in gases, even the waveforms met with in automatic telephone installations or in vehicular traffic control systems, are all instances in which similar waveforms are produced repeatedly, but separated by an unpredictable and probably random time interval.

When signals of this type are to be investigated it is usually possible to arrange matters so that the arrival of each one produces, in some subsidiary circuit, a single sharp pulse. This can then be fed to a pulse operated time-base so that a single time-base stroke of any convenient length is produced and on it the signal may conveniently be displayed.

* *Mullard Research Laboratory.*

This is illustrated diagrammatically in Fig. 1 in which (a) represents the series of randomly spaced similar signals which are to be examined. Each signal is made to produce a pulse (b) which in turn produces a time-base deflexion of the cathode-ray spot represented by (c). Only the linear sweep of each time-base waveform is made visible on the tube, since the flyback and waiting period (shown dotted) are normally blacked out. The resulting display on the tube is shown in Fig. 1(d).

So long as the pulse producing circuit is sensitive and quick in operation, little of each signal is lost before each time-base stroke starts, but in order that a steady, measurable image may be produced it is important that each trace on the tube should lie precisely over the similar images of previous signals. The position of the spot on the tube screen at the start of each time-base sweep must, therefore, be precisely maintained however long the interval that has elapsed since the last trace was produced. This means that in the quiescent state a stable system is essential.

If the waveforms being examined recur frequently, it may happen that the fast time-base circuits which are used to examine them cannot recover their original condition after each sweep ("re-set") soon enough to show the next similar part of the waveform on the tube screen.

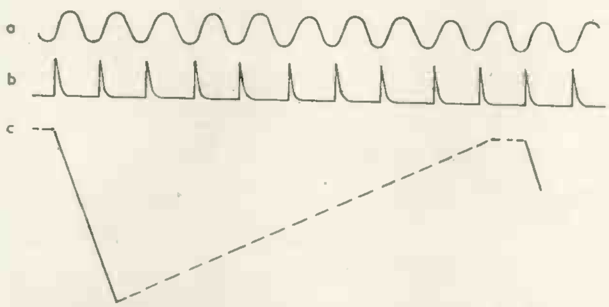


Fig. 2. Display of a 1Mc/s sine wave using a pulse-operated time-base

(a) Normal sine wave.
(b) Trigger pulses produced by sine wave (a).
(c) Time-base waveform initiated by first of triggering pulses (b).

For example, it is not difficult to sweep a cathode-ray tube spot across the screen, linearly with time, in 1μsec. It is, however, very difficult to do this more frequently than about once every 10μsec.

When a 1Mc/s sine wave is to be shown by such a time-base, the condition illustrated in Fig. 2 exists. Here (a) represents the sine wave and (b) the triggering pulses produced by it. The first of these initiates the time-base waveform (c). Nine other pulses will occur before the circuit is ready to produce a second sweep. Only one cycle in ten of the original waveform will therefore be seen.

The time-base circuits must be carefully designed to ensure that triggering impulses, which may occur after the impulse initiating the sweep, have no effect on the system until it is ready to produce a time-base again.

Delay Circuits Used with Time Bases

When transient waveforms are long and complex, another difficulty arises. A pulse-operated time-base which starts at the beginning of each transient can only be of a low speed if it is to show all of the waveform on the tube screen. Faster time-bases can only show the beginning of each transient, so that it is impossible to examine a detail which occurs late in the transient on a more extended time scale.

This is illustrated in Fig. 3 where (a) represents the transient which carries a detail of some interest x. The starting pulse produced by the transient is represented by (b) and (c1) and (c2) represent respectively the slow and fast time-base deflexions available. The detail at x occurs too late to be seen on the time-base (c2) and the best trace available is that using the time-base (c1). The appearance of the cathode-ray tube screen is illustrated in Fig. 4(a).

This difficulty can be overcome by inserting a variable delay circuit after the pulse producing circuit. A new time-base starting pulse delayed by a convenient interval from the beginning of each transient may then be generated, so that a fast time-base can be used to examine

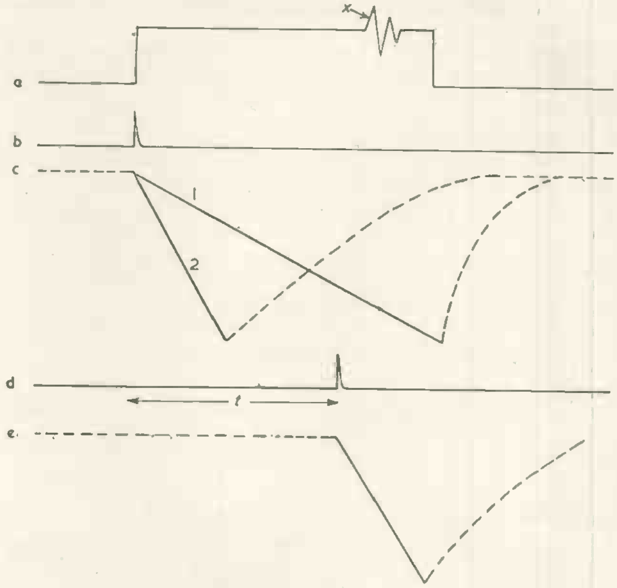


Fig. 3. Pulse-operated time-bases for examining long and complex transient waveforms

(a) Transient carrying a detail of some interest "x."
(b) Starting pulse produced by the transient (a).
(c) Time-base deflexions available (1) slow deflexion, (2) fast deflexion.
(d) Delayed pulse.
(e) Time-base initiated by delayed pulse (d).

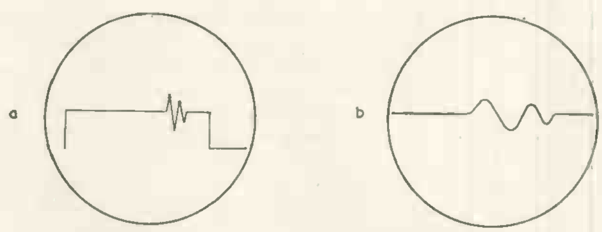


Fig. 4. Transient waveforms appearing on screen of cathode-ray tube

(a) Using time-base 1 shown in Fig. 3(c).
(b) Fast trace using delayed pulse.

any part of the waveform. This delayed pulse is represented in Fig. 3(d), and the time-base which it starts by (e). The appearance of this fast trace on the tube screen is illustrated in Fig. 4(b).

It is important to make sure that the time delay circuit is precise in operation, as any unwanted variation in delay (t) between successive pulses makes a repetitive trace blurred. This is particularly so when a fast time-base is used at an extended delay.

On signals with negligible time uncertainty, such delay circuits make it possible to examine minutely the whole of a repeated waveform. This is an advantage if synchronization can only be carried out effectively at one point of each cycle.

Practical Time-Base Circuits

A schematic diagram of the components of an oscilloscope time-base incorporating the features that have been

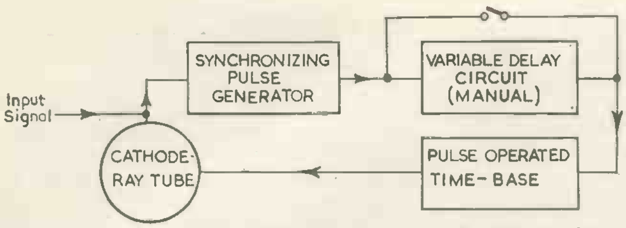


Fig. 5. Schematic diagram of the components of oscillograph time-base

described is shown in Fig. 5. The principal units, all of which must be suitable for operation from regularly repetitive or randomly spaced signals, are as follows:—

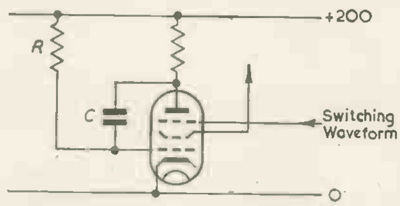
Synchronizing Pulse Generator

The waveform to be examined is first fed into a circuit which generates a series of synchronizing pulses from it. In the case of a regular cyclic waveform (as a pure sine wave), one pulse must be generated for each cycle. It does not matter at what point in the cycle the pulse is generated, but it must be stable in phase. In the case of irregularly spaced transient signals, one pulse must be generated for each separate transient, and it must lie as near as possible to the beginning of each one.

Variable Delay

The synchronizing pulses are then fed to a variable

Fig. 6. Typical circuit of a suppressor-switched time-base



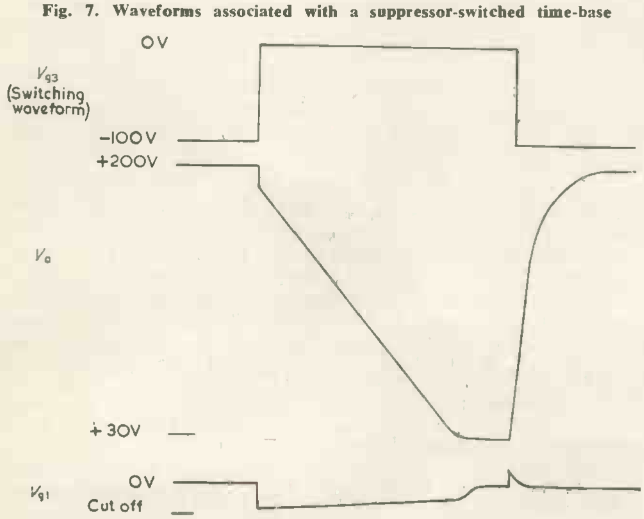
delay circuit which generates a second series of pulses. These are the time-base starting pulses and each lies at some controllable predetermined time after each synchronizing pulse.

Time-Base

The time-base starting pulses are fed to a time-base generator which produces a waveform to deflect the cathode-ray tube spot at the desired speed. The circuit also provides a brightening pulse to feed to the intensity modulating electrodes of the cathode-ray tube.

None of the circuit techniques necessary to make an oscilloscope on these lines is entirely new, and many possible variants of pulse generators, time-bases and delay circuits have already been described[1]. There are

Fig. 7. Waveforms associated with a suppressor-switched time-base

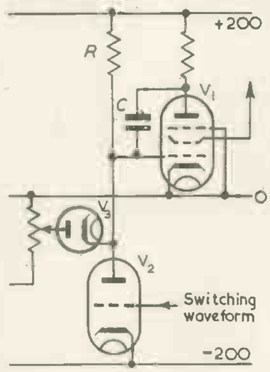


some details in connexion with the time-base proper and the delay circuit, however, which are of interest here.

Linear Time Base Circuit

The generation of linear sawtooth voltages by the Miller integrator circuit is well known. However, one feature of this method which is sometimes overlooked, becomes a limiting factor when high speeds are considered. A typical circuit, called a suppressor switched time-base, is illustrated in Fig. 6, and the waveforms associated with it in Fig. 7.

Immediately before the arrival of the switching waveform on the valve suppressor grid, the control grid is at or near the cathode voltage, and grid current flows. Immediately after the waveform has arrived, the rise in anode current brings the anode voltage down. This fall is passed to the grid via the capacitor C, so that the grid voltage falls to a point near to grid cut-off. Only after this does the true Miller charging condition set in and the linear part of the anode excursion begin. The anode waveform thus includes an initial drop of about 5-8 volts which is followed by the linearly falling potential which is required.



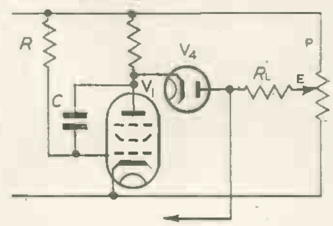


Fig. 8 (left). Circuit for preventing "initial drop" on sawtooth potential

Fig. 9 (above). Circuit in which the length of the switching wave is controlled by the sawtooth itself

Due to stray capacitances, this initial drop is not instantaneous and it usually takes at least $\frac{1}{4}$-$\frac{1}{2}\mu$sec to drive the grid down. When time-bases having total sweep times of only 1μsec are used, this drop therefore merges into the beginning of the linear sweep so that no simple voltage-time relationship exists.

A convenient way of preventing this drop is to arrange that the anode current of the Miller valve is never completely cut off. In the resting condition the control grid is held at some negative potential within the grid base of the valve. This circuit is illustrated in Fig. 8. The Miller valve is V_1 and in the quiescent condition the control grid is held at, say, -4V with respect to the cathode by the flow of current to the anode of V_2 through R. The cathode of V_2 is fed from a negative line.

If V_2 is now cut off suddenly by a negative-going square wave on the control grid, the Miller valve V_1 at once allows the capacitor C to be charged in the usual way. The descending sawtooth at the anode of V_1 thus begins. There is no initial drop as the valve V_1 is conducting the whole time.

Practically, it is important to ensure that the anode of V_2 does not fall to too low a potential before a sweep begins. This can be arranged by a subsidiary diode V_3 (Fig. 8) which prevents the anode of V_2 from falling more than a fraction of a volt below that of the anode of V_3. Excess current to V_2 then flows through V_3 rather than R.

In use, the switching waveform fed to the grid of V_2 (Fig. 8) must last long enough for the anode of the Miller valve V_1 to reach the lower limit of its travel. It must not, however, be too long, or this will lower the maximum repetition rate. A possible method of controlling the length of the switching wave by the sawtooth itself is illustrated in Fig. 9.

An additional diode V_4 and load resistor R_L is connected between the anode of V_1 and a steady potential E (of about 100V) derived from a potential divider P. While the anode voltage of V_1 is high, the diode will not conduct, so that no signal will appear across R_L. After the time-base has started the voltage on the anode of V_1 falls, and a time comes when it equals E. At this moment the diode begins to conduct and the sawtooth voltage is fed to R_L. The appearance of this signal can be used to terminate the switching waveform on the grid of V_2 (Fig. 8).

A complete circuit embodying these features is illustrated in Fig. 10. The Miller valve, at the anode of which the sawtooth voltage appears, is V_1. The switching valves which control it are V_2 and V_5. They are in the form of

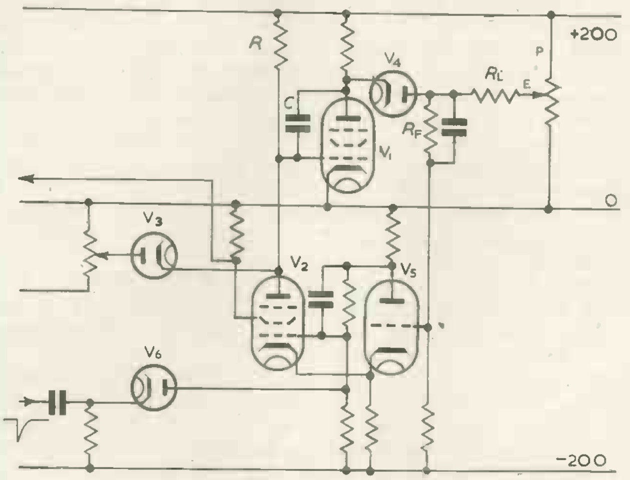


Fig. 10. Complete circuit of time-base

a D.C. connected, cathode-coupled multivibrator which has two stable positions of rest. Normally V_2 is conducting, and V_5 is biased to cut-off, the anode voltage of V_1 being high.

A negative pulse fed into the grid of V_2 reverses this condition so that V_2 is cut off and the potential of the Miller valve anode begins to fall. V_2 and V_5 are stable in this condition until some of the sawtooth voltage, fed through the diode V_4 appears at the grid of V_5. This resets the multivibrator, so that V_5 is cut off and V_2 brings down the grid voltage of V_1. The circuit is then ready to receive another triggering impulse.

The diode V_6 is provided to isolate the multivibrator circuit from triggering impulses when V_2 is cut off.

It is profitable to make V_2 a pentode valve with a small resistor in the screen-grid circuit. Across this resistor a potential, suitable for feeding to the control grid of the cathode-ray tube, is developed. This brightens the trace on the tube during the run-down of the Miller valve anode.

Delay Circuit

When a variable delay circuit is to be made, the accurate voltage-time relationship of the Miller time-base can be used to advantage. In the time-base, the limit of the excursion of the Miller valve anode is fixed by the voltage E on the anode of the diode V_4. If E is varied, the fraction of the time-base which elapses before the diode starts to conduct can be varied. This is therefore a convenient form of delay circuit.

The D.C. coupling resistor R_F is replaced by a purely A.C. coupling circuit. A rectangular pulse is then produced on the anode of V_5 which is precisely determined by the speed of the run-down of the Miller valve anode and by the setting of E. When the front edge of this delay pulse is produced by the synchronizing signal, the back edge can be used to start a time-base circuit. The delay in starting can be varied by the setting of E, which can thus be calibrated for any given Miller circuit.

Acknowledgments

The author wishes to acknowledge with thanks the work which Mr. M. N. Goldsmith, Mr. C. H. J. Beaven and others of his colleagues have put into the development of the apparatus described, and their assistance in the preparation of this paper.

Permission to publish was kindly given by the directors and management of the Mullard Research Laboratory.

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A New Magnetic Tape "Memory"

By M. Lorant

A computor-controlled external auxiliary "memory", using magnetic tape as the recording medium, is extending the problem-solving capacity of SEAC, the U.S. National Bureau of Standards Eastern Automatic Computor.

In the new system, the magnetic tape rests lightly on two smooth-surfaced rollers that rotate continuously but in opposite directions. Between these two rollers the tape passes through magnetic heads for recording, pick-up, and erasing. When either of two control solenoids is energized, a low-inertia rubber-covered roller presses the tape against one of the smooth rollers. This quickly starts the tape moving in the desired direction. Tape inertia is kept low by letting each end fall in loose folds into a tank or bin. Each tank consists of two plates of glass spaced just a little more than the width of the tape.

While computor-controlled auxiliary memory systems using magnetic pulses have been employed successfully in computors other than SEAC, most of these systems have required complex and expensive mechanisms to start, stop, and reverse the magnetic tape or wire with the necessary speed.

The speed and simplicity of the new system result from the successful elimination of reels and servomechanisms. Only two small masses need to be accelerated in starting the tape, the small jam roller and a few feet of tape hanging into the tank. The tanks are enclosed on all edges and have slots in the top for the tape to enter. Because the tanks are just wide enough to clear the tape, the loose folds in which the falls have no tendency to turn or to become tangled.

Several problems have been encountered in developing the tape memory mechanism. For one thing, the tape tends to acquire an electrostatic charge as it passes through the drive mechanisms. This can become quite troublesome at higher speeds, causing the tape to cling to the walls of the tank as soon as it leaves the drive mechanism. The present solution—satisfactory at moderate speeds (up to 8 feet per second)—is to ionize the air where the tape leaves the drive unit, using strips of alphapolonium. The ideal solution would probably be to make the tape base material sufficiently conductive, so that a charge could not collect.

A second limitation on the operating speed of the tape is imposed by the need to erase information from the tape. New heads with powdered iron magnetic circuits now being developed should solve this problem.

A third major problem is the presence of flaws in the magnetic tape. Commercially available tape has many small imperfections in the magnetic oxide coating, which are quite undetectable in ordinary audio work. In the recording of computor pulses, however, the loss of magnetic signals over a very small area in the tape may mean the loss of one or more digits of information. Some specially-treated tapes now available are nearly free from flaws; and improved manufacturing techniques may soon eliminate such difficulties. Meanwhile a process of removing the imperfections by scraping the recording surface over a suitably shaped knife-edge has been developed.

The Noise Factor of Centimetric Superheterodyne Receivers

By J. H. Evans*, A.M.I.E.E., A.M.Brit.I.R.E.

THE superheterodyne receiver comprising a crystal frequency changer, klystron local oscillator, and an intermediate frequency amplifier, is well established for the reception of centimetric waves, and it seems likely to remain the standard for some time to come. Only at decimetre wavelengths do valve amplifiers or electron multipliers seem likely to give a noise factor even approaching those obtainable at present. Yet by a suitable arrangement of standard components the noise factor of a centimetric receiver can be made to vary over very wide limits. This article is intended to assist equipment designers to make a reasonable estimate of the noise factor to be expected from these various arrangements, and of the performance deterioration to be expected from circuit simplification.

Definition

There are several definitions of the expression "receiver noise factor", but, numerically there is no significant difference between them. The one favoured by the author is:

$$N = 10\log_{10}\frac{\text{Signal-to-noise ratio at aerial}}{\text{Signal-to-noise ratio at receiver output}}$$

Another definition useful for calculation purposes is:

$$N = 10\log_{10}\frac{\text{Actual noise power output from receiver}}{\text{Theoretical (KTB) noise output from an equivalent perfect receiver}}$$

It can be shown that the noise factor N of the receiver of the type under consideration is:

$$N = L_R + L_{TR} + L_C + 10\log_{10}(N_C + N_{LO} + N_{IF} - 1)$$

where L_R is the mismatch and transmission loss in the centimetric receiver

L_{TR} is the transmission loss of the T.R. cell

L_C is the conversion loss of the crystal

N_C is the noise/temperature ratio of the crystal (i.e. the ratio of its noise power output to that of an equivalent resistor)

N_{LO} is the apparent increase in N_C caused by local oscillator noise

N_{IF} is the noise figure of the I.F. amplifier

$$\text{(i.e.)}\ \frac{\text{signal-to-noise ratio at amplifier input}}{\text{signal-to-noise ratio at amplifier output}}$$

(It will be noticed that N_{IF} is expressed as a ratio, not in decibels as this is more convenient, in this application.)

Effect of Individual Parameters

It will be shown that, assuming all components to be used under optimum conditions, N may vary from better than 6db up to worse than 20db, corresponding to a ratio of 2 to 1 in range of a primary radar equipment. Thus the importance of understanding the effect of the individual parameters cannot be overestimated.

The "Direct Loss" Parameters

The first three terms in Equation (1) L_R, L_{TR} and L_C may be termed "Direct Loss" parameters as any improvement in any of them results in the same improvement in N. Their effect is thus immediately apparent.

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The Mismatch and Transmission Loss in the Centimetric System (L_R)

Transmission losses in the waveguide are not usually included in the receiver noise factor, and, in any event, are usually well known to the designer. A.T.R. or equivalent losses may well become appreciable, however, and also signal loss into the local oscillator channel. In a simple receiver this may well be directly derived from the attenuation in the coupling between the oscillator and crystal, due allowance being made for frequency sensitivity.

The mismatch loss due to the crystal itself is usually significant. Reference to the manufacturers' published data gives for the CV 253/103 types a figure of 2 or 2·5:1 corresponding to a mismatch loss of 1 to 1·5db. However, in later types considerably lower values may be expected, around 0·2 to 0·3db.

Thus L_R may easily amount to 3db and is rarely less than 1·5db unless particular attention is paid to it during design. This reduction is normally paid for by increased complexity.

In this connexion it may be of interest to note that it is common practice to separate the T.R. cell from the crystal by such a distance that reactive variations in the crystal may be tuned out by mistuning the cell in an endeavour to reduce L_R. This arose when the admittance spread on crystals was very much greater than is now permitted and was mainly in one direction. A gain in performance on current types is only possible on 50 per cent of crystals and this is obtained at the expense of T.R. cell protection on all crystals. The maximum of 1db is only possible on, at most, a few per cent and is rarely worthwhile.

The Transmission Loss of the T.R. Cell (L_{TR})

The specification for tunable T.R. cells gives a variation in this figure ranging from some 0·5db for the best 10cm cells to 1·5db for standard 3cm cells. However, the writer is of the opinion that a somewhat higher figure may be expected under operating conditions during life—possibly as high as 3db.

The loss of the wideband cells now coming into use is considerably less than this. 0·2 to 0·3db can be expected throughout life.

The Conversion Loss of the Crystal (L_C)

Again referring to the manufacturers data, the conversion loss of crystals lies between 4·5 and 7db at 10cm and 6 to 9db at 3cm according to type. The spread between crystals of the same type is approximately ± 1db.

This figure includes the 3db loss by half the signal being transformed to sum frequency and lost.

There is a second order effect on conversion loss, which, as far as the author is aware, has not been investigated. This is the increase in conversion loss when the crystal is subject to overload from a magnetron pulse. The effect may well be thermal, but its magnitude seems difficult to determine. Its most obvious indication is a decrease in local oscillator current when the magnetron is switched on. The current recovers to its usual value after the magnetron power has been removed.

The "Indirect Loss" Parameters

The terms N_C, N_{LO}, and N_{IF} in Equation (1) are obviously interdependent and the effect of an improvement in any one of them is more difficult to assess.

THE NOISE TEMPERATURE RATIO OF THE CRYSTAL (N_C)

The specification limit for this parameter is either less than 2 or less than 2·5 times according to type. In the case of CV364/291 crystals, a numerical value is not quoted; but N_C must be of the same order, or other clauses in the specification cannot be complied with.

Occasionally N_C is considerably larger, usually because of mechanical instability in the unit. Similarly, excess T.R. cell leakage causes a rise in N_C before any other centimetric parameter is affected. (The D.C. characteristic changes first when the crystal is subjected to overload.)

THE NOISE FIGURE OF THE I.F. AMPLIFIER (N_{IF})

This figure may be made to vary considerably according to frequency and circuit design. Table 1 shows some typical theoretical values that may be obtained with various input circuits. Figures within a few per cent of these values can be obtained by valve selection, but in any event the spread is not very great, a mean about 10 per cent worse than the theoretical is generally obtainable.

The figure quoted for N_{IF} is independent of bandwidth Δf up to the point where:

$$\Delta f \text{ approaches } \frac{1}{2\pi C R_a}$$

(Where C is the total valve input capacitance including transformed mixer capacitance; and R_a is the optimum valve input load.)

provided that the gain of the first stage is large enough to make second stage noise negligible.

TABLE 1
Typical Noise Figures

VALVE COMBINATION		NOISE FIGURE	
INPUT	2ND STAGE	TIMES	DECIBELS
EF50	EF50	3·7	5·7
EF91	EF91	2·9	4·6
EF54	EF50	2·2	3·4
EF91 Triode connected	EF91	2·1	3·2
EF91 & EC91 in Cascode	EF91	1·6	2·0

These values are calculated for an input coil Q of 100 and a first-stage anode load of 1kΩ.

R_a varies with valve type and frequency, but is normally around 1 to 2kΩ, to which value the crystal I.F. impedance is transformed. Δf may be increased by a factor approaching 2 by using a bandpass input transformer. If a still wider bandwidth is required the degradation in noise figure may be calculated fairly readily. The method is described elsewhere[6].

The design procedure to be followed for low noise amplifiers is beyond the scope of this article, in any event it is amply covered in the literature. Broadly speaking it is to mismatch the crystal to the input valve grid by means of a transformer to an extent determined by the transit time loading and equivalent noise resistance of the valve. Circuit losses in the input have also to be considered and, where space is a consideration, may become the limiting factor.

THE APPARENT INCREASE IN N_C CAUSED BY LOCAL OSCILLATOR NOISE (N_{LO})

This is a somewhat complex quantity that does not lend itself readily to numerical analysis. The local oscillator generates shot noise in a similar manner to any other thermionic valve and this, attenuated by the crystal conversion loss, is fed into the I.F. amplifier.

As the noise is generated in a resonant cavity, at frequencies remote from the local oscillator frequency, the noise will decrease by an amount dependent inversely on the Q of the cavity. Therefore N_{LO} becomes progressively less as the I.F. is increased, and increases with increasing

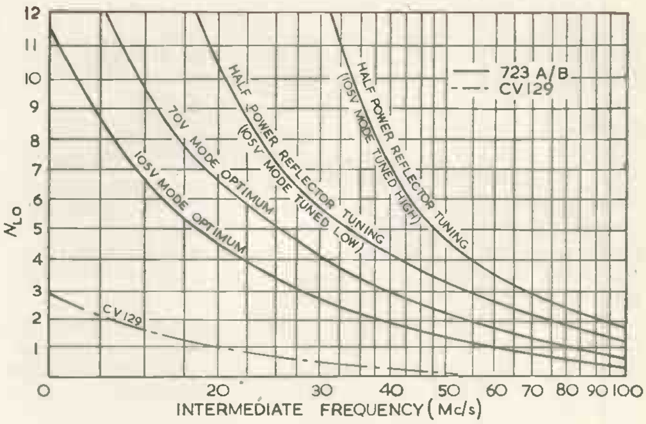


Fig. 1. Typical 3cm klystron noise figures

signal frequency. N_{LO} also varies considerably from mode to mode of the klystron. Typical curves are shown in Fig. 1. It is advisable to measure the value of N_{LO} for the case under consideration before deciding on a particular oscillator.

Additional complications to N_{LO} are introduced by another effect. As is well known, for any given resonator setting there is an optimum reflector voltage corresponding to maximum power output. At this setting N_{LO} is minimum, but at the extremes of the reflector tuning range it may be increased many times. For example, the 723 A/B gives a value of N_{LO} at optimum of the order 1 to 2 for a 45Mc/s I.F. At the extremes of the tuning range values of the order 10 may be expected. This case may occur if an A.F.C. system is controlling the oscillator.

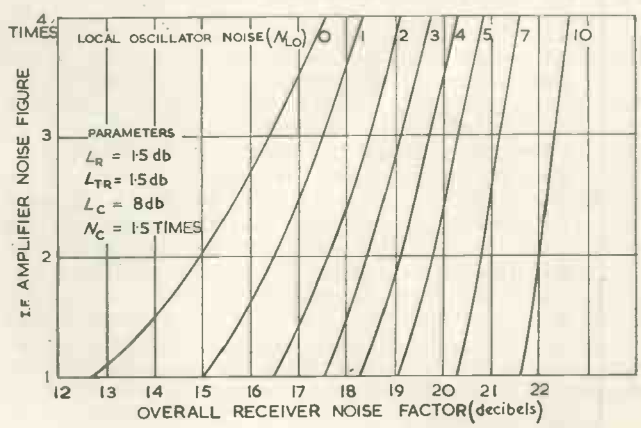
Local oscillator noise may be reduced to a negligible amount by:

(1) Using a high intermediate frequency
(2) Using a high Q oscillator
(3) Feeding a local oscillator signal via a high Q filter,
(4) Using a balanced mixer.

There are, of course, disadvantages attached to each method, the main points being:

(1) The improvement is counteracted by the rise in N_{IF}.
(2) These usually require much higher electrode voltages than standard low Q oscillators, and have a smaller electronic tuning range.
(3) Makes A.F.C. impossible without complicated mechanical servo systems.
(4) Complicates the waveguide and mixer design, and

Fig. 2. The effect of I.F. noise figure and local oscillator noise on receiver noise factor



the increase in stray capacitance causes difficulty when applied to very broad band systems.

Summary

It will be seen from the foregoing that the parameters most closely under the designer's control are local oscillator noise and I.F. amplifier noise figures. The effect of these on the receiver noise factor may be seen in Fig. 2. If direct loss parameters are changed their effect may be estimated directly by addition or subtraction from the right-hand scale.

It must be realized that the figures quoted apply only to receivers working under approximately optimum conditions. If appreciable mismatch or frequency sensitive reflexion is present in the waveguide considerable variations in the noise factor can occur. Normally, of course, these reflexions are degrading but an improvement of 3db is theoretically possible by reflexions at sum or image frequency, and 2db has been realized in practice. This is not thought to have any practical application at present however, and, as far as the author is aware, no details have, as yet, been published.

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A Note on Phase-Angle Measurements Using a Cathode-Ray Tube

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

TWO methods for determining the phase-angle between two sinusoidal quantities with the aid of a cathode-ray tube have been given by Benson and Carter[1]. The methods involve the production of two voltages proportional to the quantities concerned and applying these to the horizontal and vertical deflector plates of the tube. In this way an ellipse is traced out on the screen of the tube and the phase-angle can be found from the dimensions of the ellipse.

For the first method, which has been shown to be the more satisfactory of the two, if the ellipse cuts the horizontal centre line at B and C (Fig. 1) then $\sin\alpha = BC/2X$ where α is the phase-angle. The errors produced by trace thickness and any harmonics present in the voltages have

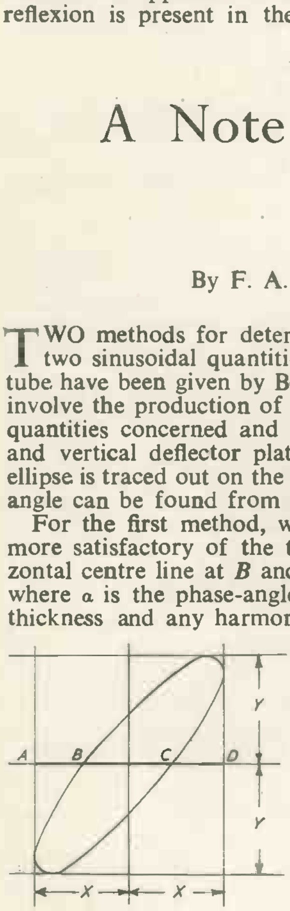

Fig. 1. The single ellipse method

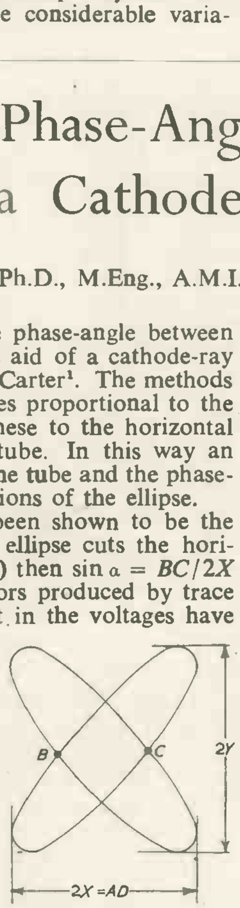

Fig. 2. The double ellipse method

been estimated by Benson and Carter, but they have neglected another source of error, namely that due to the correct location of the horizontal axis of the figure on the screen. In practice, this axis is usually located by adjusting the magnitude of the voltage applied to the Y plates so that the top and bottom of the figure coincide with rulings on a graticule in order that the axis lies along another ruling. A similar procedure is usually adopted to measure $2X$. This method suffers from the disadvantage that the rulings on the graticule may not be parallel to the horizontal axis of the figure. Thus, errors will be introduced in determining phase-angle, additional to those investigated by Benson and Carter, if the graticule is used to measure the distances BC and $2X$. It is shown below that using a double-beam cathode-ray tube, the accurate position of the centre line can be determined quite easily.

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The "horizontal" signal deflects both beams of the tube simultaneously. The "vertical" signal is applied so as to deflect only one beam. Before applying the "vertical" signal, however, the vertical-shift controls are adjusted to bring the two beams into coincidence. Thus, when the "vertical" signal is applied the "inactive" beam forms the horizontal centre line, cutting the ellipse at B and C (Fig. 1). The length of the centre line AD is equal to $2X$ since, in general, the two beams are deflected horizontally an equal amount (by the same X plates). Hence, in this way, the points A, B, C and D are fairly precisely located and the distances AD and BC can be measured with good accuracy.

A further advantage of this method of determining phase-angle is that the extremities of the ellipse can be off the screen, providing the line AD is completely on. Thus, a greater accuracy is possible for a given cathode-ray tube screen than by the previous method.

An alternative method, which has been tried but is not quite so successful, is to apply the "vertical" signal to both sets of vertical deflector plates, so that one set of plates deflects the beam upwards when the other beam is deflected downwards. Under these conditions two ellipses result, cutting each other at points B and C (Fig. 2). AD can still be found accurately. This method gives good results, but, in general, it is found that the two ellipses are not always exactly of the same size, owing to slight discrepancies in the vertical deflexion sensitivities of the two beams of a given tube.

In the second method described by Benson and Carter α is calculated from the expression $\sin\alpha = (2a)(2b)/(2X)(2Y)$, where $2a$ and $2b$ are the lengths of the major and minor axes of the ellipse respectively and $2X$ and $2Y$ are the lengths shown in Fig. 1.

The present methods using a double-beam tube also offer some advantages in this case. The traces of Fig. 1 enable $2X$ to be determined more accurately than by the original method. The traces of Fig. 2 allow both $2X$ and $2Y$ to be measured more accurately than before. Neither of the present methods, however, gives any greater accuracy in measuring $2a$ or $2b$.

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Time-Division Multiplex Systems

(Part 3)

Pulse-Length and Pulse-Position Modulated Systems

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

THE improvement in signal-to-noise ratio which can be obtained by using constant-amplitude pulses instead of amplitude-modulated pulses led to the development of systems using pulse-length modulation (P.L.M.) and pulse-position modulation (P.P.M.). A. H. Reeves[43] carried out tests on P.P.M. systems in Paris in 1938. Experimental work on pulse-length modulation was started by the Ministry of Supply in 1941 in order to obtain the security inherent in the use of centimetre waves and highly directional aerials. This work led to the development of the Army No. 10 set[45,46,47] which operated on 4 540 and 4 870Mc/s and provides an eight-channel pulse-length modulated T.D.M. system. The No. 10 set came into operational use just before D-day and was extensively used on the continent during the subsequent advance. Information regarding the No. 10 set project reached the U.S.A. in 1942 and led to the development of similar equipment[48]. The AN/TRC-5 set[49] operating in the range of 1 350 to 1 500Mc/s was designed by the RCA and the AN/TRC-6 set[50] operating in the range of 4 350 to 4 800Mc/s was designed by the Western Electric Company. The AN/TRC-6 appeared in Europe late in 1945. Both these sets use P.P.M. and provide eight channels using T.D.M. Later systems[30,31,38,51] have continued to use P.P.M. because of its improved signal-to-noise performance compared with P.L.M.

Modulation and Demodulation

In the usual method of producing pulse-length modulation, the modulating signal voltage is added to a sawtooth voltage waveform as shown in Fig. 22(a) and the composite signal is applied to a valve which is biased beyond cut-off. The valve begins to conduct when the voltage applied to its grid is about half the peak voltage of the sawtooth. When a modulating voltage is added to the sawtooth, the time at which the valve conducts is made earlier by a positive signal and later by a negative one. The valve is switched off again by the rear edge of the sawtooth, so its anode current consists of a train of pulses whose rear edges occur at fixed instants and whose front edges are modulated as shown in Fig. 22(b). If, however, the shape of the sawtooth is as shown in Fig. 22(c) the front edges of the pulses will occur at fixed instants and the rear edges will be modulated. If a symmetrical triangular waveform as shown in Fig. 22(d) is used, the time of the centre of the pulse is fixed and both front and rear edges are modulated.

Because of the length of its grid base a valve will not turn fully on or off immediately, so the pulses of anode current have appreciable rise- and fall-times unless the sawtooth and modulating voltages are much greater than the grid base of the valve. The operation is improved by replacing the valve with a circuit which uses positive feedback to obtain a rapid transition between the conducting and non-conducting conditions. One such circuit is the Schmitt trigger circuit[52] shown in Fig. 23(a). When its grid is near earth potential valve V_1 is cut off because of the large P.D. across the common cathode resistor R_K caused by the cathode current of V_2. As soon as the grid of V_1 is made sufficiently positive for it to begin

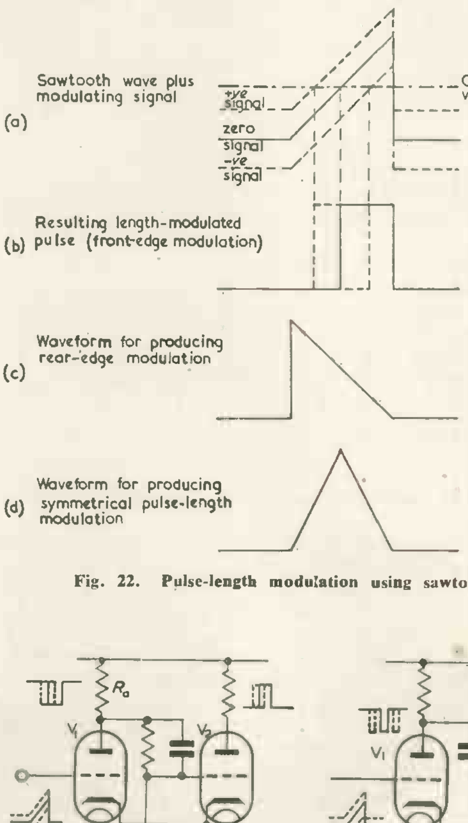

Fig. 22. Pulse-length modulation using sawtooth wave

Fig. 23. Trigger circuits

to conduct, the P.D. across R_a lowers the grid potential of V_2, thus reducing the cathode current of V_2 and the bias applied to V_1 by R_K. The change-over of current from V_2 to V_1 is thus accelerated and is completed during a very small increase in the grid voltage of V_1. A similarly small

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decrease in grid voltage of V_1 is required to turn V_1 off, the change-over again being accelerated by the feedback. A negative output pulse can be obtained from the anode of V_1 or a positive pulse from the anode of V_2. The D.C. coupling between V_1 and V_2 which is shown in Fig. 23(a) can often be replaced by an A.C. coupling of appropriately long time-constant.

In order to produce position-modulated pulses the modulating signal is also added to a sawtooth waveform, but the sum is applied to a trigger circuit which is turned on when the sum exceeds its triggering voltage but, instead of being turned off by the rear edge of the sawtooth, it turns itself off after a shorter time. In this way a train of pulses is generated, each of which has a fixed length, but whose time of occurrence with respect to the start of each sawtooth is modulated by the signal. A suitable pulse-trigger is shown in Fig. 23(b). Initially valve V_2 is conducting and V_1 is cut off, but as soon as the grid of valve V_1 is made more positive than the triggering potential the current flow changes from V_2 to V_1 as described in relation to Fig. 23(a). Because the time-constant C_1R_1 is short, however, the grid of V_2 rises rapidly towards the H.T. voltage as C_1 charges through R_1, until valve V_2 conducts again turning off V_1. Each time the grid potential of V_1 becomes more positive than the triggering potential the circuit therefore generates a pulse whose length is determined by the time-constants R_1C_1.

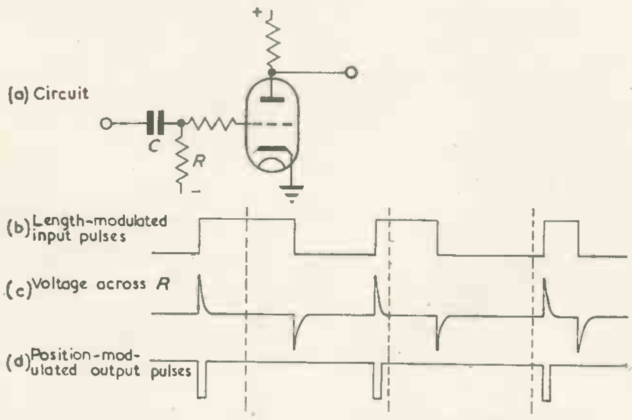


Fig. 24. P.P.M. to P.L.M. conversion by differentiating circuit

Another method of producing P.P.M. is first to generate a train of length-modulated pulses and then to convert these to position-modulated pulses by means of a differentiating circuit as shown in Fig. 24(a). If the circuit has a time-constant CR which is small compared with the length of the input pulses its output will consist of short spikes of opposite polarity at the beginning and end of each pulse as shown in Fig. 24(b). The polarity of the input pulses is chosen so that the positive spike coincides with their modulated edges. The valve is biased beyond cut-off and only conducts during the positive spikes, thus giving position-modulated output pulses as shown in Fig. 24(d).

In a T.D.M. system it is essential that each pulse shall occur only during its allotted time interval. With the methods of modulation described above, the trigger circuit can be operated by the modulating signal in the absence of the sawtooth if at any time it exceeds the triggering voltage. A voltage limiter must therefore be fitted at the input to the modulator. It is also possible to obtain a form of mis-operation known as pulse splitting[53]. If the modulating signal is changing rapidly, the sum of the modulating and sawtooth voltages can cut the triggering level twice with the result that two pulses are generated by the trigger circuit.

Because the modulators required for P.L.M. and P.P.M. are more complex than those for P.A.M. it is often economical for a T.D.M. system to use a pulse-amplitude modulator for each channel and for the outputs from the channel modulators to be connected to common apparatus for converting the amplitude modulation to P.L.M. or P.P.M.[38] There is no need for voltage limiters because each output pulse can only occur during the corresponding amplitude-modulated pulse. One method uses channel modulators which produce amplitude-modulated pulses which are trapezoidal in shape, as shown in Fig. 25. The trapezoidal pulses are fed to a pulse-trigger whose times of operation therefore depend on the amplitudes of the trapezoidal pulses, thus giving position-modulated output pulses. Systems based on the use of trapezoidal pulses have been described by Levy[32] and Kirby[31].

A method of converting P.A.M. to P.L.M. or P.P.M. using the circuit shown in Fig. 26 has been described by Potier[54]. The gates of all the channels are connected to produce

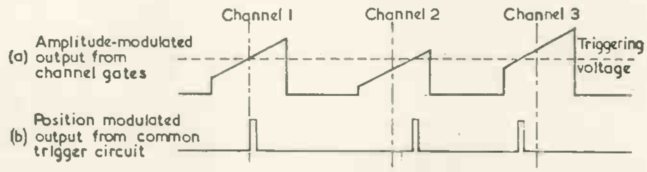


Fig. 25. P.A.M. to P.P.M. conversion using trapezoidal pulses

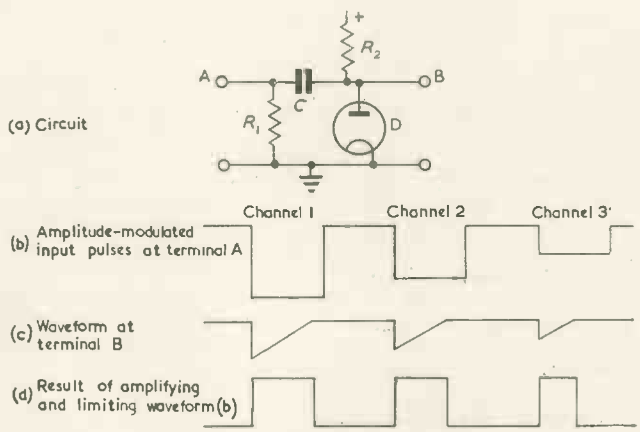


Fig. 26. P.A.M. to P.L.M. conversion

rectangular amplitude-modulated negative pulses across the common load resistor R_1. In the absence of an input pulse, a constant current from the positive supply is flowing through the high value resistor R_2 and the diode D. Each pulse which appears across R_1 causes the potential of point B to become negative, cutting-off the diode. Current from the constant current source then flows to charge C and the potential of point B rises linearly until the diode conducts again clamping the potential of point B to earth. The output comprises a train of triangular pulses whose durations are proportional to the heights of the rectangular pulses applied to terminals AB. By amplifying and limiting the triangular pulses a train of rectangular length-modulated pulses can be obtained as shown in Fig. 26(d). P.P.M. can be obtained by differentiating the length-modulated pulses as shown in Fig. 24.

When P.P.M. or P.L.M. is produced by adding the modulating signal to a sawtooth waveform, the resulting displacement of the edge of the pulse is proportional to the value of the modulating signal at the time at which the edge occurs; the times at which the signal waveform is sampled thus depend to some extent on the signal itself When P.P.M. or P.L.M. is obtained by conversion from pure P.A.M., the position of each modulated pulse edge is determined by the values of the modulating signal at the periodic sampling times of the P.A.M. system. Parks and Moss[55] have called the latter method periodic scanning and the former method synchronous scanning.

The frequency spectrum of a train of pulses modulated in length or position by a single audio frequency contains the audio frequency (f_m), the pulse repetition frequency (f_r), its harmonics (nf_r) and upper and lower sidebands about the P.R.F. and each of its harmonics. Unlike P.A.M., in which each sideband contains only a single frequency ($nf_r \pm f_m$), the sidebands in P.L.M. and P.P.M. contain an infinite series of harmonics of the audio frequency ($nf_r \pm qf_m$, $q = 1, 2 \ldots$) as shown in Fig. 27. Some of the

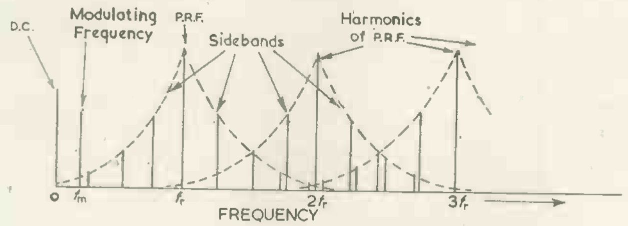


Fig. 27. Spectrum of a modulated pulse train

components of the lower sideband on the P.R.F. will be of lower frequency than f_m; these sideband components cannot be separated from the modulating signal by means of a low-pass filter and so are present as intermodulation distortion components in the output signal. Several authors[58,59,60,61,62] have made exact or approximate calculations of the spectra of modulated pulse-trains and have thus estimated the distortion produced. The distortion is usually negligible if the P.R.F. is greater than three times the highest audio frequency and may not be serious even when the P.R.F. is only double[58,61]. The distortion might be serious, however, if the modulating signal were a group of channels multiplexed by frequency-division as inter-channel interference would then be caused. When P.P.M. or P.L.M. is obtained by conversion from pure P.A.M. and is subsequently reconverted to pure P.A.M. at the receiver this inter-modulation distortion is eliminated[55] because of the simple sideband structure of the P.A.M.

The spectrum of a train of length-modulated pulses contains a component at the frequency of each component in the modulating signal with an amplitude which is independent of the frequency. The pulse-train can therefore be demodulated by passing it through a low-pass filter which cuts off at the highest frequency in the modulating signal, thus removing all components at higher frequencies. A train of position-modulated pulses, however, contains little energy at the modulating frequencies and the ratio between the amplitude of each component and the corresponding component of the modulating signal varies with the frequency of the component. A position-modulated train of pulses cannot therefore be demodulated correctly by means of a low-pass filter.

One method of demodulating position-modulated pulses is to convert them to pulse-length modulated pulses which are then demodulated by a low-pass filter. P.P.M. can be converted to P.L.M. by means of a trigger circuit which is turned on by each incoming P.P.M. pulse and switched off by a pulse which occurs at a fixed time after the unmodulated position of each incoming pulse, thus generating pulses whose front edges are modulated. Alternatively the trigger circuit can be switched on by pulses which occur at fixed times and turned off by the incoming position-modulated pulses, thus generating pulses whose rear edges are modulated. When a circuit for converting P.P.M. to P.L.M. is provided for each channel of the system it can also be used to perform the gating function for its channel. The trigger circuit of Fig. 23(a) can be modified for this purpose by replacing V_1 (which is normally non-conducting) by a pentode valve with a gating pulse applied to the screen. This circuit was used by Reid[28]. Another combined gating and P.P.M. to P.L.M. converting circuit has been described by Kirby[31]. Alternatively, the trigger circuit which is used to convert P.P.M. to P.L.M. can be made common to all the channels[32]. Each channel still requires an individual gating valve to select the correct P.L.M. pulse train, but an overall saving in the number of valves can be obtained.

Pulse-position modulation can also be converted to amplitude modulation. McGuire and Nowacki[56] have described a circuit in which each channel has a gate which is opened by the incoming pulse and applies a sawtooth waveform to a pulse lengthening circuit. The capacitor in the pulse lengthener is charged to the voltage of the sawtooth wave at the time at which the incoming pulse occurs and thus provides an output voltage proportional to the position of the modulated pulse. The output from this circuit contains some position-modulation as well as amplitude-modulation, but negligible distortion results if the depth of the amplitude-modulation is large and the peak displacement of the position-modulated pulses is small compared with the pulse period (as it usually will be in a T.D.M. system). A more complicated circuit which converts position-modulation to pure amplitude-modulation has been described by Moss and Parks[55]. Other authors have described modulating and demodulating circuits for P.L.M. and P.P.M. which make use of specially designed secondary emission valves[30], cold-cathode gas-filled valves[67] and cathode-ray tubes[11].

Noise

The effect of noise introduced during transmission can be minimized by the technique known as slicing. Taking a narrow slice from the middle of the received pulse substantially removes both the noise present between the pulses and the noise present on the tops of the pulses as shown in Fig. 28. One way to obtain a narrow slice is to use an overloaded valve, but the pulse applied to the valve must greatly exceed its grid base. An improvement can be obtained by using a trigger circuit such as that shown in Fig. 23(a) which requires a much smaller voltage change to turn it on and off. If several links are connected in tandem, the pulses can be reshaped by slicing at each intermediate repeater station.

After the slicing process, the only noise which is effective is that which is superimposed on the rise and fall of each pulse and becomes apparent as a fluctuation of the times of occurrence of the front and rear edges, unless the amplitude of the noise itself is sufficient to cut the slicing level. If this occurs the slicing circuit may generate false pulses when no input pulse is present or may obliterate

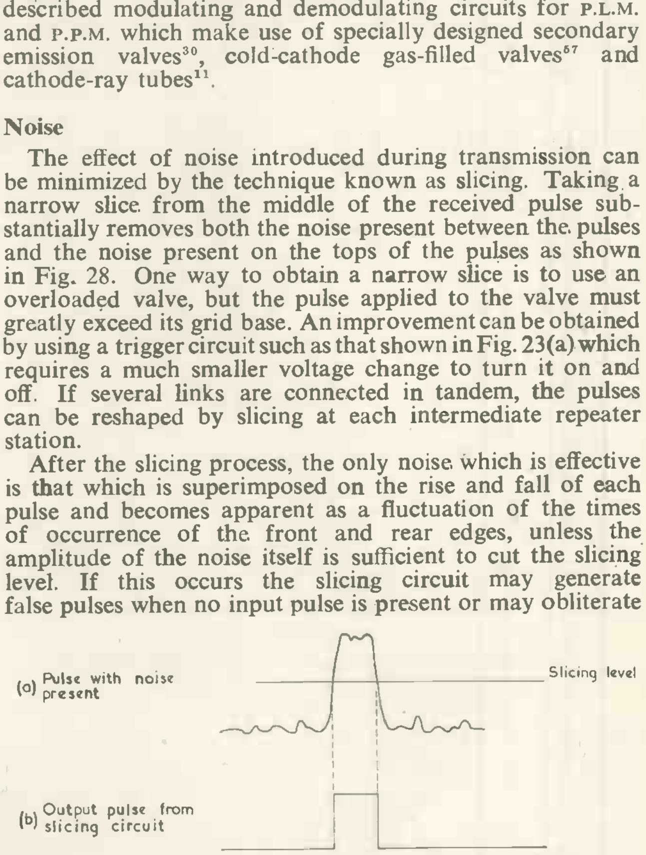


Fig. 28. Removing noise by means of a slicing circuit

pulses which are present, thus giving a very poor signal-to-noise ratio at the output of the receiver. The two types of noise present on radio links are impulsive noise and fluctuation noise. Impulsive noise is usually the result of electrical storms or man-made interference and can occur at a high level; it is, however, of less frequent occurrence at the high frequencies used for T.D.M. links than at lower frequencies. Similar effects to those caused by impulsive noise can be caused by interfering transmissions or by multi-path transmission[63]. Fluctuation noise originates as thermal noise and valve noise at the input to the receiver which is apparent as a "hiss" in the output. Several authors[18,43,44,61,65,66,67] have investigated its effects on pulse transmission.

Because of its random nature, fluctuation noise has an almost uniform frequency spectrum and the probability distribution of its amplitude with time closely follows the Gaussian law[68]. The rate of occurrence of peaks of noise which cross the slicing level thus decreases rapidly with improvement in the signal-to-noise ratio at the input as shown in Fig. 29. The curve shown in this figure is derived in the Appendix. When the slicing voltage is four times the R.M.S. noise voltage it will be exceeded for only 63 millionths of the time. As the signal-to-noise ratio is increased still further, mis-operation of the slicing circuit directly by the noise becomes a negligible source of output noise. There is thus a pronounced threshold effect. When the signal-to-noise ratio at the input of the receiver is made worse than about 15db the signal-to-noise ratio at the output deteriorates rapidly, but when the signal-to-noise ratio at the input is improved beyond the threshold the output signal-to-noise ratio improves more slowly. The signal-to-noise ratio at the output therefore varies with that at the input as shown in Fig. 30.

When the signal-to-noise ratio exceeds the threshold, some noise will still be present in the output because of fluctuation of the times of occurrence of the edges of the output pulses from the slicing circuit (jitter). If the pulses had zero rise- and fall-times they would be immune from interference, but this would require infinite bandwidth. Fig. 31 shows a pulse $v(t)$ in the presence of a disturbing voltage $v_x(t)$. Let the front edge of the pulse cut the slicing level V at time t_1 when there is no disturbance and let the slope of the front edge of the pulse at the slicing level be $(dv/dt)_{t_1}$. If there is a small disturbing voltage $v_x(t)$, the time at which the front edge of the pulse cuts the slicing level is $t_1 - \delta t_1$, where:

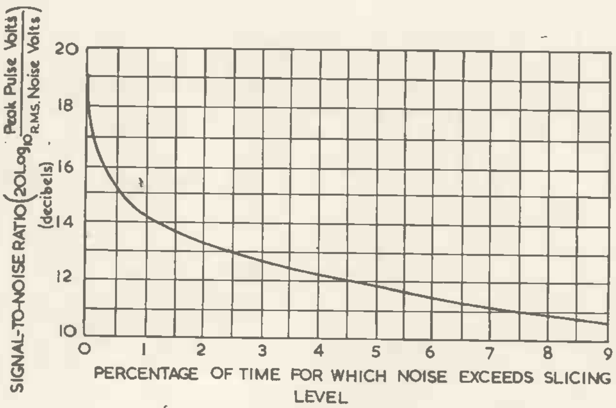


Fig. 29. Probability of fluctuation noise exceeding slicing level (*slicing level equals half pulse height*)

$$\delta t_1 = \left(\frac{dt}{dv}\right)_{t_1} v_x(t_1) \quad \ldots\ldots\ldots\ldots\ldots (1)$$

Let the rise-time of the pulse be defined as:

$$T_r = V_p \left(\frac{dt}{dv}\right)_{t_1} \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

where V_p is the peak height of the pulse. Then:

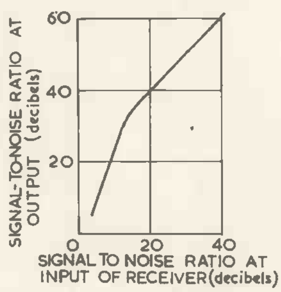
$$\delta t_1 = T_r \frac{Vx(t_1)}{V_p} \quad \ldots\ldots\ldots\ldots\ldots\ldots (3)$$

If the peak displacement of the front edge of the pulse caused by the modulation in a P.P.M. system is T_D and the R.M.S. noise voltage is V_x, then:

$$\frac{\text{R.M.S. displacement due to noise}}{\text{Peak displacement due to modulation}} = \frac{T_R}{T_D} \cdot \frac{V_x}{V_p} \quad \ldots (4)$$

so the signal-to-noise ratio at the output of the slicing circuit is equal to that at the input (V_x/V_p) multiplied by T_D/T_R, thus giving a considerable improvement. The signal-to-noise ratio at the output of the receiver therefore varies directly with the signal-to-noise ratio at the input if this is above threshold.

In a P.L.M. system the greatest possible peak displacement T_D is equal to the unmodulated pulse length, whereas in P.P.M. the same displacement can be obtained using a much shorter



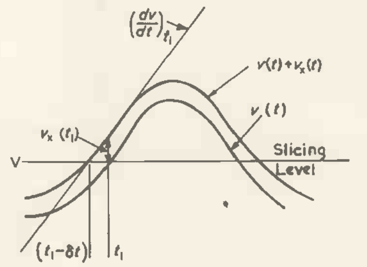


Fig. 30. Signal to noise ratio in typical P.P.M. system

Fig. 31. Pulse $v(t)$ in presence of disturbing voltage $v_x(t)$

pulse. For the same mean transmitted power a P.P.M. system can therefore have a much greater peak pulse power and so give a much better signal-to-noise ratio than a P.L.M. system with the same peak time displacement.

If the bandwidth of a P.P.M. system is increased n times the pulse length and the rise-time of the pulse can each be reduced by a factor of n. The ratio T_R/T_D is thus divided by n. If the pulse length is reduced n times, but the mean transmitted power is kept constant, the peak pulse power transmitted is increased n times. Increasing the bandwidth of the receiver n times, however, will increase the noise power at its input n times so the ratio V_x/V_p will remain constant. The net result of the n-fold increase in the bandwidth is therefore an n-fold improvement in the signal-to-noise ratio at the output by virtue of the n-fold reduction in the rise-time of the pulses. The improvement in signal-to-noise ratio (compared with A.M.) which can be obtained by the use of P.P.M. is thus proportional to the bandwidth of the system.

Crosstalk in P.P.M. and P.L.M. Systems

Crosstalk occurs between the channels if the transfer characteristics of the transmission path vary sufficiently with frequency to cause the pulse waveforms to spread in time so that there is present at the allotted time of one channel a disturbing voltage due to another channel which causes a displacement of the pulse as shown in Fig. 31. The ratio between the wanted and unwanted signal voltages present at the output (the crosstalk ratio) can then be calculated from Equation (3).

A small amount of attenuation and phase distortion at low frequencies causes crosstalk between channels which is independent of their separation in time. The crosstalk attenuation for P.L.M. increases with the frequency of the modulating signal, but that for a P.P.M. system is independent of frequency[69]. The crosstalk caused by a given amount of distortion is less for P.L.M. than for P.A.M.; P.P.M. is still better because its waveform has much smaller components at low frequencies. When the amount of low-frequency distortion is large, however, the pulse shape is distorted and the crosstalk is no longer independent of time; immediately after the pulse the crosstalk voltage is larger, but at a considerable time later, the crosstalk voltage may be smaller than that present when the distortion is much less. If the distortion is made very severe, resulting in a "differentiated" pulse, the crosstalk can be made very small even to the adjacent channel. Pulse transformers can therefore sometimes be used in P.P.M. systems although they are usually impracticable for P.L.M. and P.A.M. systems.

The crosstalk caused by typical networks introducing distortion only at high frequencies has been calculated from their known transient responses[69,70,71,72]. The crosstalk is independent of the modulating frequency and decreases rapidly as the time separation between the disturbing and disturbed channels is increased. Usually the crosstalk from any channel is negligible to all except the next succeeding channel. The crosstalk caused by a given amount of distortion is usually worse for P.P.M. than for P.A.M., but that for P.L.M. can sometimes be better than for P.A.M.[69]. High-frequency distortion also causes harmonic distortion and intermodulation of the signals of the disturbed and disturbing channels, but these are negligible if the intelligible crosstalk is adequately attenuated.

Chatterjea and Reeves[73] have proposed a method of substantially reducing the crosstalk caused by high-frequency distortion in a P.P.M. system. In a normal P.P.M. system the modulating signal of each channel varies the time separation between its pulse and the synchronizing pulse which thus acts as a fixed reference point in the cycle. In their system the position of the pulse of the first channel is modulated with respect to the synchronizing pulse, but the modulating signal of every other channel varies the separation between its pulse and that of the preceding channel. In this way the time separation between the n^{th} pulse in the cycle and the $(n + 1)^{th}$ pulse depends only on the signal of the $(n + 1)^{th}$ channel, thus preventing crosstalk from the n^{th} to the $(n + 1)^{th}$ channel. This method of multiplexing should enable a P.P.M. system to have a smaller bandwidth or pulse spacing than is possible with the conventional method. If, however, the bandwidth or pulse spacing is too small, crosstalk will become appreciable from each channel to the preceding one.

Because the signals of the different channels are very unlikely to reach their peak values simultaneously, the peak displacement of each pulse can be made greater than the channel-pulse repetition period divided by the number of channels. Thus, for a given P.R.F. and number of channels, the peak displacement can be several times that for a conventional P.P.M. system with consequent improvement in signal-to-noise ratio. A similar use of the fact that the modulations of the various channels add statistically instead of arithmetically is made in the design of multi-channel carrier systems whose amplifiers usually overload on a signal which is much smaller than the sum of the signals of the individual channels.

Twin-Index

A method of modulation called twin-index pulse-position modulation has recently been described by C. W. Earp[74]. First a train of pulses is position-modulated by say ± 10 microseconds as shown in Fig. 32(a). Each pulse of this train is converted to a "pulse-comb" of duration rather greater than 20 microseconds as shown in Fig. 32(b). This may be done by exciting a resonant circuit tuned to 500kc/s and squaring and differentiating the resulting damped wave-train. The spacing between the pulses of this comb is therefore 2 microseconds. The successive pulse-combs are, of course, position-modulated. A second modulated train of pulse-combs is derived from the same P.P.M. train as the first, but the pulse spacing in this case is 2·2 microseconds as shown in Fig. 32(c). One pulse is selected from each comb by approximately adjacent fixed-time gating pulses of 2 and 2·2 microseconds duration (the spacing of the pulses in the corresponding comb). The positions of the gating pulses, shown in Fig. 32(d), correspond roughly to the centre portions of the combs when these are unmodulated. The pulses selected from the combs by gating are thus modulated ± 1 microsecond and ± 1·1 microsecond. They are transmitted together with a

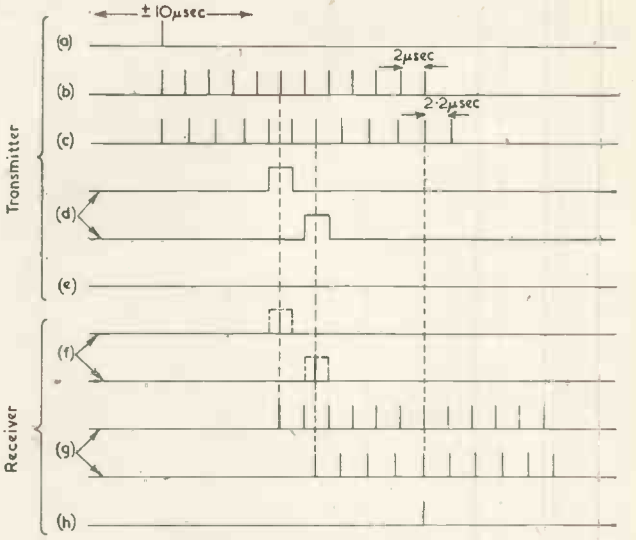


Fig. 32. Twin-index P.P.M. transmission

(a) High deviation P.P.M.
(b) First comb; modulated ±10 μsec.
(c) Second comb; modulated ±10 μsec.
(d) First and second fixed gates.
(e) Transmitted pulses
(f) First and second index pulses selected.
(g) First and second combs.
(h) Pulse selected by mutual gating.

synchronizing pulse and pairs of pulses corresponding to other channels.

At the receiver, gating pulses similar to those used at the transmitter are derived from the synchronizing pulse and used to gate the two transmitted pulses into separate circuits. The selected pulses are used to produce new pulse combs, each of at least 20 microseconds duration as shown in Fig. 32(g). Adjacent pulse spacing for these combs is made 2 and 2·2 microseconds, exactly as in the transmitter. One pulse in the first comb coincides in time with a pulse in the second comb because each occurs exactly 22 microseconds after the original position-modulated pulse. By applying the pulse combs to separate electrodes of a gating valve the coincident pulse is selected. The single pulse train produced by the mutual gating is modulated with exactly the same time displacement as the original P.P.M. pulse-train and can be demodulated by a method such as those described previously.

The significant feature is that from each sample of the modulating signal two distinct representations are transmitted, hence the term twin-index P.P.M. Because each position of one of the index pulses can correspond to several signal values, the correct value being determined by the position of the other pulse, the system is also called an ambiguous-index system. It is claimed that the system gives approximately 23db better signal-to-noise ratio than simple P.P.M. This results from the ten-fold (20db) magnification of time displacement (from approximately 1 to 10 microseconds) and a further 3db from the mutual gating process in the receiver which averages the noise jitter of the two received index pulses.

Appendix

Fluctuation Noise

Fluctuation noise, being of a random nature, has an almost uniform frequency spectrum and the probability distribution of amplitude with time is Gaussian[68]. The probability of the noise voltage lying between v and $v + \delta v$ is therefore:

$$p = \frac{1}{\sigma\sqrt{(2\pi)}} e^{-v^2/2\sigma^2} \delta v \quad \ldots\ldots\ldots\ldots \quad (5)$$

where σ is the root mean square noise voltage. The probability of the noise voltage exceeding the slicing level V in either direction is therefore:

$$P = \frac{1}{\sigma\sqrt{(2\pi)}}\int_{-\infty}^{-V} e^{-v^2/2\sigma^2}\,dv + \frac{1}{\sigma\sqrt{(2\pi)}}\int_{+V}^{\infty} e^{-v^2/2\sigma^2}\,dv$$

$$= \frac{2}{\sqrt{\pi}}\int_{V/\sqrt{2}\sigma}^{\infty} e^{-x^2}\,dx$$

$$= 1 - \text{erf}\left(\frac{V}{\sqrt{2}\sigma}\right) \qquad (6)$$

where erf y is the normal error function given by:

$$\text{erf}\, y = \frac{2}{\sqrt{\pi}}\int_0^y e^{-x^2}\,dx \qquad (7)$$

The curve shown in Fig. 29 was plotted from Equation (6), by assuming that the slicing level is half the pulse height.

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SPECTRA AND DISTORTION

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(To be continued)

A Television Waveform Monitor

By J. E. Attew

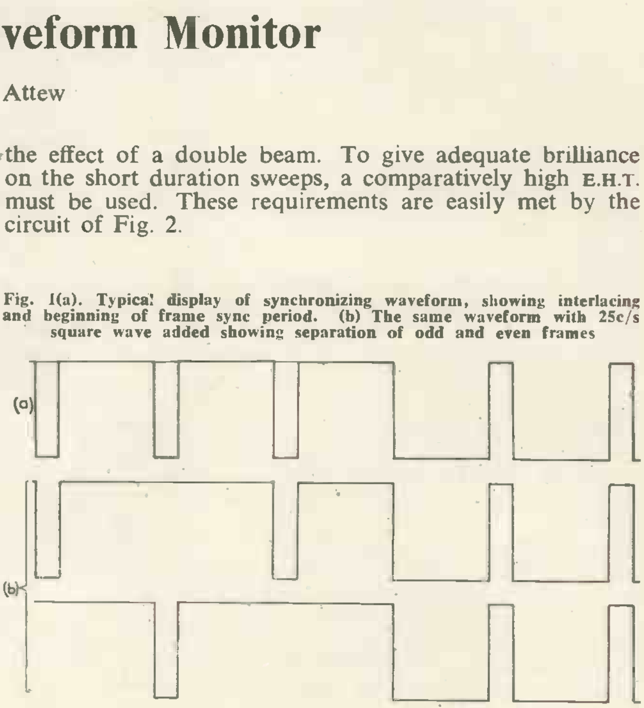
TO overcome the problem of studying the complex television waveform, the writer has recently developed a monitor, which is easier to use than the more conventional oscilloscope.

Measurements of a television waveform require time-base durations allowing the examination of the following periods:

1. Frame.
2. Frame synchronizing pulse.
3. Frame blanking.
4. Line.
5. Line blanking and line synchronizing pulse.

It is convenient to make the repetition frequency of the monitor time-base equal to the frame frequency and to provide a variable phase shift circuit, so that the time-base sweep can be triggered to start at any line or part of a line in the frame period. Then by varying the time-base duration the above five periods can be easily displayed, with the odd and even frames interlaced. (Fig. 1(a).) By adding to the display a square waveform at half frame frequency, odd and even frames can be separated (Fig. 1(b)), giving the effect of a double beam. To give adequate brilliance on the short duration sweeps, a comparatively high E.H.T. must be used. These requirements are easily met by the circuit of Fig. 2.

Fig. 1(a). Typical display of synchronizing waveform, showing interlacing and beginning of frame sync period. (b) The same waveform with 25c/s square wave added showing separation of odd and even frames



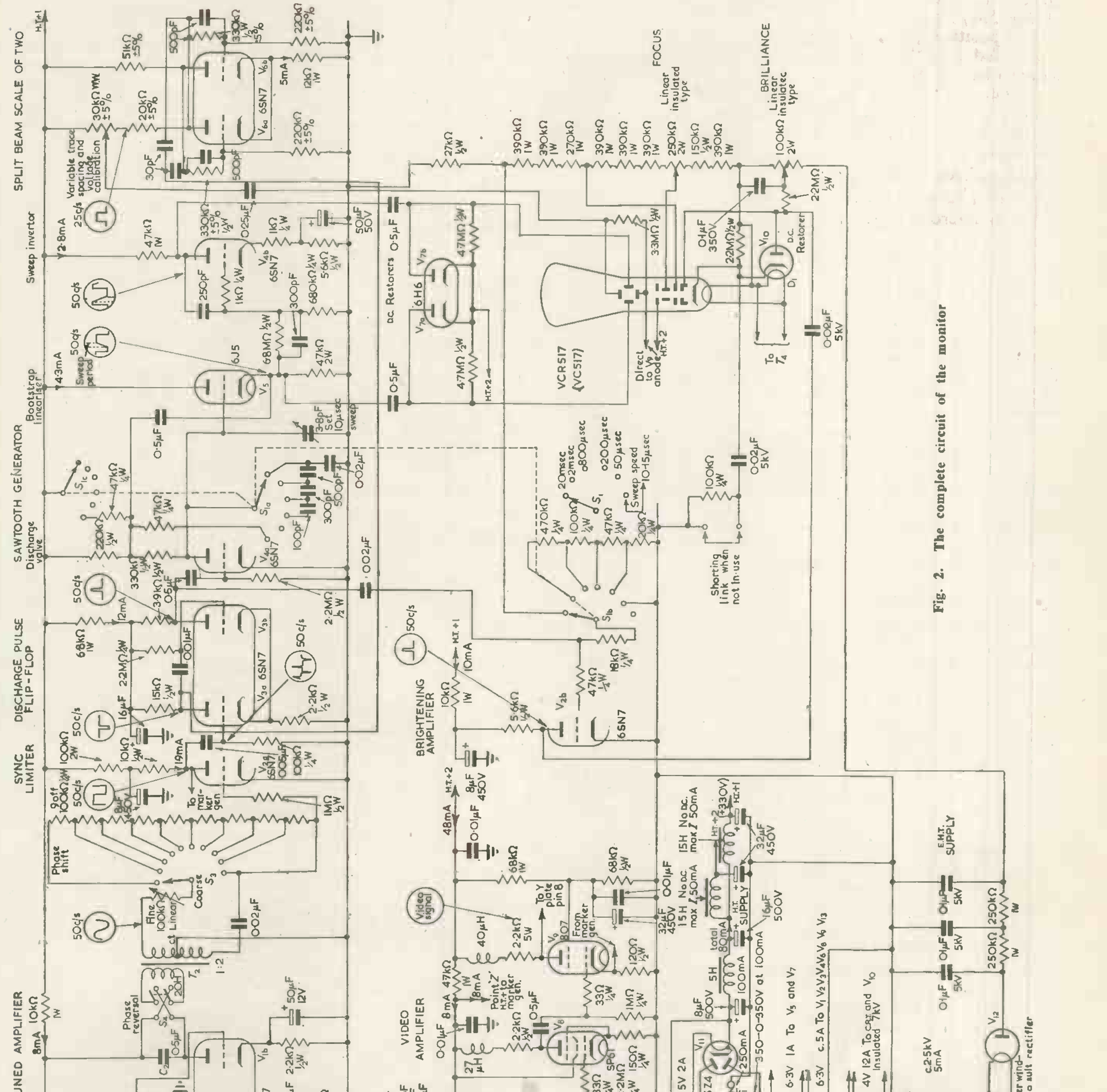


Fig. 2. The complete circuit of the monitor

Time-base

The discharge pulse flip-flop (V_3) is triggered via a limiter (V_{2a}) and a variable phase shifting circuit, which allows the beginning of each sweep to occur at any part of the 50c/s sine wave.

The positive going short duration pulse at the anode of V_{3b} is fed into the grid of the saw-tooth generator (V_{4a}), where the waveform is D.C. restored, so that the valve conducts only during the positive pulse period. During the *off* period the selected capacitor (by means of S_{1a}) at anode of V_{4b} charges up, the time taken to charge being the sweep period.

The periods chosen are:

1. 10 microseconds.
2. 50 microseconds.
3. 200 microseconds.
4. 800 microseconds.
5. 2 milliseconds.
6. 20 milliseconds.

The amplitude remains constant while the sweep duration is proportional to the capacitance. The sweep is linearized by positive feedback from a cathode follower (V_5), while symmetrical deflexion is achieved by a sweep invertor of unity gain (V_{4b}). D.C. restorers are necessary at the deflector plates to ensure that the trace remains in the centre of the screen for all the sweep durations.

During the sweep period the trace is brightened by a positive pulse from the brightening amplifier (V_{2b}). The input to the grid being the differentiated output of V_2. The brightening amplifier is cut off during the frame period sweep, as the coupling time-constants are too short to allow an efficient brightening pulse.

50c/s Tuned Amplifier

This circuit is used to select the fundamental from the frame time-base waveform of a television receiver, or the frame synchronizing waveform of a television synchronizing generator. A suitable synchronizing point in a television receiver being the anode of the frame output valve; or alternatively, the time-base can be synchronized from the mains supply via switch S_2.

For transformers T_1 and T_2 any available transformers can be used providing that they are of similar ratios and that the D.C. through the primaries does not effect their characteristics appreciably. Both anode circuits resonate at 50c/s.

If required, a conventional synchronizing separator and frame interlace filter could be used ahead of the amplifier, if it is necessary to synchronize from the video waveform, but this has not been found necessary in the writer's case.

Split Beam Scale of Two

This circuit provides a 25c/s square wave at either anode of V_6 and is triggered by the negative going pulse from anode of V_{3a}. The output waveform is fed into one Y plate, the amplitude being adjustable, so varying the beam spacing. By calibrating the output control against input volts to the video amplifier, and using one trace as reference for the other, a convenient means of pulse measurement is obtainable.

Video Amplifier

The two valve amplifier V_8 and V_9 has sufficient gain to produce a useable waveform, with the input connected directly to the detector diode of a television receiver. The anode circuits are shunt compensated by chokes to increase the high frequency response, which is substantially linear up to 3Mc/s. The chokes have a value which introduces no overshoot of a square waveform.

A frequency compensated attenuator is fitted to preserve this bandwidth.

The cathode-ray tube used is a VCR517. The longer persistence of this tube being helpful in preserving the brightness of the trace on the shorter sweeps.

An input is provided into the cathode of the C.R.T. to allow external calibration pulses to intensity modulate the beam for time measurements. This circuit should not be omitted, even if unused, as by making the grid time-constant of the C.R.T. equal to the cathode time-constant, any small amount of 50c/s ripple voltage left in the E.H.T. supply is fed equally to the grid and to the cathode, and as a result there is no intensity modulation of the beam at the mains frequency.

Marker Generator

This circuit (Fig. 3) consists of a triggered delay line oscillator[1], which provides square waves at a frequency of 1Mc/s and is triggered by the 50c/s square wave from anode of V_{2a}. This waveform is differentiated at the cathode of V_9, producing alternate positive and negative going pulses, spaced 0·5 microseconds apart. These appear on the trace superimposed with the video signal, so enabling pulse duration measurements. Fig. 4 gives details of the delay line used.

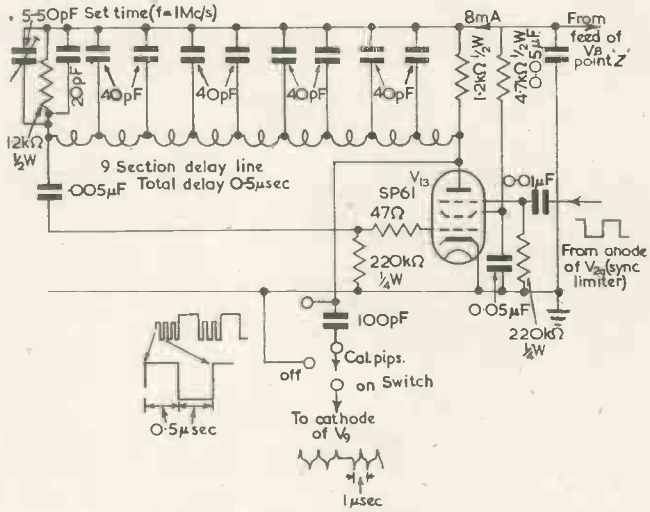


Fig. 3. Pulse duration calibrator

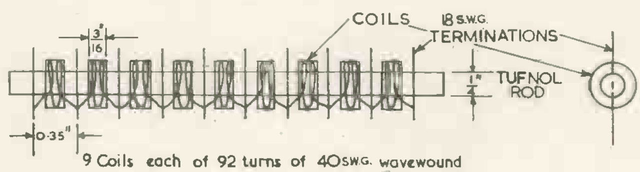


Fig. 4. Delay line

Construction has no special difficulties, but it is advisable to build the power supply as a separate unit to eliminate unwanted deflexion of the C.R.T. beam due to stray magnetic fields. The video amplifier should be screened from the rest of the circuits, with the anode of V_9 close to the base of the C.R.T. to allow a short direct lead to the Y plate, so reducing the stray circuit capacitance.

While this oscilloscope allows easier measurements of the television waveform, it has also been found invaluable for instructional purposes when demonstrating television fundamentals.

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A Note on Gas Discharge Tubes for E.H.T. Stabilization

By J. G. G. Hempson*, A.M.I.Mech.E.

THE use of gas discharge tubes as voltage controlling elements has been largely confined to moderate voltages and currents in the range 5 to 40mA. However, in connexion with C.R.O. tubes, photo-multiplier tubes and Geiger counters it is frequently necessary to stabilize voltages in the range 1 to 3kV at currents under 1mA. It is also convenient to have intermediate fixed potential points without resort to resistance chains.

The problem may be exemplified by considering the design of a special C.R.O. where a stable final anode voltage and complete lack of interaction between controls such as focus, shift and brilliance was needed. The voltage required was about 2kV. Three solutions were considered, a corona regulator, a hard valve stabilizer circuit and a chain of gas discharge tubes. The first was discarded as available corona tubes had a maximum current of 100μA, which was insufficient for the expected range of beam current, and intermediate voltages were not provided. This last objection also applied to the hard valve circuit which had further disadvantages of bulk, complexity and power consumption. Thus attention was concentrated on the last alternative, a chain of gas discharge tubes.

The maximum voltage drop per tube in normally available types is about 150V and thus 14 tubes would be required in the present case. Even if these were of the B7G type such as the QS/150/15 (CV 287) the cost and space required would be considerable. Further, the desired current of approximately 1mA is below the recommended minimum for stable operation of these tubes. Attention was then given to the Hivac NT2 tube. This is a subminiature gas discharge tube weighing 1gm and measuring about 27mm long by 7mm diameter. The nominal maintaining potential is 60V and maximum current 1mA. No special claims are made as to their behaviour as precision stabilizers so it was decided to make some simple tests. An average voltage/current relationship is given in Fig. 1. This represents an apparently stable condition reached after about 3 or 4 hours operation. With new tubes the initial voltage might be about 1½ per cent higher, slowly falling to the figures shown. The impedance derived from the slope of the curve in the region of the optimum operating point is about 1 000Ω.

Although the tube operating voltage is low, this is compensated by their small size and low cost, and thus it was decided to proceed with the construction of a C.R.O. using these stabilizers. The small space actually needed for the 36 tubes used is shown by Fig. 2, which shows them as mounted in the complete unit. The weight and space are less than the smoothing and bypass capacitors eliminated and, in one lightweight portable C.R.O., the savings on these were the chief reasons for employing the tubes. A chain current of about 0·7mA is used.

The oscillograph unit has a most satisfying lack of interaction between controls, due to auxiliary potentials being derived from constant potential points provided by the chain.

Although several tube chains have now been made up, no case of failure of a tube has been experience, nor appreciable change in characteristics during use.

Another use is stabilizing the stage potential steps in electron multiplier photo-tubes such as the 931A. Here considerable economy in current is achieved, as where resistance chains are used, the drain must be made large compared with the last stage photo current if a linear response is required.

The ability of these tubes to stabilize reasonably well at currents of 50μA or even less provided a solution to one rather difficult application of the photo-multiplier tube. It is well known that these photo-tubes develop bad fatigue effects or actual permanent damage if an accidental excess of light reaches them when they are being operated at high sensitivity. This is due to excessive current in the last stages of the multiplier which can flow where a high current chain and smoothing capacitor is employed.

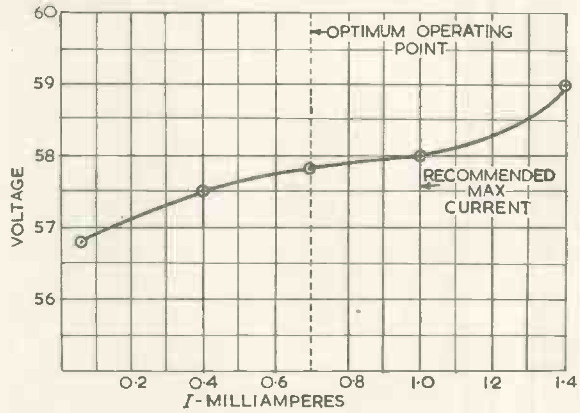

Fig. 1. Current/voltage relationship for NT2 discharge tubes (*mean of ten tubes*)

Fig. 2. The complete unit of 36 tubes

In a special application it was necessary to measure accurately very low light levels, but it was also impossible to prevent occasional bursts of light of very high intensity. By supplying the multiplier tube from a chain of 20 NT2 stabilizers, i.e. about 112 volts per stage, passing about 50μA via very high value feed resistors, the limiting current was kept to a safe level yet linear response to low values of radiation was achieved, and no damage resulted to the photo-tube even if exposed to full daylight.

For Geiger counters, a chain of tubes forms an alternative to corona regulators, which are not yet widely available in this country, with the added advantage of being able to adjust the voltage in steps of 60V instead of having a single fixed value.

Although no new principles are disclosed, it is felt that these applications of miniature gas discharge tubes may help other workers in allied fields.

* *Ricardo & Co. (1927) Ltd.*

Visual Method for the Determination of Electronic Conductivity

By J. K. Grierson*, B.Sc., A.M.Inst.E.

A visual method is described for the determination of electrolytic conductivity, in which the associated capacitance of the cell is balanced out during the determination. The main application is the determination of conductivities in the range 0·02 mhos to 0·001 mhos, to an accuracy of 0·1 per cent.

THE subject of electrolytic conductivity has an extensive history. The first attempts to determine the conductivity of an electrolyte were carried out using straightforward D.C. Wheatstone bridges, and the earlier workers went to great pains to try to eliminate the effects of polarization, but they were not entirely successful. A real advance was made by Kohlrausch et al.[1] who first used A.C. determinations. His method made use of an induction coil oscillator and telephone detectors. Later workers pointed to defects in his system, the poor waveform of the oscillator voltage caused a D.C component of voltage and consequent polarization. Poor design of resistors employed as bridge elements, leakage capacitances and absence of electromagnetic and electrostatic screening were further disadvantages. A further advance was made when Jones and Joseph[2] introduced the use of valve oscillators, amplified detection, and specially constructed resistors to this problem. Radio frequency bridges were investigated by Dowling and Preston[3], but leakage capacitance effects proved troublesome and the conductivity figure obtained appeared to depend on the frequency. Medium audio frequencies are now regarded as the best regions in which to work. Parallel with these advances work has been carried out on the design of the cell and electrode system, however, few of the designs described are applicable to latex. Most modern instruments use a valve oscillator and telephone detector, but few commercial models attain much more than 1 per cent accuracy, although in general the sensitivity is rather higher than this. Although many of the earlier workers were aware of capacitance effects in the cell, most of the modern workers deem this effect unimportant. Some work has already been done on electrolytic conductivity using cathode-ray oscillography by P. A. Einstein[4], but there seems to be a need for information on an indicating device which is simple to use. The instrument described uses a specially constructed cathode-ray oscillograph.

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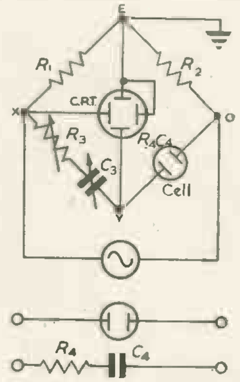


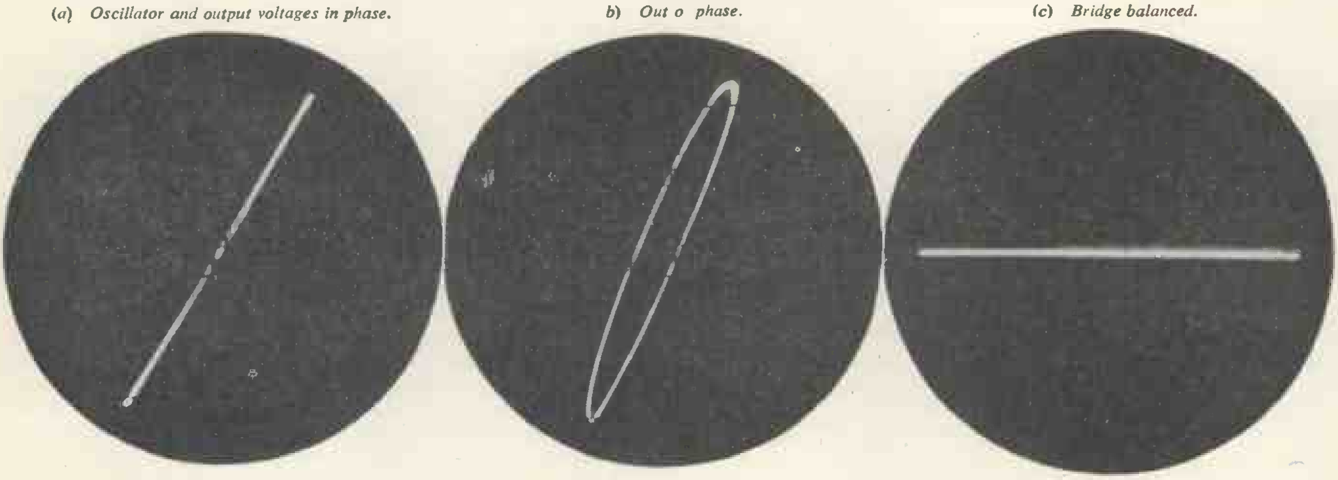
Fig. 1. Basic bridge circuit and the equivalent circuit of the cell

Basic Bridge Circuit

An electrolytic cell behaves as a resistance in series with a capacitance. The resistance is dependent on the main body of the electrolyte and on the geometry of the cell, while the capacitance is determined mainly by the boundary between the electrodes and the electrolyte[4], the geometry of the cell being less important for a constant surface area of electrode. The bridge comprises the cell, a series capacitance-resistance balancing arm (both elements variable) and the usual resistance ratio arms. Fig. 1 shows the ratio arms R_1 and R_2, the variable resistance and capacitance elements R_3 and C_3, and the fourth element the cell, R_4 and C_4. The junctions of the elements are E, from which

Fig. 2. Screen traces

(a) Oscillator and output voltages in phase. *b) Out o phase.* *(c) Bridge balanced.*



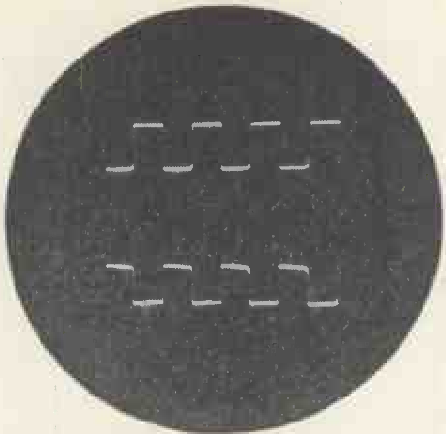


Fig. 3. Square wave test of Y amplifier
(Upper trace — input ; lower trace — output).

all voltages are measured, X and O, the oscillator connexions, and Y, the final junction. The circuit is conventional to this point, as in a normal bridge headphones would be connected between E and Y. However, in the circuit used, the X amplifier of the cathode-ray oscillograph is connected between X and E, and the Y amplifier connected between Y and E. The voltage existing across the resistor R_1 is in phase with the driving voltage from the oscillator, since R_1 and R_2 are purely resistive. The voltage existing across the terminals E and Y may, or may not, be in phase with the driving voltage, depending on whether the equation

$$R_3 + 1/jpC_3 = k(R_4 + 1/jpC_4) \quad \ldots\ldots\ldots \quad (1)$$

is satisfied or not (where p is the angular velocity of the oscillator, and k is a non-complex constant). If this equation is satisfied, and in addition $R_1 = kR_2$, then no voltage appears across the points E and Y. These equations may be rewritten:

$$R_1/R_2 = R_3/R_4 \quad \ldots\ldots\ldots\ldots \quad (2)$$

and

$$R_1/R_2 = C_4/C_3 \quad \ldots\ldots\ldots\ldots \quad (3)$$

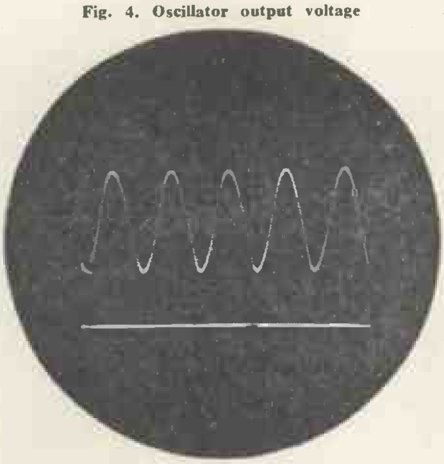
Equation (2) may be said to be the D.C. condition of balance, while Equation (3) is the A.C. condition of balance.

From these two equations a fourth may be derived:

$$C_4/C_3 = R_3/R_4 \quad \ldots\ldots\ldots\ldots \quad (4)$$

This equation, corresponding to the first equation quoted, indicates a voltage in phase with the driving voltage.

Fig. 4. Oscillator output voltage



Function of the Cathode-Ray Tube

The cathode-ray tube indicates whether the voltage appearing across the terminals E and Y is in phase with the voltage appearing across X and E. If this voltage is in phase with the reference voltage, then a straight line appears on the screen, Fig. 2(a). If not, then an ellipse appears on the screen, Fig. 2(b). If a horizontal line appears on the screen, Fig. 2(c), there is no voltage between Y and E and the bridge is balanced. The bridge is therefore simple enough to be operated by relatively unskilled personnel.

Ancillary Equipment

A cathode-ray tube alone is not sensitive enough to produce a pattern without amplifiers. Two amplifiers are provided in the instrument. The Y amplifier consists of a three-stage resistance-capacitance coupled amplifier with

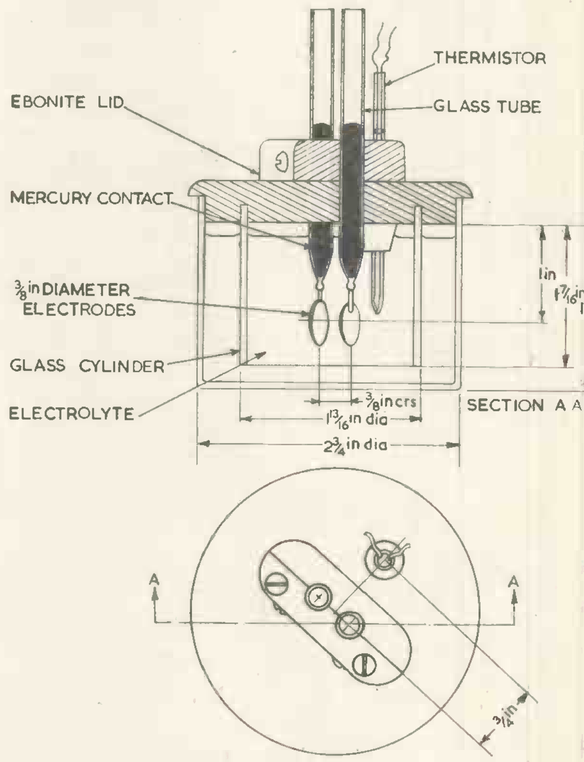
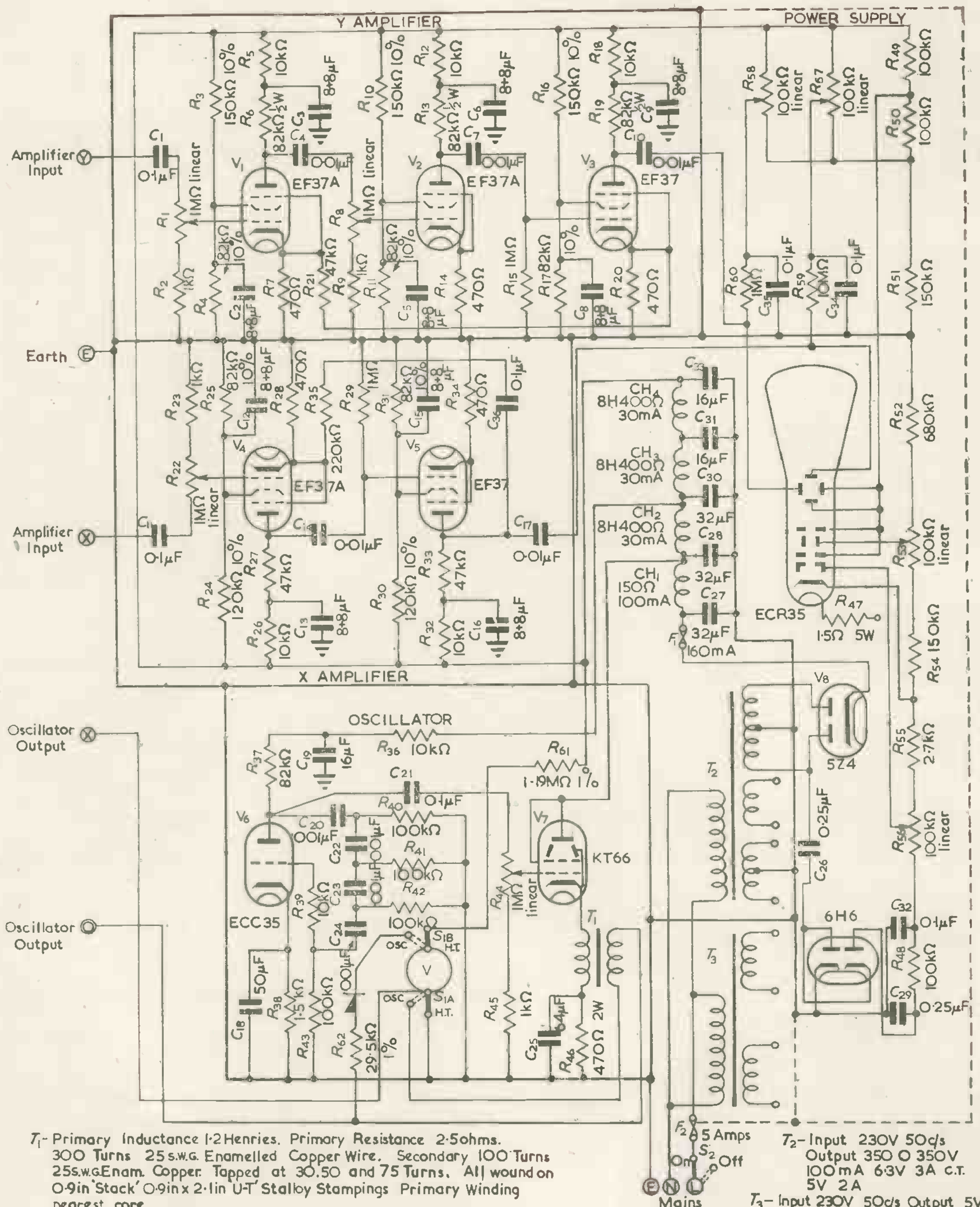


Fig. 6. Electrolytic cell assembly

negative feedback. The overall sensitivity of the Y amplifier and tube is 2 centimetres per millivolt, with negligible phase shift at 1 250c/s. A square wave test, Fig. 3, shows the response of the amplifier. For this test the oscillator voltage was squared, fed into the amplifier and traces of the input and output voltages recorded photographically on a Cossor model 1049 oscillograph. The X amplifier is a similar two-stage amplifier. Power supplies follow conventional lines with a voltage doubling circuit supplying the E.H.T. for the tube. The oscillator is a resistance-capacitance coupled positive feedback amplifier, coupled to a cathode-follower output stage, giving up to 17 volts of good sine wave form, Fig. 4, with an output impedance of approximately 25 ohms. The circuit diagram, Fig. 5 (page 112) shows the complete detector and oscillator unit.

Description of Cell

The cell used in the experiment comprised a glass evaporating dish, into which is inserted an electrode system consisting of two glass tubes into which are sealed two platinum wires. On the ends of the wires are welded two



T_1 – Primary Inductance 1·2 Henries. Primary Resistance 2·5 ohms. 300 Turns 25 s.w.g. Enamelled Copper Wire. Secondary 100 Turns 25 s.w.g. Enam. Copper. Tapped at 30, 50 and 75 Turns. All wound on 0·9 in 'Stack' 0·9 in x 2·1 in 'U-T' Stalloy Stampings. Primary Winding nearest core

T_2 – Input 230V 50c/s Output 350 0 350V 100mA 6·3V 3A C.T. 5V 2A

T_3 – Input 230V 50c/s Output 5V 2A 6·3V 2A

All resistors are ¼ W, 20 per cent tolerance, unless otherwise stated

Fig. 5. The complete circuit

circular electrodes of platinum. The free ends of the wires within the tubes are immersed in mercury, into which the connecting wires are placed. The glass tubes are held by an ebonite plate, which is recessed to fit the evaporating dish. A glass cylinder is placed around the electrodes and is fixed to the ebonite plate, Fig. 6. It is essential for the satisfactory operation of the apparatus that the same dish containing the same volume of liquid is used in each experiment, and that the temperature of the electrolyte is controlled to 0·01°C. This is done by means of a thermostatically controlled water bath.

Thermal Control

In the present system, the cell is placed in a water bath which is controlled at 25°C by a mercury-toluene regulator, in conjunction with a 250 watt immersion heater. A thermistor is used in determining the temperature within the cell, the resistance being measured by a Wheatstone bridge. The electrical resistance of the thermistor depends on temperature according to the law: $R = R_0 e^{T_0/T}$, where R_0 and T_0 are constants, T is the absolute temperature and R the resistance. This method presents numerous advantages, the only disadvantage being the subsequent calculation of temperatures, and the need to calibrate the thermistor. However, charts can be prepared for small temperature ranges, and temperatures read from these.

Results

Initial results on latices have indicated that the present apparatus is superior in accuracy and sensitivity to methods formerly used. It also shows a saving in time for routine work on batch samples of similar conductivities. On the basis of determinations carried out, further work is proceeding on (a) calibration and use of thermistors, (b) temperature control, (c) use of magnetic stirrers within the cell, (e) further control of leakage capacitances and use of electrostatic screening.

Acknowledgments

The author wishes to acknowledge help freely given by the staff of the Dunlop Research Centre, and the Directors of the Dunlop Rubber Co., Ltd., for permission to publish this work.

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DOPPLER RADAR SYSTEMS

By R. C. Coile*

DOPPLER radar systems have many applications where their characteristics permit better performance than that obtainable from conventional pulse radar systems, such as in the detection of moving objects in the presence of large amounts of ground clutter and the measurement of velocities of projectiles.

When a target moving with radial velocity V_r is illuminated by a radar signal of frequency f, the echo frequency is given by the doppler formula[1]

$$f' = \frac{C + V_r}{C - V_r} f$$

The doppler frequency is

$$f_D = f' - f = \frac{2V_r}{C - V_r} f$$

Since V_r, the radial velocity is small compared with C, the velocity of light, the doppler frequency shift may be written

$$f_D = \frac{2V_r}{C} f = \frac{2V_r}{\lambda}$$

For V_r in miles per hour and λ in centimetres, the doppler frequency shift is

$$f_D = \frac{89{\cdot}4\, V_r}{\lambda}$$

The nomogram allows calculation of the doppler frequency knowing the wavelength in centimeters or frequency in megacycles per second of the radar and the radial velocity of the target in miles per hour.

To find the doppler frequency of a 3·5cm radar detecting a target of 400 M.P.H radial velocity, lay a straightedge between 3·5cm on the wavelength scale and 400 M.P.H. on the velocity scale and read the answer of 10 000c/s on the doppler frequency scale.

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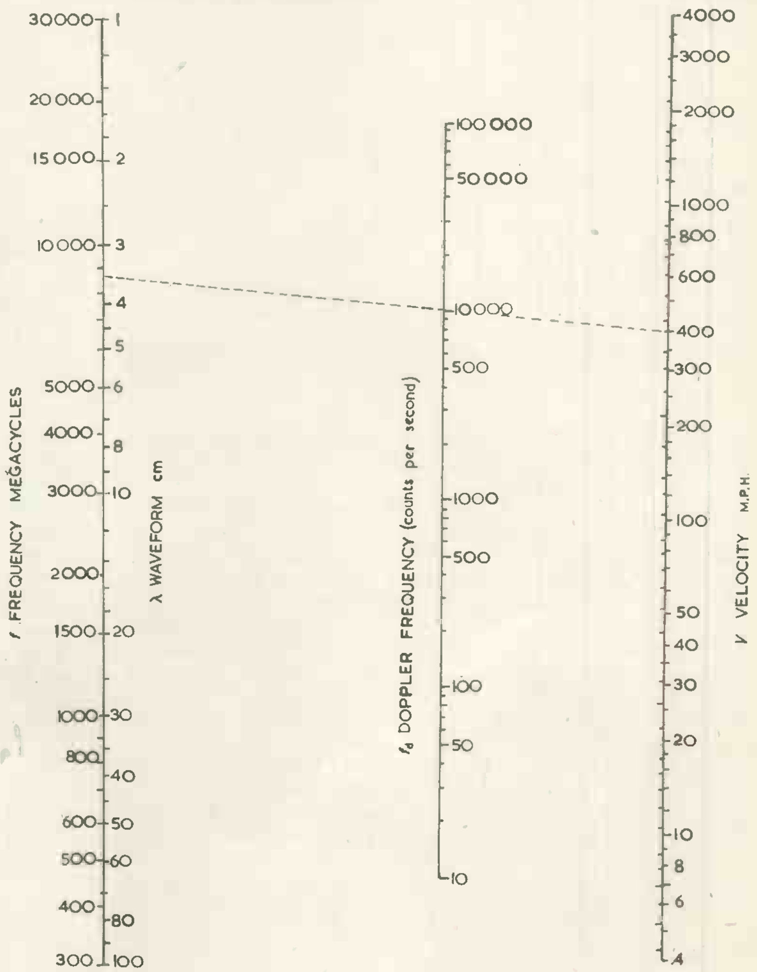

* *Massachusetts Institute of Technology.*

Television Sound Reception

The Critical Capacitance Coupling System

By S. L. Fife*

In a television superheterodyne receiver, it is current practice to make use of common R.F. amplification together sometimes with one I.F. stage, for the amplification of both sound and vision signals. The chief requirement for the reproduction of the accompanying sound with vision signals is a high signal-to-noise ratio, where the denominator includes:

(1) Vision signal breakthrough.

(2) Interference spikes—ignition, etc.

(3) The products of cross-modulation.

Selective circuits are therefore necessary following the sound take-off from the common I.F. to eliminate (1), while an adequate bandwidth is desirable to preserve the steep fronted noise spikes in order that the noise limiting action following demodulation may be efficient (see Appendix C). A bandwidth of 600kc/s (at 6db down) is a good compromise for these requirements. Cross-modulation is avoided by limiting the number of common stages in order

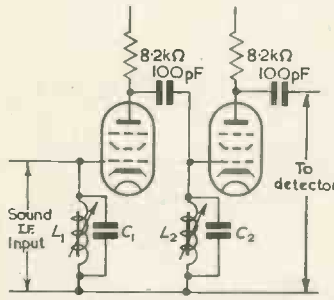

Fig. 1(a). Sound I.F. amplifier using single tuned circuits

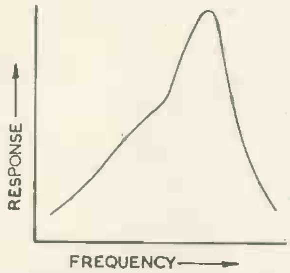

Fig. 1(b). Overall sound response using common R.F. amplification followed by sound I.F. of poor selectivity

that neither signal is of sufficient amplitude at the grid of the last common stage to swing the valve over a non-linear part of its characteristic.

It is an advantage to have all iron-dust tuning cores available from one side of the chassis. This will generally necessitate circuits as shown in Fig. 1(a). The overall sound response to be expected may then be that in Fig. 1(b), the asymmetry being due to the common amplification of vision and sound signals by the early stages, followed by the sound amplifiers proper of poor selectivity. Alternatively, the anode is sometimes tuned, whereby the resistance effectively shunted across the tuned circuit is then increased. However, it is desirable to avoid high D.C. grid to cathode leaks and Fig. 1 has that merit.

The adoption of an I.F. within the G.P.O. recommended band of 34-38Mc/s means appreciable damping by valve input impedance being shunted across the tuned circuit, which may reduce the Q by some 60 per cent when in circuit, which results in a wide I.F. bandwidth which may exceed 1Mc/s with single circuits only (see Appendix A). The vision frequency " bulge " in the sound curve, clearly, can only be eliminated by critical coupling throughout the sound I.F. amplifiers. This may be achieved by link coupling or by the simple expedient of positioning the primary and secondary coils side by side and finding the correct geometrical positions of each for critical coupling. Both these coupling systems, however, allow no simple adjustment of coupling.

A method, which the writer has used, employing critical capacitance coupling provides flexibility of adjustment of coupling, while the coils remain screened in their cans, also both circuits may be tuned from the same side of the chassis.

The sound I.F. amplifier (tuned at 37·5Mc/s) following the common I.F., is shown in Fig. 2. The sound absorber trap following the common I.F. stage is critically coupled to the grid circuit. This capacitive coupling by C_{c1} and C_{c2}, when of the right order for single peak response, results with little damping on the high Q trap, and thus neither the vision bandwidth nor sound rejection is

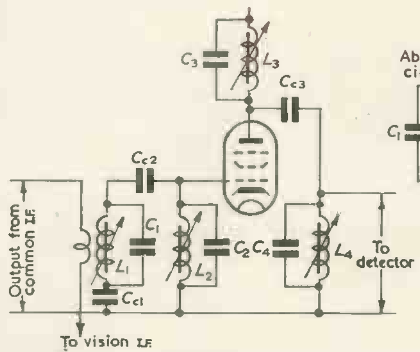

Fig. 2. Sound I.F. amplifier using capacitance coupled circuits

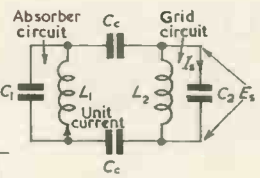

Fig. 3. Grid input circuit using capacity coupling

seriously impaired by the take-off system. With the small capacitor required for critical coupling, it is found that the tuning positions of the trap for maximum sound feed to the sound I.F., and maximum sound rejection through the vision I.F. stages is made more coincident. This defect is inherent in some sound take-off systems when the tuning points become appreciable[1].

The input circuit of Fig. 2 uses " top and bottom " capacitance external impedance coupling, and is reproduced in Fig. 3. The analysis of this coupling system may be carried out with reference to Terman[2] who gives the following rules to apply in the analysis of any coupled circuit system (see Appendix B).

1. Any two circuits that are coupled by a common impedance, have a coefficient of coupling that is equal to the ratio of the common impedance to the square root of the product of the total impedance of the same kind as the coupling impedance that are present in the two circuits, that is:

$$k = \frac{Z_m}{\sqrt{Z_1 \cdot Z_2}} \quad \ldots\ldots\ldots\ldots \quad (1)$$

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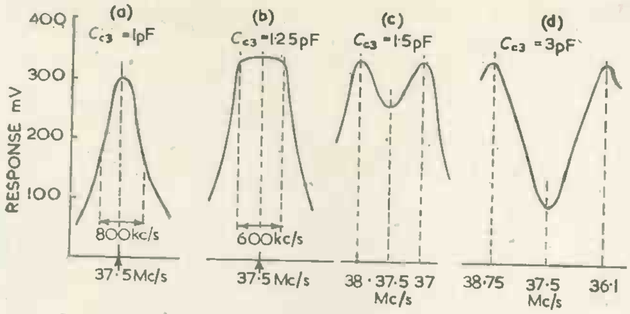


Fig. 4. Response curves of the capacitance coupled amplifier

where Z_m is the impedance common to the two circuits, and Z_1 and Z_2 are the total impedances of the *same kind* in the two circuits.

2a. The equivalent primary impedance Z_p of the equivalent circuit is the impedance that is measured across the primary terminals of the actual circuit when the secondary circuit has been opened.

b. The secondary impedance Z_s of the equivalent circuit is the impedance that is measured by opening the secondary of the actual circuit and determining the impedance between these open points when the primary is open-circuited.

c. The equivalent mutual inductance M is determined by assuming a current I_p flowing into the primary circuit. The voltage which then appears across an open-circuit in the secondary is equal to $-j\omega M \cdot I_p$.

It may be shown that the coupling factor for the circuit of Fig. 3 is given by:

$$k = \frac{-C_c}{\sqrt{(2C_2 + C_c)(2C_1 + C_c)}} \quad \ldots\ldots\ldots (2)$$

C_1 and C_2 will be chosen sufficiently large to swamp valve and stray capacitances, and C_c then adjusted for critical coupling. The coupling capacitor (C_c) may conveniently be formed by twisted P.V.C. wire, which in the writer's case was made 1pF.

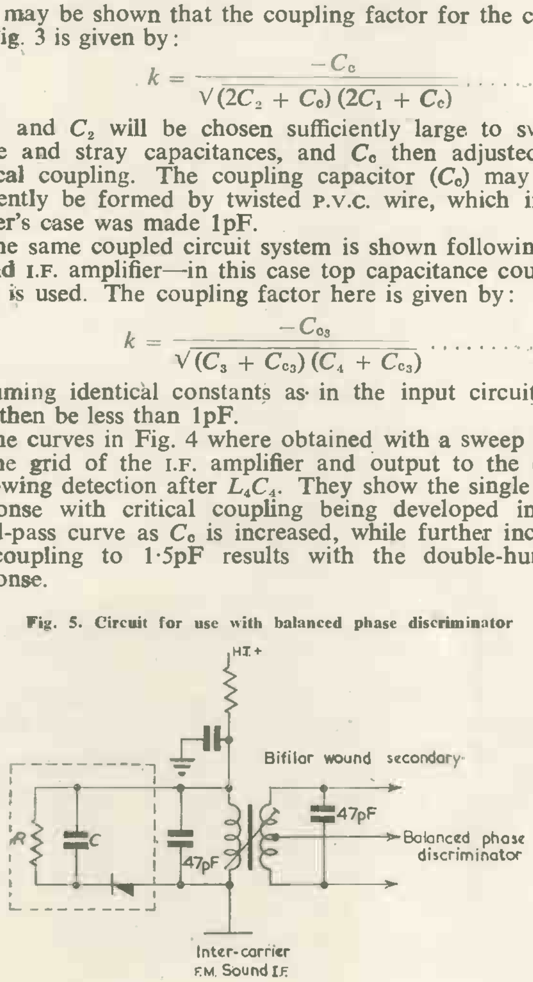
The same coupled circuit system is shown following the sound I.F. amplifier—in this case top capacitance coupling only is used. The coupling factor here is given by:

$$k = \frac{-C_{c3}}{\sqrt{(C_3 + C_{c3})(C_4 + C_{c3})}} \quad \ldots\ldots\ldots (3)$$

Assuming identical constants as in the input circuit, C_{c3} will then be less than 1pF.

The curves in Fig. 4 where obtained with a sweep input at the grid of the I.F. amplifier and output to the C.R.O. following detection after L_4C_4. They show the single peak response with critical coupling being developed into a band-pass curve as C_c is increased, while further increase of coupling to 1·5pF results with the double-humped response.

Fig. 5. Circuit for use with balanced phase discriminator



With this form of coupling the energy transfer has been found comparable with mutual inductive coupling. So that while the coupling capacitance may appear to be alarmingly small, both the theoretical analysis and performance would commend its use both for critical coupling through the sound channel, or overcoupling in the vision circuits, when tuning from one side of the chassis is a condition to be met in the design and where critical magnetic coupling is an undesired quantity to adjust *in situ*.

Inter-Carrier F.M. Sound

Both the American and European (C.C.I.R.) television standards lend themselves advantageously to the incorporation of an inter-carrier F.M. sound circuit in receivers designed for operation in those countries where the above standards have been adopted. Local oscillator frequency stability assumes less importance, in the placing of the sound carrier at the centre of the linear part of the discriminator response, since the inter-carrier frequency stability is that of the transmitters only. This fact is of particular importance in television F.M. sound reception for use on bands higher than 175Mc/s.

The system involves the amplification of both the vision and frequency-modulated sound signals through common R.F./I.F. stages, with the difference frequency component

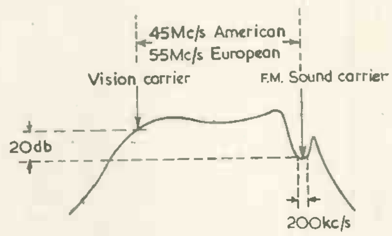


Fig. 6. Frequency response necessary for inter-carrier F.M. sound

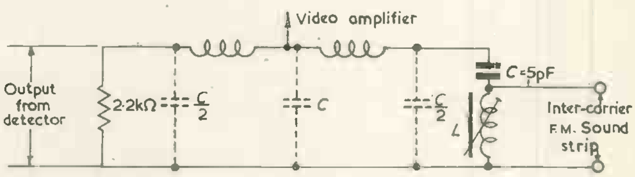


Fig. 7. Sound take off circuit

selected at the vision detector and coupled to an F.M. sound I.F. strip tuned to the inter-carrier frequency, i.e. 4·5Mc/s or 5·5Mc/s for the American or European standards respectively, and finally, demodulation by a ratio detector or the Seeley-Foster phase discriminator. The choice here is a personal one, since the advantage of an unresponsive ratio detector to A.M. is offset by its poor sensitivity as compared with the phase discriminator. The latter must, however, be associated with a limiter stage, which, when of the grid limiting type, implies a low gain I.F. stage. Results (to which the reader is referred), of comparison tests between the two systems have been made known elsewhere[3]. However, an economy has been attempted by avoiding the conventional low gain limiter while retaining the sensitivity of the balanced phase discriminator, and preliminary results with the type shown in the primary circuit of Fig. 5, have been good, particularly in the suppression of impulsive noise. The circuit is effective only in the suppression of amplitude modulation, which means that impulsive noise peaks with a phase-angle, relative to the carrier, other than 0° or 180°, will reach the discriminator as F.M. noise and there a noise output will be produced. The time-constant CR is made long compared with the reciprocal of the lowest frequency to be suppressed, and for this a low voltage electrolytic capacitor is used.

Sound rejection preceding the vision detector is not of prime importance insofar as it serves to attenuate sound signals reaching the video stage, but attenuation through the common stages has been found necessary to avoid distortion of the inter-carrier F.M. component which occurs when both vision and sound are of the same level at the vision detector. To avoid this, the sound carrier must be some 20db below the vision carrier (Fig. 6), and in the writer's case one trap was used, suitably designed to give a passband in the trough of 200kc/s, which is adequate for the F.M. spectrum of the two standards mentioned (see Appendix D).

The sound take-off is made at the π-type low-pass video filter and is selected by the LC acceptor circuit (Fig. 7), resonant at the inter-carrier frequency, and thus effects rejection of this component from the video stage to avoid the inter-carrier R.F. pattern appearing on the picture tube.

Acknowledgments

The writer is pleased to acknowledge the aid of his colleague, Mr. J. Harrison, and is grateful to the English Electric Company—particularly Mr. D. W. Heightman for permission that the relevant data obtained in the laboratory may be published.

Appendix A

Referring to the single tuned circuits as in Fig. 1(a)

Total tuning capacitance = 30pF

Valve input damping at 37·5Mc/s = 21kΩ

Thus the total resistance across the tuned circuit by anode load and the input resistance becomes 5·9kΩ

Q of isolated tuned circuit = 200

Resistance within isolated tuned circuit (R_1) = 0·708Ω

Resistance injected into tuned circuit is given by $X^2/R_p\Omega$, where X is the reactance of a branch at resonance, R_p is the shunt resistance and R_s the injected series resistance, then:

$$R_s = \left(\frac{10^{12}}{2\pi \times 37{\cdot}5 \times 10^6 \times 30}\right)^2 \cdot \frac{1}{5{\cdot}9 \times 10^3} = 3{\cdot}4\Omega \quad \ldots\ldots (4)$$

$\therefore$ the effective Q =

$$\frac{1}{\omega C(R_1 + R_s)} = \frac{10^{12}}{2\pi \times 37{\cdot}5 \times 10^6 \times 30\,(0{\cdot}708 + 3{\cdot}4)} = 34{\cdot}4 \quad \ldots\ldots (5)$$

it follows then that:

$$\frac{37{\cdot}5}{f_2 - f_1} = 34{\cdot}4 \quad \ldots\ldots (6)$$

where f_1 and f_2 are the frequencies at which the response is $1/\sqrt{2}$ down from that at resonance,

i.e. Bandwidth at 3db down = 1·1Mc/s.

Appendix B

Applying rule 2a to the circuit in Fig. 3.

Open-circuiting L_2 in the secondary:

$$Z_p = j\omega L_1 - j\,1/\omega C' \quad \ldots\ldots (7)$$

where $C' = C_1$ paralleled by C_{c1}, C_2 and C_{c2} in series,

let $C_{c1} = C_{c2} = C_c$

$$\text{thus } C' = \frac{2C_1C_2 + C_cC_1 + C_cC_2}{2C_2 + C_c}$$

$$\therefore Z_p = j\omega L_1 - j\frac{(2C_2 + C_c)}{\omega(2C_1C_2 + C_cC_1 + C_cC_2)} \quad \ldots\ldots (8)$$

Applying Rule 2b.

Open-circuiting L_1 in the primary:

$$Z_s = j\omega L_2 - j\frac{1}{\omega C''} \quad \ldots\ldots (9)$$

where $C'' = C_2$ paralleled by C_{c1}, C_1 and C_{c2} in series;

$$\text{thus } C'' = \frac{2C_2C_1 + C_cC_2 + C_cC_1}{2C_1 + C_c}$$

$$\therefore Z_s = j\omega L_2 - j\frac{(2C_1 + C_c)}{\omega(2C_1C_2 + C_cC_2 + C_cC_1)} \quad \ldots (10)$$

Applying Rule 2c.

Assuming unit current in L_1, then P.D. across C_1

$$= -j/\omega C' \quad \ldots\ldots (11)$$

open-circuiting L_2, then current through C_{c1}, C_{c2} and C_2 will be:

$$I_s = \frac{-j}{\omega C'} \times \frac{\omega C_cC_2}{-j(2C_2 + C_c)} \quad \ldots\ldots (12)$$

$$= \frac{C_cC_2}{C'(2C_2 + C_c)} \quad \ldots\ldots (13)$$

$\therefore$ Voltage across secondary

$$E_s = \frac{C_cC_2}{C'(2C_2 + C_c)} \times \frac{-j}{\omega C_2} \quad \ldots\ldots (14)$$

Substituting for C':

$$E_s = \frac{-jC_c}{\omega(2C_2 + C_c)} \times \frac{2C_2 + C_c}{(2C_1C_2 + C_cC_1 + C_cC_2)} \quad \ldots\ldots (15)$$

$$= \frac{-jC_c}{\omega(2C_1C_2 + C_cC_1 + C_cC_2)} \quad \ldots\ldots (16)$$

$\therefore$ In the equivalent circuit referred to in the rules, the value assigned to the mutual coupling is then given by:

$$j\omega M = \frac{-jC_c}{\omega(2C_1C_2 + C_cC_1 + C_cC_2)} \quad \ldots\ldots (17)$$

$$\therefore M = \frac{-C_c}{\omega^2(2C_1C_2 + C_cC_1 + C_cC_2)} \quad \ldots\ldots (18)$$

Applying Rule 1

Z_m from Equation (18)

$$= \frac{-C_c}{\omega(2C_1C_2 + C_cC_1 + C_cC_2)} \quad \ldots\ldots (19)$$

Z_1 from Equation (8)

$$= \frac{-(2C_2 + C_c)}{\omega(2C_1C_2 + C_cC_1 + C_cC_2)} \quad \ldots\ldots (20)$$

Z_2 from Equation (10)

$$= \frac{-(2C_1 + C_c)}{\omega(2C_1C_2 + C_cC_2 + C_cC_1)} \quad \ldots\ldots (21)$$

$$\therefore k = \frac{-C_c}{\sqrt{(2C_2 + C_c)(2C_1 + C_c)}} \quad \ldots\ldots (22)$$

Appendix C

The effect of bandwidth on the noise limiting action by the typical circuit (Fig. 8(a)) following demodulation is examined below.

Fig. 8(b) shows random noise on A.F. as applied to the limiter inset (c) a noise pulse of 5μsec duration which is to be suppressed.

Assuming for simplicity that the pulse waveform is a periodic function of ωt, and the period is 2π, then the pulse is represented by the series:

$$y = A\sin\omega_1 t + A/3\sin 3\,\omega_1 t + A/5\sin 5\,\omega_1 t + \ldots$$

$$\therefore dy/dt = A\,\omega_1(\cos\omega_1 t + \cos 3\,\omega_1 t + \cos 5\,\omega_1 t + \ldots)$$

when $t = 0$ $dy/dt = A\,.\,\omega_1 n$

where ω_1 = angular frequency of the fundamental.

n = harmonic content of the pulse.

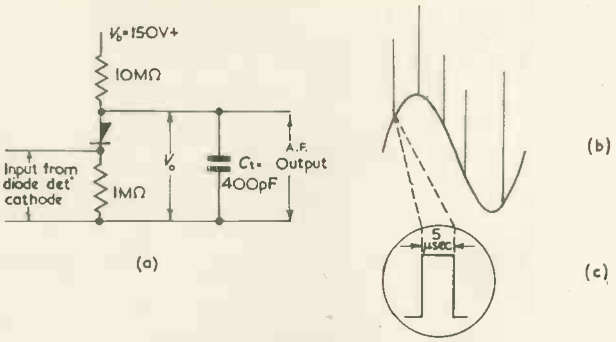


Fig. 8. Typical sound noise suppressor

The frequency components of the noise pulse shown are, fundamental 100kc/s, 3rd harmonic 300kc/s, 5th harmonic 500kc/s . . . to ∞.

Assuming an ideal receiver response the maximum rate of change produced by the noise pulse at the cathode is dependent on the receiver bandwidth as follows, assuming $A = 10$ volts.

Bandwidth 200kc/s $dy/dt = 2\pi \times 10^5 \times 10^{-6} \times 10 = 2\pi$ volts/μsec

„ 600 „ $dy/dt = 2 \times 2\pi = 4\pi$ „ „

„ 1Mc/s $dy/dt = 3 \times 2\pi = 6\pi$ „ „

The maximum rate of change which the anode circuit of the limiter can follow, is given by

$$\frac{\Delta V}{\Delta t} = \frac{\text{D.C. change across } C_t \text{ with diode non-conducting and conducting } (V_b - V_o)}{\text{time-constant of output circuit } (C_t \, . \, R_t)}$$

where $V_b = 150\text{V}$ $V_o = 1/11 \times 150 = 14\text{V}$.

R_t = effective resistance across C_t of the paralleled resistive components 10MΩ, 1MΩ + back resistance of diode, input resistance of A.F. amplifier (500kΩ) + R.F. stopper (33kΩ).

let $R_t = 500\text{k}\Omega$ and $C_t = 400\text{pF}$.

$$\frac{\Delta V}{\Delta t} = \frac{150 - 14}{400 \times 10^{-12} \times 5 \times 10^5} \, 1/10^6 = 0{\cdot}68\text{V}/\mu\text{sec}$$

thus, a signal producing a change at the cathode $> 0{\cdot}68\text{V}/\mu\text{sec}$ will be suppressed at the anode.

It is required that the maximum amplitude of the highest A.F. to be reproduced should pass undistorted, i.e. $dv/dt < 0{\cdot}68\text{V}/\mu\text{sec}$

if this A.F. component be given by $E_g = E_m \, . \sin\omega_m t$

$$deg/dt = E_m \, . \, \omega_m \cos\omega_m \, . \, t$$

$$\text{At } t = 0 \quad deg/dt = E_m\omega_m$$

Assuming a maximum A.F. of 15kc/s is to be reproduced and a 1·25V (peak) signal is required for maximum undistorted output from an ECL80 for example

$$\text{then } deg/dt = 1{\cdot}25 \times 2\pi \times 15 \times 10^3 \times 10^{-6} = 0{\cdot}12\text{V}/\mu\text{sec}.$$

The maximum rate of change of the 15kc/s tone for a pentode such as the PL82 requiring a 8·5V (peak) drive from the limiter for a maximum undistorted output of 4 watts becomes:

$$deg/dt = 8{\cdot}5 \times 0{\cdot}094 = 0{\cdot}8\text{V}/\mu\text{sec}.$$

The limiting action must therefore be reduced in the circuit given when it precedes a directly driven PL82.

Appendix D

Assessing Required F.M. Bandwidth for the European (C.C.I.R. System)

Pre-emphasis	50μsec
Frequency swing (ΔF_c)	± 50kc/s
Highest modulating A.F. (M_f)	15kc/s

Modulation index $(M_i) = \Delta F_c / M_f = 50/15 = 3{\cdot}3$

Required bandwidth $(B) = 2 \, . \, n \, . \, M_f$

where $n = M_i + 1 = 4{\cdot}3$ assuming that sideband currents $< 0{\cdot}1$ of the carrier current can be ignored,

$B = 2 \times 4{\cdot}3 \times 15 = 130\text{kc/s}$.

Similarly it can be shown that the required bandwidth for the American system is 80kc/s.

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Hopper Contents Control

Storage hoppers are vital parts of many processing plants, being employed for such purposes as the storage of raw, partly-processed, or completely processed materials, or as "reservoirs" to ensure that subsequent processes or machines are not "starved" in the event of a temporary hold-up during earlier stages of production.

An example of this is provided by the large-scale production of a well-known powdered chemical compound. The earlier stages of manufacture are concerned with abstracting certain solids from liquids, and then grinding and mixing them to produce the finished material, which is discharged on to a belt conveyor feeding a bucket elevator in the packing department. In this department there are two hoppers, one being 8ft deep and of 5-ton capacity, and the other considerably smaller. The first acts as a bulk storage hopper, and the second feeds an automatic weighing and packing machine.

This process is both continuous and automatic, and thus some means of controlling the amount of material in the hoppers is essential. This has been done in a simple and foolproof manner by using the Tyne B.L.49 bunker level control developed by Radiovisor Parent Ltd. This unit operates from the electrical conductivity of the material, and is quickly and easily installed. It incorporates a projecting rod-type electrode of any length to suit the requirements of the installation, which is electrically connected to a small control unit comprising a single-stage thermionic valve amplifier incorporating red and green lights, and terminals for connexion to an audible alarm.

Because "full" and "empty" conditions have to be indicated it was necessary to employ two electrodes, one at the top and the other near the bottom. To avoid chances of damage the position for the top electrode was chosen so that it is not in the direct line of flow of the material as it discharges into the hopper. Except when the hopper is full, a green light shows on the control panel, indicating that filling can proceed. The material level then rises until it touches the electrode and completes a circuit through the material to earth, thus closing a relay and causing the red light to show. At the same time, the motor responsible for feeding the material to the hopper is automatically stopped. To provide an even output of powdered material, Locker vibratory equipment is fitted at the outlet of the bulk hopper, this discharging the material into a second bucket elevator feeding the smaller hopper. In the case of this hopper it has been possible to arrange the two electrodes vertically side by side; one is short and the other longer to control respectively the maximum and minimum levels of material. As before, these are fitted away from the direct line of flow of the material as it enters the bunker. A vibratory plate is provided at the outlet, and this discharges the material into bags on an automatic bagging and weighing machine situated directly below. As each sack is filled with the correct amount, the supply is cut off temporarily while the operator seals the top and transfers the sack to a belt conveyor.

This system is, of course, equally applicable to the control of liquids in tanks etc., or to larger solids such as coke, sand, crushed stone, etc., and one of the special advantages claimed for the Radiovisor equipment is that it is designed to operate with materials which behave as insulators in other systems operating on a similar principle.

A Magnetic Impulser

By C. A. Routledge *

IN punched card operated electrical calculating and printing equipment it is customary, for the sake of reliability and ease of service, to control the make and break of current in a circuit by a small number of quickly replaceable contacts or circuit breakers, having tungsten or other similar points capable of withstanding the load imposed when switching circuits on or off one or more times every machine revolution; the many relays and other devices with light duty contacts only rearrange the circuits but do not themselves normally make or break current.

These circuit breakers have functioned satisfactorily on equipment used until now, but with the advent of higher machine speeds involving the use of more sensitive relays, or electronic devices, requiring short, accurate impulses free from distortion or breaks due to contact bounce, and also because of the much increased wear of mechanical circuit breakers at higher speeds, it has become necessary to look for alternative methods, such as the magnetic impulser to be described.

The function of the impulser is to provide a source of electrical impulses, suitable both in magnitude and time, for the direct control of valve circuits similar to those shown in the accompanying diagrams, the valves being used instead of circuit breakers for high speed switching.

The device consists of an energizing coil with an iron core, to the ends of which are screwed two iron yokes. A brass strap is screwed to the yokes so that the core, strap and yokes form a rigid framework, and at the free ends of the yokes four iron pole pieces are attached on which are wound the valve control coils A, B, C, D (Figs. 1 and 2). A and B are wired as one pair and C and D as the other, and the pole pieces are arranged so that there is a gap between them through which the teeth of the timing wheel pass. This wheel, which is also made of iron, has a number of square cut teeth around the periphery; the example shown, which is for a 16 point machine, has 16 teeth, but as only 12 points are to be pulsed, these points are selected by a contact in the load circuit.

The method of operation of the device is as follows:—

The energizing coil is permanently operated by a direct current, and the resultant magnetic field creates a flux through the magnetic circuit made by the core, the two yokes and the two parallel paths formed by the pole pieces and their air-gaps.

As the wheel rotates and its teeth pass between the pole pieces, the circuit reluctance decreases and the flux density increases, the change of flux causing a voltage to be induced in coils A and B, or C and D, according to which pair of pole pieces have a tooth passing between them; actually two successive impulses of opposite polarity are induced in a pair of coils as each tooth is passed between their pole faces, a pulse in one direction as the leading edge of a tooth passes across, and a pulse in the opposite direction as the trailing edge of a tooth passes across; for the present purpose only the first impulse is used, being applied in a positive direction to fire a valve, the negative impulse having no effect other than to increase the bias. To obtain the maximum effect coils A and B are wired in series, as also are coils C and D.

The pulse-to-space ratio is determined by the relative widths of a wheel tooth and pole piece thickness and by the spacing of the pole pieces around the circumference; in the example shown a pulse-to-space ratio of between 1 to 3 and 1 to 2 is obtainable, the tooth width at maximum diameter being approximately $\frac{1}{4}$in., the pole piece thickness being $\frac{1}{8}$in., and the angular spacing between the pairs of pole pieces being equal to $1\frac{1}{2}$ teeth. The pole piece fixing holes are relatively large, so it is possible to vary the position of the pole pieces along the yokes, and the whole framework containing the energizing coil and the pole pieces is adjustable along a radius of the wheel, so by combining these two sets of adjustments it is possible to vary the pitch of the pole pieces around the wheel and thereby obtain a fine adjustment of the pulse length.

A further point which has been observed in the impulser

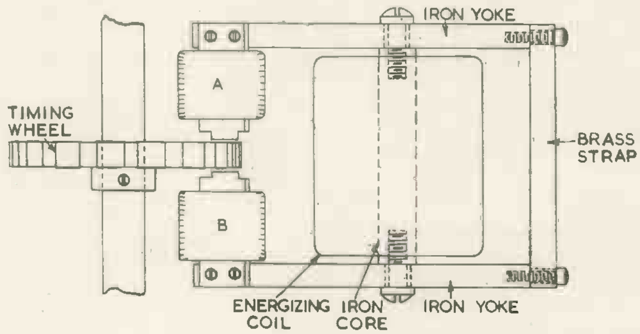

Fig. 1. The magnetic impulser

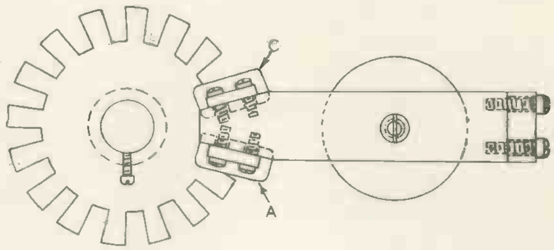

Fig. 2. The timing wheel

design is that the ratio of pole piece width to tooth width, and the angular spacing of the pole pieces, has been so arranged that as a tooth is entering between one pair of pole pieces another tooth is leaving the other pair, and this arrangement ensures that the total flux in the circuit is practically constant, merely being switched from one pair of pole pieces to the other by the teeth of the wheel. The advantage of this arrangement is that there is no feedback on to the energizing coil due to variations in the flux, and the load of the energizing coil on its supply circuit remains constant; as the field generated by the coil is quite large, interference from this source was marked before the above conditions were observed.

One application of the device is in the circuit shown in Fig. 3, in which two gas-filled triodes V_1 and V_2, of the GT1C type, are wired together to form a flip-flop circuit, the coils A and B, C and D, being wired in the grid circuits

* *British Tabulating Machine Co. Ltd.*

as shown. The resistor R_2 is a permanent load in the anode circuit of valve V_2, and the resistors R_1 are the various machine circuits for which valve V_1 is acting as a circuit breaker; with resistor R_2 and capacitor C_1 of fixed value, the total value of resistors R_1 can be varied to some extent. Cam 1 is a method of providing the contact mentioned earlier for selecting the 12 points to be impulsed out of the 16 available; an alternative method which was tried for providing pulses at 12 out of the 16 points is to cut only 12 teeth on the timing wheel instead of the full 16, and the cam 1 can then be dispensed with, resistors R_1 being wired straight to H.T. This method was dropped as it puts the timing wheel out of mechanical balance.

The operation of the circuit is as follows:—

Suppose the machine to be at rest with cam 1 open, the bias and H.T. supplies are switched on, and valve 2 may or may not fire according to whether the bias or H.T. is switched on first, and if the bias is first, according to whether a tooth of the timing wheel is between the poles of coils C and D or not, if a tooth is present the rising flux in the magnetic circuit of the core, yokes, pole pieces, and a tooth of the timing wheel, caused by the rising field of the energizing coil, will generate an impulse in coils C and D, which by the wiring of these coils will give a positive

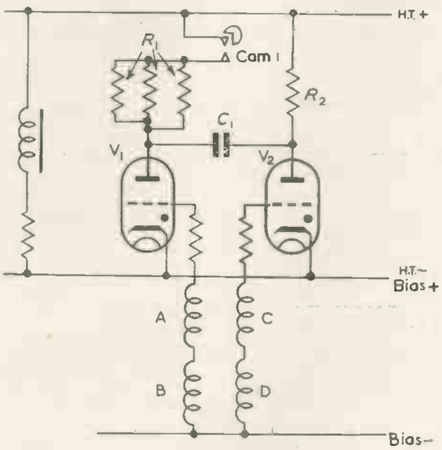

Fig. 3. A circuit using the magnetic impulser

impulse to the grid of V_2, which will then fire; V_1 cannot fire at this stage as cam 1 is open.

When the timing wheel rotates a series of impulses will be generated in coils A, B, and C, D, the wiring of these coils being such that positive impulses arrive at the grids of the valves when a tooth is entering the gaps between the pole faces of A, B, and C, D. V_2 will stay fired until after cam 1 makes, which it will do just after a positive impulse has arrived at V_2 and before the next one arrives at V_1. With cam 1 closed, the next positive pulse on the grid of V_1 will fire this valve, which will pass current through the machine circuits shown as resistors R_1, and at the same time pass an impulse through capacitor C_1 to the anode of V_2 causing it to stop conducting for long enough to allow the grid to regain control, in the usual manner of a flip-flop circuit. Similarly, when the next tooth enters the gap between the poles of A and B, a positive impulse will be given to the grid of V_2, causing it to fire and in so doing stop V_1 from conducting, thereby stopping the current flow through the machine circuits represented by resistors R_1. In this way a series of impulses will be given to resistors R_1 by V_1, the duration and timing of these impulses being controlled by the timing wheel in conjunction with coils A, B, and C, D.

A variation of the circuit in Fig. 3 is shown in Fig. 4, where the valves V_3 and V_4 take the place of valves V_1 and V_2, and perform the same circuit breaking functions, but V_3 and V_4 are of much larger capacity, being of the Mullard CT10-12 type, capable of controlling several amperes. The grid current of these valves is, however, also large, and to avoid the attendant difficulties of controlling them directly by the magnetic impulser, a pair of GT1C valves are interposed as shown by valves V_5 and V_6. The magnetic impulser controls V_5 and V_6 directly, causing them to fire in turn under the control of the timing wheel, and the alternate firing of these valves in turn causes V_3 and V_4 to fire at the same rate. Thus, when V_5 is fired and V_6 is not, V_4 will be fired and V_3 not; when V_6 receives a positive impulse and fires, V_5 will cease conducting, the voltage on the grid of V_3 will rise and V_3 will conduct, which will in turn cause V_4 to cease conducting, and V_4 will be held non-conducting because its grid voltage is lowered by the firing of V_6.

V_5 and V_6 will flip-flop continually under the control of the timing wheel, but the selection of the required 12 impulses will be made by cam 2, as was done by cam 1 in the circuit of Fig. 3.

Additional relay controls are shown in Fig. 4, and these

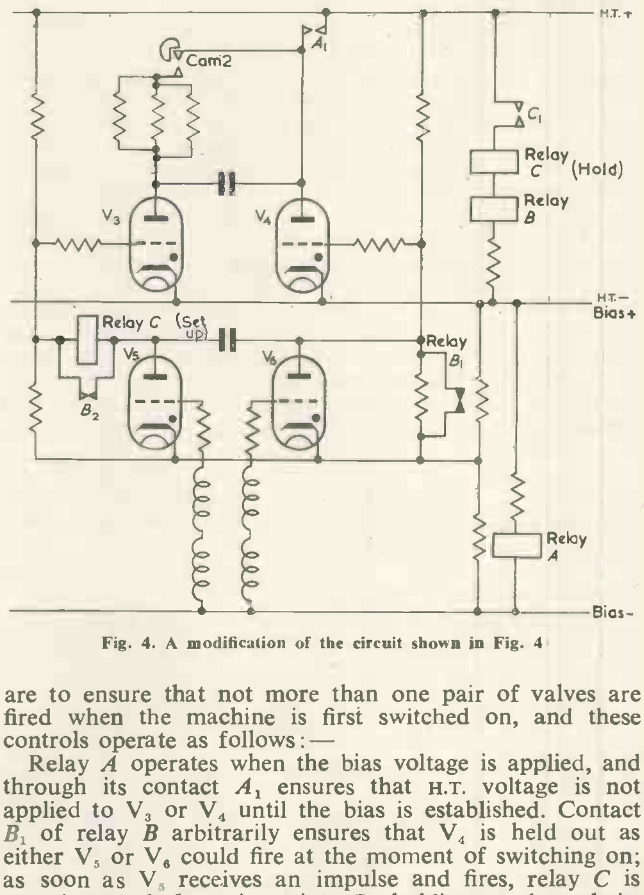

Fig. 4. A modification of the circuit shown in Fig. 4

are to ensure that not more than one pair of valves are fired when the machine is first switched on, and these controls operate as follows:—

Relay A operates when the bias voltage is applied, and through its contact A_1 ensures that H.T. voltage is not applied to V_3 or V_4 until the bias is established. Contact B_1 of relay B arbitrarily ensures that V_4 is held out as either V_5 or V_6 could fire at the moment of switching on; as soon as V_5 receives an impulse and fires, relay C is energized and closes its points C_1, holding on through its second coil until the machine is switched off; also when contacts C_1 close, relay B is energized, opening its contacts B_1 to allow V_4 to fire, and closing its contacts B_2 to short out the set-up coil of relay C, thereby equalizing the anode loads of V_5 and V_6.

A further simple use of the impulser is to control the pulsing of one or more hard valves, but as in this case there is no flip-flop action, only one pair of coils, such as A and B, are required.

Acknowledgments

Acknowledgment is given to the British Tabulating Machine Co., Ltd., for permission to publish this article.

A Phase Shifting Pulse Generator for Thyratron Control

By J. C. West*, Ph.D., A.M.I.E.E., and D. K. Partington*, M.Sc.

A CIRCUIT is given for producing positive pulses of constant amplitude at the supply frequency repetition rate of 50c/s, but with variable phase relative to the supply voltage. The amount by which the pulses lag behind the supply voltage is controlled by a small D.C. potential applied to the grid of a high vacuum valve. This feature is of importance in servomechanisms, in which thyratrons are controlled by a D.C. signal.

A range of 128° has been achieved over which the phase may be controlled, the control being linear for a large proportion of the range, with a sensitivity of 51·5 degrees/volt. The amplitude of the pulses is 235 volts and the rise time is 40μsec.

In systems using several grid-controlled gas-filled valves in parallel[1], and when it is important that a thyratron fires at exactly the same point of the cycle in successive cycles, it is necessary to use impulse or "hard" control.

The impulses may be obtained by mechanical methods, using a synchronously driven commutator, or statically, using electromagnetic or electronic devices. It is, however, generally preferable to avoid using auxiliary rotating machinery, so that mechanical methods are now little used.

Several systems have been developed, using saturable reactors[2], but these fail to give sharp impulses, and better results are obtained using impulse transformers[3,4]. The latter, however, require careful manufacture and shunts of Mumetal have to be used in their construction, so making them expensive. Purely electronic methods for producing phase shifted pulses are generally too complicated to be suited for normal systems[5], but have the advantage that there is no appreciable time-constants involved in their operation.

The circuit described below is relatively simple and inexpensive, requiring only two valves and an ordinary audio-frequency inter-valve transformer, instead of the special impulse transformer.

Principle of Operation

The basic circuit is shown in Fig. 1, in which r is a variable resistor in one arm, R_1 and R_2 are of fixed value and

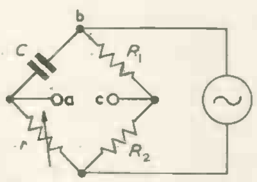

Fig. 1. Basic phase shifting circuit

C is a fixed capacitor. If r is assumed to be infinite, then a and b are at the same potential, and the voltage between a and c is in phase with the supply voltage. If r is zero, then a is connected to d and the voltage between a and c is 180° out of phase from the supply voltage. For intermediate values of r, the voltage across a and c is out of phase with the supply voltage by a corresponding amount. Hence, by varying r, a variable phase shifted voltage may be obtained between a and c.

Instead of r being varied manually, a valve may be used in place of the resistor, the impedance of which is controlled by the amplitude of a D.C. signal fed to its control grid. This is of advantage in automatic control systems, in which the controlling signal is in the form of a voltage rather than as a mechanical displacement. Because the valve is a non-linear device, the phase shifted output across a and c is no longer sinusoidal and is applied immediately to a circuit which converts the output to a series of pulses occurring each time the voltage waveform passes through zero. This avoids the direct use of the phase shifted waveform for control purposes, which would be unsatisfactory, because the wave shape varies with the amount of phase shift. If the capacitor C is made sufficiently small, pulses are obtained only in one direction, occurring once every cycle of the supply voltage.

Theory of Circuit

The operation of the circuit is shown in more detail by a mathematical analysis:

The voltage between points a and c, when the supply voltage $E\cos\omega t$ is applied to points b and d is,

$$E_{ac} = \left[\frac{jrC\omega}{1 + jrC\omega} - \frac{R_2}{R_1 + R_2}\right] E\cos\omega t$$

$$= \left[\frac{(R_1 + R_2)\,jrC\omega - R_2 - R_2 jrC\omega}{(R_1 + R_2)(1 + jrC\omega)}\right] E\cos\omega t$$

$$= \left[\frac{R_1 jrC\omega - R_2}{(R_1 + R_2)(1 + jrC\omega)}\right] E\cos\omega t$$

If G is the amplitude and ϕ the phase of E_{ac}, then:

$$\phi = \tan^{-1}\frac{(R_1 + R_2)rC\omega}{r^2R_1C^2\omega^2 - R_2}$$

$$G = \frac{\sqrt{(R_2^2 + r^2R_1^2C^2\omega^2)}}{\sqrt{[(R_1 + R_2)^2(1 + r^2C^2\omega^2)]}}$$

If $r = 0, \phi = 180°, G = \dfrac{R_2}{(R_1 + R_2)}$

If $r = \infty, \phi = 0°, G = \dfrac{R_1}{(R_1 + R_2)}$

When $\phi = 90°$, $r^2R_1C^2\omega^2 = R_2$

$$r = \sqrt{\frac{R_2}{R_1C^2\omega^2}} \text{ and } G = \frac{\sqrt{(R_1R_2)}}{(R_1 + R_2)}$$

Thus, by making the two fixed resistor arms of equal value, the amplitude of the output signal is independent of the phase shift, over the complete range of 0 to 180°. The use of a transformer connected between a and c modifies the action of the system, so that the maximum range over which the phase can be varied, is limited to 130°.

Details of Circuit

Originally, it was intended to use a pentode as the phase controlling valve, with both constant H.T. voltage from a power pack and alternating voltage from the network applied to its anode. By using suitable values for the anode and screen grid resistors to the H.T. supply it was hoped to reduce the equivalent resistance of the valve to a minimum of 200Ω. Owing to the range over which

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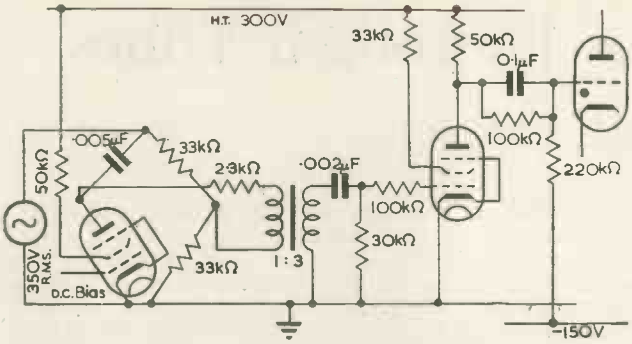


Fig. 2. Complete circuit of the phase shifting pulse generator

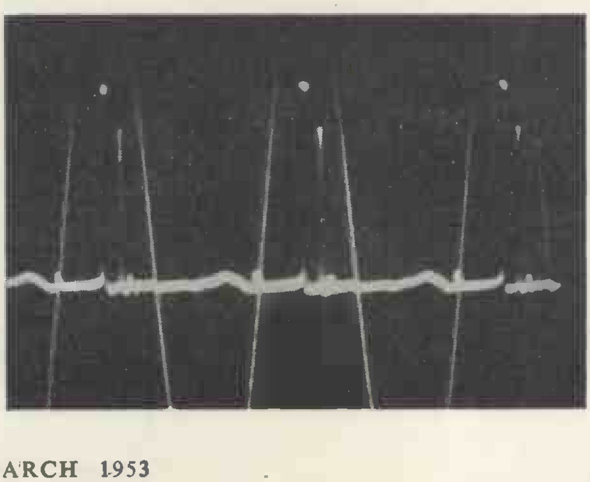
the phase may be controlled being limited by the transformer connected between a and c, it was found unnecessary to use such an elaborate system, only the screen grid being held at a constant potential.

Fig. 2 shows the complete circuit, with the values of the components. The phase shifted output is applied to a 1 : 3 audio-frequency intervalve transformer, of which the core is nickel alloy. This becomes magnetically saturated in a similar manner to an impulse transformer, and the peaky output waveform passes through a differentiating circuit of 0·002μF and 30kΩ. The time-constant of this circuit is so small (60μsec) that impulses are only produced if the rate of change of voltage is sufficiently high. The signal from the differentiating circuit is damped oscillatory, the beginnings of the oscillations being governed by the phase shifted output to the transformer, and the amplitude and the damping vary with the actual phase shift. To obtain pulses of constant amplitude irrespective of phase shift, this signal is applied to the control grid of a second pentode. The first peak of each oscillation is arranged to be negative by correct connexion of the secondary of the transformer, and is of sufficient magnitude to cut off the pentode. This results in the anode of the pentode rising from approximately 100V to H.T. potential, and it remains at this value until the negative peak has fallen almost to zero.

It is these pulses that are of importance in thyratron control, as pulses applied to the grid of a thyratron subsequent to triggering have no effect on its operation. At low values of phase change the second negative peak is also large enough to cut off the pentode, but this is unimportant as the thyratron will have been triggered already. The anode of the second pentode is D.C. connected to the grid of the thyratron, which is held normally at a negative potential relative to earth because of the potential dividing action of the 220kΩ and 100kΩ between the negative line and the anode of the pentode.

The damping is so great that subsequent oscillations,

Fig. 3. Pulses with a phase shift of 80°



when they exist, fail to cut off the pentode. Therefore positive pulses are produced once every cycle of the mains supply, the phase of the pulses being controlled by the D.C. voltage applied to the grid of the first pentode. When the pentode is cut off, the pulse is applied to the grid of the thyratron, the capacitor speeding up the forward edge of the pulse. The pulses are large enough to cause the grid of the thyratron to rise to high positive voltage, and so triggering of the thyratron is almost instantaneous and independent of fluctuations of its triggering potential.

Results

A photograph of the pulses obtained is shown in Fig. 3.

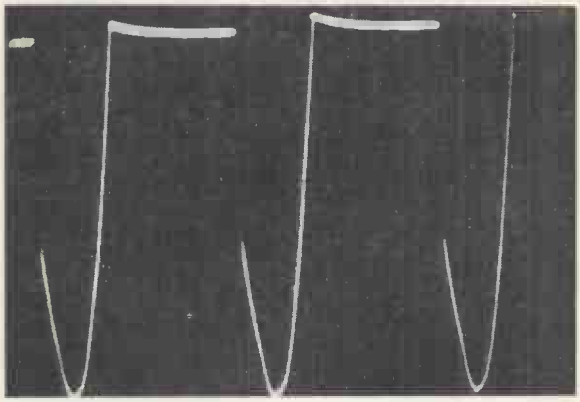


Fig. 4. The current waveform of a thyratron triggered by the generator

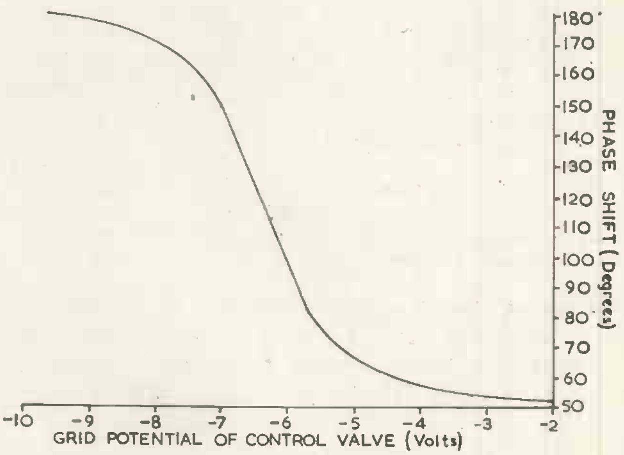


Fig. 5. The phase shift of the output pulse against the grid potential of the control valve

The amplitude is 235V and the time for this value to be reached is 40μsec. The circuit was used to trigger a thyratron feeding a pure resistive load and the current wave-shape is shown in Fig. 4. The graph of the grid potential of the control pentode against the phase shift of the pulses is shown in Fig. 5, and it will be seen that control is linear between phase shifts of 80° and 150°. The slope of the graph is 51·5 degrees/volt and, therefore, the system is very sensitive to changes in the control voltage.

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Repetitive Working of Photoflash Tubes

By A. S. V. McKenzie*, B.Sc. and D. B. Cleland*, B.Sc.

THE simplest method of using a photoflash tube for repetitive working is to charge up a capacitor C through a resistor R and discharge it through the tube by triggering the latter (Fig. 1). This circuit, however, has considerable limitations when high repetition rates are required, as the following considerations will show.

The power-pack voltage V is limited by the voltage rating of the tube, and above this voltage the tube is liable to be self-triggering. The resistor R must be as low as possible to allow C to charge up to as high a voltage as

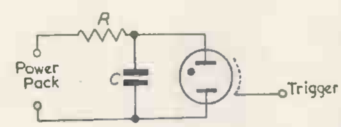

Fig. 1. Simple circuit for photoflash tube

possible between flashes, but there is a minimum value of R below which the tube is liable to go into a continuous discharge. This minimum value is a function of the power-pack voltage V, for example with the 100 joule tube used in this case the smallest safe value of R was 20 000 ohms when V was 2 500 volts. The value of the capacitor C is obviously important, since it has to charge up through R between flashes, and it may be readily shown that there is an optimum value of C for maximum electrical energy per flash at each operating speed, given by the expression

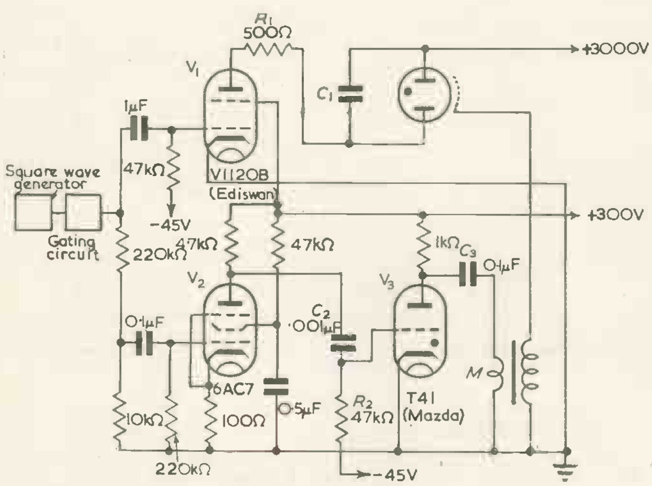

Fig. 2. The complete circuit

$T = 1{\cdot}26CR$, where T is the time between successive flashes. The light output from the tube is not directly proportional to the electrical energy stored in C, as it is a function also of C and the voltage across the capacitor at the time of flash, so that a full analysis would be difficult. However, it is clear that there is an optimum value of C for maximum light output with a given T, but even if the optimum value is chosen in each case, the light output of the individual flashes decreases approximately linearly with T, and for most photographic purposes the highest feasible repetition rate is a few hundred cycles per second.

One method of obtaining larger light outputs at higher speeds is to control the flashes by means of a hydrogen-filled thyratron[1], but the following circuit is a considerable improvement on the simple one, and involves the addition of only one hard valve to the latter.

Essentially the circuit (Fig. 2) uses a hard valve V_1 (a V1120B was found to be suitable) to control the charging of the reservoir capacitor C_1 from the power-pack. The circuit is controlled from a square-wave generator, the output of which is gated to limit the flashes to the number required; the type of gating circuit used depends, of course, on the particular application. This square wave is applied to the grid of V_1, and is of sufficient amplitude to switch

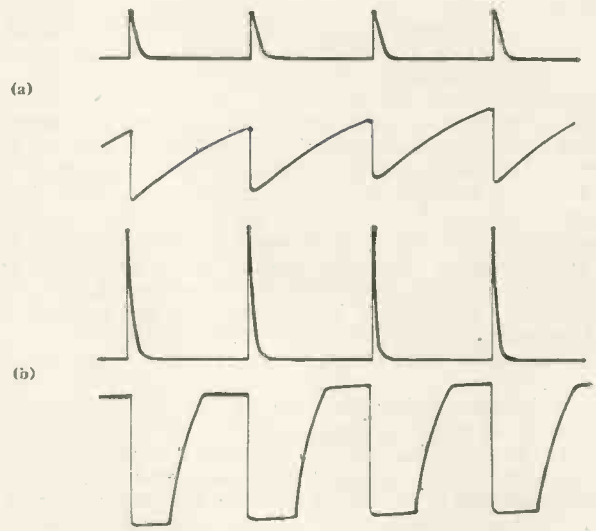

Fig. 3. Oscillograms of the voltage across the reservoir capacitor

this valve on and off. The same square wave is also inverted by V_2 and differentiated by C_2R_2. The positive pulse of the differentiated output is then used to trigger the thyratron V_3, which discharges a capacitor C_3 through the primary of the spark coil M, thus providing a damped oscillatory triggering pulse in the usual way.

Suppose V_1 is in a conducting state and C_1 is charged to the power-pack voltage. When the triggering pulse occurs, C_1 is discharged through the tube, and at the same time V_1 is cut off completely for half a cycle, so that no voltage at all appears across the tube for this time (apart from the residual voltage), and the discharge is completely quenched. After half a cycle, V_1 is again rendered conducting, and C_1 is recharged rapidly to the full power-pack voltage, the charging current being limited only by R_1, which is only a few hundred ohms. Fig. 3 shows traces of oscillograms of the voltage across the reservoir capacitor (in this case $0{\cdot}1\mu F$), using (a) a 20 000 ohm resistor for charging, and (b) the present type of circuit. The same power-pack voltage (2 500 volts) and the same repetition rate (600 flashes per second) were used in each case. The upper traces show the corresponding integrated light out-

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put, recorded by a photo-multiplier. Note that in case (a) the reservoir capacitor only has time to charge to a fraction of the power-back voltage, whereas in (b) it is fully charged before each flash. Consequently in (b) the light output is considerably greater.

In addition to quenching the discharge after each flash this circuit allows the tube a full half-cycle after each flash during which deionization can take place with only the residual voltage across the tube, so that the possibility of spurious flashes occurring during this time, due to build-up of voltage across the tube before it is sufficiently deionized, is eliminated. The highest repetition rate obtainable using this circuit is, in fact, limited by the deionization time of the tube. At higher speeds the tube does not have time to deionize sufficiently during the half-cycle after a flash, and when the full power-pack voltage is applied after half a cycle, spurious flashing may occur. The effective deionzation time is, of course, a function of C_1, and the maximum speeds obtainable with various values of C_1 were approximately:

C_1 (μF)	V (volts)	$\frac{1}{2}CV^2$ (joules)	MAX. SPEED (flashes/sec.)
0·15	3 000	0·67	1 200
0·3	—	1·35	600
0·6	—	2·70	300

Better performances than this can be obtained by using a hydrogen-filled thyratron in series with the tube to control the flashes, as is described in the reference given. However, the present circuit has some advantages. It is simpler and uses only familiar valves. Also, since there is no thyratron in series with the tube, the energy from the capacitor is dissipated entirely in the tube, and not partly in the thyratron. There is thus no trouble about having to use a flash tube which will match the impedance of the thyratron.

The circuit was used to enable photographs of cam systems to be taken at speeds of up to 1 200 frames per second. Photography was done by a simple drum camera

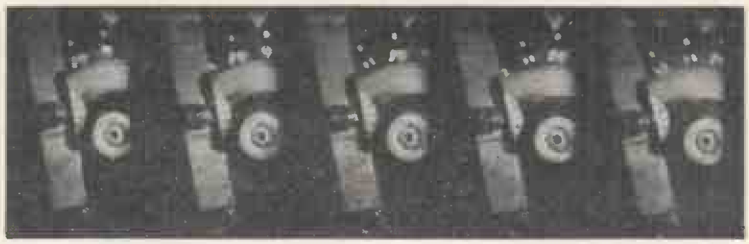

Fig. 4. Part of a series of photographs taken at 1 000 frames per second, using the circuit described

(without shutter), using an $f/2{\cdot}8$ lens. The number of exposures which could be taken in one sequence was, of course, limited by the rating of the tube. In general, it can be said that the addition of the valve V_1 to the simple circuit allowed the light output per flash to be increased by some three or four times at the higher repetition rates, though the performance of individual tubes of the same type varied somewhat.

Some possible improvements to the circuit suggest themselves; for example, a longer time could be allowed for deionization by using a non-symmetrical multivibrator, of ratio about four to one, to control the charging instead of a square-wave generator. It could then be arranged that V_1 is cut off for four-fifths of a cycle, instead of for half a cycle. Even higher speeds might be obtained by incorporating a charging valve such as V_1 in the thyratron-controlled type of circuit, since in the latter type of circuit the top speed is still, presumably, limited by the deionization time of the thyratron. Those workers who are familiar with the circuits associated with Geiger-Müller counters will recognize the similarity between the present problem and the problem of obtaining high counting rates, and it is possible that some more of the techniques used in counter circuits might be applied to this problem.

Acknowledgments

The work described above formed part of an investigation carried out at the Engineering Department of the University of Edinburgh for the Mechanical Engineering Research Organization of the D.S.I.R.

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A Simple Wide Range Sine Wave Generator

By A. D. Booth,* D.Sc., Ph.D., F.Inst.P.

IT is the purpose of this article to describe the circuit details of an extremely simple wide range sine wave generator which has been constructed in this laboratory and which has been found very satisfactory in operation.

No great originality is claimed for the general principle of operation, but the circuit has been found to be easy to set up, stable in frequency, and accurate in its amplitude output. The latter virtue was notably absent from two earlier circuits which were culled from the literature[1,2]; in particular they would either not oscillate at all over part of the range, or alternatively would act as square wave generators.

General Principle

The circuit used depends upon the inter-connexion of a linear amplifier and a phase delay element as shown in Fig. 1.

Suppose the amplifier to have gain A and the phase shifter to produce a delay with attenuation $1/A$, then under steady state conditions the functional relationship to be satisfied by the amplifier input, $f(t)$, is:

$$f(t) = f(t - \phi)$$

an equation, the fundamental solution of which is:

$$f(t) = \sin(2\pi\, t/\phi) \quad \text{.............} \quad (1)$$

thus the system shown will oscillate with frequency $1/\phi$. The above elementary derivation assumes that the amplifier gain is amplitude and frequency invariant. Furthermore, frequency stability will only occur if the delay element has a sharp minimum in its attenuation for the given frequency.

The Phase Shift Network

This is what has been (somewhat incorrectly) described

* *Birkbeck College Computation Laboratory.*

as a Wien bridge, it is shown in Fig. 2.

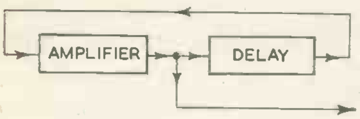
The frequency of oscillation as predicted by (1) is easily shown to be[3]:

$$f = 1/2\pi\, RC \text{ cycles per second} \quad \ldots\ldots\ldots (2)$$

The arrangement has the practical advantage of requiring the simultaneous variation of only two elements either R or C, in order to produce the desired range of frequency, and also that each pair of elements consists of identical components.

The Complete Circuit

This is given in Fig. 3. Power supply and output stage are included for completeness, but any laboratory power unit capable of producing the required voltages may be substituted. Similarly the output pentode V_4 and its



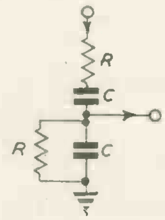


Fig. 1. The principle employed. Fig. 2. The phase-shift network

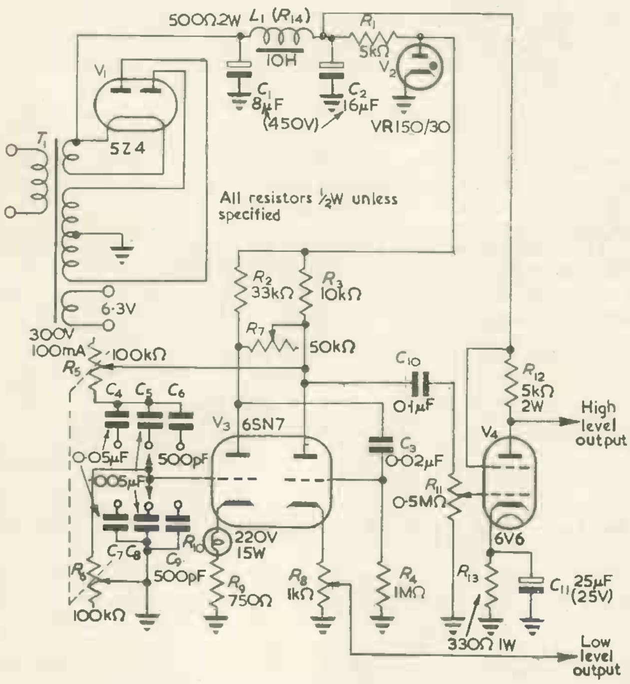


Fig. 3. The complete circuit

associated circuit, may be omitted if only low level output is required.

T_1 is a standard 300V, 100mA centre tapped transformer, it supplies a full wave rectifier V_1 feeding a filter network $C_1L_1C_2$. Note that if the high level output is not required the choke L may be replaced by a simple resistor R_{14}. Stabilized 150V D.C. is produced from voltage regulator V_2 fed by the limiting resistor R_1.

The amplifier consists of the two halves of V_3. Straightforward RC coupling is adopted via the grid network $R_2C_3R_4$. The feedback path and Wien bridge is via $R_5C_4..C_6$; $R_6C_7..C_9$. R_5 and R_6 are ganged and operate simultaneously. $C_4..C_9$ are switched to give the required frequency range.

The twin triode sections are biased by R_9, R_{10} and R_8 respectively. In order to obtain a constant amplitude output, independent of frequency, no by-pass capacitors are used and R_{10} is a 15W, 220V lamp. This component is indispensable, since the large positive temperature coefficient of the tungsten filament is the main, frequency invariant, amplitude stabilizer.

Further gain control and stabilization is given by R_7 and when the circuit is built this is set so as to give the required output amplitude at the highest frequency, other lower frequencies will be found to be of the same amplitude.

A low level, low impedance, output is to be had at R_8 and further attenuation can be inserted as required.

V_4 is a straightforward power amplifier whose gain is controlled by R_{11}.

Should a sonic output be required, L_1 may be replaced by the field coil of a speaker, the transformer input to which is substituted for R_{12}.

Frequency Range

A considerable advantage of the above circuit lies in its predictable frequency output. This can be calculated from Equation (2) and a range of values is given in Table I.

TABLE I Frequency/Resistance

$C = 10^n \times 0.0005\ \mu F$

$f(c/s)$	$R.(ohms)$
50 ($\times 10^n$)	63 600
75	41 400
100	31 800
150	21 200
200	15 900
300	10 600
400	7 950

It has been found that an adequate means of calibrating the generator is to adjust the ganged variable resistors R_5 and R_6—by means of a meter or bridge—to the values given in Table I, and to mark a scale on the front of the generator at these values. At the low end of the scale a check can be made by comparing the output with the standard 50c/s mains frequency.

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3. VALLEY and WALLMAN. Vacuum Tube Amplifiers, p. 384 (McGraw Hill Pub. Co., 1948).

New Civil Aviation Communications Centre

AN Aeronautical Fixed Telecommunications Network of teleprinter, radio-teletype and hand-operated morse circuits radiating to all parts of the world has been set up within and between the member states of the International Civil Aviation Organization. The new Croydon Communications Centre is the heart of the United Kingdom's part of this network. It replaces the former Signals Centre at London Airport and marks the beginning of a major rationalization of the Aeronautical Fixed Telecommunications Network which has grown up piecemeal in the transition period between war and peace.

The Centre is housed in one of the Croydon Airport buildings which has been entirely reconstructed internally. Its radically new layout has been designed in the light of experience to contain the latest automatic teleprinter equipment, message tubes, etc., arranged in such a way as to streamline the functions of the individual operator. The greatly increased speed of the latest types of aircraft makes it more important than ever before to pass messages, e.g., reporting their time of arrival, well ahead of the aircraft themselves.

Initially about 1 000 transactions will be handled in the peak hour over some 30 circuits; the need for expansion has however been taken into account and the Centre so designed as to deal ultimately with 3 000 transactions an hour at peak over more than 60 circuits. Croydon Airport has been chosen as the most suitable site in the London area as a point easily connected with G.P.O. circuits and at the same time near the major users of the network.

Notes from the Industry

The Radio Industry Council have announced that H.M. Queen Mary has graciously consented to be patron of the National Radio Show to be held in London in September. Her Majesty has been patron on each occasion since the war. Provisional dates for the Radio Show are September 2 to 12, with a preview for overseas visitors and other special guests on September 1.

Dr. Walter Cawood, C.B.E., is to be the Ministry of Supply's Principal Director of Scientific Research (Defence). He succeeds Dr. O. H. Wansborough-Jones, C.B., O.B.E.

Isotope Developments, Ltd. have been awarded a substantial contract for the equipment of the Belgian Civil Defence with electronic instruments for use in case of atomic warfare. Two types of portable equipment are concerned, each housed in a service pattern haversack and designed for use by Civil Defence personnel.

Wickman, Ltd. announce that their interests in industrial high frequency heating equipment under the name of Applied High Frequency, Ltd. are now run directly as a division of the company under the title of Wickman, Ltd., Applied High Frequency Division, Actarc Works, Goldhawk Road, London, W.12.

The Regional Advisory Council for Higher Technological Education has, for the fifth year in succession, prepared a summary of the applied research in electrical engineering in progress in university colleges and technical colleges in London and the Home Counties. In issuing this document the Council hopes to stimulate and further the undertaking of applied research and to assist, where required, in establishing contacts between firms and colleges concerned.

E. K. Cole, Ltd. announce that production of the EKCO Solder Pencil has been suspended through pressure of other work. It is regretted that no further supplies are available. Users of existing solder pencils are, however, assured of full spares facilities for the maintenance of these tools.

Telcon Telecommunications, Ltd. Mr. R. R. C. Rankin, O.B.E., A.M.I.E., A.R.I.C., a director of Mullard Equipment, Ltd., has been appointed by Mullard, Ltd., to be a director of Telcon Telecommunications, Ltd., owned jointly by Mullard, Ltd., and the Telegraph Construction and Maintenance Co., Ltd., in place of Dr. C. F. Bareford, who has resigned from the board following his appointment as Chief Superintendent of the Long Range Weapons Establishment at Salisbury and Woomera, South Australia.

The Royal Society announce that the Bakerian Lecture for 1953 will be delivered on May 7 by Professor N. F. Mott, F.R.S., Professor of Physics in the University of Bristol and Director of H. H. Wills Physics Laboratory. The Bakerian Lecture of the Royal Society was founded by Mr. Henry Baker, F.R.S., for a yearly oration or discourse by one of the Fellows on some part of natural history or experimental philosophy. The first lecture was delivered in 1775.

The Atomic Energy Research Establishment, Harwell, are inviting applications from physicists and electronic engineers holding a degree, or equivalent qualification, who wish to attend a specialized course on the design, use and maintenance of electronic instruments used in nuclear physics, radiochemistry, and in work with radioisotopes. The course, to be held at the Isotope School, Harwell, will be from Monday, March 16, to Friday, March 20, 1953. The fee for the course is 12 guineas and living accommodation, morning and evening meals will be provided at a charge of 7 guineas. Application forms may be obtained from the Electronics Division, A.E.R.E. Harwell, near Didcot, Berks.

The National Institute of Oceanography announce their change of address to Wormley, near Godalming, Surrey. The nearest railway station is Witley.

The Council of the Institution of Electrical Engineers have elected Sir Harry Railing, D.Eng., to Honorary Membership of the Institution for his services to the electrical engineering profession and to the science; and for his services to the Institution. The Council of the Institution have also made the thirty-first award of the Faraday Medal to Colonel Sir A. Stanley Angwin, K.B.E., D.S.O., M.C., T.D., B.Sc.(Eng.), for his outstanding contributions to the development of telecommunication in Great Britain and in the international and intercontinental fields.

Dr. Thomas E. Allibone, F.R.S., has been appointed to the Board of the Edison Swan Electric Co., Ltd., in the capacity of Director of Research. Dr. Allibone will retain his position as Director of A.E.I. Research Laboratory, Aldermaston.

Temporary Television Station in Northern Ireland. As already announced, the BBC proposes to install a temporary low-power television station in Northern Ireland before the Coronation. A site for this station has now been acquired on the Glencairn Road, some 2½ miles from the centre of Belfast. The station will be known as Glencairn and will serve the city of Belfast and its immediate surroundings. It will share the same frequencies as Alexandra Palace, vision 45·0Mc/s and sound 41·5Mc/s.

PUBLICATIONS RECEIVED

THE AMATEUR'S GUIDE TO VALVE SELECTION has recently been issued by Mullard Ltd. and should prove of considerable interest to radio amateurs. Not only does this book assist amateurs in the selection of suitable valves and tubes from the Mullard range, but it also indicates under what conditions they should be operated in order to achieve optimum performance. The book, therefore, has special value for radio amateurs who wish to construct their own equipment. Copies can be obtained from radio and television retailers at a price of 1s. 6d. per copy.

TELEPHONES AND TELEGRAPHS—PUBLIC AND PRIVATE SERVICES are two booklets issued by the Council for Codes of Practice for Buildings. They were drawn up by a Committee convened on behalf of the Council by the Institution of Electrical Engineers. The Code dealing with Public Services describes the principal telephone and telegraph systems supplied and maintained by the Post Office to provide intercommunication within a building or group of buildings, and between remote buildings via the public telephone or telegraph services or via private circuits supplied by the Post Office. The Code dealing with Private Services describes telephone and telegraph systems other than those connected to the public system or supplied by the Post Office. Copies of the booklets may be obtained from the British Standards Institution, 24/28 Victoria Street, London, S.W.1, price 7s. 6d. each, post free.

PHILIPS INDUSTRIAL CATALOGUE is the first combined industrial catalogue issued by the industrial group of this company. The catalogue sets out concisely the full range of Philips industrial products under section headings. Its main object is to bring to the notice of all sections of industry the wide range of products manufactured and also to provide a reference book for the trade. A revised catalogue on Arc Welding and an Arc Welding Electrode Wall Chart designed for display in factories and workshops are also available on application. Philips Electrical Ltd., Century House, Shaftesbury Avenue, London, W.C.2.

BICALEX WINDING WIRES and BESTOS AVC CABLES are two publications issued by British Insulated Callenders' Cables Limited. The former booklet describes the more important features of "Bicalex" winding wires including mechanical characteristics, thermal stability, technical data, electrical strength, etc. The latter brochure gives information and details of the "Bestos" asbestos varnished-cambric cables which are superior in construction to ordinary taped or braided asbestos coverings, both electrically and mechanically, and provide a degree of flexibility unobtainable with cables having a hard metallic sheath. Copies of both booklets are obtainable free on application to British Insulated Callender's Cables Limited, Norfolk House, Norfolk Street, London, W.C.2.

TELESYN SYNCHROS, publication No. 552, of the Sperry Gyroscope Co., Ltd., Great West Road, Brentford, Middlesex, refers principally to the A.C. rotary inductor type of Synchro, the Sperry range of which are known as "Telesyns". The booklet describes the three categories of synchros, torque elements, control elements and computing elements. Information on practical applications is also given, with diagrams.

DUBILIER CAPACITORS FOR THE SERVICE ENGINEER is a catalogue containing details of recent developments in Dubilier capacitors. It covers drilitic electrolytic capacitors, including television types and the ear mounting method. Moulded mica and silvered mica capacitors are described, as well as nitrogol, tubular paper, metallized paper and ceramic types. Dubilier Condenser Co. (1925) Ltd., Ducon Works, Victoria Road, North Acton, London, W.3.

EDISWAN MICROFILM READERS describes the equipment manufactured by this firm for individual viewing of films. The Ediswan microfilm readers employ an optical system with a film gate a few inches above table level and the screen just above the gate. The readers are fully described in the booklet, together with operating instructions and maintenance details. The uses and advantages of microfilm are also included. The Edison Swan Electric Company, Ltd., 155 Charing Cross Road, London, W.C.2.

LETTERS TO THE EDITOR

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

Beta-Particle Thickness Gauges

Dear Sir,—I have read with great interest the article on "Beta-Particle Thickness Gauges" by M. G. Hammett and H. W. Finch in your November issue. There are one or two points, however, which I feel may be ambiguous to potential users of such equipment. The most important of these is that while I entirely agree with the advantages of the two chamber method, which are more fully described in British Patent 677 451, it is important to realize that these advantages are independent of the D.C. measuring system with which they are associated. Since we first developed this system at Harwell three years ago, it has been clear that the vibrating reed system is an excellent method of measuring small currents, but the balanced chamber method is equally advantageous when associated with a well designed D.C. amplifier, and it is possible by this alternative means to provide equipment which meets the requirements of the problem in a simpler and cheaper way.

Secondly, it should be made clear that the measurement ranges associated with different isotopes given in Table 1 are the end points determined by careful experimenting and are not the ranges over which they can be used in a beta-ray thickness gauge. These latter are usually about two-thirds of the values given.

Thirdly, the data given in Fig. 1 and Table 3 depends very much on the type of ionization chamber and source and the spacing of components which were used in making the measurements quoted, and will not necessarily apply to beta gauges other than the type described.

A fourth and important point, is that the standard piece of material shown in Fig. 4 can be replaced by an appropriately calibrated diaphragm, without loss of accuracy. This is convenient in many industries, where it would otherwise be necessary to keep a large range of standards on hand.

A fifth and minor error is the statement that the window of the ionization chamber must be electrically conducting to maintain a uniform field in the chamber. The objection to an insulating window arises from the fact that in the absence of radiation electro-static charges can build up on this window and so distort the field within this chamber. In the presence of radiation being measured, however, these charges are rapidly dispersed and we have found it preferable to use a plastic window, which is considerably tougher than the aluminium foil for the same weight per unit area and introduces no extra errors because of its non-conducting nature.

Yours faithfully,

K. Fearnside,

Technical Director,

Isotope Developments Limited.

The Author Replies

Dear Sir,—I welcome Mr. Fearnside's comments on our article on "Beta-Particle Thickness Gauges," especially where he draws attention to a certain lack of clarity in the presentation of information. In this category are the points raised in paragraphs 2 and 3 of his letter and I propose to deal with these first.

Referring to para. 2, I entirely agree with Mr. Fearnside that the practical ranges are approximately 2/3^{rds} of the values given in the fourth column of Table 1. However, while it undoubtedly would have been helpful to include the additional information, the column as headed is correct. Referring to para. 3, the curves of Fig. 1, as drawn, are only approximate and are limited to those portions of the characteristics with which the user of a thickness gauge would normally be concerned. It is true that, under practical conditions of measurement, the curve shapes would be affected by the factors mentioned by Mr. Fearnside but these variations would largely be confined to those portions of the curves omitted from Fig. 1.

As stated in the text, the figures in Table 3 are intended to indicate the *possible* accuracies attainable for various thicknesses of sheet material using types of beta emitter that are known to be suitable for thickness measurements and, what is at least as important, are already available in adequate quantities.

There happens to be an error in the placing of a decimal point in the "20 sec" low limit figure for aluminium (i.e. 0·005 should read 0·0005) but apart from this, we are adhering to our view that (bearing in mind their admittedly approximate nature) the figures given are generally applicable in the sense that the low limit figures indicate the limitation imposed by the random nature of beta particle emissions (thus being quite fundamental) and the high limit figures by the signal-noise ratio with the best of the known types of detector amplifier. In short, we strongly doubt that the ranges given could be significantly extended but are always open to conviction if evidence can be produced to the contrary.

We now come to the more controversial parts of Mr. Fearnside's letter. His statement in the fourth paragraph that "the standard piece of material shown in Fig. 4 could be replaced by an appropriately calibrated diaphragm, *without loss of accuracy*" is one which I cannot accept. There can be no question that, under varying ambient conditions of temperature, humidity, etc., met with in a factory, the best standard of comparison is an identical sample of the material to be matched. No other arrangement of comparable simplicity is capable of the same accuracy under these conditions.

Mr. Fearnside's statement that plastic windows introduce no extra errors in relation to metal windows is quite wrong because he has only considered part of the problem, and once again one suspects that this may be due to a rather too academic approach.

The purpose of an ionization chamber in a thickness gauge is to produce an electric current which is directly and exclusively related to the amount of radiation, originating in the isotope and attenuated by the absorber, which enters the chamber. It is therefore of the utmost importance that the electric field be confined to the interior of the chamber, because otherwise any random movement of objects in the near vicinity of the chamber is liable to produce a significant change of chamber current. Furthermore, and far more important, a material being measured as it is produced is almost always in motion and it may therefore accumulate very large electric charges. These may lead to spectacular changes of chamber current in the absence of very thorough electrostatic screening. Any other electric field of comparable strength would be equally troublesome if unscreened chambers were used, and I would regard it as very unsafe to assume that a two-chamber arrangement would automatically balance out such effects.

I have reserved until last my comments on Mr. Fearnside's first paragraph because of its contentious nature. I have no wish to contest here the validity of any patent, or to detract from the value of any work carried out within the past five years or so by Mr. Fearnside and his former colleagues at Harwell, but I am bound to say that there is nothing fundamentally novel about any of that referred to in his letter, in so far as the profession as a whole is concerned.

I am, however, bound to criticize Mr. Fearnside's approach to the technique of measurement.

In order to shorten as far as possible a letter which is perhaps already too long for the correspondence columns of this journal, may I enumerate a number of items of information with which Mr. Fearnside may not already be acquainted.

1. "Beta" particle thickness gauges were described in some detail by Rutherford, Chadwick and Ellis in "Radiations from Radioactive Materials" as long ago as 1930. There have been several patent applications on this subject since that date.
2. The "two-chamber" method is simply an adaptation of a device long familiar to "measurements" men, namely the use of a comparator bridge in place of an attempted measurement of a fundamental nature, the accuracy of which is in doubt. This method is practically as old as the art of electrical measurement itself and is a very obvious way of handling factory measurements.
3. The statement that "the advantages of the two-chamber method are independent of the D.C. measuring system with which they are associated" is tantamount to saying that a detector which generates spurious signals of varying amplitude and the same frequency as the output of the bridge with which it is associated, does not affect the accuracy of the bridge. Of

course, it will not affect the accuracy of the bridge elements themselves or the actual bridge output, but the only means one has of knowing when a bridge is balanced is the detector, and any bridge in association with such a detector will obviously only give the right answer fortuitously.

Zero drift in D.C. detector-amplifiers when used as a means of measuring the D.C. output of a bridge will obviously appear as direct error in apparent bridge output. No known purely electronic D.C. amplifier is capable of the same long-term zero stability as the vibrating reed electrometer unless fitted with elaborate means for combating zero drift, which would render the whole device at least as complicated as the vibrating reed instrument. Occasionally one may obtain exceptionally good results in electronic amplifiers by careful (and lucky) valve selection, but this procedure will not commend itself to factory users of such equipment. Since Mr. Fearnside refers to "meeting the requirements of the problem in a simpler and cheaper way" I am assuming that this is what he has in mind.

4. The method of modulating unidirectional currents in order to permit amplification without introduction of errors is not new and was, to the best of my recollection, used by colleagues of mine in the E.M.I. Laboratories 15 or more years ago. The vibrating reed instrument presumably referred to in Mr. Fearnside's letter was developed by my present company for A.E.R.E. some two or three years ago and the modulator, which is probably the most novel and important part of the instrument, is due, in certain important details, to my co-author, H. W. Finch, and is the subject of British Patent 681 214.

In conclusion, I would like to say that it was the intention of Mr. Finch and myself in our article to carry out an objective and fair survey of the present state of the art. We must leave it to the reader to judge whether or not we have succeeded. If we have sometimes appeared to be biased, dare I suggest that the facts themselves may be to blame and not our method of presenting them?

Yours faithfully,

M. G. HAMMETT,
E. K. Cole, Ltd.
Electronics Division.

Cold-Cathode Voltage Stabilizer

DEAR SIR,—In the article on "A Variable Voltage Stabilizer Employing a Cold-Cathode Valve" in the November issue, Mr. Goulding describes how a cold-cathode triode may be employed as a D.C. amplifier. This is an important and valuable contribution to the art of cold-cathode tube design, as it allows these tubes to be used in applications requiring continuous control rather than just on-off applications. The development of the parallel stabilizing circuit from this amplifier by analogy with the conventional hard valve and reference tube circuit is a good example of its use. It is, however, possible by considering this stabilizer circuit from another point of view to develop a different mode of operation for the circuit, and to show that both modes of operation are possible, dependent on the frequency at which the cold-cathode tube is triggered. The circuit operates in this second mode at frequencies below a certain frequency, which is dependent on the circuit values; above this frequency it behaves as already described. It is proposed to describe this second mode of operation as it is felt that it is of interest and adds to the understanding of the circuit. It will be referred to as the "output rise controlled" mode of operation.

Briefly, in this mode of operation the frequency, and therefore the mean current through the valve, is determined by the rate of charge of the output capacitor C_1 (see Fig. 1) between two fixed potential e_H and e_L, e_H the upper potential, being such that when C_1 is charged to this value the valve is triggered, and e_L being the voltage to which C_1 is discharged as a consequence of triggering.

Consider the circuit given, assuming for the present that the capacitor C_3 is absent, and that it is merely necessary to raise the trigger potential to the breakdown voltage e_{st} to trigger the valve. The resistors R_2 and R_3 are chosen so that

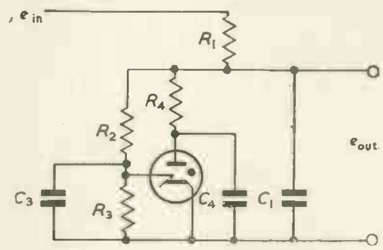


Fig. 1. The cold-cathode stabilizer

when the potential of the capacitor C_1 is e_H, the potential at the trigger is equal to e_{st}. Hence, each time the capacitor C_1 is charged to e_H, the valve is triggered.

The resistor R_4 and the capacitor C_4 as already described by Mr. Goulding are chosen so that the valve is extinguished when the capacitor C_4 has been discharged through the valve to the extinguishing potential e_t. The capacitor C_4 is then recharged to the potential of C_1, with a time-constant R_4C_4. The charge Q required to recharge the capacitor C_4 from e_t to e_L is taken from the output capacitor C_1 reducing its potential from e_H to e_L

where $e_H - e_L = Q/C_1$

The output capacitor is then recharged to the higher potential e_H at a rate determined by the input voltage and the load current, and the cycle is repeated.

Thus the output voltage oscillates between two fixed voltages determined entirely by the valve and the circuit components; changes in the input voltage and the load current merely produce a change in the frequency of operation. The mean output voltage is therefore independent of the input voltage and load current. The stabilization ratio is thus infinite and the output impedance zero.

In practice the small capacitor C_3 connected between trigger and cathode is necessary to ensure that the valve is triggered each time the trigger potential is raised to e_{st}. This introduces a delay in the return of the trigger potential from the potential e_g to the fraction of the instantaneous output voltage $\alpha\ e_{out}$, (where e_g is the potential down to which the trigger is carried during a discharge and $\alpha = R_3/(R_2 + R_3)$).

This capacitor C_3 introduces the possibility of two modes of operation. The operation of the circuit is substantially as just described, provided the trigger potential has been returned to the potential $\alpha\ e_{out}$ before the output capacitor C_1 has been charged to the potential e_H. If, however, the capacitor C_1 is charged to the potential e_H before this delay period has elapsed, then the output voltage will continue to rise to some higher value until the capacitor C_3 has been charged to e_{st}. In this manner the output voltage rises to some higher value dependent on the rate of charge of the trigger capacitor C_3 between the two potentials e_g and e_{st}. This is the mode of operation described by Mr. Goulding, which may be referred to as "trigger transient controlled."

It will be seen that there is a frequency, dependent on the circuit components, at which the trigger is returned to the potential $\alpha\ e_{out}$ at the same time as the output voltage is returned to e_H. At frequencies above this frequency the circuit operates in this "trigger transient controlled" mode, and below, in the "output rise controlled" mode.

It is now possible to consider the performance of the circuit when it is operated in the "output rise controlled" mode. In practice there are two factors which cause the performance to depart from the ideal of infinite stabilization ratio:

1. The charging current of trigger capacitor C_3 due to the rise in output voltage introduces an error voltage.
2. The trigger breakdown voltage is not entirely independent of the frequency of operation.

The trigger capacitor introduces a small error voltage due to the voltage drop produced across the resistors R_2 and R_3 by the charging current of this capacitor. This error voltage is a function of the rate of charge of C_1 (and therefore of e_{in} and i_{out}) but it may be made small by making the ratio of the capacitors C_1/C_3 large. Theoretical values for the stabilization ratio of the order of 2 000 may be readily obtained using components of convenient values. The fact that the trigger breakdown voltage e_{st} changes slightly with frequency limits the stabilization ratio of the circuit to values of the order of 100 to 500. This performance is, however, still very good for so simple a circuit.

The circuit may be designed to operate in this manner over a wide range of input voltage and load current variations.

The ripple voltage ($e_H - e_L$) may be objectionable in certain applications, but by making the ratio of the capacitor C_1/C_4 large it may be reduced to a value limited only by valve effects. In the case of VX8086 this limit gives a peak to peak ripple of approximately 0·1V.

Yours faithfully,

G. O. CROWTHER,
Valve Measurement and Application Laboratory, Mullard, Ltd.

BOOK REVIEWS

Electronic Measurements

By F. E. Terman and J. M. Pettit. 707 pp., 450 figs. Royal 8vo. McGraw-Hill Publishing Co., Ltd. 2nd edition. 1952. Price 72s. 6d.

THIS book is a successor volume to Terman's "Measurements in Radio Engineering." As the new title suggests the scope of the book is greatly increased and might very well be considered as a new book. Terman is well known as the author of two excellent books, "Radio Engineering" and "The Radio Engineers' Handbook." This new book certainly lives up to the high standard of his earlier books.

The book covers an extremely large field and starts with a chapter on the measurement of current and voltage. Although this may sound rather elementary this chapter contains much useful and, often, not well known information not only on normal measuring instruments but also on valve voltmeters of various types. The next chapter deals with the measurement of power, including measurements on microwaves. Measurements on systems with lumped constants and on systems with distributed constants is then considered in two chapters. The authors then cover measurements of frequency, waveform, phase and time intervals and then proceed to chapters dealing with measurements on valves, amplifiers, receivers, antennas and radio waves. Chapters are then included on laboratory oscillators, generators of special waveforms, reactance and resistance standards and, finally, attenuators and signal generators. The book covers such a large field that it is difficult to think of a topic that is not included, the book being well up to date.

In common with other Terman books the style of the book is such that although of high standard it can be easily understood and may be read for considerable periods without fatigue. When a formula is quoted the proof is usually given at the bottom of the page, an idea which has many advantages and makes for easier reading. In a book of this nature covering such a large field it is obvious that a detailed description cannot be given of all the many electronic measurements, but this deficiency is overcome by the inclusion of a large number of references (the author index contains over 1 000 entries). These references are conveniently arranged at the bottom of the page on which they occur, a system that has much to recommend it and saves continual reference to the back of the book. As is to be expected a large number of references are to American journals, but the references are not exclusively American. The reproduction of the book is to the usual high standard of the publishers with almost complete absence of printing and other errors. The diagrams are all clear and well drawn.

To the electronic or radio engineer the book should prove a real goldmine of information on all types of measurements from those at low audio frequencies to measurements on microwaves. It will be found that nearly all the information included in the book is of direct value in actual practical measurements and does not include details of methods which are of little practical importance or obsolete. Much of the information is also most useful to the general electrical engineer who has to deal more and more with electronic devices and measurements using electronic devices. To the student taking electronics, radio engineering or measurements the book will be found to be excellent, since the book, although going to degree standard, explains the subject in such a way that it may be understood by those not having a detailed knowledge of the subject.

G. N. PATCHETT.

Television

By F. Kerkhof and W. Werner. 434 pp., 360 figs. Crown 8vo. Philips Technical Library, Holland. Cleaver Hume Press Ltd., London. 1952. Price 50s.

THIS book is a translation of the work of two Dutch authors from the Television Development Laboratory of the Philips' organization and has also been published in Dutch and German; a French edition is in preparation. Occasionally the phraseology appears to be unusual but on the whole nothing seems to have been lost in the translation and very few errors have been noticed.

Although its title is "Television" this volume is by no means a complete reference book on the subject. The authors state in the preface that their general purpose is to explain television receiver circuits and also in a limited way the important circuits used in studios and transmitters. As far as receivers are concerned this aim has been well achieved but specialized information on television transmission, and studio practice is almost entirely lacking except for explanations of fundamental pulse techniques, which are common to many branches of electronic engineering, and for a short section on camera tubes.

The first chapter reviews in general terms the transmission and reception of television and includes a useful section on U.H.F. propagation.

In Chapter 2 the treatment of the physical principles of electronic scanning forms an excellent preparation for the following short chapter on pickup (camera) and picture tubes.

Under the heading of "The Transmission and Separation of Information", the standard signals of the British, American, French and European systems are explained but no critical comparison is made of their various characteristics. The removal of the D.C. component by finite time constants of A.C. couplings and the restoration of D.C. for receivers are discussed but no mention is made of the special problems of D.C. restoration and clamping for the transmission chain. The worked example of D.C. restoration time constant deals with only one aspect and no really practical result is obtained.

There are some 100 pages in Chapters 5, 6 and 7 devoted to pulse techniques (whose applications are by no means confined to television) and include the principles of pulse generators, multivibrators, pulse mixing, frequency dividers, sawtooth generators, electrostatic and electromagnetic time base generators; the latter are examined in considerable detail including discussion on the use of booster and efficiency diodes. There is a very useful chapter on fly back, high frequency, and pulse e.h.t. generators for cathode-ray tubes.

A long chapter (109 pages) on wideband amplifiers forms the most valuable section of the book. Design requirements are given for various forms of compensation in video frequency amplifiers, comparisons are drawn between transient response and amplitude and delay characteristics for these stages, and there is a survey of staggered and band pass filter circuits for radio and intermediate frequencies. Sound rejection, the sound receiver, the limitation of interference on sound and vision, video detection for positive and negative modulation, and frequency changing all receive attention; there is an interesting section on special valve circuits for wide band amplifiers.

Transmission lines and aerials are dealt with in two short chapters. Transmitting aerials are treated very briefly but there is a useful section on receiver aerial termination.

Chapter 9 on Picture Synthesis is mainly concerned with optical systems for projection television and includes information on photometric units, lenses, aberrations, stigmatism and distortion.

The CBS and RCA colour television systems are very briefly outlined in Chapter 12.

In the last chapter, circuit details and component values are given for two television receivers, one suitable for the British system and the other for the European system of 625 lines with negative modulation.

An Appendix includes a glossary of television terms, tables of the units of the rationalized Giorgi system (which is used throughout the book) with conversion tables to other systems and also conversion tables of units of brightness and illumination.

A literature list containing a large number of references, a short index, information on Philips' Technical Library (to which this book belongs) and 16 photographs showing the effects on the received picture of incorrect receiver adjustment complete the book.

It should be mentioned that there are 20 photographs dispersed throughout the book showing components and apparatus not specifically mentioned in the text.

The main criticism of this book is that very few of the circuit diagrams have been provided with component values. There are a few numerical solutions to problems but in general practical guidance in the form of representative values is lacking and for students this is a serious drawback. This is surprising in view of the fact that the work has been based on a course of lectures for television development technicians.

Apart from these and other shortcomings mentioned earlier, however, the book will be useful for engineers and students interested in pulse techniques and television receivers and is a guide to the requirements of the various television systems.

The mathematical standard required for the full understanding of the general argument is not high; the more difficult proofs have been inserted in smaller print and may be ignored by the reader less well founded in mathematics without detriment to continuity.

H. V. SIMS

Electrical Contractor's Annual 1952-1953

Edited by J. Rosslyn Stuart. 288 pp., 74 figs. Demy 8vo. E. & F. N. Spon Ltd. Price 12s. 6d.

THIS is a collection of new articles on subjects of interest to contractors, with tables, data, references, etc. While most of the articles will appeal to the experienced reader, a section on the calculation of resistances in series and in parallel gives the impression that it is a beginners' book. Articles include:—estimating, installing water heaters, the history of the ring main system, lighting in shops and stores, lamp developments, installation of television aerials, etc. Of the tables, 20 pages are extracted from the I.E.E. Regulations. Since all contractors should have these regulations these pages seem to be redundant. There are useful tables on purchase tax, ratings of blankets, irons, etc., and a list of trade names in 58 pages. Some British Standards are listed, but the very useful Handbook No. 9, a collection of those most useful, is not quoted, neither is Code of Practice C.P.321 (1948), Electrical Installations, General, which could well be in the hands of all contractors. The book unfortunately contains several non-standard terms and symbols and shows a lack of co-ordination.

E. H. W. BANNER

Data and Circuits of Receiver and Amplifier Valves

Compiled and edited by N. S. Markus and J. Otte. 500 pp., 505 figs. Royal 8vo. Philips Technical Library, Holland. Cleaver Hume Press Ltd., London. 1952. Price 40s.

THIS book is the third in a series by Philips Technical Library and contains data and circuits on radio receiver and amplifier valves developed during the period 1945-1950. It also contains descriptions of measuring instruments and auxiliary equipment for use in the laboratory, testing department and factory, as at December 31, 1950.

Dictionnaire Anglais-Francais

(des Termes Relatifs a L'Electrotechnique, L'Electronique et aux applications Connexes)

By H. Piraux. 296 pp. Demy 8vo. Editions Eyrolles, Paris. Price 1850 fr.

THIS dictionary contains most of the electric and electronic technical terms for acoustics, atomics, wave guides, radio, television, etc.

Commercial A.C. Measurements

By G. W. Stubbings. 377 pp., 183 figs. Demy 8vo. Chapman & Hall Ltd. 3rd edition. 1952. Price 55s.

THE electronics engineer who considers that the book has no application to his work is mistaken. Far from all measurements in electronics are at high frequencies or at very low power, and basic A.C. measurements at power frequency are necessary either directly or as a starting point for more specialized books on specific measurements. The name of G. W. Stubbings is well known as a writer on the subject of electrical measurements, and this third edition is of the usual standard. Both theory and practice are covered. The book is equally suited to the advanced student studying for a degree or Institution examination, and to the engineer as a daily guide in the laboratory, test-room, etc. The mathematical standard is not too high for the purpose and the text is 'readable.' The book opens with basic A.C. theory, including those factors of importance in measurement, such as distorted waveforms, which are sometimes optimistically termed 'rich in harmonics.' This is followed by a full discussion on balanced and unbalanced three-phase circuits. The measurement of current and voltage is described with reference to the instruments available, but instrument design itself is excluded. Wattmeters and watt-hour meters are fully covered, with the theory and practice of the many variations in the measurement of polyphase energy. Other versions of the subject not always dealt with in other books include summation metering, reactive metering and the measurement of maximum demand. The various power factor meters and frequency meters which are commercially available for power frequencies are well treated, and there is a chapter on the involved subject of kVA measurement. Instrument transformers are discussed both from first principles and with respect to commercial practice as covered by the British Standard. The final chapter on miscellaneous apparatus is a wide one, of use to all electrical laboratory workers, and including descriptions of A.C. galvanometers, oscillographs, phase-shifters, and power supplies, etc. Within the scope of the subject, the book can be recommended without qualification.

E. H. W. BANNER

Vector Analysis

By Earl C. Rex. 88 pp., 45 figs. Royal 8vo. Wm. C. Brown Company, Iowa. 1952. Price $3·25.

THIS book is intended as the basis for a course on vector analysis and is founded on the author's own experience in teaching to mathematics and physics students. A sound knowledge of the calculus is assumed while some knowledge of differential equations and solid analytic geometry will prove of the utmost use.

It starts with a definition of vector and scalar quantities and works, in a logical and practical manner, right through four vector additions to dyadics, vector differential calculus and vector integrals. A large number of worked examples is included in each section though from the electrical engineer's point it is, perhaps, a little unfortunate that these are mainly based on mechanics.

ELECTRONIC EQUIPMENT

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

Portable Load Measuring Set

(*Illustrated below*)

THE L4000 portable load measuring set has been designed for the purpose of measuring the forces acting on control rods, operating levers and other linkages of road transport vehicles, engineering structures, etc. Measurements can be made simply and rapidly during field trials, and in this way it is possible to obtain complete information on the forces acting upon the component under investigation under all required operational conditions.

To incorporate the device in a particular test rig, it is only necessary to prepare a special coupling rod, which will be available for all further tests on rigs of the same type. The rod is made in two sections, each with a threaded end

which screws into the load measuring ring. This is a specially protected electrical strain gauge device, designed to be weatherproof and insensitive to temperature changes and to bending forces; it is connected to the measuring set by a length of screened or armoured cable which is terminated in strong and durable metal-clad plugs and sockets. A check calibration may be carried out on the ring at any time, by means of a standard load testing machine.

Strain gauge ring units can be supplied to suit the range of the loads required to be measured, and the recording instrument requires no modification when a ring having a different load range is fitted.

In operation, the loads are read directly from a pointer and scale having 100 divisions, with an accuracy of 1 per cent of the full scale reading.

The meter unit is resiliently mounted inside the instrument case, so that damage due to shock can be avoided and a steady meter deflexion obtained, and the outer case is supplied with shock absorbing mountings.

The load measuring set can also be used with any bonded resistance strain gauge device for measurement of pressure displacement, strain, etc., provided that a suitable calibration test is carried out beforehand. For this reason, each bridge circuit is provided with a pre-set sensitivity control, which may be locked into position after adjustment and is employed to set the instrument scale to a convenient value by calibrating the transducer (i.e., load, pressure, or displacement gauge) against a standard laboratory gauge. If supplied complete with a set of type L4003 load gauges, the sensitivity controls are all pre-adjusted before delivery to provide full scale deflexion of the indicator at maximum load.

**Boulton Paul Aircraft, Ltd.,
Wolverhampton.**

Mechanical Time-Delay Relay

(*Illustrated below*)

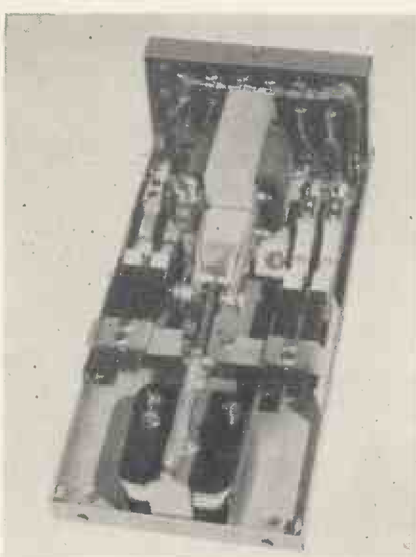
THIS time-delay relay is a robust, compact equipment comprising a mechanical escapement and a solenoid.

The escapement, which is actuated by the solenoid, is directly coupled to four insulated fingers which, in turn, control the action of four sets of change-over contacts. Each finger can be independently adjusted to operate its respective set of contacts at any period within the instrument's time-cycle.

The solenoid consists of a rigidly constructed circuit of low-loss silicon-iron laminations, a fully impregnated coil and a shaded-pole plunger. The shading-ring is accurately positioned and ensures a high sealing pull and noiseless operation on A.C. For D.C. operation auxiliary contacts are provided to facilitate the inclusion of an economy-resistance into the circuit.

The mechanical time-delay relay is supplied complete with metal cover; and all electrical connexions are made to a tag-panel located externally at one end of the instrument.

All contacts are of fine-silver which ensures long service-life and low contact-



resistance. In operation, the contacts undergo a wiping action which promotes high pressure and continuity even at low voltage on small loads.

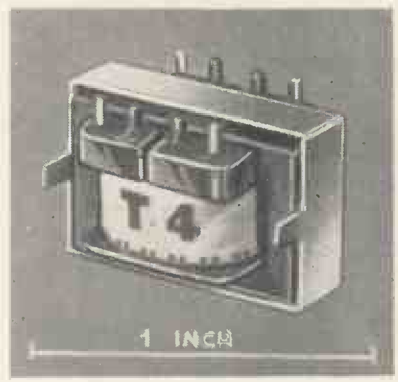
The contacts are rated at 5A at 250V A.C., while the time delay range is 1sec to 4min (in 2 ranges) with an accuracy of 10 per cent or better.

**Electro Methods, Ltd.,
220 The Vale,
London, N.W.11.**

Sub-miniature Output Transformer

(*Illustrated below*)

THIS transformer has been designed for use as an output transformer in hearing aids and similar miniature equipment. The windings of the transformer are brought out to terminals moulded into the thermo-setting plastic material of the bobbin. A laminated core of high permeability magnetic alloy is used.



Fixing is normally effected by means of a small cover which holds the transformer to the mounting panel. Two lugs on the cover pass through the mounting panel and are bent over on the other side. Rubber packing pieces fit between transformer laminations and mounting panel or clamp.

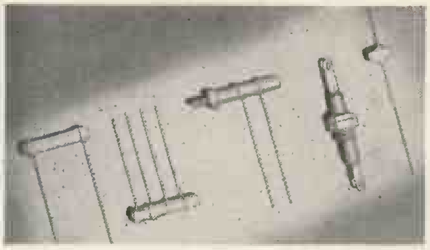
Insulation resistance between primary and secondary at 500V D.C. is better than 100MΩ, while the ratio of primary turns to secondary turns is normally 31·6 : 1 although other ratios are available on request. The primary inductance measured at 5V R.M.S., 1 000c/s varies from 21H with zero D.C. to 6H at 1mA D.C.

**Fortiphone, Ltd.,
247 Regent Street,
London, W.1.**

Hi-K Ceramic Capacitors

(*Illustrated top right*)

THE latest T.C.C. receiver type ceramic capacitors have their pure silver electrodes fired on to a special ceramic of the barium-titanate group which permits high capacitance to be combined with small physical size. This particular material has a very high dielectric constant (3 000) and high insulation resist-



ance (more than 5 000MΩ) at all working temperatures, and these figures are maintained after extended life tests.

An important parameter of these capacitors is the variation of capacitance with temperature changes, and it is an inherent feature of Hi-K ceramics that this variation can be quite substantial for comparatively small changes in temperature. In view of this characteristic, these capacitors are principally used for decoupling purposes, where reasonable variations will not prejudice the stability of the apparatus.

The low inductance of the tubular and miniature bead types is of especial advantage when they are used as by-pass capacitors in radar, television and other H.F. equipment.

The lead-through type serves as an efficient R.F. filter to leads passing through a compartment screen, in which position it acts as an H.F. T-section low-pass filter. It can also be used as a filter for H.T. and L.T. feeds in F.M. receivers and in radar I.F. amplifiers.

A new insulating finish (Type "F") is now available for the wire ended tubular and miniature bead types, consisting of a black synthetic coating, applied by a dipping process. This finish provides insulation for these capacitors against other components and/or live chasis, and gives excellent protection at all temperatures up to 100°C. (max).

The working voltage of all types is 500V D.C. or 250V R.M.S. A.C. The wire ended tubular types are available as single, double and triple elements in a range from 500pF to 0·01µF, while the maximum capacitance of the screw base tubulars is 0·002µF. The lead-through and miniature bead types are available as single elements only the former in a range from 1 000pF to 0·01µF and the latter from 68pF to 470pF.

The Telegraph Condenser Co., Ltd.,
N. Acton, London, W.3.

Stabilized Power Supply

(Illustrated below)

THIS stabilized power supply unit has been designed as a source of D.C. supply for general laboratory use. The type

100 is housed in a steel case finished in grey enamel, while the type 100B is designed for mounting in a standard 19in. rack.

The reference voltage is provided by a high stability regulator tube connected in a bridge circuit, the out-of-balance voltage of which is fed, via a high gain pentode, to the series control valve.

The D.C. output is continuously variable from 200 to 350V at up to 120mA; either pole can be connected to earth. An A.C. output of 6·3V, 3A is also provided (this is tapped at 4V).

The stability is such that a variation in mains input of ±10 per cent produces an output voltage variation of less than ±0·5 per cent and a change from no load to full load produces a change of the same order. The source impedance is 5Ω.

Harvey Electronics, Ltd.,
273 Farnborough Road,
Farnborough, Hants.

A Portable V.H.F. Field Test Set

(Illustrated below)

THE Radio-Aid portable test set has been designed to provide a rapid overall check of the performance of most

types of V.H.F. multi-channel communications equipment, as fitted in civil aircraft.

The equipment which is operated from an external 24-28 volt D.C. supply, incorporates a low powered oscillator which, in the "Transmit" position, is continuously swept in frequency over the 115 to 135Mc/s band at an approximate rate of 200 times per second by a motor-driven variable capacitor. The radiated signal thus covers the civil aeronautics V.H.F. band of 118 to 132Mc/s, and will be audible on all normally operative channels of the aircraft receiver under test. Care has been taken to reduce the output to a level where no interference occurs to neighbouring services, and which will not operate a defective receiver channel when the test set is placed at the appropriate distance from the aircraft concerned.

In the "Receive" position, the oscillator contained within the test set is switched off, and a broadband receiver substituted, working from the same quarter-wave aerial rod. For simplicity and remote observation by the user, the tests for the aircraft V.H.F. transmitter output are indicated by two $2\frac{1}{4}$in. diameter coloured lights, mounted at an angle on the front panel of the test set, which can be seen easily from the cockpit when set up near the aircraft. The left-hand green light operates when the aircraft transmitter carrier exceeds a pre-determined pre-set level, and the right-hand red light, when modulation amplitude is above a pre-determined and pre-set miniumum.

When placed at a distance from any particular aircraft which has previously been determined by trial to suit the type of communications equipment and aerials fitted, the test set thus provides a "go or no-go" gauge, showing positively if each channel is working normally, or is defective.

Below the indicator lamps are situated two $2\frac{1}{4}$in. moving coil meters, the left-hand indicating carrier level and the right-hand modulation level. The receiver and transmitter performances of the test set are substantially linear over the band, and therefore the receiver section can be used in conjunction with the meters for relative channel checks. Provision is also made for monitoring the aircraft's radiated signal through headphones plugged into the test set, when distortion of telephony is suspected.

Radio-Aid, Ltd.,
29 Market Street,
Watford, Herts.

Radar Display Tubes

A RANGE of flat-faced radar display tubes has recently been introduced by the Communications and Industrial Valve Department of Mullard Ltd. These tubes give pictures of very high definition and contrast. They are also notable for their low deflexion defocusing, low astigmatism and long persistence.

All the tubes employ magnetic focusing and the screens are metal-backed magnesium fluoride giving an easily visible orange afterglow. Their heater voltage is 6.3 volts and current consumption 0·3A.

The range includes a 5-inch tube MF13-1, a 12-inch tube MF31-55, and a 16-inch tube MF41-15.

The 5-inch MF13-1 tube is designed for small marine and airborne radar displays where high performance coupled with saving in space is required. It is fitted with an octal base and can be used in place of the 5FP7A under most conditions.

The 12-inch tube MW31-55 is modelled on the armed services Preferred Type CV429. It has a B12A base, and a 6·3V, 0·3A heater.

The MF41-15 is a 16-inch metal-coned, wide-angle tube which was, in the first instance, specially developed for use in harbour radar systems. It will, however, find wide application in mobile equipments wherever larger displays are required.

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

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: March 10. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1.
Discussion: The Classification of Electronic Circuits and the Standardization of Symbols.
Opened by: L. H. Bainbridge-Bell, M.C., M A.

Scottish Section

Date: March 5. Time: 7 p.m.
Held at: Lecture Theatre, Glasgow University.
Lecture: Electronics in Nuclear Research.
By: J. M. Reid.

West Midland Section

Date: March 24. Time: 7.15 p.m.
Held at: Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: The Principles of Electronic Computing Machines.
By: B. V. Bowden, Ph.D.

BRITISH SOUND RECORDING ASSOCIATION

Date: March 20. Time: 7 p.m.
Held at: The Royal Society of Arts, John Adam Street, London, W.C.2.
Lecture: The Synchronization of Magnetic Tape and Film for the Amateur and Professional.
By: N. Leevers, B.Sc., F.B.K.S., and E. W. Berth-Jones, B.Sc.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Date: March 5.
Lecture: Design Features of Certain British Power Stations.
By: S. D. Whetman, B.Sc., and A. E. Powell, B.Sc.(Eng.).
(Joint meeting with the Institution of Mechanical Engineers.)

Measurements and Radio Sections

Date: March 3.
Lecture: The Study of a Magnetic Invertor for Amplification of Low Input Power D.C. Signals
By: E. H. Frost-Smith, B A., Ph.D.
And The Parallel-T D.C. Amplifier: A Low Drift Amplifier with Wide Frequency Response.
By: P. S. T. Buckerfield.
(Proceedings I.E.E., Part II, October, 1952.)

Radio Section

Date: March 11.
Lecture: Low-Level Modulation Vision Transmitters, with special reference to the Kirk o'Shotts and Wenvoe Stations.
By: E. McP. Leyton, E. A. Nind. B.Sc.(Eng.), and W. S. Percival, B.Sc.(Eng.).

Supply and Measurements Sections (Joint Meeting)

Date: March 25.
Lecture: Transformer-Analogue Network Analysers.
By: M. W. Humphrey Davies, M.Sc., and G. R. Slemon, Ph.D., M.A.Sc.

Cambridge Radio Group

Date: March 17. Time: 8.15 p.m.
Held at: The Cavendish Laboratory, Cambridge.
Paper on Electronic Equipment in Aircraft.

Mersey and North Wales Centre

Date: March 16. Time: 6.30 p.m.
Held at: The Town Hall, Chester.
Lecture: Radio Telemetering.
By: E. D. Whitehead, M.B.E., and J. Walsh, B.Sc.

North-Eastern Centre

Date: March 23. Time: 6.15 p.m.
Held at: Neville Hall, Westgate Road, Newcastle-on-Tyne.
Lecture: Post-Graduate Activities in Electrical Engineering.
By: W. J. Gibbs, M.Sc.(Eng.), D. Edmundson, B.Sc., R. G. A. Dimmick, B.Sc., and G. S. C. Lucas, O.B.E.

North-Eastern Radio and Measurements Group

Date: March 2. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Lecture: An Improved Scanning Electron Microscope for Opaque Specimens.
By: D. McMullan, M.A.
Date: March 16. (Time and place as above.)
Annual General Meeting.

Sheffield Sub-Centre

Date: March 4. Time: 6.30 p.m.
Held at: City Hall, Sheffield.
Faraday Lecture: Light from the Dark Ages, or the Evolution of Electricity Supply.
By: A. R. Cooper.

North-Western Radio Group

Date: March 18. Time: 6.30 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: A Method of Designing Transistor Trigger Circuits.
By: Prof. F. C. Williams, O.B.E., D Sc., D.Phil., F.R.S., and G. B. B. Chaplin, M.Sc.

North-Western Centre Supply Group

Date: March 24. Time: 6.15 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: The Electrolytic Analogue in the Design of High-Voltage Power Transformers.
By: D. McDonald, B.Sc.

South Midland Radio Group

Date: March 23. Time: 5.30 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Electronic Motor Control.
By: S. H. Dale.

Southern Centre

Date: March 4. Time: 6.30 p.m.
Held at: The University, Southampton.
Lecture: Electronic Telephone Exchanges.
By: T. H. Flowers, M.B.E., B.Sc.

Farnborough District

Date: March 18. Time: 7.30 p.m.
Held at: The Royal Aircraft Establishment Technical College, Farnborough.
Lecture: Electronic Telephone Exchanges.
By: T. H. Flowers, M.B.E., B.Sc.

INSTITUTION OF ELECTRONICS

North-Western Branch

Date: March 27. Time: 7 p.m.
Lecture: Modern Developments in the Technique of High Vacuum Measurements.
Held at: The College of Technology, Manchester.
By: J. Blears, B.Sc.

INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: March 31. Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2.
Lecture: Alternatives to the Conventional P.C.L.C. Telecommunications Cable.
By: J. Gerrard, A.M.I.E.E.

THE PHYSICAL SOCIETY

Spring Provincial Meeting

Dates: March 30, 31 and April 1.
Held at: The Department of Physics, The University, Leeds.
Meeting on: Aspects of Solid State Physics, divided into three sessions.
Visitors should apply to the offices of the Physical Society, 1 Lowther Gardens, Prince Consort Road, London, S.W.7. Closing date for applications March 9.

RADIO SOCIETY OF GREAT BRITAIN

Date: March 20. Time: 6.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London. W.C.2.
Lecture: V.H.F. Aerial Developments.
By: F. Charman, B.E.M.

SOCIETY OF INSTRUMENT TECHNOLOGY

Date: March 31. Time: 7 p.m.
Held at: Manson House, Portland Place, London, W.1.
Lecture: The Presentation of Control Theory and the Training of Control Engineers.
By: G. L. d'Ombrain, Ph.D., B.Sc.(Eng.), D.I.C., A.C.G.I., M.Am.I.E.E., M.I.E.E.

THE TELEVISION SOCIETY

Date: March 12. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: Television Aerial Equipment.
By: N. M. Best.

RECENT BRITISH STANDARDS

British Standard for Dimensions of Circular Cone Diaphragm Loudspeakers. (B.S.1927:1953).

This standard for dimensions of circular cone diaphragm loudspeakers specifies the mounting dimensions which affect the interchangeability of such loudspeaker units.

During the preparation of the standard it was considered neither practicable nor advisable to attempt to include all types of loudspeakers, and the standard is therefore limited to that type which is at present produced in the greatest quantity and for which the feature of interchangeability has greatest importance.

Although the standard does not deal with performance characteristics, certain stipulations are made regarding nominal resonance frequency and nominal impedance, in view of their importance in interchangeability. Copies of this standard cost 2s.

British Standards for Reels for Covered, Solid, Round Winding Wire for Electrical Purposes. (B.S.1489:1952).

Reels and Wooden Drums for Bare Wire, Stranded Conductors and Trolley Wire for Use in the United Kingdom. (Ref:PD.1472 Amendment No.1 to B.S.1559).

Precision Reels for Bare and Oxidized Resistance Wires (0·0005-0·0048in. diameter inclusive). (B.S.1888:1952).

Four years ago when the British Standard for reels for covered winding wires was published, it was thought that it would be a long time before it could be fully adopted in view of the alterations in machinery, etc., that would be required. In practice, however, the standard has proved so successful that it is now possible still further to reduce the range of sizes of reels and a revision of B.S.1489, "Reels for covered, solid, round winding wires for electrical purposes", has now been issued. In the new standard a system of reference numbers has been introduced, marking requirements have been added and other modifications made.

These modifications to B.S.1489 have made it necessary to make some adjustment to B.S.1559, "Reels and wooden drums for bare wire, stranded conductors and trolley wire for use in the United Kingdom", and an amendment slip has been issued.

To preserve the dimensions of fine resistance wires during de-reeling the reels must be of low inertia and precise dimensions.

To meet this need, the British Standards Institution has also published B.S.1888, "Precision reels for bare and oxidized resistance wires (0·0005-0·0048in. diameter inclusive)". The standards cost 2s. 6d. each, and the amendment slip is available free.