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IN APPRECIATION OF LOCAL SECTIONS

The 1949/50 programme booklets of the Sections have recently been issued and it is therefore appropriate to pay tribute to the Section Committees and particularly the local Honorary Secretaries for the work they have accomplished. In each case an interesting programme of meetings has been arranged, largely by local endeavour and at all times in co-operation with the Papers Committee. Sufficient tribute can never be paid to the local sections of the Institution and to the enormous amount of voluntary labour which is put into this work. They represent the endeavour of all active members in fulfilling the objects of the Institution.

Comparison does not always give useful or reliable information, but it is interesting to note that 15 years ago there were four established branches or Sections of the Institution ; following reorganisation of Section activities some 10 years ago there are now seven Sections and at least three more are contemplated. In forming additional Sections it is necessary to stipulate two desiderata : firstly, *increase in membership and particularly the local membership and secondly, the reading of a larger proportion of papers on subjects of everyday importance to the radio engineer before Sections by members.*

Further Sections of the Institution are most likely to be formed in Nottingham and Cardiff during 1950, in readiness for the 1950/51 session. Other centres like Belfast will no doubt follow if there is a sufficient nucleus of members in each of those areas. Within this network of branches, it is hoped to form a large number of informal groups of members in other areas, each group being sponsored by the nearest branch.

The formation of Overseas Sections is also contemplated and local secretaries have already been appointed in South Africa and New Zealand. The Council of the Institution is also considering the formation of local sections in India, where there is a growing concentration of members.

In many districts plans have been made for the formation of Graduate and Student Sections. Scotland already has a Student and Graduate Section and this has been entirely due to the initiative of the main Committee. Whilst the General Council has to approve any such arrangements, it is hoped that other Sections will follow this example and submit appropriate plans to the Council for ratification.

In this connection the publication of a Student's Section of the Journal has been discussed. This would include papers submitted by students which are not of a sufficient standard to warrant inclusion in the Journal. Papers written specially for students would be published and, of course, the proceedings of Student Section meetings. It is proposed that, in the first instance, it would be published quarterly.

All this work for Sections and/or Graduate and Student Groups is finally dependent on the support given by every member of the Institution. Indeed, the programme of all meetings relies almost entirely on the effort contributed by individual members—a fact which can be too easily overlooked.

In the July Journal, the retiring Chairman of the Papers Committee pointed out the need for yet more members to contribute to the common pool of radio engineering knowledge from their first-hand experience. No apology is necessary for reiterating this plea, since the *raison d'être* of the Institution is to facilitate the exchange of information and ideas. The preparation of an informative and stimulating paper is an undertaking that can be a worth-while task. It produces the satisfaction of a job well done, at the same time enabling the qualification and crystallization of one's thoughts, and is a useful contribution to a pool of knowledge. Will you share your experience and ideas with other members ?

G.D.C.

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A TECHNIQUE FOR THE DESIGN OF PULSE TIME MULTICHANNEL RADIO SYSTEMS*

by

M. M. Levy† (*Member*)

SUMMARY

The description of a technique for the design of pulse time multichannel radio systems, the application of the design to an automatic 24-channel system, the description of a first 18-channel experimental system and its performance in radio links are given.

The fundamental parts in a pulse multichannel system are: the distributor (transmitter) and the synchronized distributor (receiver) producing the correct time distribution between channels, the modulator (transmitter), the demodulator (receiver) and the mixing and radio frequency transmitting and receiving circuits.

Simple conceptions are introduced in the design of each of these parts.

The distributor uses a delay line, in a special feedback circuit, and a square wave. Each channel is selected by a correct tapping on the delay line and the channel boundaries are accurately timed by the square wave. The whole scheme is entirely automatic and within large limits it is unaffected by the pulse shape distortions appearing at the tappings along the line. In the distributor, and generally in the system, a special three-valve multivibrator is used as pulse generator. It has the property of producing pulses of great peak power in low impedances.

The modulation is obtained by means of trapezoidal pulses triggering the short pulse multivibrator generators. When a signal is added to the trapezoidal pulses, the instant of triggering is effectively time modulated. The advantage of using trapezoidal pulses is that the pulse time modulation never crosses the boundaries of the channel.

A special type of short pulse multivibrator has been designed. It uses an inductance or, more exactly, a shock-excited tuned circuit, to define the length of the pulse. A three-valve multivibrator of this type has the advantage of producing pulses of great peak power. This type is called in the paper a "power modulator."

The demodulation process consists of transforming the time modulation into variable length pulse modulation and then to filter the signal with a low-pass filter.

To change phase modulation into length modulation a special type of multivibrator is used. The circuit is so adjusted that the multivibrator becomes sensitive only during the time allocated to the corresponding channel and it triggers only when the corresponding channel pulse appears. This continues until the time allocated to the channel comes to an end. Here also "power multivibrators" are used to give a considerable gain in power.

An important feature of the modulators and demodulators is the use of one circuit for many channels. This appreciably reduces the number of valves required. Such circuits are called in the paper "multiplex modulators" and "multiplex demodulators." It is shown that if the power multivibrator is used, the cross-talk in these circuits is negligible.

The mixer circuit comprises cathode followers with cathodes connected together. When multiplex modulators or demodulators are used, about three valves are required per two channels including mixing valves.

This technique is particularly valuable for systems with a large number of channels. The optimum seems to be between 20 and 40 channels. The maximum number appears to be one hundred.

Block and simplified schematic diagrams are given for a 24-channel system, as an example.

Finally, a very early experimental 20-channel system, built in 1943, is described and its performance is discussed. This system was satisfactorily tested up to distances of about 36 miles in mobile vans.

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† G.E.C. Research Labs., Wembley. Late of Standard Telephones and Cables Ltd.

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1. Introduction

This paper was written in its first version in 1944. Although its publication has been deferred up to now, the material is still of interest because there have been few publications dealing with the intimate technique of pulse modulation.

It is surprising how simple the design of pulse systems becomes once the proper technique is applied and how easily the usual multiplex difficulty, that is inter-channel interference, can be reduced to a negligible amount.

Pulse technique employs simple systems particularly when it is compared with other methods of modulation such as shifted carrier methods.

The advent of Pulse Code Modulation (P.C.M.) seems to have reduced the importance of Pulse Time Modulation (P.T.M.) for multiplex radio links. However, analysis shows that P.T.M. can give approximately the same efficiency provided that advantage is taken of the improvement of signal-to-noise ratio introduced by the use of compression at the transmitter and expansion at the receiver. Then both systems compare favourably particularly for radio links of not more than 300 to 500 miles. P.T.M. has also the advantage of simple repeater design mainly because of the simplicity of pulse reshaping, either by the use of a multivibrator controlled by the incoming pulses or by other pulse generating devices.

The paper describes a multiplex system completed in 1943 with transmitter and receiver connected by a short radio link. It seems to be the first successful system of its type. Later the system was tested with good results over a distance of 36 miles.

It should be mentioned in this connection that some of the details making up the multichannel pulse system to be described in the following pages have been discussed at length in a previous paper by the author, namely in "Some Notes on Pulse Technique," *Journal Brit.I.R.E.*, Vol. VII, No. 3, p. 99. It was felt, however, that, for the sake of completeness of the present paper, it would be desirable to include those items again, at least in a cursory form.

2. Fundamental Principles

Pulse technique is very different from the usual A.C. technique because it is based on intermittent work. Instead of continuous waves there are very sharp, narrow pulses appearing and disappearing suddenly. Between pulses the whole system should be completely at rest. Furthermore, because the valves are used intermittently, the same valves can be used for many different functions provided they do not overlap in time. In multichannel work, for instance, one

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valve can be used as the pulse modulator for many independent channels. Because of these conditions this technique has the following requirements: the pulses must travel through delay lines and circuits with no appreciable distortion; the valves must be strongly overloaded to give full efficiency because, when pulsed, they work intermittently; the circuits must be designed in order to take account of this overload with its consequent high grid current; the circuits must produce very short pulses with no after-effects when they disappear; the modulation and demodulation methods are of a new type since information on the signal must first be translated into time shift and then re-translated into signal amplitude, without introducing harmonic distortion; and finally, the circuits must be designed so that valves are used where possible, by many channels in succession without introducing interference from channel to channel, that is, without producing any cross-talk.

These requirements need special designs and new conceptions. Typical examples will be given, showing how, by combining these principles, one can design a simple and efficient multichannel system for any number of channels between 10 and 50.

2.1. *Composition of a Typical Multichannel Pulse System*

Figure 1 represents a block schematic diagram of the transmitter and receiver of a typical multichannel pulse system.

At the transmitter a distributor supplies the same number of phase pulse outputs as there are channels. One phase modulates the radio transmitter directly giving the synchronizing pulse. The other phases go to respective *trapezoidal generators*, each of which produces the appropriate channel trapezoidal modulator pulse train; each trapezoidal train is combined with the corresponding channel signal to control the triggering time of a *modulator multivibrator* producing a phase modulated train of pulses. All these trains are first combined in a *mixer circuit* and then used to modulate the radio transmitter. The modulated multivibrators can be either independent of one another as shown in Fig. 1, or combined together so that one multivibrator can be used for many channels thus introducing a saving in the circuit design and valves. This type will be called *multiplex modulator*. The R.F. transmitter consists of a very stable R.F. oscillator followed by a modulated R.F. amplifier feeding the aerial.

In the receiver, the synchronizing pulses are separated from the channel pulses by amplitude or length discrimination and used to synchronize a *pulse distributor generator* which delivers the same number of phase pulse outputs as the transmitter, both being in phase.

These pulses are used to render *gating valves* successively sensitive. The channel pulses from the radio receiver are used to trigger *multivibrator demodulators* each rendered sensitive during the corresponding channel allocated time by a gating valve. These multivibrators convert the phase modulated pulses into length modulated pulses. From these pulses the intelligence is filtered out by individual low-pass filters. As in the transmitter, many multivibrator demodulators may be combined together in one, thus forming a *multiplex demodulator*.

The main features of this equipment are :—

At the Transmitter

1. The use of a delay line distributor producing an accurate time division and requiring no adjustments;[1, 2, 3, 4, 5]
2. The use in the distributor and in most parts of the circuit of *power multivibrators* producing pulses with sharp edges and great peak power;[6]
3. The use of trapezoidal pulses for modulation, giving definite boundaries for each channel and avoiding overlapping between adjacent channels;[7]
4. The use of multiplex multivibrators combining the modulation of many channels in one circuit; and the use of *power multiplex multivibrators* producing channel phase modulated pulses of sufficient power to modulate directly the R.F. transmitter.[9]

At the Receiver

1. The use of a circuit suppressing noise interference and giving an accurate and stable time division synchronized with the transmitter;[12]
2. The use of multivibrator demodulators, combining the features of *power multivibrators* and *multiplex multivibrators*, giving great power outputs, and demodulating many channels all in the same circuit;[8]
3. The elimination of harmonic distortion by

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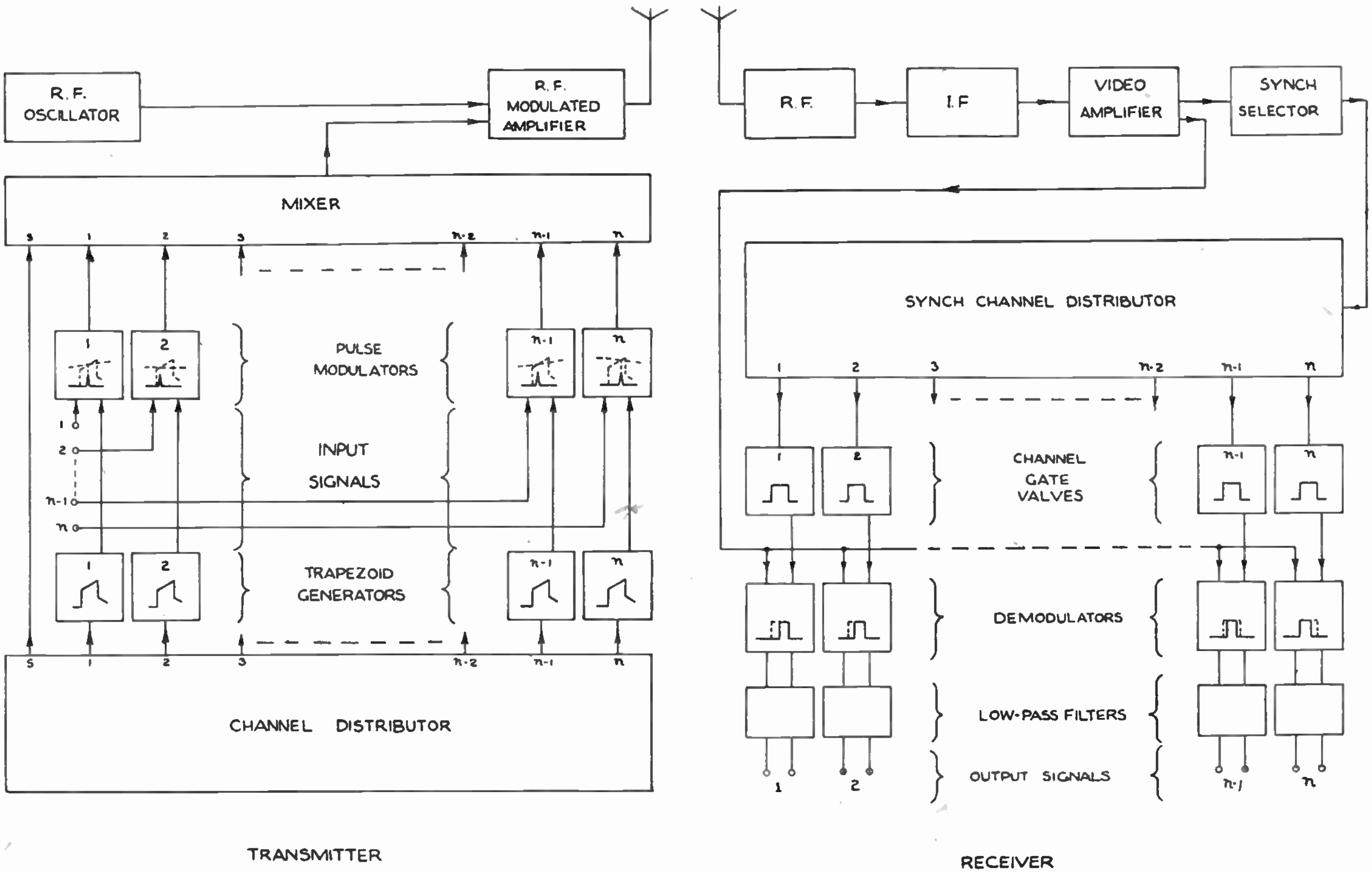

Fig. 1.—Block diagram of a multiplex pulse system. The main components are: (a) for the transmitter, the time distributor, the trapezoidal pulse generators, the pulse modulators and the channel mixer; the synchronizing pulses are generated by the channel distributor. (b) For the receiver, the synchronizing pulse selector, the synchronized time distributor, and the individual channel demodulators.

convenient design of modulator and demodulator circuits.[7]

These features will be described in the next section and will be followed by the description of a typical 24-channel system, and an early experimental 18-channel system.

2.2. *Principles Used in the Transmitter*

2.2.1. *Distributor*

The main features of the distributor (Fig. 2) are :—

1. The use of a delay line for time division ;
2. The use of a special *power multivibrator* producing pulses of great amplitude, to feed the delay line ;[6]
3. The use of feedback between the delay line and the pulse generator, producing a very steady distributor ;[1]
4. The use of a special technique in order to reduce the number of sections of the line to a minimum ;[5]
5. The combination of the pulses at each tapping of the line with a fundamental square wave in order to obtain pulses whose boundaries are accurately determined and practically independent of distortions produced in the line.[5]

The idea of using a delay line as a distributor is not new. Sandeman had suggested its use more than 20 years ago for time division telegraph systems. It was also considered by many research workers but was never applied because pulses travelling through the line are progressively distorted, and this distortion is considerable unless a great number of sections is used.

Assuming that it is desired to build a delay line distributor for a 24-channel system, then if the recurrence frequency is $f_r = 8$ kc/s, or the recurrence period $T_2 = 125$ microseconds and the time allocated to each channel is 125/24 = 5·2 microseconds, the total delay of the line must be 125 microseconds and the recurrent pulse applied to the line must have a width of about 5 microseconds. If, at the end of the line, leading and trailing edges with a slope of 0·5 microseconds, that is 10 per cent. of the channel width, are accepted, an ideal line should have a cut-off period of about 1 microsecond. In practical lines, because of the great number of sections, the attenuation starts much earlier than the cut-off frequency, and the phase characteristic is not linear beyond 0·3 to 0·6 of the cut-off frequency.

For these reasons, the cut-off frequency should be at least three times the value for ideal lines. This gives a cut-off frequency of 3 Mc/s, and a delay of 0·4/3 = 0·13 microseconds per section.[1] The total number of sections is thus at least 125/0·12, that is approximately 1,000.

It does not seem to have been realized that a distortionless line is not necessary. The writer has shown elsewhere that when pulses travel through a delay line, the length at mid-height remains practically the same even if the pulse has been considerably distorted. Figure 3 shows the distortion produced by an ideal low-pass filter to a square pulse when the cut-off frequency varies. It can be seen clearly that, so long as the cut-off frequency is not smaller than the inverse of the pulse length, the length at mid-height remains practically the same. If the pulse applied to the line has a very large amplitude, it is easy to cut a small slice off the pulses at each tapping with a high slope valve and obtain on the anode or cathode a square pulse of required width.

Further improvement can be obtained by bridging alternate sections either with capacitors or a suitable mutual inductance. Bridging by mutual inductance or by capacitors can improve the phase linearity in delay lines. Some experiments with mutual inductance bridging are reported in "Pulse Technique," loc. cit., page 105-107.[10] Calculation shows that maximum improvement is obtained if the mutual inductance with successive sections is alternately positive and negative and decreases rapidly, following a definite law. However, because the mutual inductance is small and is decreasing, its value becomes very small, of the order of 1 per cent. or less, after the second or third mutual inductance. It is very difficult to produce these values in practical lines. If these mutual inductances are neglected, delay variations of 1 per cent. or more per section occur. Then delays, which are negligible for one or more sections, become important for a great number of sections, such as 20 or 40.

Fig. 2.—Principle of a delay line time distributor with approximate or with accurate time division. (a) Block diagram of the distributor. When only approximate time division is required use the parts represented in thick lines. When accurate timing is required add the parts represented in dotted lines. (b) Showing how each channel selector pulse is produced. The boundaries of the channel are very accurately determined by these pulses. (c) Showing how the master multivibrator is synchronized by the master oscillator (see Fig. 2a).

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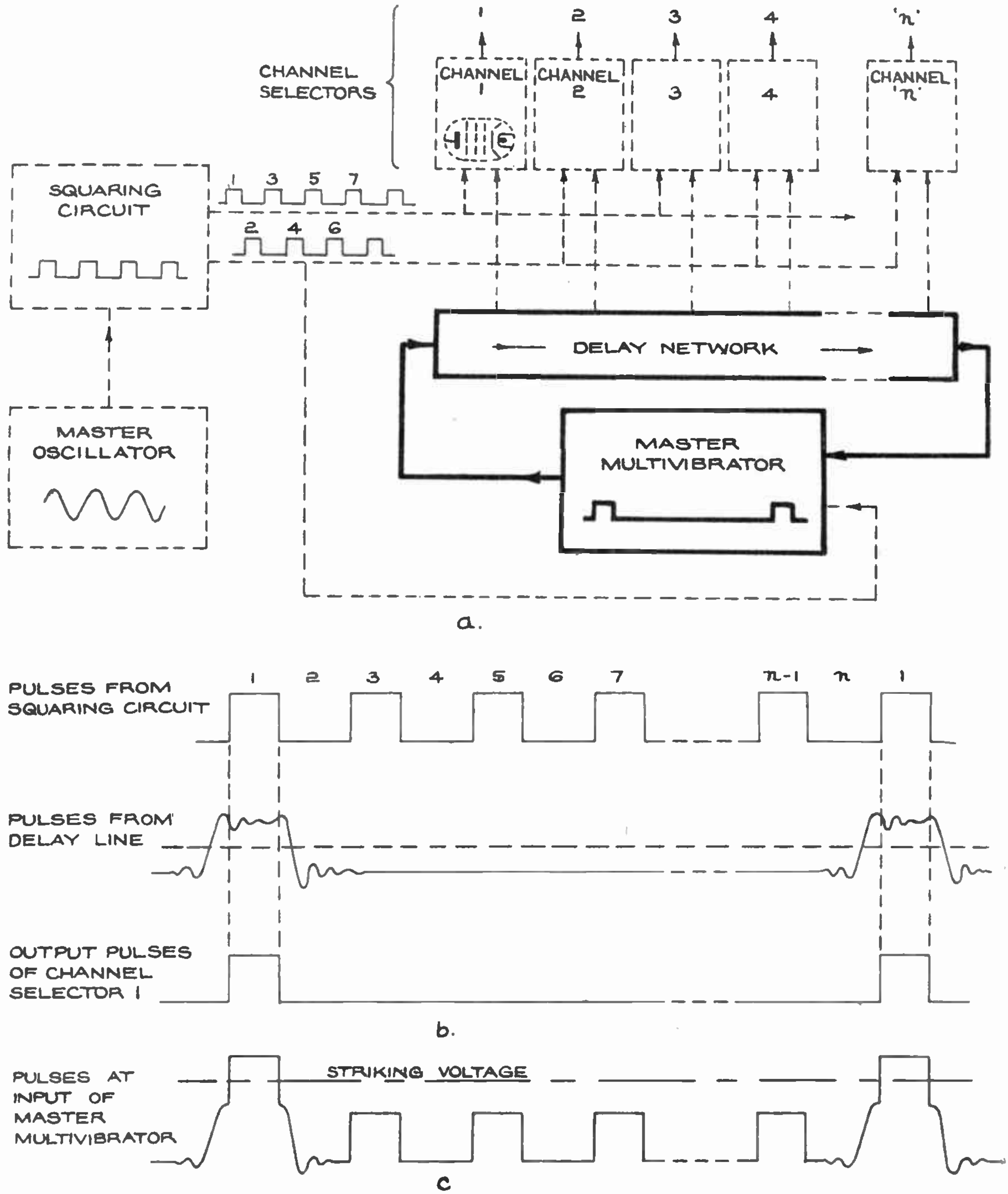

See caption, page 390.

Fig. 2.

Bridging improves the phase characteristic of the line to very near the cut-off frequency. On the other hand the attenuation characteristic is slightly modified, with, in good conditions, a peak near the cut-off. This helps to improve the response shape.

A complete account will be given in another publication. This method permits the number of sections to be reduced from 1,000 to approximately 100 or even less.

The pulse generator feeding the delay line is a multivibrator suitably designed to produce pulses of great amplitude in the low characteristic impedance of the delay line. To obtain this result a special three-valve circuit is used (Fig. 4). The circuit comprises a normal multivibrator circuit with the addition of a cathode follower between the plate of the first valve and the grid of the second. If the multivibrator is producing sharp pulses and, if the second valve works only during the duration of each pulse, the cathode follower enables this valve to be driven with considerable grid and peak cathode current. The valve then produces peak pulses of considerable power, although the mean power is the rated one.

To show this, consider first a normal multivibrator circuit as exemplified by valves V_1 and V_2 of Fig. 4. Assuming that valve V_1 is saturated and that no current is flowing in valve V_2 then if a negative pulse is applied on the grid of V_1, a positive pulse will appear on the grid of V_2 and current will flow in this valve. In order that the pulse may have a sharp edge at this instant, R_{p1} and the amplifying factor of V_2 must be as great as possible. But, if R_{p1} is great, the grid-cathode capacity of V_2 affects the operation, and if the slope of V_2 is great, the grid current will become very important thus introducing a shunt resistance across R_{p1}. In addition, this circuit is not very efficient when it produces pulses of a

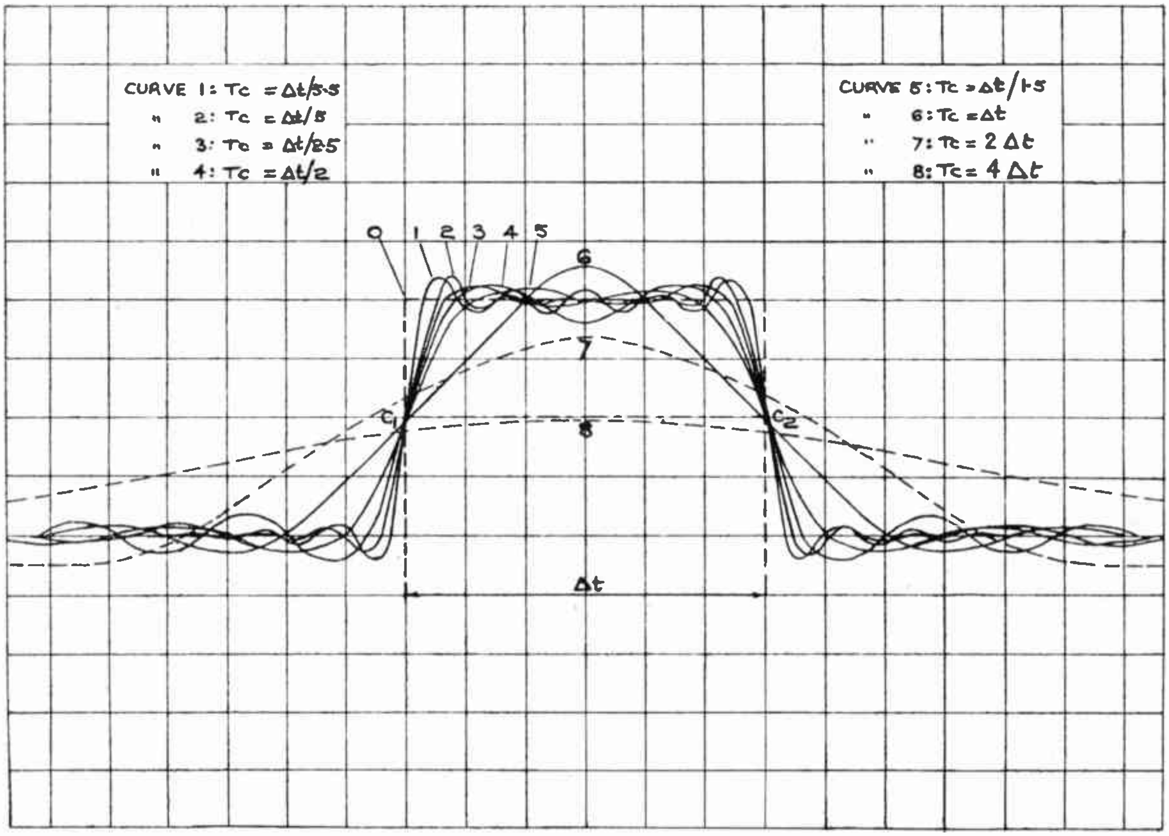

Fig. 3.—Response of an ideal low pass filter to a square pulse for various values of the cut-off frequency. For $T_c < \Delta t$, the width at mid-height is equal to the pulse width.

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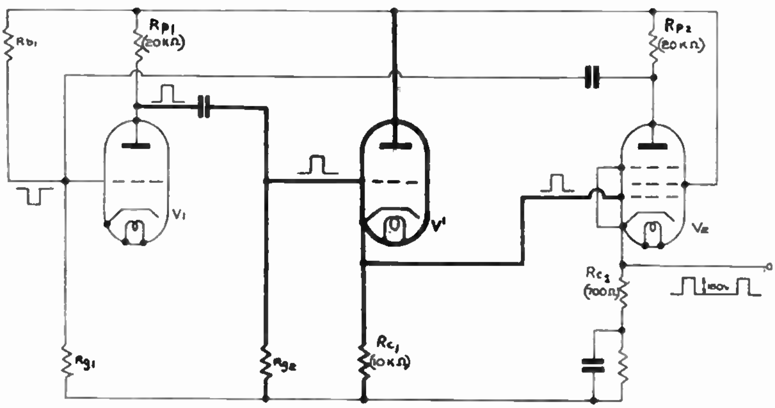

Fig. 4.—Power multivibrator—by adding a cathode follower between the first and third valve of a multivibrator, the peak power of the pulses generated is considerably increased.

duration shorter than half the repetition period. In this case, one of the valves is working all the time, except during the occurrence of the pulses, and the other only during the occurrence of the pulses. If the current flowing in the first valve is nearly equal to (or not too much greater than) the normal rated value, the valve will be used at about its normal dissipation rate. The second valve, working only during a short fraction of time, will be using only a small fraction of its normal dissipation unless the current during the occurrence of the pulses is much greater than the normal value. But this results in great grid current and it has been shown that this is not possible with the normal multivibrator circuit.

If a cathode follower is now added between the anode of the first valve and the grid of the second, it will transfer the pulses appearing across R_{p_1} to the grid of V_2. The load on R_{p_1} will be very small because of the cathode resistance R_{c_1}, and the power applied to the grid of V_2 will be great because the cathode follower impedance is usually much smaller than R_{p_1}. Besides, the capacity shunting R_{p_1} and that due to the connection on to the grid of the next valve will be negligible because of the negative feedback introduced by the high value of the cathode load of valve V′. If V′ is an EF50 or another high slope valve, the impedance looking back from the grid of V_2 is of the order of from one to some hundred ohms. This impedance is comparatively small and enables the grid of V_2 to be driven with high grid current. Thus, the peak current in valve V_2 can be made very great, easily equal to 10 or 20 times the normal rated current.

Examples of practical values are given in Fig. 4. The mean cathode current in V_1 is equal to about 13 milliamps and the peak current is nearly equal to the mean current. The peak cathode current of V′ is about 15 milliamps and the mean current less than 2 milliamps (the pulse length was about 1/20th of the repetition period). Valve V_2 had a peak cathode current of about 200 milliamps and a mean current of about 10 milliamps. Peak cathode currents as high as 300 milliamps can easily be obtained in valve V_2. Resistance R_b is used to bias positively the grid of V_1 so that this valve becomes sensitive only to negative pulses.

This circuit has been used extensively in the Multichannel Pulse System described in the next section. For example, the master-multivibrator feeding the delay network distributor is a power multivibrator. It produces pulses of about 150 volts amplitude in an impedance of about 3,000 ohms, that is a peak current of 50 milliamps. Also the demodulator multivibrators are of this type in order to get the required audio output without audio amplification. Nearly all multivibrators in the circuit are of this type.

Another feature of the distributor is the feedback coupling between the multivibrator pulse generator and the delay line (Fig. 2a). When a pulse is produced by the multivibrator, it travels through the delay line, returns at the input of the multivibrator and triggers it again. The result is that the recurrence frequency produced by this circuit is very stable. In a practical circuit it was found that the repetition frequency varied by less than 0·05 per cent. when the H.T. supply voltage

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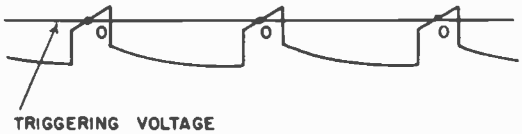
varied by ±8 per cent. This, in itself, is very good but if a higher degree of stability is required, the circuit can be very accurately controlled by a crystal oscillator tuned to a high harmonic of the recurrence frequency, its wave having preferably been squared. Figure 2c shows why the addition of the square wave to the feedback pulses will produce perfect stabilization if applied correctly. It is clear that the triggering line can be moved upwards or downwards within certain limits without altering the triggering time when the square wave is added. It is also evident that, even if the multivibrator is synchronized with a harmonic of high order, it will not unlock and lock on another harmonic because of the selection introduced by the delay line feedback pulse.[2]

However, even with all these features, the distributor is not perfect because the pulses through the delay line are distorted. An example of the type of distortion obtained in a practical case is shown on Plate VI. Other examples are given in "Pulse Technique" (Figs. 8 and 9).[10] To obtain perfectly shaped selector pulses and an accurate automatic time division, the circuit of Fig. 2a in dotted lines should be added. A crystal oscillator, whose frequency equals the recurrence frequency multiplied by half the channel number, is followed by a squaring circuit supplying two square waves in phase opposition. One wave corresponds to the odd channels and the other to the even channels. The squaring circuit is arranged so that the width of the positive pulse is slightly less than the channel width in order to leave a small gap between channels (see Fig. 2a). The first wave is applied to the suppressors of the odd channel selector valves and the other to the suppressors of the even channel selector valves. The taps on the delay line are connected to the grids of the corresponding valves. Fig. 2b shows clearly the considerable improvement introduced by the square waves. One of the square waves is used to lock the master multivibrator.

Such a distributor has the advantage of producing channel selector pulses of perfect shape and accurate timing with a relatively small number of sections in the delay line. It has been used extensively in other equipments where time distribution is required.

2.2.2. *Method of Modulation*

The main feature of the method of modulation is the use of trapezoidal pulses (Fig. 5a and b). These pulses are applied at the input of a multivibrator in series with the signal. In this multivibrator one grid is biased negatively far below the cut-off voltage so that normally no current flows in the valve. The multivibrator is then non-sensitive until a pulse appears on the biased grid. Then, if the pulse has a sufficient amplitude, the grid may cross the cut-off point, current will appear in the corresponding valve and the multivibrator will trigger.



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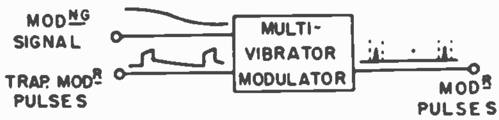

(*b*)

Fig. 5.—Pulse time modulation. (a) A multivibrator strikes when the trapezoidal pulse crosses the triggering level. This level is modulated by the signal. (b) Showing how the time modulation is produced.

For simplicity, it will be assumed that the multivibrator strikes at the moment when the grid voltage attains, or becomes greater than, a certain voltage which will be called the "triggering voltage." The variations of the grid voltage with respect to the triggering voltage are represented in Fig. 5a. This figure shows what happens in the absence and in the presence of a modulating signal. For simplicity, it has been assumed that the triggering line is modulated and the trapezoidal pulses are fixed in position.

It is assumed that the bias voltage is adjusted so that the multivibrator triggers on the axis of symmetry of each pulse when there is no modulation (point 0, Fig. 5a). Then, if the modulating wave is applied, the triggering point will move to the left or right of the axis of symmetry within the boundaries defined by the leading and trailing edges of each pulse. The modulated multivibrator is designed to produce very sharp pulses whose amplitude and length are practically unaffected by the modulation.

The great advantage of obtaining definite boundaries for the modulation is that no cross-

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talk is possible between adjacent channels if the adjacent space is occupied by other channels. The modulation is linear so long as the triggering point remains inside the boundaries.

Any type of multivibrator can be used. The simplest type, so far as number of valves is concerned, is the blocking oscillator. This type of multivibrator has the advantage of producing pulses of great peak power energy with small valves. Peak currents of many amperes can be obtained with small valves supplying normally only a few milliamperes. Blocking oscillators have been extensively used during the war and circuit details and information on the design of the special transformer required can be found in many publications.

In a simplified version of a blocking oscillator modulator the channel selector pulses are applied to the suppressor of the oscillator valve, and a sawtooth waveform to the cathode through another valve. The signal is applied to the grid through the secondary of the blocking transformer. A diode is used to introduce a small impedance in the cathode circuit when the first valve is triggered. Alternatively a trapezoidal pulse could be applied to the cathode of this valve. Then the suppressor need not be pulsed.

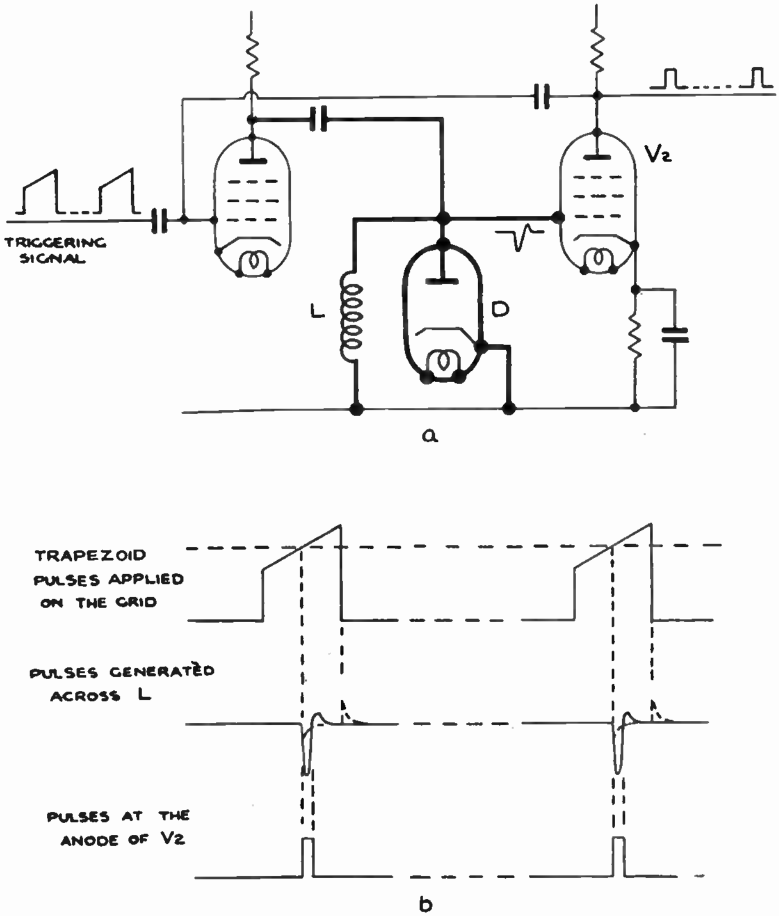

Fig. 6.—Principle of a new multivibrator modulator producing very short time modulated pulses. This is obtained by introducing a small inductance anywhere in the feed-back loop. (a) Simplified circuit of a modulator with inductance. (b) Diagram showing the very short time modulated pulses produced by the circuit.

The advantage of this circuit is that it requires only one valve per channel. Its disadvantage is the accumulation of many functions in a single valve. In addition, since the outputs of all channels must be mixed together, one mixer valve per channel is also required. It appeared that this second valve could also be used to improve the efficiency of the multivibrator by introducing it in the feedback circuit. A detailed analysis showed that a power multivibrator modulated by many channels simultaneously has the advantage of efficiency without an increase in the total number of valves. The difficulty was to produce a multivibrator producing very short pulses and introducing no cross-talk when modulated by many channels simultaneously.

To produce very short pulses accurately timed, a new concept was introduced, the use of an inductance in the feedback circuit; and to avoid cross-talk a three-valve multivibrator was used.

2.2.2.1. *Multivibrator with Inductance Coupling*

Figure 6 shows a two-valve multivibrator with an inductance coupling between the

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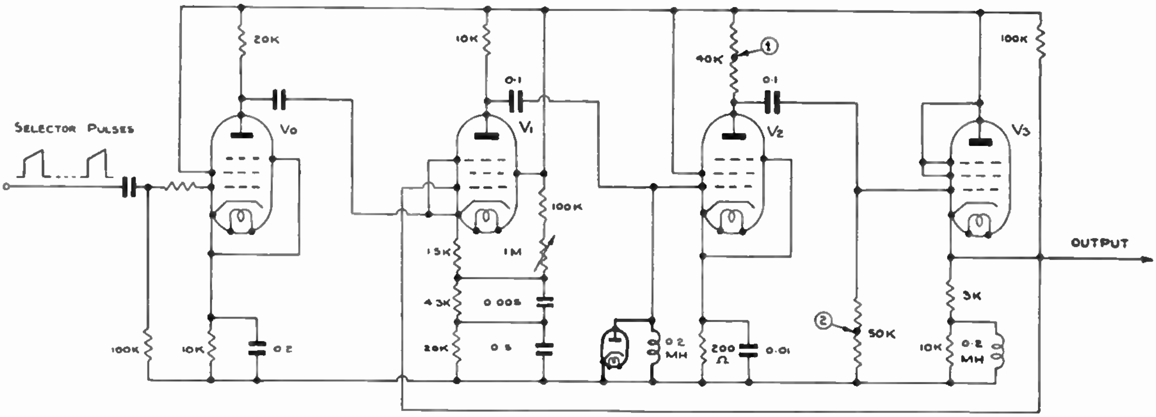

Fig. 7.—Circuit of a power modulator producing very short time modulated pulses of great peak power. In this circuit, the inductance is connected between the anode of V_1 and the grid of V_2.

anode of the first valve and the grid of the second. This inductance L is shunted by a diode D. Normally no current flows in the first valve. When this valve is triggered, the inductance is shock excited by the anode current and starts to oscillate. The first oscillating pulse is negative and is not damped by the diode. The next pulse is positive and therefore strongly damped, thus preventing further oscillations. Figure 6b shows the shape of this pulse. It will be observed that the grid of V_2 swings sharply from cut-off to saturation. This sharp swing produces a square pulse of definite length in the anode of V_2, equal approximately to half the period of oscillation of the tuned circuit formed by inductor L shunted by the capacitance across it. With inductances of small values, very short pulses can be produced. Figure 6b shows in dotted lines the type of pulses produced, using a resistance capacity coupling instead of an inductance, on the grid of V_2. It is obvious that this kind of pulse is not satisfactory.

Figure 7 shows a three-valve multivibrator using the inductance principle. This circuit was designed in order to study its properties when used as a pulse modulator. The trapezoidal pulses were applied to the cathode and the modulating signal was replaced by an adjustable D.C. cathode bias obtained from an adjustable resistor connecting the cathode to the H.T. voltage. The shapes of the pulses were recorded at the main points of the circuit for two extreme positions of the generated short pulse (Fig. 8). The pulses were observed with a monitoring equipment having an appreciable input capacity, so that they represent worse conditions than would be obtained in practice. Furthermore, they were copies taken with pencil and transparent paper from the monitoring C.R.T. However, they show clearly that the tails of the pulses are, in most cases, of a duration of some microseconds only. From Fig. 7 the time constants can be easily calculated. They are everywhere smaller than one microsecond. Thus, 10 microseconds after the appearance of a pulse at any point of the circuit, the residual pulse voltage is at least 100 db below the pulse peak voltage. Hence the circuit can be used to modulate many channels without cross-talk.

The addition of the third valve (cathode follower valve V_3) enables pulses of great peak power to be obtained. In this case pulses of more than 120 volts in 3,000 ohms are generated. As will be explained in next section, it also eliminates any cross-talk by direct coupling between channels when the circuit is used to modulate many channels.

2.2.2.2. *Combined Modulation of Many Channels (Multiplex Modulators)*[9]

Figure 9 represents an example of this type of multichannel modulator. It is a three-valve multivibrator of the type represented in Fig. 7 but with the first valve duplicated for each additional channel. Thus, V'_9 will correspond to channel 9, V'_{13} to channel 13, V'_{17} to channel 17 and so on. All these valves have their anodes connected together as well as their grids. The signals and corresponding trapezoidal pulses are applied to the respective cathodes. During the time allocated to channel 9, valve V'_9 is sensitive, and the other V′ valves are insensitive. The three

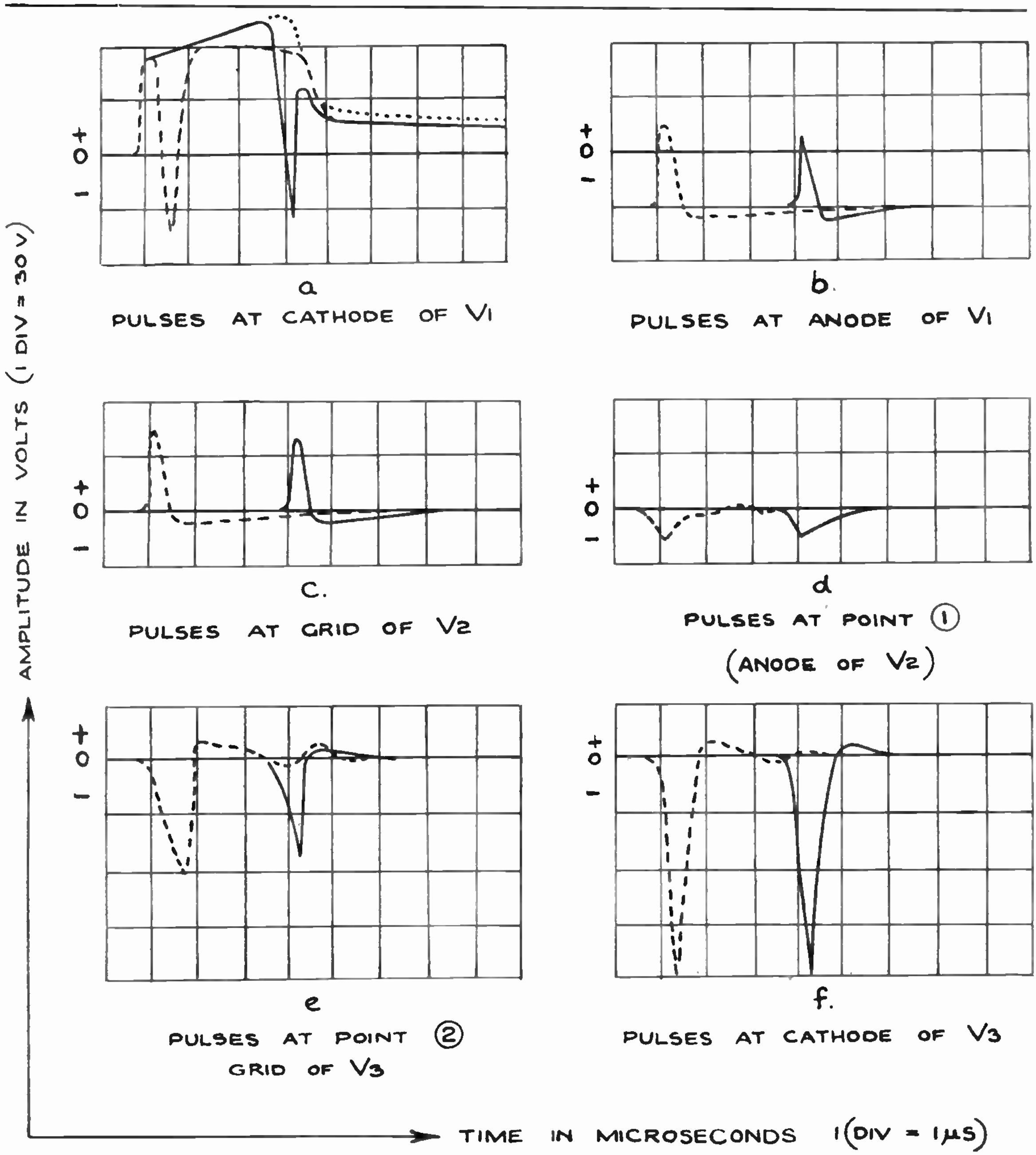

Fig. 8.—Waveform diagram recorded at various points in the circuit of Fig. 7 for the two extreme positions of the generated short pulses. These pulses were observed with an oscilloscope having an appreciable input capacity, which explains the long tails.

valves V'_9, V_2, V_3 form a multivibrator circuit and will respond to channel 9. When the time allocated to channel 13 arrives, valve V'_{13} becomes sensitive and all other V′ valves are not sensitive. Thus the circuit V'_{13}, V_2, V_3 will respond to channel 13, and this process will

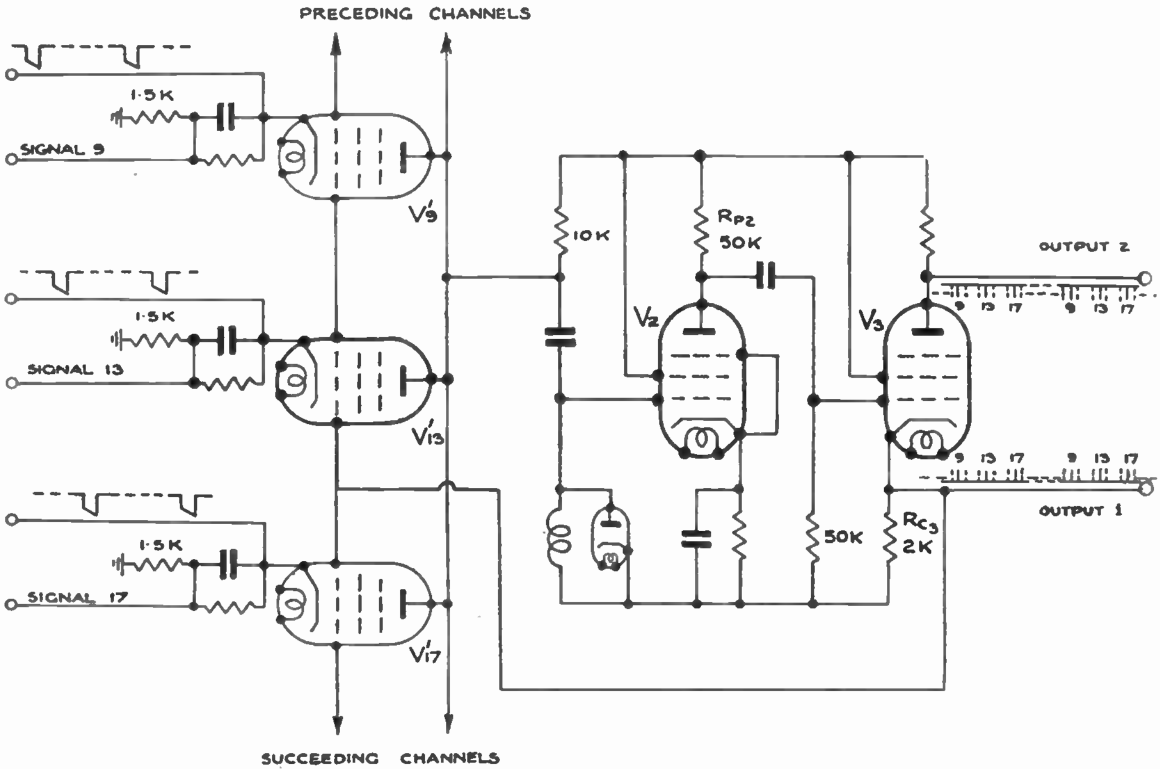

Fig. 9.—Multiplex multivibrator modulator. This circuit is obtained from the circuit of Fig. 7 by multiplexing valve V_1 into V'_9 V'_{13} V'_{17}. . . . This circuit modulates many channels and supplies all the time modulated pulses together.

continue in logical sequence for the following channels.

It is obvious that in such a system cross-talk is possible between the channels modulated by the circuit. Cross-talk will come mainly from two different sources, namely from trailing voltages which have not disappeared before the next channel pulse appears and from direct coupling between the audio signals by capacity through the circuit.

It has been shown for the circuit of Fig. 7 that all time constants are smaller than one microsecond. This still applies to the circuit of Fig. 9 provided that the number of V′ valves is not greater than about 10.

Thus, the cross-talk from this source will be smaller than 100 db if the successive channels have gaps of at least 10 microseconds between them. In a 24-channel system with a recurrence frequency of about 8 kc/s, the time allocated to each channel is 5·2 microseconds. Thus, this type of modulator can easily modulate 6 channels, the spacing between two successive channels being 15·6 microseconds.

It is important to note that this result is obtained only if valve V_3 is introduced in the circuit. Without this valve, the anode of V_2 is loaded by the grid-to-earth capacitance of all V′ valves. Since the anode impedance of V_2 is 50 kΩ, which is more than 10 times the cathode impedance of V_3, the time constant across this impedance is at least 10 times greater and could introduce appreciable cross-talk.

Cross-talk may also appear by direct coupling. For instance, a fraction of the signal 9 applied to the cathode of V'_9 may travel through the cathode-grid capacity of this valve to the grid of V'_{13} (or V'_{17}) producing cross-talk. The amount of cross-talk appearing through this path depends on the value of the impedance connecting the grids of valves V′ to earth. With a two-valve circuit (V_1 and V_2), this impedance is of the order of 40 kΩ and the amount of cross-talk can be calculated and is found to be about

60 db. With valve V_3, the grids of valves V' are connected to earth by a resistance at least 20 times less and the cross-talk is smaller than 86 db, which is completely negligible.

By reducing the value of R_{c3}, still smaller cross-talk values can be obtained without reducing the peak output pulse amplitude too much.

Summarizing the above study, it is clear that this circuit introduces no cross-talk and produces channel modulated pulses of great power (about 100 V in 2,000 ohms) and short length. In addition it requires one and a fraction of one valve per channel (1 and $\frac{1}{3}$ of a valve per channel in a 24-channel system).

2.2.3. *Mixer Circuit*

The mixer circuit is very simple and introduces no cross-talk. An example of a 24-channel system is shown in Fig. 10a. The pulses coming from the modulators are applied to the grids of cathode followers which have the cathodes connected together. The correct cathode bias helps to clip the pulses appearing on each grid and to suppress the trails on the common cathodes, thus suppressing any possibility of cross-talk provided the common cathode resistance is small enough to introduce no tail. This clipping is clearly shown in Fig. 10b.

Advantage has been taken of Fig. 10a to show, as an example, a complete mixer circuit for a 24-channel system. It is assumed that 4 modulators are used, one for every 6 channels.

The odd channels are combined together with the synchronizing pulse in resistance R_o. The even channels are combined together in resistance R_e. Then all the pulses are combined together in resistance R. The advantage of this progressive grouping is that the shunt capacity on each resistance is smaller than if all the grouping had been done in one operation. Furthermore R_o and R_e can have a greater value than R because the spacing between pulses is greater for R_o and R_e, and because the last cathode followers will clip the tails appearing from R_o and R_e.

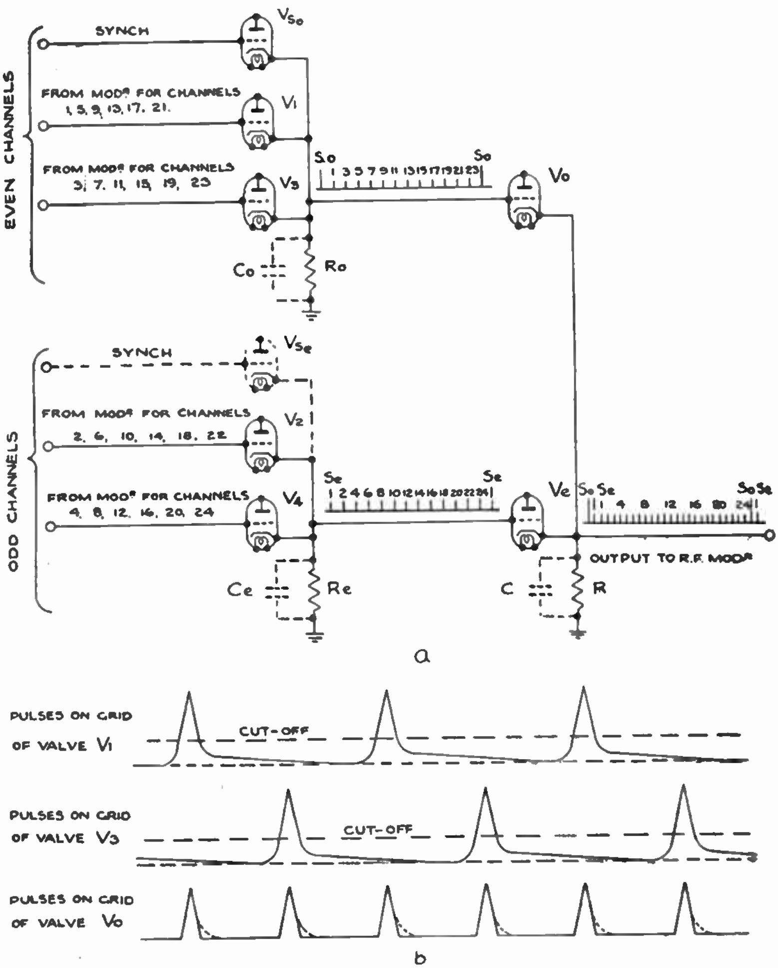

Fig. 10.—Mixer output circuit.

2.3. *Principles used in the Design of Receivers*

2.3.1. *Distributor*[1, 2, 3, 4, 5]

The distributor for the receiver is very similar to the distributor for the transmitter. They both use the same type of power master multivibrator, delay line, and accurate time division by the use

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of square waves (see Fig. 2). The differences are that the master multivibrator in the receiver is synchronized by means of synchronizing pulses selected from the channel pulses by amplitude or length discrimination, and the square waves are obtained by selection, amplification and squaring of the correct harmonic existing in the pulses generated by the master multivibrator (Fig. 16). If the selectivity of the amplifier is very great, noise modulation on the synchronizing pulses is filtered by the amplifier and not transmitted to the square waves.[12]

The channel selectors are identical with those used for transmitters except that their function is to produce gating pulses of square shape defining the time allocated to each channel.

2.3.2. *Method of Demodulation, and Multiplex Demodulators*[8, 9]

In order to demodulate a train of phase modulated pulses, they are converted into a train of variable length pulses which can be demodulated very simply by means of a low-pass filter. The process can be followed in Fig. 11, where a multichannel train is represented in (b) a train of selector pulses corresponding to channel 2 in (c) and the conversion from phase modulation to width modulation in (d). This conversion can be obtained in the following way (Fig. 11a). The train of selector pulses is applied in series with the train of channel pulses to the grid of a multivibrator valve which is negatively biased so that normally the valve is not sensitive. When a selector pulse appears, the grid bias is greatly reduced, but the bias voltage is so adjusted that the multivibrator does not strike (Fig. 11d) until a channel pulse appears. Then the multivibrator strikes and remains in the new state until the channel selector pulse disappears. To avoid accidental triggering by channel pulses of great amplitude they are amplitude limited before being applied to the multivibrator.

A practical example of this type of demodulator multivibrator is shown in Fig. 12. A three-valve multivibrator is used. The grid of valve V_1 is biased positively so that normally current flows continuously in this valve. The cathode of V_3 is biased positively so that normally no current flows in V_3. A selector valve V_4 produces negative anode current pulses flowing during the time allocated to the corresponding channel. This current flows in the cathode resistance R_{c3} of valve V_3 and reduces its positive bias just sufficiently to make the valve ready to trigger when the corresponding channel pulse appears on the grid of V_3 and drives this grid positive. The channel pulses are applied negatively to the grid of V_1. They are amplified by valve V_1 and transmitted to the grid of V_3 through the cathode follower valve V_2. After V_3 strikes, it remains in

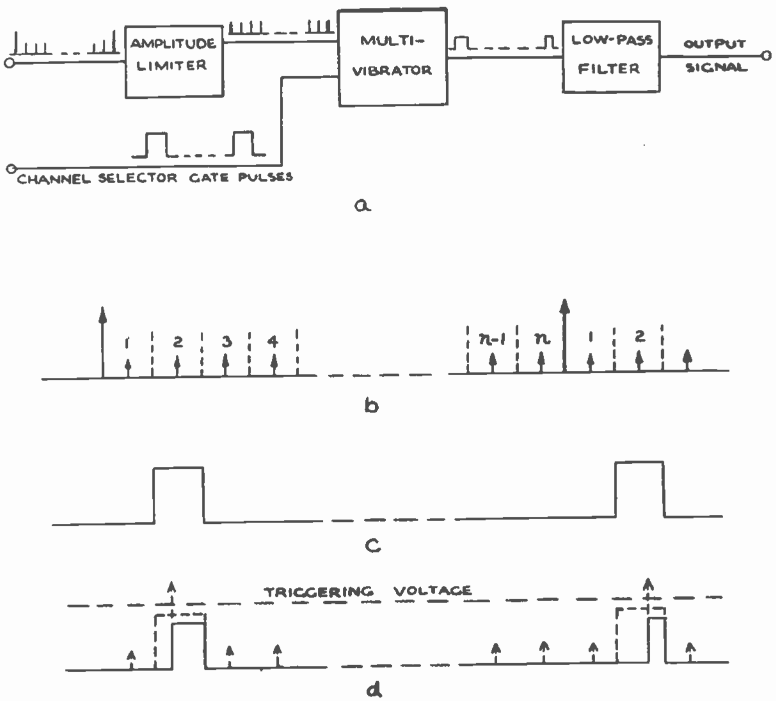

Fig. 11.—Multivibrator pulse demodulator. (a) Principle of the method. The channel selector pulses gate a multivibrator. When a pulse appears in the channel time interval, the multivibrator strikes. At the end of the channel the multivibrator reverts to its original condition. (b) (c) and (d) represent respectively time modulated pulses, selector gate pulses for channel No. 2 and multivibrator output pulses. In (d) it is shown how the channel 2 pulse is lifted by the corresponding gate pulse and strikes the multivibrator.

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this state until the end of the time allocated to the channel, when the current supplied from the anode of V_4 to the cathode of V_3 disappears and V_3 becomes insensitive. Thus the channel phase modulated pulses, applied to the grid of V_1 and corresponding to the selected channels, are converted into length modulated pulses in valve V_3, the proper channel being selected by the selector valve V_4 connected to V_3. The variable length pulses are demodulated by a low-pass filter connected to the screen or anode of V_3. Because the multivibrator is of the power type, the current flowing in V_3 is considerably greater than that in a two-valve circuit, and enough power can be collected in the low-pass filter to avoid audio amplification, particularly if this filter is in the anode circuit. The coupling between V_3 and the grid of V_1 can be made from the anode or the screen of V_3.

This circuit can be easily adapted to demodulate many channels simultaneously.

It is next assumed that instead of one output valve V_3, there are as many output valves as the number of channels required (some are represented in Fig. 13). All these valves will be fed in parallel by the cathode follower and the screens will be connected together. A selector train of pulses, corresponding to the required channel, will be applied to each cathode and the output will be taken from the anodes. Each of these valves will work alternately and, when one is working, no current is flowing in the others. A circuit of this type can work very efficiently for as many as 6 channels. For a 24-channel system, 4 complex demodulators are required, each one handling 6 channels. It is wise to interlace these channels, i.e. to distribute channels successively to the demodulators and repeat the operation as many times as required. Thus, in a 24-channel system, channels 1, 5, 9, 13, 17, 21 will be demodulated by the first multiplex demodulator, channels 2, 6, 10, 14, 18, 22 by the second and so on. With this precaution no appreciable cross-talk will appear in any demodulator, provided the circuit is carefully designed. A system of this type will require for demodulation about one and a fraction of a valve per channel.

3. Design of Multichannel Systems

We are now in a position to design a multichannel system.

It must be observed that, since the distributor uses square waves for channel timing, the number of channels must be even. Furthermore, the channel spacing must be equal and the synchronizing pulse or pulses must occupy one or more channel spaces. Another interesting observation is that, since the odd channels are selected by one square wave and the even channels by another, it is practical to design the circuit for odd channels only and to duplicate for even channels, with the exception of the distributor. This also advantageously leaves an appreciable gap between channels in each system and reduces the amount of possible cross-talk. The odd and even channels can be mixed by means of cathode followers. It has already been explained (see section 2.2.3 and Fig. 10) why this method of mixing eliminates cross-talk.

Accurate timing is important only when a

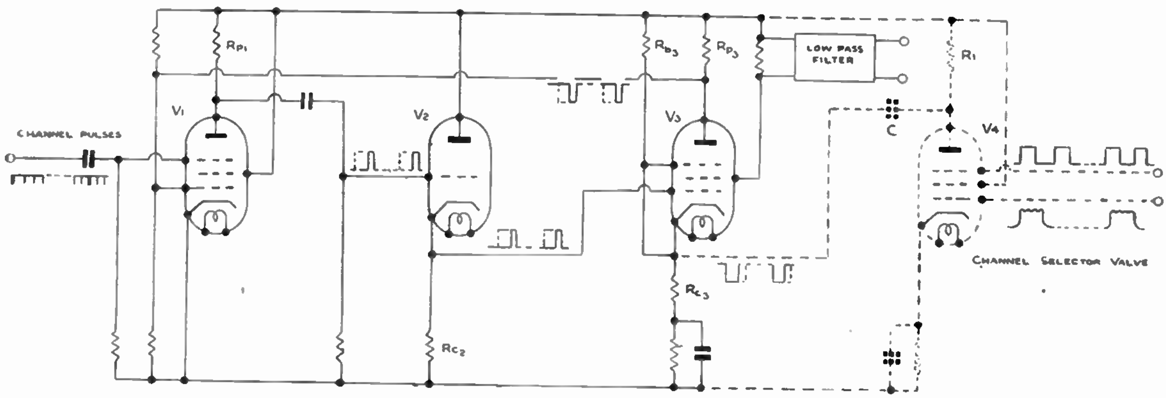

Fig. 12.—Power multivibrator used as demodulator. Valve V_3 works at high efficiency and supplies great peak power. The output is taken on the screen but could be taken conveniently on the anodes.

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large number of channels is required. The delay line distributor as shown in heavy lines in Fig. 2, that is without combination with square waves, is practical and satisfactory when the number of channels is not greater than 20. The improved distributor is recommended for more than 20 channels.

The modulators and demodulators work satisfactorily so long as the channel length is not much smaller than one microsecond. For smaller values, the time constants must be reduced accordingly and this affects the amplification efficiency of each valve.

Extra amplification is required when the number of channels is increased with the consequent increase in the number of valves per channel required. A typical example is the demodulator circuit (Figs. 12 and 13). If the modulation length is reduced, the audio power obtained at the output of the filter drops, and eventually extra audio amplification is required which means the addition of an audio amplifier valve per channel.

A 24-channel system will now be described. No appreciable modifications in the system are required up to about 36 channels.

Two recurrence frequencies each of 8 kc/s are considered, this being roughly the lowest value giving a good intelligibility for speech. The 24 channels are divided into 12 odd and 12 even channels. In each group one channel space is added for the synchronizing pulse. The system is

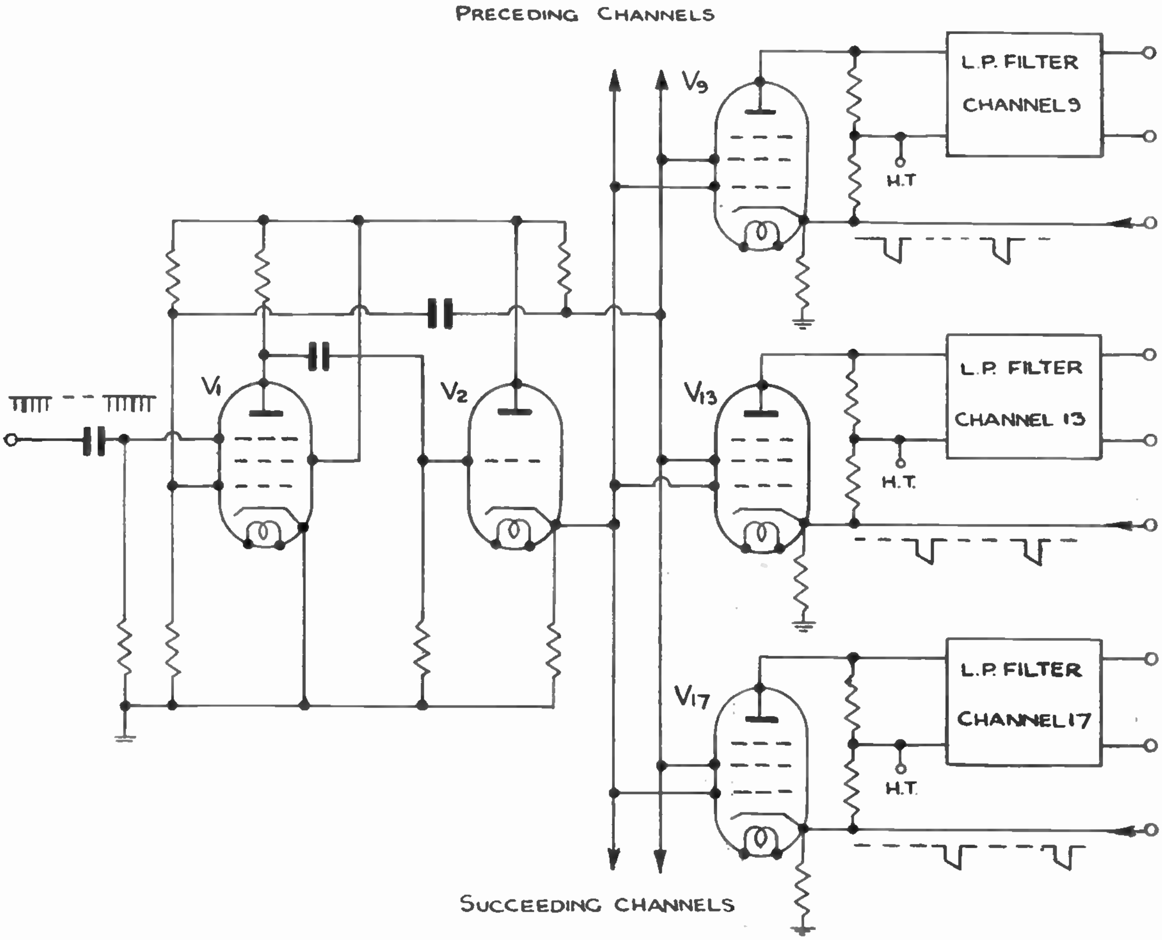

Fig. 13.—Example of multiplex power demodulator demodulating many channels simultaneously.

effectively a 26-channel system. Each channel spacing is equal to 4·8 microseconds. A maximum modulation length of ± 2 microseconds is taken and a safety gap of 0·8 microsecond between successive channels.

3.1. *Design of the Transmitter*

The block schematic of the transmitter is shown in Fig. 14 and a simplified circuit schematic in Fig. 15. The distributor is controlled by a crystal oscillator adjusted to 104 kc/s, that is

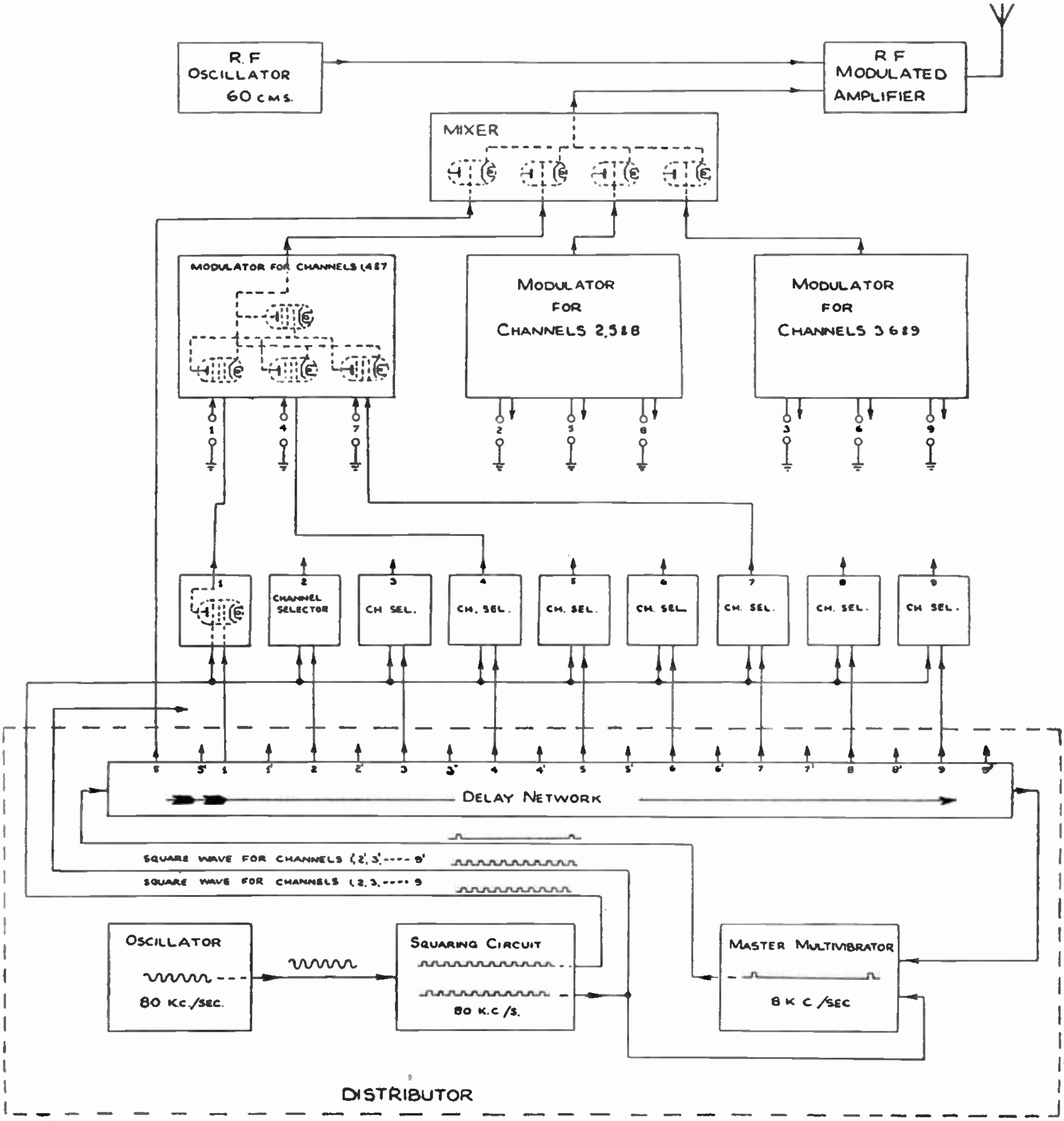

Fig. 14.—Block schematic of a 24-channel pulse transmitter showing the distributor, channel selector valves, multiplex modulators and mixers.

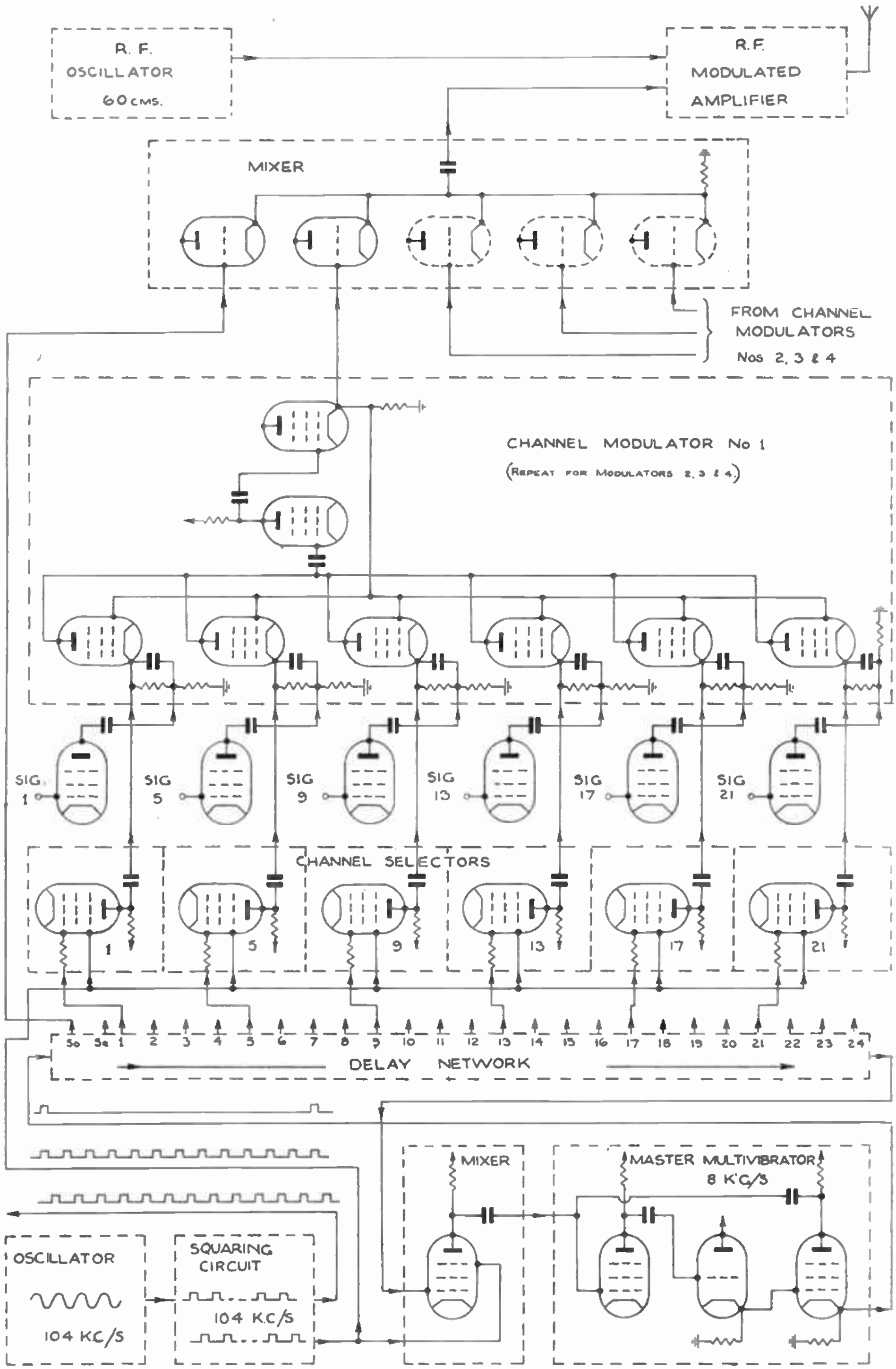

Fig. 15.—Simplified schematic of the 24-channel pulse transmitter. This circuit requires in all 3½ valves per channel. An earlier circuit required 2½ valves per channel but this circuit is preferred for a system with a large number of channels.

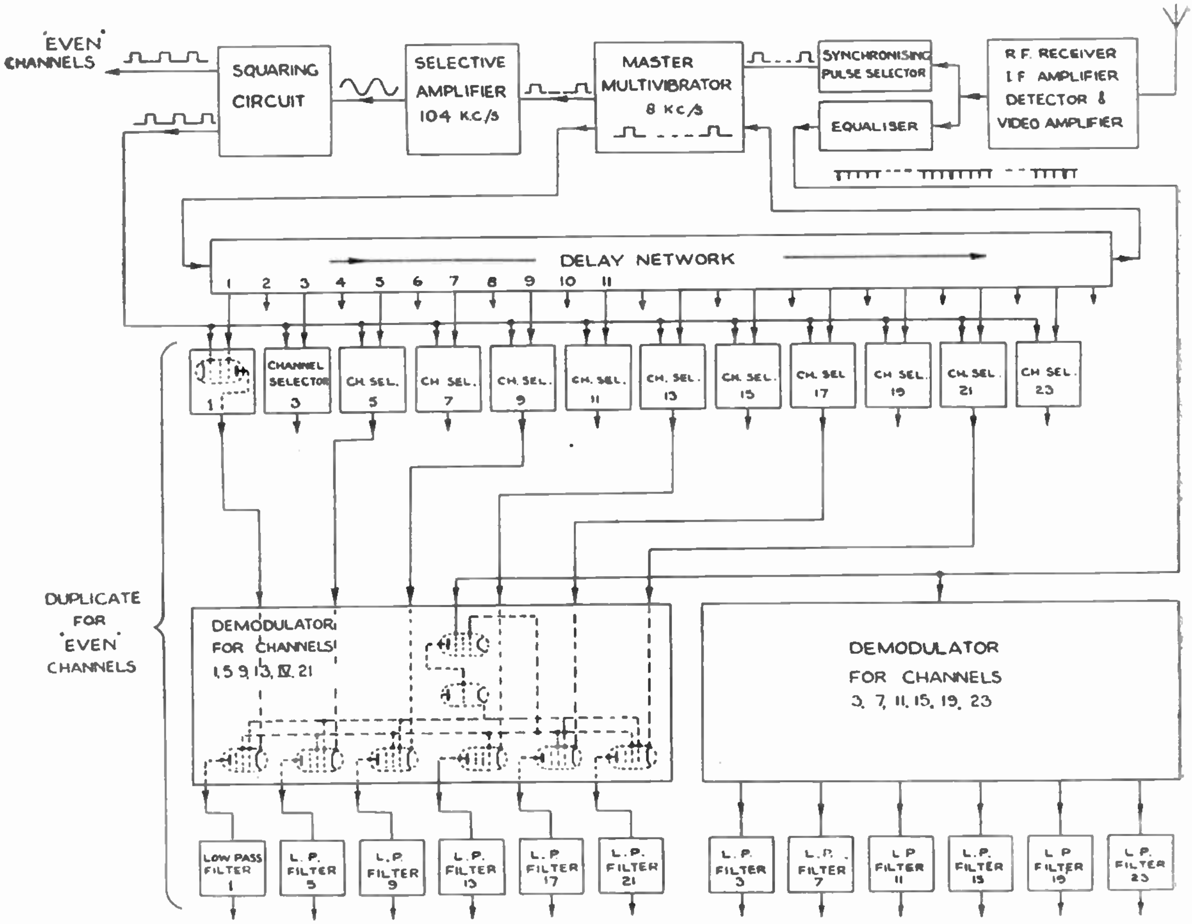

Fig. 16.—Block diagram of a 24-channel receiver showing the distributor, the channel selector valves and the multiplex demodulators, and the method used for synchronizing the time distributor at the receiver. The synchronizing pulses selected from the incoming pulses lock the master multivibrator. A selective amplifier (harmonic generator) produces the square wave which defines accurately the time boundaries of each channel. Most of the noise carried by the synchronizing pulses is filtered by this selective amplifier.

13 times the recurrence frequency. Two multiplex modulators are used for the odd channels and two for the even channels with interlaced channel distribution to the modulators.

The first modulator will deal with channels 1, 5, 9, 13, 17, 21, the second with channels 3, 7, 11, 15, 19, 23 and so on. In each modulator, the gap between successive channels is 14·4 microseconds. It has already been shown that, with such a gap, no appreciable cross-talk appears, providing power type modulators are used. This type of multiplex modulator is chosen. It also has the advantage of producing channel pulses of great power and short length. Pulses of less than 0·5 microsecond length and about 100 V in 2,000 ohms can be easily obtained (section 2.2.2.1, Figs. 7 and 9). These pulses are transmitted to the mixer and from there to the R.F. transmitter. For small power transmitters and short radio links no further pulse amplification is required.

A simplified schematic diagram is shown in Fig. 15. The signals are supplied to the modulators preferably through valves, which are not shown in the circuit, and except for one valve in the master multivibrator and one valve in each multiplex modulator, all work intermittently and no appreciable current flows in them when they are at rest.

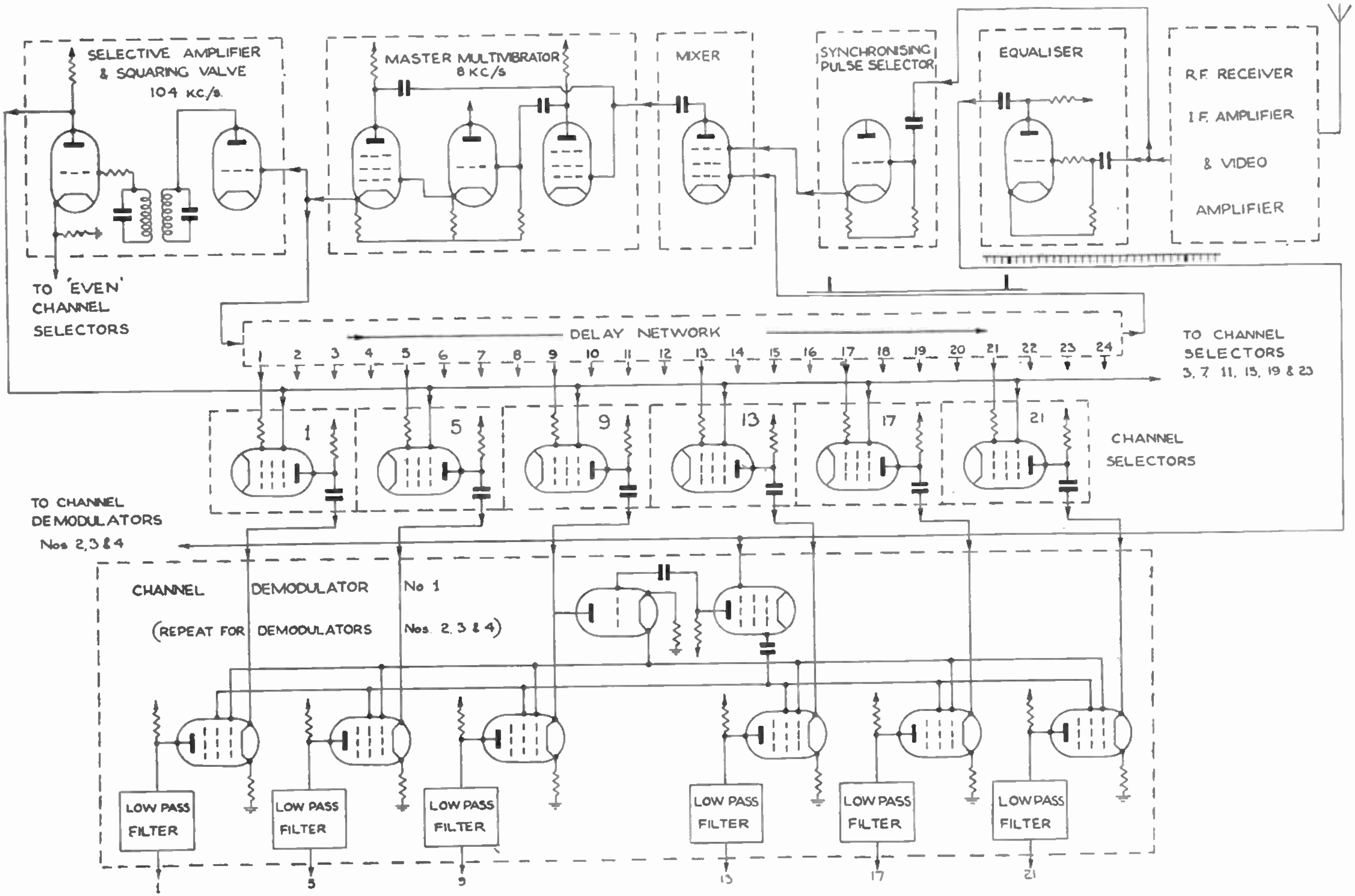

Fig. 17.—Simplified diagram of the 24-channel receiver. This circuit requires in all about 2½ valves per channel.

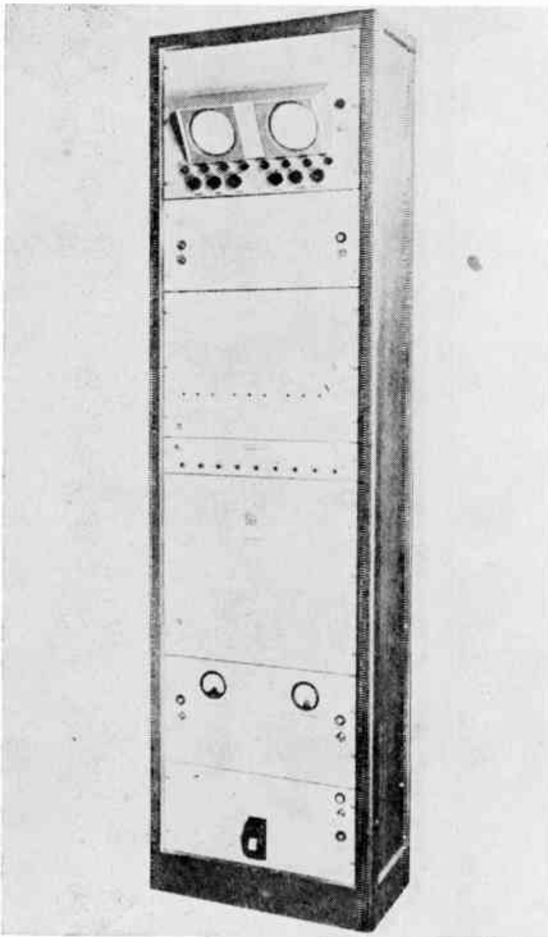

PLATE I.—Front view of the experimental transmitter demonstrated in 1943. Designed on a 20-channel basis, it contains only 9 channels fully equipped: 3 at the beginning of the channel scale, 3 at the middle and 3 at the end. The equipment is entirely automatic and contains a double monitoring system. One C.R.T. is used to monitor all the channels simultaneously and the other to monitor any particular channel required.

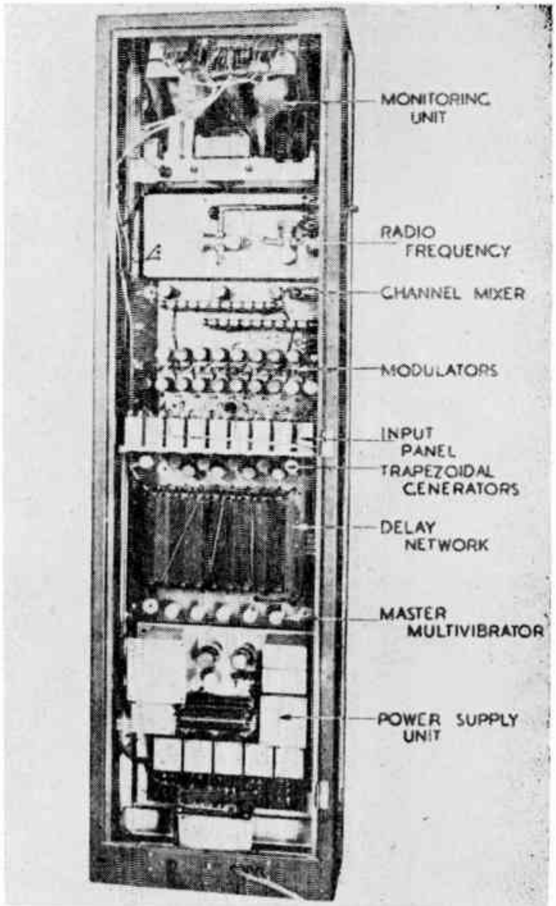

PLATE II.—Rear view of the experimental transmitter. The delay line is visible in the middle of the rack. On top of the rack, below the double C.R.T. monitoring equipment, is the R.F. transmitter chassis which is connected directly to the aerial by a cable (not shown).

This circuit requires three to four valves per channel.*

3.2. *Design of the Receiver*

The block schematic of the receiver is shown in Fig. 16 and a simplified circuit schematic in Fig. 17.

The incoming signals are first amplified and detected, then they are sent simultaneously through a pulse selector, and an equalizer. The pulse selector separates the synchronizing pulses by virtue of their special characteristic (greater length). These pulses are used to synchronize the distributor as explained in section 2.3.1. The equalizer gives a constant amplitude level to all channel and synchronizing pulses. This equalization is required before the pulses are applied to the complex demodulators (section 2.3.2).

Two multiplex demodulators are used for the

* A simple circuit using less than three valves per channel and no audio valves was initially designed and used in the 18-channel experimental system with good results (see Plates I and II). The above circuit is preferred for a system with a large number of channels.

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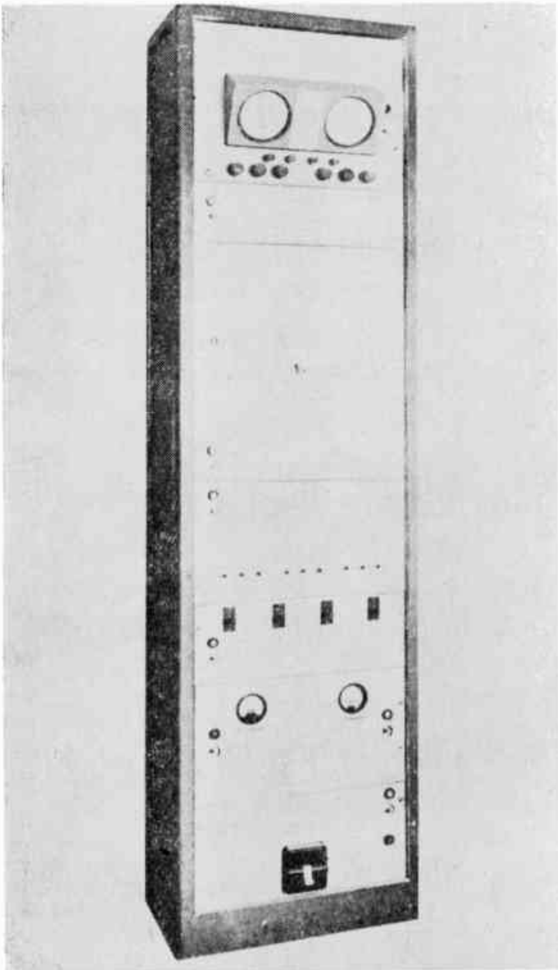

PLATE III.—Front view of the experimental receiver. The equipment is self contained and need only be connected to an aerial by a cable.

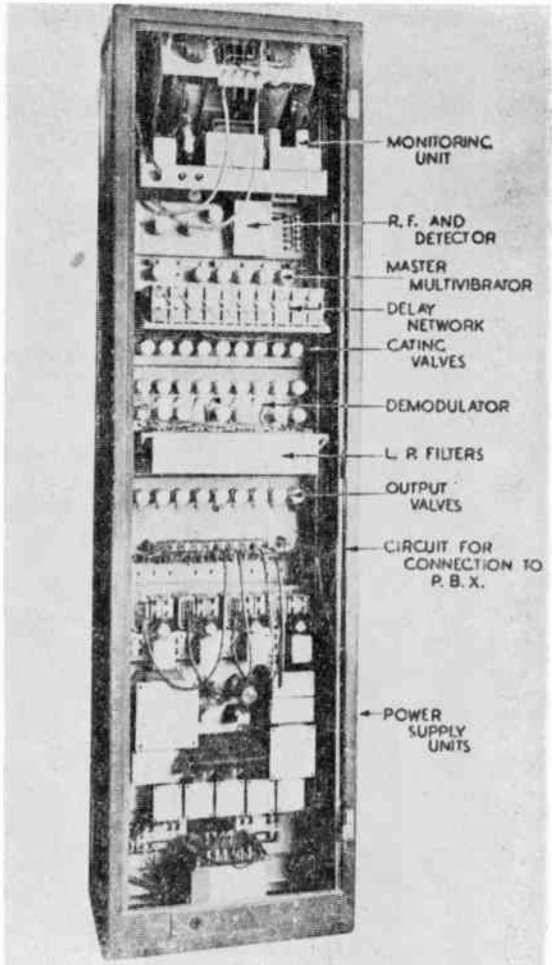

PLATE IV.—Rear view of the experimental receiver. An improved version of the transmitter delay line is used.

odd channels and two for the even channels, with interlaced channel distribution. The first demodulator deals with channels 1, 5, 9, 13, 17, 21, the second with channels 3, 7, 11, 15, 19, 23 and so on. In each demodulator the gap between successive channels is 14·4 microseconds. With such a gap, no appreciable cross-talk appears if the demodulator is of the power type. With a good design the cross-talk can be made much smaller than 70 db. The power type also has the advantage of producing a large output so that no audio amplification is necessary.

A simplified schematic diagram is shown in Fig. 17. Excepting for the valves of the selective amplifier, one valve in the master multivibrator and one valve in each multiplex demodulator, all valves work intermittently and no appreciable current flows in them when they are at rest.

The whole circuit requires less than three valves per channel.

The main advantages of this system are, apart from its simplicity, the absence of any manual control, the relatively small number of valves required, its low power consumption, its low cross-talk ratio (smaller than 70 db) and its high signal-noise efficiency due to the possibility of

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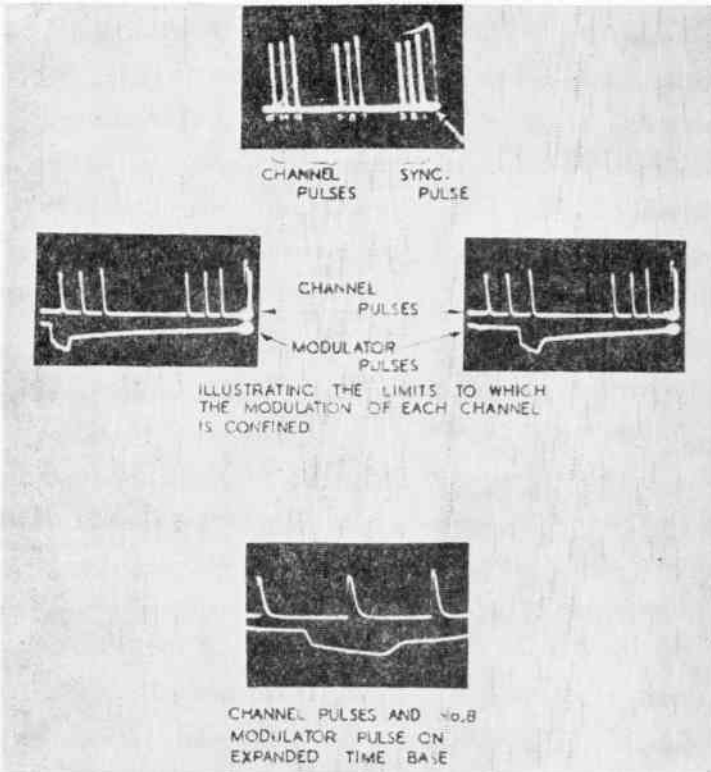

PLATE V.—Typical oscillography of channel and modulator pulses—Transmitter.

reducing the time gap between channels to a small fraction of the time allocated to each channel.

4. Description and Performance of an Experimental System

Two experimental equipments were developed by the writer. The first was started in July 1943, completed in January 1944, and was tested over a distance of some hundred yards. It gave excellent performances and it was decided to divide it up into smaller units for mounting in a mechanically propelled vehicle. The writer took advantage of this situation to partly redesign the transmitter and improve the receiver.

Tests were made at a distance of 36 miles with satisfactory results.

4.1. *First Experimental Equipment*

4.1.1. *General Description*

The first experimental equipment was contained in two cabinets, one for the transmitter and one for the receiver. Front and rear views of these cabinets are shown in Plates I to IV.

Transmitter and receiver each have a monitoring unit. For this purpose, two double-beam cathode ray tubes are provided, one displaying the synchronizing, and all the channel pulses. On the other can be selected and displayed any channel pulse and its correspondent trapezoidal or selector gating pulse. On the second trace of the first tube a sharp pulse appears pointing at the selected channel pulse. Plates V to VII show the views which can be obtained on these tubes.

The equipment was designed for 18 channels but only 9 channels were incorporated by groups of 3, leaving three gaps each of 3 channels space. Three multiplex modulators and three multiplex demodulators were used. For 18 channels the total number of valves was approximately 65, that is 3·6 valves per channel. The main object when designing this equipment was to produce a system as soon as possible. No attempt was made to reduce the number of valves if this had any temporary disadvantage or required further experimental investigations.

The power consumption was 300W each for transmitter and receiver.

The equipment was entirely automatic and no adjustments were required during operation.

4.1.2. *Performance*

It was tested at some hundred yards and gave the following results.

1. Distortion

(*a*) Overall Amplitude Distortion

About $\pm$ 1 db from 300 to 3,000 cycles. A typical overall audio response characteristic is given in Chart 1. The theoretical amplitude distortion appears at 3 kc/s and is equal to $\pm 0{\cdot}1$ db. The actual distortion was due probably to the audio circuits in the transmitter and receiver.

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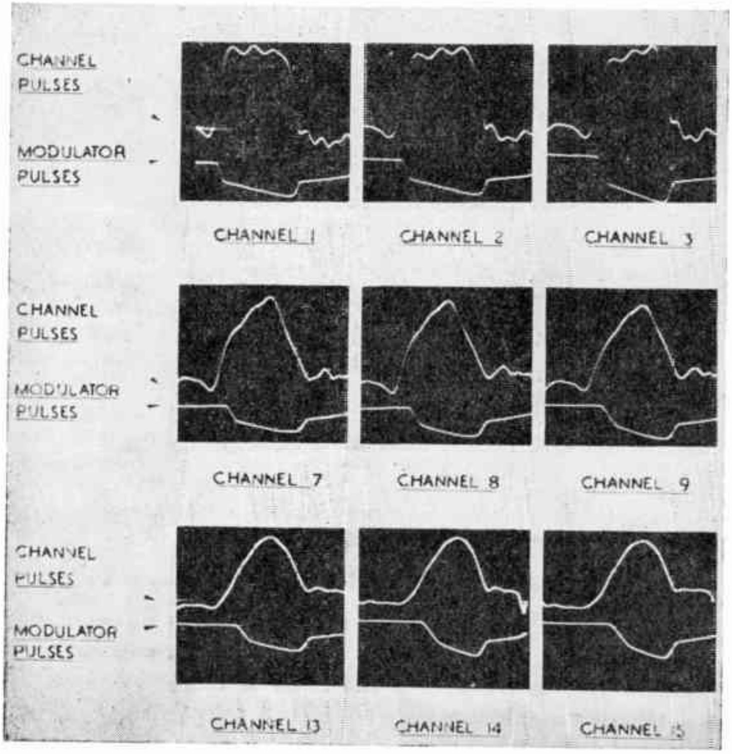

PLATE VI.—Pulses obtained at tappings on the delay line and trapezoidal pulses obtained from the former. When a rectangular pulse travels through the delay line, it is progressively distorted. This is seen clearly by inspection of the oscillograms from channel 1 to channel 15. The delay line was one of the very first used and the distortion is appreciable. For better results see "Pulse Technique," loc. cit., Figs. 8 and 9, where it is shown how a correct amount of mutual inductance between successive coils combined with mutual inductance with distant coils improves considerably the shape of the pulses. Theoretically by alternate positive and negative mutual inductance between successive coils, a perfect phase linearity can be obtained but the values required are so small that in practice it is difficult to obtain a real improvement.

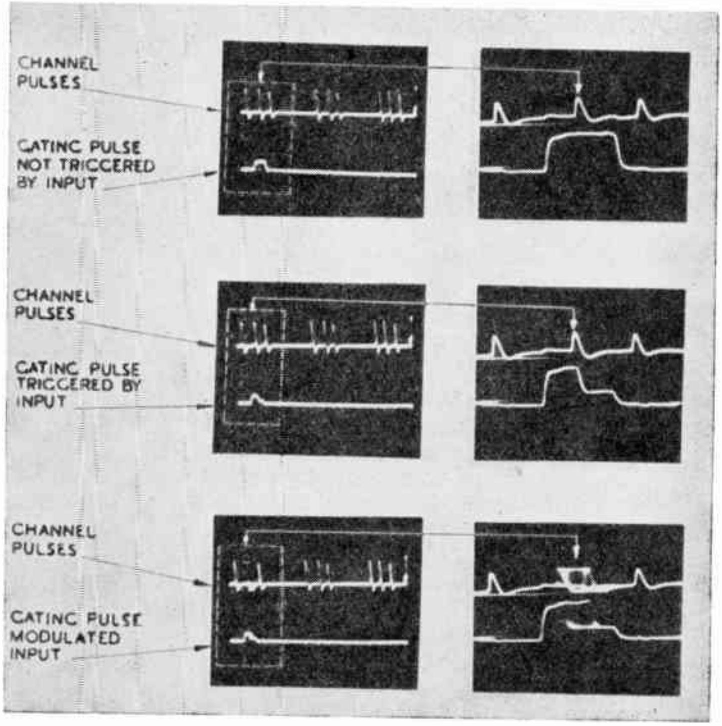

PLATE VII.—Typical oscillographs of channel and gating pulses—Receiver. The left-hand column of oscillographs shows the synchronizing pulse followed by three groups of three-channel pulses. These correspond to the channels fully equipped in the system. The empty spaces correspond to three other groups of three channels each. The gating pulse in the top photographs indicates the channel space.

The right column shows, for each case, some part of the left-hand photograph on an enlarged time scale. These pictures are obtained directly on the monitoring cathode ray tubes shown in plates III and IV.

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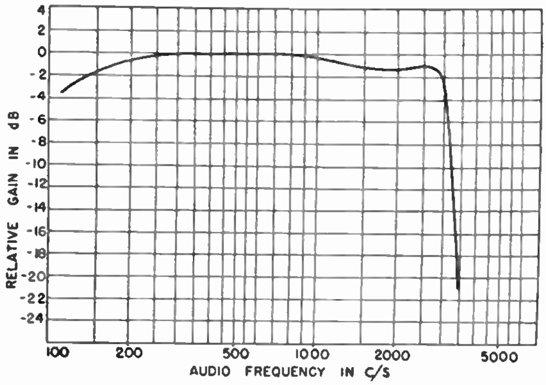

CHART I.—Typical overall frequency response of a single channel.

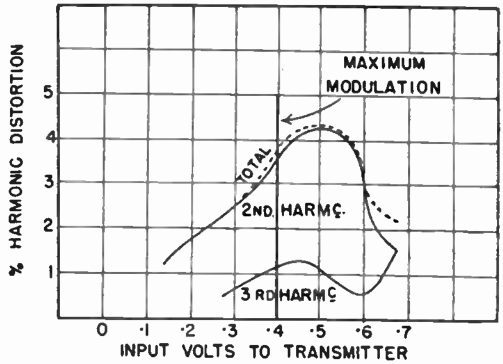

CHART II.—Typical overall harmonic distortion. These curves were obtained with the very first version of multiplex modulators, a version using a very small number of valves (PLATE II).

(*b*) Overall Harmonic Distortion

About 3 to 5 per cent. at full modulation.

The amount of distortion varied from one channel to another. No attempt was made to obtain the best adjustment in this first experimental circuit. A typical curve, corresponding to one of the best channels is shown in Chart II.

(*c*) Cross Modulation

Very small and comparable to the theoretical expectations, that is to say, less than 0·5 per cent. For instance, for an input of 2·5 kc/s and approximately full modulation, a component of 900 cycles was found with an amplitude of about 0·35 per cent. of the 2·5 kc/s amplitude.

4.2. *Mobile Equipment*

The first equipment had a simple delay network distributor at transmitter and receiver. In the mobile equipment the square wave circuit and complex modulators with inductance were used.

The final adjustments and tests were made over a distance of 36 miles. It was found that the signal-to-noise ratio was always better than 50 db and the cross-talk usually inaudible.

During all the tests the system worked perfectly and no adjustments were required during operation.

5. Patents

British Patents only are mentioned below. They refer to the system described.

(*a*) *Delay Line Distributors*

1. 578.690 (B. B. Jacobsen and M. M. Levy)
2. 581.328 (M. M. Levy)
3. 581.330 (M. M. Levy)
4. Applic. 10179/44 (C. W. Earp and M. M. Levy)
5. 587.939 (M. M. Levy)

(*b*) *Power Multivibrators*

6. 587.940 (M. M. Levy)

(*c*) *Trapezoidal Pulses*

7. 587.941 (M. M. Levy)

(*d*) *Multivibrators, Modulators and Demodulators*

8. 587.942 (M. M. Levy)
9. Patent pending (M. M. Levy)

6. Bibliography

10. Levy, M. M., " Some Notes on Pulse Technique," *Jour.Brit.I.R.E.*, Vol. 7, 1947, No. 3, pp. 99 and 105.
11. Levy, M. M., " Some Theoretical and Practical Considerations of Pulse Modulation," *J.I.E.E.*, Vol. 96, 111.A, No. 13, 1947, p. 565.
12. Discussion on " Pulse Communication," *J.I.E.E.*, Vol. 94, Part 111.A, No. 11, March 1947, p. 585.
13. Kirby, H. D. B., " A Time Sharing System of Multiplex," *Electronic Engineering*, Vol. 21, No. 260, October 1949, pp. 360-368.

7. Acknowledgments

The writer is indebted to Messrs. Standard Telephones & Cables, Ltd., for permission to publish this record of work done whilst in their employment, and under their direction, during 1943-1944.

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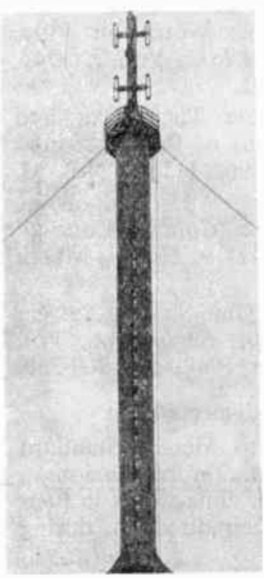
SUTTON COLDFIELD TELEVISION STATION

The new television station for the Midlands is now being built on a site about 550 feet above mean sea level at Sutton Coldfield, some ten miles north of Birmingham, and is to open on December 17th, 1949.

The vision transmitter has a peak power rating of 35 kilowatts and will operate on a carrier frequency of 61·75 Mc/s, which corresponds to a wavelength of 4·86 metres, or about 16 feet. Asymmetric sideband transmission will be used, with the upper sideband partially suppressed, in order to conserve the limited band of frequencies allocated to television, and so enable more transmitters to operate in this band of frequencies.

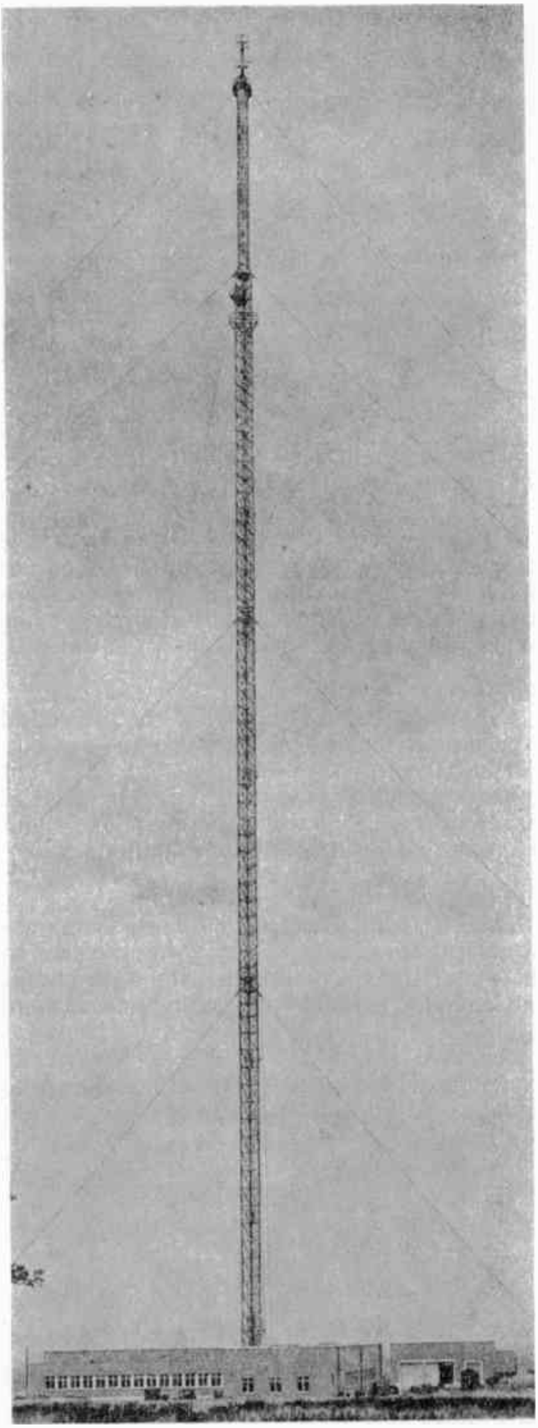
The sound transmitter has an unmodulated carrier power rating of 12 kilowatts. Conventional amplitude-modulated double-sideband transmission will be used on a carrier frequency of 58·25 Mc/s, corresponding to a wavelength of 5·15 metres, or 17 feet approximately.

The vision signals will conform to the standards adopted for the British Television Service, and will be sent from Alexandra Palace to Sutton Coldfield over a radio link or a coaxial cable, both of which are being provided by the General Post Office.



LEFT
The television transmitting aerial 750 feet above the ground. A single wide-band aerial of novel design will radiate both the sound and vision signals. It consists of two tiers of four vertical folded dipoles arranged in cruciform.
The cylindrical structure below the television aerial is connected with BBC proposals to broadcast on very high frequencies.

RIGHT
The transmitter building and the 750-ft. mast.



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TRANSFERS AND ELECTIONS TO MEMBERSHIP

Subsequent to the publication of elections to membership which appeared in the October issue of the Journal, a meeting of the Membership Committee was held on October 25th, 1949. Twenty proposals for direct election to Graduate or higher grade of membership were considered, and thirty-four proposals for transfer to Graduate or higher grade of membership.

The following list of elections was approved by the General Council : eighteen for direct election to Graduate or higher grade of membership, and thirty-one for transfer to Graduate or higher grade of membership.

Direct Election to Full Member

Pezopoulos, Kyriakos N., D.Sc. — Athens, Greece

Direct Election to Associate Member

Burt, Sidney John George — Hong Kong
Jones, Ioan Myrfyn Aveston — Newcastle
Lee, William Clayton, B.Sc. — Wellington, New Zealand
Ramamurti, Tiruvadi Venkatraman — Bombay, India

Direct Election to Associate

Charnock, John — Purley, Surrey
Jack, Ian Munro — Hargeisa, British Somaliland
Lane, Albert, Capt. — London, S.E.25
Mawer, Cecil Ernest — Newcastle, 3
Mitchell, Francis Albert — Stockport
Sherrard, Joseph Lister, F/O — Carlisle, Cumberland
Wallis, Kenneth Douglas — Leicester
Woodacre, Albert — Gt. Malvern, Worcs.

Direct Election to Graduate

Everett, Athelstan Frith, B.Sc.(Eng.) — Kew Gardens, Surrey
Fuller, Frederick Reginald — Accra, Gold Coast
Khoury, Khalil Ibrahim — Southampton, Hants.
Medhurst, Philip John — Hobart, Tasmania
Old, William Desmond — Redruth, Cornwall

Transfer from Associate Member to Full Member

Kapur, Brahm Dev, B.Sc. Brigadier, — London, W.1

Transfer from Associate to Associate Member

Brandt, Joseph Charles — Ware, Herts.
Draper, James Krishen — Hayes, Mddx.
Driffill, Joseph — Hull, Yorks.
Shipton, Harold William — Bristol, 6
Sukhadia, Pratapchandra U. — Bombay, India
Walker, Clarence Henry, B.Sc. — Mansfield, Notts.

Transfer from Student to Associate Member

Asquith, Stanley Francis William — Hockley, Essex

Transfer from Student to Graduate

Acharya, Khadri Samachar S., B.Sc.(Hons.) — Trichinopoly, India
Beecroft, William Douglas — Bradford, Yorks.
Cavanagh, Edward Joseph — Brentford, Middlesex
Coleman, William Frank, B.Sc.(Eng.) — London, S.W.1
Eldridge, Dennis Arthur George — Enfield Wash, Middlesex
Eve, Godfrey Arnold — London, N.8
Fellows, Horace — Wolverhampton
Franks, Percy Ronald — Luton, Beds.
Gifkins, Geoffrey Charles — Hertford, Herts.
Goodings, Harry Arthur — Ilford, Essex
Halsall, James Richard — Liverpool, 9
Jarman, Eric — Warrington, Lancs.
Justice, James William Henry — London, N.5
Kameswararao, Gummuluru Satyam, M.Sc. — Trichinopoly, India
Kinally, Dennis Raymond — Morden, Surrey
Land, Leonard Ernest, Lieut., R.N., B.Sc. — London, N.21
Page, John Guy — Reading, Berks.
Pawlus, Jan — London, E.12
Plowman, John Antony — Luton, Beds.
Pringle, John — Wembley, Middlesex
Sayers, John Francis — Barnehurst, Kent
Simpson, Mackenzie Adams — London, S.W.7
Spinks, Harry Roy — London, N.8

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THE INTEGRATION METHOD OF LINEARIZING EXPONENTIAL WAVEFORMS

by

A. W. Keen* (*Associate Member*)

SUMMARY

The exponential response of the simple CR " integrator " to a step impulse e.m.f. is shown to differ from the exactly linear rise required in many electronic applications by an " error " voltage which is proportional to the integral of the exponential output. Three methods are given of adding a second CR integrating section in such a manner that a close approximation to the error voltage is produced at the correct amplitude and superimposed on the output of the basic integrator. Each corrected network has an LR equivalent and may be arranged for voltage or current excitation. A number of suitable practical arrangements are given for each of the following applications :—

(i) Linear sawtooth voltage wave generation,

(ii) Waveform linearization in time base voltage amplifiers,

(iii) Linear sawtooth current wave generation, as in electromagnetic time bases used for television reception.

In the last application effective, superposition of the principal current and its integral is achieved by the use of split deflecting coils.

1.0 Introduction

1.1. It is frequently necessary in electronic circuit design to produce voltage or current varying linearly with time, an obvious example being the time base needed for waveform observation on the cathode ray oscillograph. In such applications it is usually essential that the degree of non-linearity in the desired variation be negligible; this requirement is particularly desirable in the case of scanning circuits for television receiving equipment, and the resumption of the B.B.C. television service has revived interest in the problem of achieving sensibly perfect time base linearity without recourse to complex circuit technique of the kind developed during the war for radar reception.

1.2. The object of the present paper is to draw attention to one of the more effective of the simpler methods of linearity correction. Existing literature contains brief accounts[1, 2, 3, 4] of an application of this method; the present purpose is to state the theoretical basis and describe the circuit technique, including new developments. Throughout, the treatment will deal specifically with scanning circuit applications, although the method is generally applicable to cases where the need arises for waveforms containing linear portions.

1.3. Practical time base systems are usually based upon the type of sawtooth generator shown in Fig. 1 (*a*), and the exponential voltage wave developed across the capacitor C is amplified by some form of push-pull voltage amplifier to provide a large symmetrical output for the deflector plates in the case of electrostatic tubes, or is used to drive a single-ended current output stage in the case of the all-magnetic tubes now universally employed for television reception.

2.0. Analytical Basis

2.1. In the present discussion, interest centres upon the degree of linearity of the output voltage, so that the action of the discharge device required to provide a recurrent variation will be ignored but see 2.5.

The output (e) of the CR network contained in Fig. 1 (*a*) to an applied e.m.f. of the form of Heaviside's Unit Function 1 and of amplitude E is shown at A in Fig. 1 (*b*). The analytical representation of this response is well known and is

$$e = (1 - \exp(-\tau^{-1}t))\,E \quad (t \nless o) \quad \text{..(1)}$$

where τ (secs), the circuit time constant, is, by

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* Formerly of R.F. Equipment, Ltd.

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definition, equal to CR (farads × ohms).

The slope of the output voltage is given by

$$\frac{d}{dt}\, e = \tau^{-1} \exp(-\,\tau^{-1}t)\, \mathrm{E} \quad \ldots\ldots\ldots(2)$$

$$= \tau^{-1}\mathrm{E}, \text{ at } t = o \ldots\ldots\ldots\ldots\ldots(3)$$

It is desired that this initial slope be maintained ((B) of Fig. 1 (*b*)); accordingly, the equation to the required response may be taken as

$$e' = \tau^{-1}\mathrm{E}\,.\,t \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(4)$$

The difference between expressions (1) and (4), preferably taken in ratio with the magnitude of the driving voltage, is therefore a measure of the degree of non-linearity of the network output. See C of Fig. 1 (*b*).

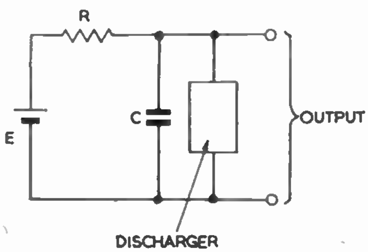

Fig. 1 (a).—Skeleton circuit of a basic form of exponential wave generator.

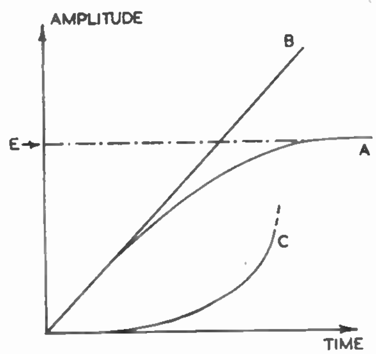

Fig. 1 (b).—Response of basic CR integrator to "step" type e.m.f.

This differential quantity

$$(e' - e)\,\mathrm{E}^{-1} = \Delta e\,.\,\mathrm{E}^{-1} = \tau^{-1}t - 1 + \exp(-\,\tau^{-1}t) \quad \ldots\ldots\ldots\ldots(5)$$

is independent of E but corresponds in form to the network configuration and to the type of applied voltage. It will be noted that this quantity increases with t for a given value of the circuit parameter τ, while for a given time interval the deviation from perfect linearity decreases with increase of τ.

Expanding the exponential term in (5), thus:—

$$\exp(-\,\tau^{-1}t) = 1 - (\tau)^{-1}t + (\tau^2\lfloor\underline{2})^{-1}t^2 - (\tau^3\lfloor\underline{3})^{-1}t^3 + \ldots \quad \ldots\ldots\ldots(6)$$

and substituting in (5) gives

$$\Delta e\,.\,\mathrm{E}^{-1} = (\tau^2\lfloor\underline{2})^{-1}t^2 - (\tau^3\lfloor\underline{3})^{-1}t^3 + \ldots \quad \ldots(7)^*$$

Inspection of (6) and (7) suggests comparison of the latter with the expansion

$$\int (1 - \exp(-\,\tau^{-1}t))\, dt$$

$$= \int \{(\tau)^{-1}t - (\tau^2\lfloor\underline{2})^{-1}t^2) + \ldots\}\, dt$$

$$= \tau\{(\tau^2\lfloor\underline{2})^{-1}t^2 - (\tau^3\lfloor\underline{3})^{-1}t^3 + \ldots\} \quad \ldots\ldots\ldots\ldots(8)$$

It follows that

$$\Delta e\,.\,\mathrm{E}^{-1} = \tau^{-1} \int (1 - \exp(-\,\tau^{-1}t))\, dt$$

$$= \tau^{-1} \int e\,(t)\, dt \ldots\ldots\ldots\ldots\ldots(9)$$

The significance of this result is that the amount of deviation of the output voltage of the simple network under consideration from the ideal linear rise desired is proportional to the integral of the actual output voltage. Accordingly, if a voltage proportional to the integral of this exponential output could be developed and superposed, in correct amplitude, on the latter, a perfectly linear total output would be obtained.

2.2. In considering the problem of electronic voltage integration suggested by the last paragraph it will be immediately apparent that the configuration analysed is the familiar CR integrating circuit.[5] In fact the attempt to develop an exactly linear rising voltage from the constant amplitude applied voltage calls for a network capable of exact integration. The implication that the simple integrator does not perform an accurate integration of its input voltage may be readily confirmed in operational notation. (An alternative treatment is also available.)[6] We have

$$e = \frac{\dfrac{1}{\mathrm{C}p}}{\mathrm{R} + \dfrac{1}{\mathrm{C}p}}\,\mathrm{E}1 = \frac{\dfrac{1}{\mathrm{CR}}}{\dfrac{1}{\mathrm{CR}} + p}\,\mathrm{E}1 = \frac{1}{\tau^{-1} + p}\,.\,\tau^{-1}\mathrm{E}1 \quad \ldots\ldots\ldots\ldots(10)$$

and the operator $(\tau^{-1} + p)^{-1}$ approaches p^{-1},

* Note: $\Delta e.\mathrm{E}^{-1} \approx (\tau^2\lfloor\underline{2})^{-1}t^2$ when τ is large, i.e. this function is then approximately parabolic in form.

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which represents exact integration, more closely as τ increases.

It would appear, however, that the association of an additional CR "integrator" with that of Fig. 1 (*a*) in the manner indicated in para. 2.2, while not resulting in perfect linearization, would at least lead to an appreciable reduction in non-linearity ; and this is borne out by experiment.

The use of two integrators in tandem to produce an approximation to the required integral correcting waveform is shown in Fig. 2. As R_2 is increased, the output waveform varies continuously through the series of forms tabulated at (*b*). The present paper deals particularly with the technique of superposing forms (i) and (v) but it may be noted in passing (see (ii)) that network Fig. 2 (*a*) is by itself an improvement on the single integrator shown in Fig. 1 (*a*).

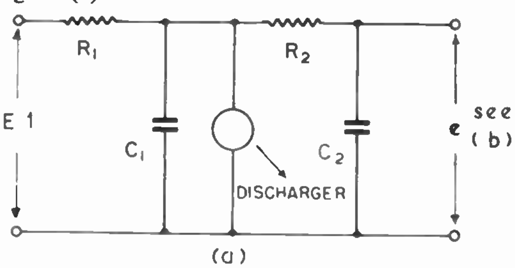

Fig. 2 (a).—Use of additional integrator (C_2R_2) to develop an approximation to the integral of the exponential output of C, R, (cf. Fig. 1 (a)).

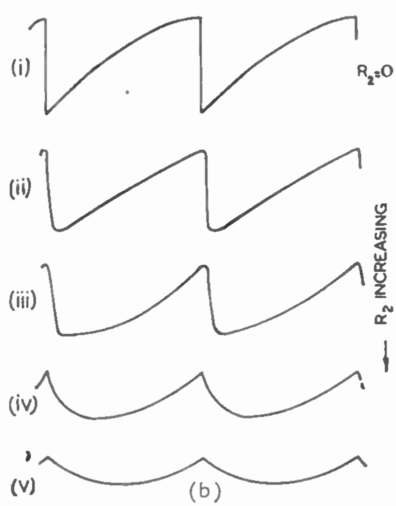

Fig. 2 (b).—Output waveform of (a) for different values of R_2.

2.3. A further property of the simple CR integrator, which will be of subsequent use, is the fact that as the parameter τ is increased the resistor voltage waveform approaches more and more closely that of the applied voltage. A simple demonstration of this point is as follows :

$$e_R = \frac{R}{R + \frac{1}{Cp}} E1 = \frac{p}{\tau^{-1}+p} E1 \rightarrow E1 \text{ as } \tau \rightarrow \infty \qquad \ldots\ldots\ldots\ldots\ldots(11)$$

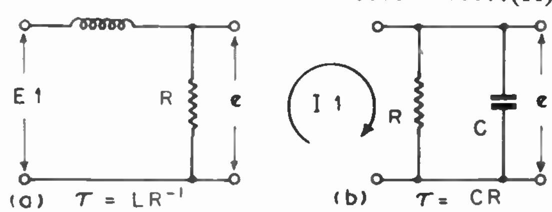

Fig. 3.—Other forms of basic integrator.
Fig. 3 (a).—The voltage-operated series LR integrator (See ref. 5).
Fig 3 (b).—The current-operated parallel CR integrator.

2.4. It will be useful to recall that the voltage-operated CR network discussed so far has two basic equivalents : (*i*) the voltage-operated LR circuit shown at Fig. 3 (*a*) and (*ii*) the current-operated parallel CR circuit of Fig. 3 (*b*).

In case (*i*) we have

$$e = \frac{R}{R + Lp} E1 = \frac{\frac{R}{L}}{\frac{R}{L} + p} E1 = \frac{\tau^{1}}{\tau^{1} + p} E1,$$

$$\text{where } \tau = LR^{-1} \qquad \ldots\ldots\ldots\ldots\ldots(12)$$

and, in case (*ii*)

$$e = \frac{R \cdot \frac{1}{Cp}}{R + \frac{1}{Cp}} I1 = \frac{R}{1 + CRp} I1 = R\frac{\tau^{-1}}{\tau^{-1}+p} I1,$$

$$\text{where } \tau = CR \qquad \ldots\ldots\ldots\ldots\ldots(13)$$

2.5. In the foregoing analysis a single exponential response to the applied step impulse has been considered in order to facilitate understanding of the basic principle of the correction method under discussion. In practice, of course, the discharge device connected into the section of the network producing the exponential charge makes the latter recurrent so that the input to the correcting network is an exponential wave rather than a single transient impulse. It is possible, however, to associate a discharge device

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with the reactive component of the corrector in such a manner as to reset the circuit to its initial state each time the discharger in the exponential generator operates, but this is not necessary unless the permissible recovery time is a very small fraction of a period. The simplest method of resetting the corrector is to connect a low-impedance diode in correct polarity across the correction resistor; another method is shown in the circuit of Fig. 5 (*c*). For simplicity, the analyses of the following sections assume resetting of the corrector circuit conditions at the end of each cycle.

3.0. Basic Networks

3.1. The first attempt to improve the linearity of the simple two-element integrator on the lines suggested by the foregoing analysis appears to have been made by G. F. Hawkins whose basic circuits are given in Fig. 4 (*a*), (*b*).

In the first arrangement, Fig. 4 (*a*), the capacitor of the basic integrator C_1R_1 is split into two series portions C_1', C_1'', and the voltage across the upper part C_1' is integrated by the additional network C_2R_2. The total output therefore consists of the sum of the exponential voltage across the lower section C_1'' of the split capacitor, and the correcting voltage developed across C_2. If the correcting CR circuit were capable of exact integration of its input voltage and if it imposed no load on the portion of the network from which it is fed, the total output would be perfectly linear for correct partition of C_1.

The relative values of the network parameters are readily derived by the following approximate analysis :—

Since C_2R_2 is large, $i_{R_2} \approx e_{C_1'}/R_2$, ...(14)

$$\therefore e_{C_2} = \frac{1}{C_2}\int i_{R_2}\,.\,dt \approx \frac{1}{C_2R_2}\int e_{C_1'}\,.\,dt \quad ..(15)$$

Neglecting the loading effect of C_2R_2 on C_1R_1

$$e_{C_1'} = \frac{C_1''}{C_1' + C_1''}(1 - \exp(-\tau_1^{-1}t))\,\mathrm{E}, \text{ where}$$

$$\tau_1 = R_1\frac{C_1'C_1''}{C_1' + C_1''} = C_1R_1 \quad ....(16)$$

$$\therefore e_{C_2} \approx \frac{C_1''}{C_1' + C_1''}\cdot\frac{1}{C_2R_2}\,.\,\mathrm{E}$$

$$\int(1 - \exp(-\tau_1^{-1}t))\,.\,dt \quad ......(17)$$

To this voltage is added

$$e_{C_1''} \approx \frac{C_1'}{C_1' + C_1''}\,.\,\mathrm{E}(1 - \exp(-\tau_1^{-1}t)) \quad (18)$$

and it follows from equation (9) that we must have

$$e_{C_2} = \tau_1^{-1}\mathrm{E}\int e_{C_1''}\,.\,dt \quad ....(19)$$

which reduces to

$$\frac{C_1''}{C_2R_2} = \frac{C_1'}{\tau_1}, \text{ (using (17), (18) in (19))}, \quad ..(20)$$

$$\text{or } \frac{C_1'}{C_1''} = \frac{\tau_1}{\tau_2}, \text{ where } \tau_2 = C_2R_2 \quad ....(21)$$

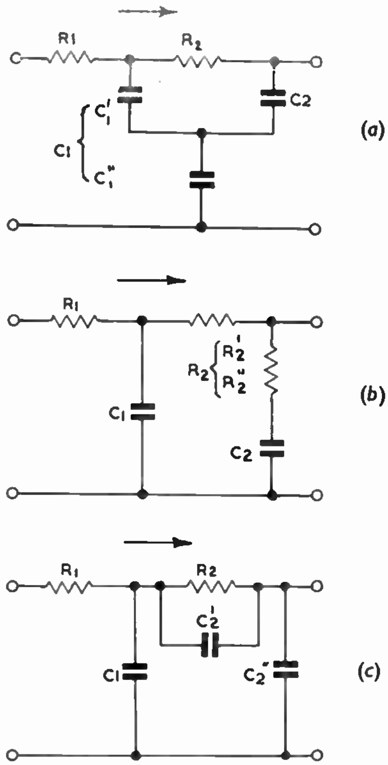

Fig. 4.—Improved integrating networks.

It will be noted that this relationship places no restriction upon the extent to which τ_2 may be increased in order to achieve a close approximation to exact integration together with negligible loading of C_1R_1; for this reason this simple application of the method is of practical value.

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3.2. The alternative arrangement, Fig. 4 (*b*), makes use of the property noted in section 2.3, and is theoretically identical in performance with Fig. 4 (*a*), for equivalent parameters, but with two practical advantages ; firstly, assuming fixed element values, it is cheaper, and secondly, if adjustment is required, it allows of easier setting up, since R_2 may take the form of a potentiometer with the slider forming the output terminal.

Proceeding as in the previous case we obtain

$$e_{R_2''} = R_2'' i_2 \approx \frac{R_2''}{R_2} e_{C_1}, \text{ since } C_2R_2 \text{ is large} \quad \ldots\ldots\ldots\ldots\ldots(22)$$

$$\text{and } e_{C_2} = \frac{1}{C_2}\int i_2 \,.\, dt \approx \frac{1}{C_2R_2}\int e_{C_1} \,.\, dt \quad \ldots.(23)$$

Using equation (9), we must have

$$e_{C_1} = \frac{1}{C_1R_1}\int e_{R_2''} \,.\, dt \ldots\ldots\ldots\ldots\ldots\ldots(24)$$

Substituting the above results in this expression, gives

$$\frac{1}{C_2R_2}\int e_{C_1} \,.\, dt = \frac{1}{C_1R_1}\,\frac{R_2''}{R_2}\int e_{C_1} \,.\, dt \,..(25)$$

which reduces to

$$\frac{\tau_1}{\tau_2} = \frac{R_2''}{R_2}, \text{ using } \tau_1 = C_1R_1, \tau_2 = C_2R_2 \,..(26)$$

3.3. A transformation of practical value arises from the fact that the three capacitative elements of Fig. 4 (*a*) form a T (or Y, or star) which suggests the possibility of deriving an equivalent circuit by applying the T-Π (or star-delta) transformation. The result is shown at Fig. 4 (*c*). This new network is also related to Fig. 4 (*b*) since the right-hand portions of the two networks have potentially identical voltage transfer constants, both correction networks being characterized by the operator :—

$$\frac{\tau^{-1} + kp}{\tau^{-1} + (k+1)p} \ldots\ldots\ldots\ldots(27)$$

For Fig. 4 (*b*) $\tau = C_2R_2''$, $k = R_2'/R_2''$.
For Fig. 4 (*c*) $\tau = C_2''R_2$, $k = C_2'/C_2''$.

3.4. Networks equivalent to the three arrangements just described may be derived by substituting inductors for the resistors, and resistors for the capacitors ; these circuits are applicable to electromagnetic time-bases, as will be shown later.

3.5. The relationship between the networks of Fig. 4 or their LR equivalents and the simple integrator is clearly shown in a comparison of the operational forms of their voltage transfer constants. For the simple integrator we have from equation (10)

$$e = \tau^{-1} \,.\, \frac{1}{\tau^{-1} + p} \,\text{E1}, \text{ where } \tau = CR \ldots.(28)$$

The corrected networks, on the other hand, are all characterized by the form

$$e = \tau_1^{-1}\,\frac{(\tau_2^{-1} + p)}{(\tau_3^{-1} + p)(\tau_4^{-1} + p)}\,\text{E1} \ldots\ldots(29)$$

where the τ_1 are functions of the network constants. Thus, for τ_2, τ_3, τ_4 all large, the expression again approaches $1/p$.

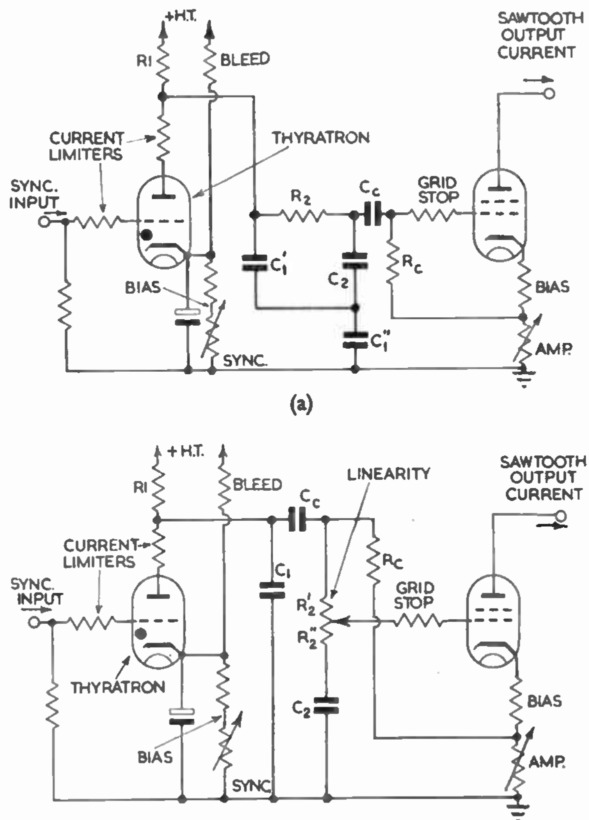
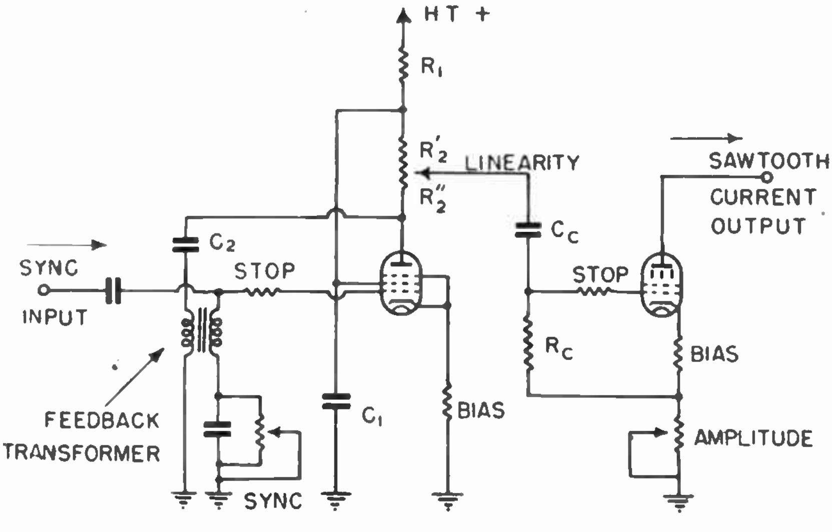
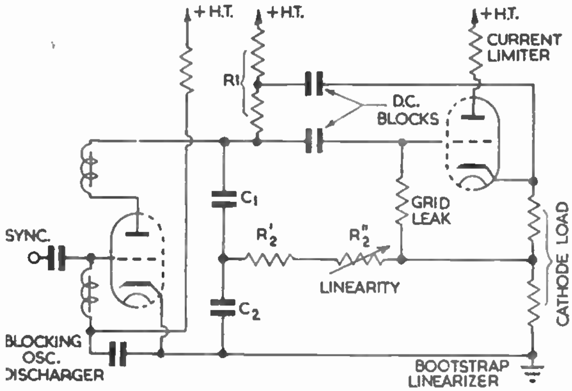


Fig 5 (a) and (b).—Practical sawtooth generators. Note : R_1, R_2, C_1, C_2 are integrator components.



(c)



(d)

Fig. 5.—Application of integrator correcting circuits to practical sawtooth generators.

4. Practical Circuit Arrangements

4.1. While the basic networks incorporating linearity correcting sections which have been discussed in the foregoing sections are not capable of providing a perfectly linear output waveform they are able to give appreciable improvement. In considering the problem of linearizing the output of a time base circuit it should be remembered that perfect linearity in the earlier stages is not essential ; in fact, the only practical method of achieving linearity in the final output of multi-stage circuits, such as are usually employed, is to arrange cancellation of the unavoidable factors giving rise to non-linearity. It is, therefore, desirable firstly that the various causes of non-linearity are comparable in magnitude, and, secondly, that each be readily controllable over an appreciable range. The technique of partial linearity correction of exponential wave sections, by means of the integrating networks at present under discussion, is of practical value in providing a simple means of reducing to reasonable proportions the non-linearity arising in the various stages of the complete time-base system, thereby facilitating the problem of overall cancellation. Moreover, as will appear later, a knowledge of the theoretical basis of the method is of assistance in minimizing distortion throughout the entire circuit.

5.0. Application to Sawtooth Generators

5.1. In considering the degree of non-linearity tolerable in the circuit in which the sawtooth wave is generated, it should be realized that, while perfection is futile (due to succeeding distortion), a large drive for the output stage is desirable since it allows the application of adequate negative feedback with corresponding advantages. Accordingly, partial correctors of the kind under discussion are useful in allowing increased output from the sawtooth generator, i.e. for a given degree of non-linearity.

5.2. Four practical time base voltage generators incorporating integrator correcting circuits are shown in Fig. 5. All are suitable for television reception, the first two representing examples of commercial design. The first arrangement, Fig. 5 (*a*), is a straightforward application

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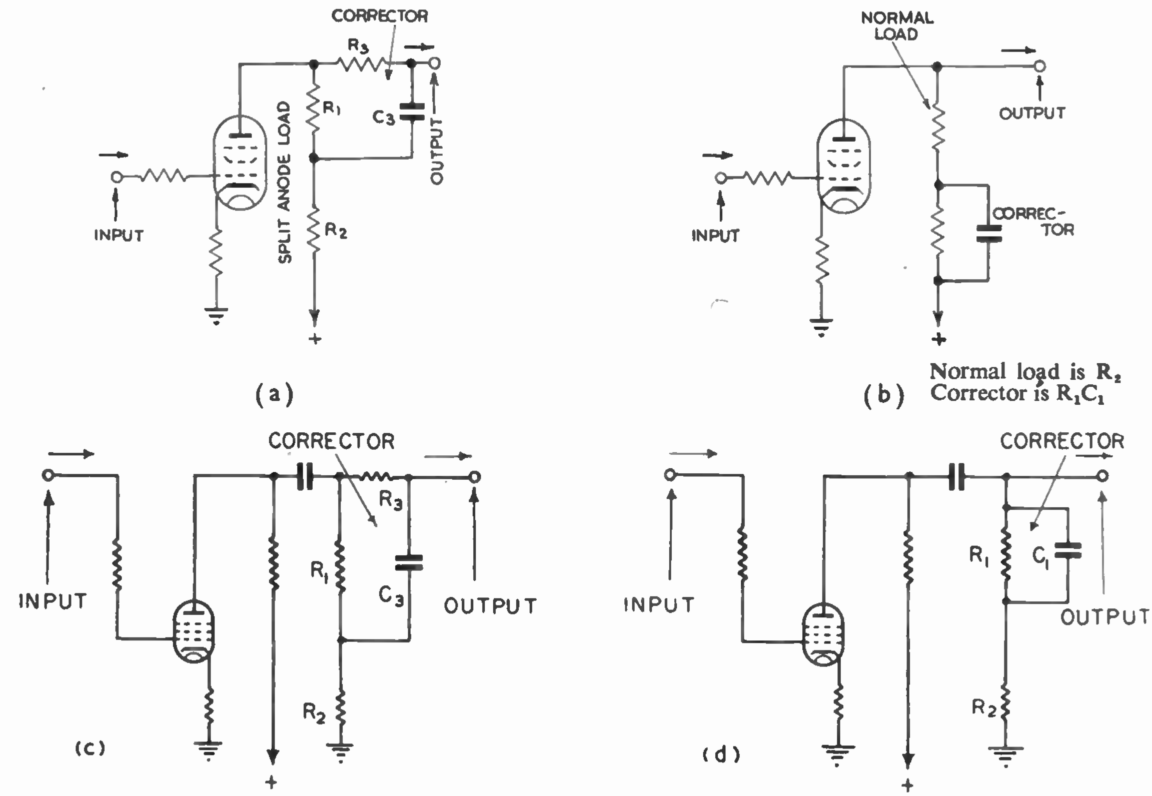

Fig. 6.—Application of integrator correcting circuits to time base voltage amplifiers.

of the network shown in Fig. 3 (*a*) to the thyratron type of scanning oscillator. In the next application (Fig. 5 (*b*)), this time using the basic network of Fig. 3 (*b*), the correcting integrator (C_2, R_2', R_2'') is separated from the main charging " time-constant " (C_1R_1) by the long time constant coupling C_cR_c. The third (Fig. 5 (*c*)) is an original adaptation based on the blocking oscillator type of sawtooth generator. An interesting feature of this circuit is the use of the oscillator valve as a double discharge device, since it discharges C_1 at its screen and C_2 at its anode duiing its brief conducting periods. It will be noted that the circuit branch containing the capacitor C_2 of the correcting circuit also includes the primary of the feedback phase-reversing transformer. This winding may, alternatively, be connected in the screen circuit.

The fourth example shows (see Fig. 5 (*d*)) how an additional CR circuit may be used to augment and control the linearizing influence of the " bootstrap " circuit on an exponential wave generator. Apart from the correcting section C_2, R_2', R_2'', the circuit components will be identified readily since the left-hand triode is operating in a conventional blocking oscillator type discharge circuit, while the right portion of the double-triode (e.g. 6 SN7GT) is connected as a normal bootstrap. The slightly exponential output at the bootstrap cathode is used to feed the integrator corrector; and the correcting wave developed across C_2 is inserted under C_1, whose lower plate would otherwise be connected to H.T.—. Alternatively, C_2R_2 may be fed from the split charging resistor R_1. The bootstrap valve may be anode-loaded to obtain an additional (anti-phase) output. The entire circuit is very suitable for driving the X plates of a small electrostatic C.R.T.

6.0. Application to Time Base Voltage Amplifiers

6.1. The integration method of linearity correction is readily applicable to time base amplifiers ; these usually take the form of two

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high-impedance triode or pentode valves operating in a resistance-capacitance coupled paraphase push-pull arrangement. Two methods of partially correcting a single CR coupled stage are shown at Fig. 6 (*a*), (*b*).

6.2. In the first case, Fig. 6 (*a*), the pentode delivers a current

$$i_a = g_m e_g \quad \ldots\ldots(30)$$

into the anode load, and it will be assumed that

$$e_g = -\mathrm{E}\left(1 - \exp(-\tau^{-1}t)\right),\ t \nless o \quad \ldots\ldots(31)$$

i.e. that the input is an uncorrected exponential voltage change.

The anode load resistor is split into two series-connected portions R_1, R_2, and the first section (R_1) is used to feed an integrator (R_3C_3) of the voltage-operated type. The output of the entire network will be the sum of the voltages appearing across C_3 and R_2, i.e. :—

$$e_o = e_{C_3} + e_{R_2} \quad \ldots\ldots(32)$$

But

$$e_{C_3} \approx \frac{1}{C_3R_3}\int e_{R_1}\,.\,dt$$

$$= \frac{R_1}{C_3R_3}\, g_m\mathrm{E}\int\left(1 - \exp(-\tau^{-1}t)\right).\,dt \quad (33)$$

and

$$e_{R_2} = R_2 g_m \mathrm{E}\left(1 - \exp(-\tau^{-1}t)\right) \quad \ldots\ldots(34)$$

Assuming that $r_a \gg z_L$, so that the anode load (z_L) has no effect on the form of the output current, and that C_3R_3 is large enough to ensure accurate integration, we must have from equation (9)

$$e_{C_3} = \tau^{-1}\int e_{R_2}\,.\,dt \quad \ldots\ldots(35)$$

i.e.

$$\frac{R_1}{C_3R_3}\, g_m\mathrm{E}\int\left(1 - \exp(-\tau^{-1}t)\right) dt$$

$$= \tau^{-1}R_2 g_m \mathrm{E}\int\left(1 - \exp(-\tau^{-1}t)\right).\,dt \quad \ldots(36)$$

which reduces to $\frac{R_1}{R_2} = \frac{\tau_3}{\tau_1}$, where $\tau_3 = C_3R_3$ ......(37)

Thus, for the case $\tau_3 = \tau$ (the time constant of the source of e_g), the anode load resistor must be divided into equal parts ($R_1 = R_2$).

6.3. The alternative arrangement of Fig. 6 (*b*) makes use of a current-operated CR integrator. The voltage developed by the latter for an output current of the same form as before is

$$e_{R_1} = e_{C_1} = \frac{1}{C_1}\int i_{C_1}\,.\,dt$$

$$\approx \int i_a\,.\,dt, \text{ since } C_1R_1 \text{ is large,} \quad \ldots\ldots(38)$$

$$= \frac{1}{C_1}\, g_m\mathrm{E}\int\left(1 - \exp(-\tau^{-1}t)\right).\,dt \quad \ldots(39)$$

This correction voltage is added to the voltage developed by the load resistor R_2 :—

$$e_{R_2} = R_2 i_a = R_2 g_m \mathrm{E}\left(1 - \exp(-\tau^{-1}t)\right) \quad \ldots(40)$$

We need $e_{R_1} = \tau^{-1}\int e_{R_2}\,.\,dt$ ......(41)

$$\text{i.e. } \frac{1}{C_1}\, g_m\mathrm{E}\int\left(1 - \exp(-\tau^{-1}t)\right) dt$$

$$= \tau^{-1}R_2 g_m\mathrm{E}\int\left(1 - \exp(-\tau^{-1}t)\right) dt \quad \ldots(42)$$

which reduces to $C_1R_2 = \tau$ ......(43)

It will be noted that this approximate solution arises from the assumption of a high value for R_1. In this arrangement, Fig. 6 (*b*), C_1R_1 forming the corrector, and R_2 the normal load may, of course, be transposed; also, when C_1R_1 is placed on the H.T. side of R_2, C_1 may be returned to H.T.− instead of H.T.+.

6.4. When the amplifier requiring correction is CR coupled to a succeeding stage, the coupling resistor may be operated upon instead of the anode resistor, giving the alternative configurations shown at Fig. 6 (*c*), (*d*).

6.5. The results of this section indicate the effects of anode and cathode decoupling CR circuits in time base amplifiers upon the sawtooth wave being amplified. Clearly the cathode C should be very large to eliminate negative feedback of the uncorrected input, while the decoupling CR may, if too small, modify the values of the correcting CR.

7.0. Application to Current-Output Stages

7.1. Time bases employed in television usually terminate in a power valve which supplies a sawtooth current wave to a split deflector coil. For frame deflection, CR coupling to the coil is usually employed; in the line scanning circuit, however, the coil is invariably fed through a

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step-down transformer. In both cases the resistive component of the deflector coil impedance necessitates a linear rise in the voltage applied to, or developed across, the coil. The output valve may be arranged either as a high-impedance source of current having the correct waveform, or as a low-impedance supply of the desired voltage waveshape, so that the complex load presented by the coil does not distort the coil current appreciably by comparison with the valve input which must be substantially linear. Both methods of operation permit the application of integrator correcting circuits.

7.2. It has already been demonstrated in section 2.4 that the current in a series RL circuit has the same form as the output voltage of the series RC network analysed in section 2.1. The inductive element may, of course, take the form of a deflector coil winding. Accordingly, it is possible to use a deflector coil having two sections in one of the RL equivalents of the linearized CR configurations already discussed, with a consequent reduction in the degree of non-linearity as compared with the conventional method of operating a simple coil.

7.3. The LṘ equivalents of the basic network of Fig. 3 (*a*) and its current-operated counterpart (cf. Fig. 6 (*b*)) are shown in Fig. 7 at (*a*) and (*b*) respectively. Taking the latter case and assuming that the time constants L_1/R_1, L_2/R_2 are both large and that the output valve delivers a current

$$i_o = g_m \, . \, E\left\{ \left(1 - \exp(-\tau^{-1}t)\right)\right\} \quad ..(44)$$

where E and τ refer to the exponential generator providing the diiving voltage at the grid. Then we have :

$$i_{L_1} = i_o = g_m \, . \, E\left\{ 1 - \exp(-\tau^{-1}t)\right\} \quad (45)$$

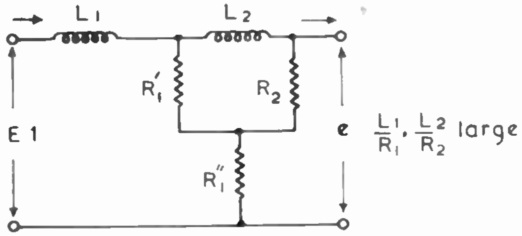
and $$i_{L_2} \approx (R_2 L_2^{-1}) g_m \, . \, E\int\left\{ 1 - \exp(-\tau^{-1}t)\right\} dt \quad \ldots\ldots\ldots\ldots\ldots(46)$$

For linearity correction we need from equation (9):

$$i_{L_2} = (\tau^{-1})\, g_m \, . \, E\int\left\{ 1 - \exp(-\tau^{-1}t)\right\} dt \quad \ldots\ldots\ldots\ldots\ldots(47)$$

which reduces to $R_2 L_2^{-1} = \tau^{-1}$...........(48)

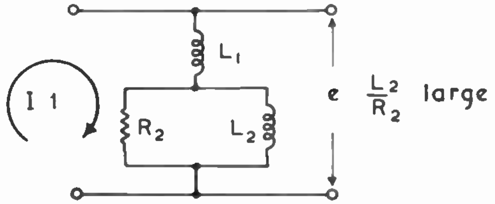
Since L_1, L_2 may both be deflection windings, the principal current and its integral add without additional coupling; and R, in the voltage-driven version of Fig. 7 (*a*), is redundant. On the other hand the two windings will be in close proximity, and mutual coupling will modify the equivalent network but without seriously modifying the nature of the response since, although magnetic deflection assemblies generally employ some form of magnetic circuit, the air gap (tube neck) is inevitably large and the mutual induction consequently restricted.



(a)

Fig 7 (a).—LR equivalent of Fig. 3 (*a*) (*q.v.*).

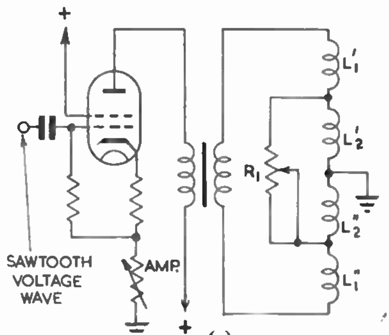
Note.—For current output with L_1, L_2, as deflector coils R_1'' and voltage output terminals are redundant.



(b)

Fig. 7 (b).—Current-operated version of Fig. 7 (a) (cf. Fig. 6 (b)).

Note.—For current output with L_1, L_2 as deflector coils omit output terminals.



(c)

Fig. 7 (c).—Output stage employing circuit of Fig. 7(b).

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7.4. A practical electromagnetic time base output stage employing the basic arrangement just described is shown in skeleton form in Fig. 7 (*c*). It is, of course, only applicable in the line case since frame deflecting coils are largely resistive. Negative current feedback is adopted to raise the output impedance of the valve. It is simpler but less effective to use a single winding and parallel resistor, with the latter adjusted to the transitional condition (c.f. the CR case—Fig. 2 (*b*)) between the two extremes represented by the exponential output current and its integral. In any case the value of R required is not usually comparable with the value needed for damping and a flyback-operated diode-switched damper is desirable.

It is assumed, of course, that the step-down transformer introduces little distortion and that the output resistance (r_a) of the driving valve is sufficiently high with respect to the effective inductance (L_{eff}) of the entire output that the time constant L_{eff}/r_a is small enough to avoid modification of the output current wave. These conditions must, in any case, be satisfied if a high degree of linearity is to be achieved. Where the flyback damping is not sufficient to substantially suppress the flyback transient by the commencement of the next sweep, as when, in " reactance " or " resonance " scanning, the transient is purposely allowed to persist in order to achieve greater efficiency, the arrangement described will be effective only during the latter part of the sweep. Under such conditions other methods of linearity correction are employed.

8.0. Conclusion

8.1. It is felt that the foregoing treatment of the method will be sufficient to provide an adequate basis for the development of practical circuits of which a great variety are feasible ; but it is hoped to treat the latter aspect in a subsequent paper, at least in the case of electromagnetic time bases.

8.2. This paper was written in 1947 and is published by permission of RF Equipment, Ltd., Langley Park, Bucks, and of *Electronic Engineering*, in which journal a modified version has already appeared.[7] The author is indebted to Mr. O. S. Puckle and Dr. H. Moss for criticism of the manuscript and suggestions for its improvement.

9.0. References

1. Murphy Radio Ltd. and G. F. Hawkins, British Patent Specification No. 511,600 (1937).
2. " Television Topics : Linear Sawtooth Oscillator," *Wireless World*, May 4th, 1939, p. 425.
3. W. T. Cocking, " Television Receiving Equipment." pp. 103-5, Iliffe (1940).
4. O. S. Puckle, "Time Bases," p. 87, Chapman and Hall (1943).
5. O. S. Puckle, ibid. Appendix iv, pp. 178-183, " Differentiating and Integrating Circuits."
6. A. C. Clarke, " Linearity Circuits," *Wireless Engineer*, June, 1944, pp. 256-266.
7. A. W. Keen, "Television Time Base Linearisation," *Electronic Engineering*, June, 1949, pp. 195-198, 223.
8. A. W. Keen, " Television Time Base Linearization," *Proc.I.R.E.*, Jan., 1949, p. 61.

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NOTICES

Obituary

Council regrets to record the death of Major Patrick Joseph BEST, recently living in New Longton, Preston.

Prior to the recent war, Major Best was employed in the Post Office Engineering Department in the Lancashire area and was engaged on radio work. Immediately on the outbreak of war, he went overseas and secured rapid promotion to the rank of Major. He was appointed a Member of the Most Excellent Order of the British Empire in the January 1946 Honours List.

In addition to his professional work, Major Best was also a very keen member of the amateur radio organization. He was elected an Associate Member of the Institution in 1941.

Television Extension Screen

It has recently been announced that Mr. J. C. G. Gilbert (Member) and Mr. R. S. Roberts (Member), both Telecommunication lecturers at the Northern Polytechnic, were joint designers of a Television Extension Screen which was recently exhibited by a well-known firm of radio manufacturers at Radiolympia.

It is hoped to publish in a subsequent issue of the Journal a short technical description of the apparatus and techniques involved.

Mr. R. S. Roberts, besides being a member of the Education and Examinations Committee, is also the senior examiner in Part III of the Graduateship Examination (Advanced Radio Engineering).

Radiolympia

It has been announced that the total paid attendance at Radiolympia was 395,465. The attendance in 1947 was 440,320 which was an all-time record. This compares with a total attendance of 144,363 in 1938.

The exhibition was undoubtedly a great success and was attended by a large number of overseas buyers.

The main emphasis of course was on television, and a number of projection models were on view. Also a very interesting exhibition of radar and electronic apparatus was displayed by the Services and the D.S.I.R.

BBC Staff Changes

The B.B.C. has decided to split the Operations and Maintenance Department of the Engineering Division into two parts. One part, consisting of the Studio, Transmitter, Recording and Lines Departments, will continue to be in charge of Mr. L. Hotine, Senior Superintendent Engineer.

Mr. M. J. L. Pulling, M.A., has been appointed to take charge of the other part, the Television Department, with the title of Senior Superintendent Engineer, Television.

Mr. D. C. Birkinshaw's post as Superintendent Engineer, Television, is not affected.

Television Servicing Certificate Examination

The Radio Trades Examination Board and the City and Guilds of London Institute announce that the first Television Servicing Certificate Examination will be held on May 2nd and 4th, 1950, and the practical examination on a date to be announced later. In this instance London will be the only centre and it may, therefore, be necessary to restrict the number of entries. Admission to the examination is limited to candidates who have passed one of the following examinations :—

(i) The Radio Servicing Certificate Examination held jointly by the City and Guilds of London Institute and the Radio Trades Examination Board,

(ii) the examination in Radio Servicing held by the Radio Trades Examination Board from 1944 to 1946,

(iii) the examination in Radio Service Work of the City and Guilds of London Institute held from 1938 to 1947,

(iv) the " third year " examination in Radio Service Work of the Union of Lancashire and Cheshire Institutes,

(v) the examination in Radio Servicing held prior to 1944, by the British Institution of Radio Engineers,

(vi) the examination in Radio Servicing held prior to 1944, by the Scottish Radio Retailers' Association.

The closing date for the examination is THURSDAY, DECEMBER 15th, 1949. Regulations and entry forms are obtainable on application to the Secretary of the Board, 9 Bedford Square, London, W.C.1.

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