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ELEMENTS
OF PULSE CIRCUITS

Elements of Pulse Circuits

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PREFACE

THIS book is addressed primarily to physicists and research workers who wish to obtain an introduction to pulse circuits. It is assumed that the reader is already familiar with radio valves and elementary receiving technique, and accordingly the fundamentals of radio practice are either taken for granted or reviewed briefly: the application to pulse waveforms is then tackled immediately.

Although mathematical statement is used occasionally in the interests of brevity and precision, the approach is mainly non-mathematical, the emphasis being on a direct understanding of the physical principles involved. It is hoped that the book will be of service to radio workers generally and, therefore, while connections are made with advanced physics they are never essential to the argument. In particular certain topics are omitted because the general reader will not have the equipment to deal with them (notably the use of transmission lines for pulse shaping).

In an introductory volume such as this it is impossible to give a detailed acknowledgement to all sources of information. No attempt is made to trace circuits to their source and references are given only for the benefit of the reader who wants to pursue the subject further. I would, however, like to express here my thanks to all those who have contributed either in conversation or in print to my studies and therefore to this book. In particular I gratefully acknowledge the assistance received in composing the manuscript from my colleagues in Auckland Mr. J. B. Earnshaw and Dr. H. A. Whale.

F. J. M. FARLEY.

Harwell,
Nov., 1955

CONTENTS

CHAPTER I

BASIC CONCEPTS

1.1. Pulse waveforms. In radio communication the usual practice has been to consider only sinusoidal waveforms. We are well aware, of course, that a typical speech waveform is far from sinusoidal; the simultaneous presence of several

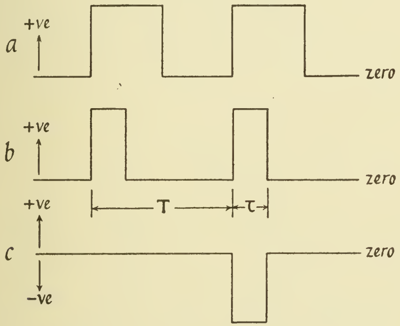


FIG. 1.—Pulse waveforms.

different frequencies makes the wave deviate from the pure sine-wave shape. Nevertheless, in analysing circuit performance it is often sufficient to confine the discussion to sine-waves. The sine-wave has been chosen as the basic waveform for circuit analysis because it has the property of passing through linear electrical networks without changing its shape. We shall see below, when we consider other waveforms, that in general their shape is changed by a linear network.

In pulse circuits we are not interested so much in the fundamental frequency of a wave and its harmonic content, if any; our attention is concentrated mainly on the exact shape of the wave and its variation from point to point in the circuit. The basic waveform is now the *square-wave* (Fig. 1). In the ideal square-wave the voltage changes in an infinitely short time from one steady level to another: that is, the wave has a perfectly flat top and bottom and infinitely steep sides. Fig. 1*a* shows a symmetrical square-wave, while in Fig. 1*b* we see an asymmetrical form in which the positive voltage regime is of shorter duration. This latter is often regarded as a square (or rectangular) *pulse* of amplitude V

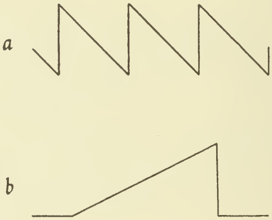


FIG. 2.—Triangular or saw-tooth waveforms.

and duration τ, rising from a steady *base-line*. In this example, the pulse recurs regularly after a time interval T, called the recurrence period. The quantity τ/T is called the *mark-space* ratio: thus a symmetrical square wave would have a mark space ratio of $\frac{1}{2}$. In many cases, however, the pulses are not regular, but occur singly, or at irregular intervals. It is often convenient, therefore, to discuss the response of a circuit to a single square pulse, which may, of course, be positive as in Fig. 1*b*, or negative as in Fig. 1*c*.

Another common pulse waveform is the triangular wave, Fig. 2*a*, often called the saw-tooth, or time base, waveform. If this wave is applied to the X-deflection plates of a cathode ray tube, it causes the spot to sweep across the screen at a uniform rate, and then to fly back infinitely quickly (in the ideal case) to repeat the process. Thus, we may use the cathode ray tube to plot the shape of another waveform applied to the Y-deflection plates. In such applications it is important that the time base waveform be exactly linear: methods of generating and handling linear saw-toothwaves will be discussed in Chap. V. Fig. 2*b* shows another common type of triangular wave.

Both the square and the triangular waves have a large harmonic content. In the ideal case of infinitely sharp

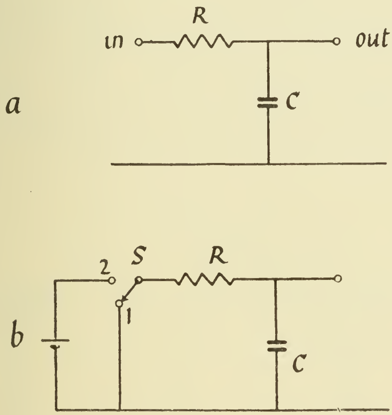


FIG. 3.—Integration circuit.

corners, harmonics of infinite frequency must be present. In practice there is some rounding of the corners, but it remains true that our circuits must handle a wide range of frequency if the waveform is to be transmitted at all faithfully. Attenuation of the higher frequencies produces a rounding of the corners; whereas attenuation of the lower frequencies results in distortion of the base-line or nonlinearity of a saw-tooth.

Postponing the general discussion of these effects to Chap. VI, we now consider the effect on the pulse shape of two simple resistance-condenser combinations.

1.2. Integration. Suppose a positive square pulse of amplitude V is applied to the circuit of Fig. 3*a*. To determine the output waveform we may suppose that the pulse has been produced from a battery of voltage V, by the operation of the switch S, as shown in Fig. 3*b*. Initially, with the switch in position 1, the output voltage will be zero. When the switch is moved to position 2, current i flows through resistance R, through condenser C, and back to the battery. The result

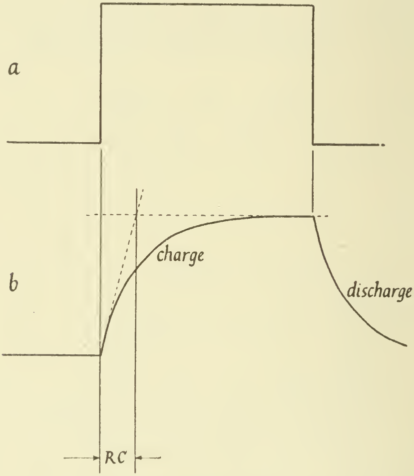


FIG. 4.—Integration of a square pulse.

is that C is slowly charged and the output voltage, v, slowly rises. The differential equation for the output voltage is:

$$\frac{dv}{dt} = \frac{i}{C} = \frac{V - v}{RC} \quad . \quad . \quad . \quad (1)$$

Hence, $\log (V - v) = - t/RC + \text{const.}$

The condition, $v = 0$ when $t = 0$, shows that the constant of integration is $\log V$, and we obtain finally

$$v = V(1 - e^{-t/RC}) * \quad . \quad . \quad (2)$$

* Using the expansion series for the exponential this gives $v \simeq Vt/RC$ at the beginning of the wave where $t \ll RC$.

The output voltage therefore rises linearly at first, but then more slowly, eventually reaching the steady value V, equal to the input voltage. This is the well-known *exponential* charging of a condenser, and is illustrated in Fig. 4*b*.

The time scale of this output waveform is entirely determined by RC, called the *time constant* of the circuit. The product of resistance and capacity is a quantity of dimensions *time* ; if the resistance is in ohms and the capacity in farads the product gives the time in seconds. It is useful to remember that the output voltage rises initially at the rate V/RC; that is as if to reach the final voltage after time RC. (See equation (1).) The percentage of the final voltage reached, and remaining, after various times is given in Table 1.

TABLE 1

Time in Units RC	*Voltage change per cent.*	*Subsequent voltage change per cent.*
0	0	100
0·2	18·1	81·9
0·5	39·4	60·6
1	63·2	36·8
2	86·5	13·5
3	95·0	5·0
4	98·2	1·8
5	99·3	0·7
7	99·9	0·1

When switch S is returned to position 1 (corresponding to the end of the input pulse), the condenser C discharges through resistance R until the output voltage, v, is again zero. If we solve the differential equation as before we find

$$v = Ve^{-t/RC} \quad . \quad . \quad . \quad . \qquad (3)$$

This waveform is shown in Fig. 4*b*. We see that it is similar in shape to the positive going waveform at the beginning of the output pulse, and Table 1 applies in this case also.*

The net result is that the ideally square input pulse of Fig. 4*a* is transformed at the output to the shape of Fig. 4*b*. We have so far supposed that the input pulse duration is much

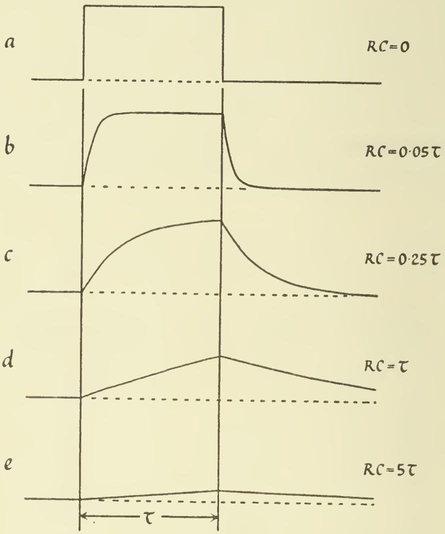


FIG. 5.—Integration of a square pulse by various time constants.

larger than the time constant *RC*. If these times are comparable, the discharge phase sets in before the final positive voltage is reached. Fig. 5 shows the output waveforms for a fixed input pulse and various values of the time constant *RC*. We see that the distortion is negligible when *RC* is very short, and increases progressively as *RC* is increased.

* This is an example of the general rule that for linear networks positive and negative pulse fronts produce effects which are the same, but inverted; and the end of a positive pulse can be regarded as an isolated negative going pulse front. The symmetry between positive and negative disappears, however, when valves and other non-linear elements are included in the circuit.

When the time constant is very long, the peak output voltage is proportional to the area under the input pulse, that is, to V. This follows from equation (1) which shows that

$$\frac{v}{t} = \frac{V}{RC} \text{ for } v \ll V. \quad . \quad . \quad . \quad (4)$$

$\therefore v = V\tau/RC$ at the end of the pulse if $\tau \ll RC$.

Physically, the result follows because the condenser C is charged at a rate proportional to V and for a time τ. More generally, for an input pulse of arbitrary shape, the peak output voltage is proportional to the area under the pulse, provided always that $RC \gg$ pulse length.

Because of this property the circuit of Fig. 3*a* is called an *integrating* circuit. The whole process, resulting in the whole series of output waveforms in Fig. 5, is known for convenience as *integration*. It is important to realize that exact mathematical integration is approached only in the extreme case of Fig. 5*e*. In the more typical case of Fig. 5*b* there is no connection with mathematical integration, but we still use this term for convenience.

We now consider the effect of the integrating circuit on a triangular wave. Suppose that the input to the circuit of Fig. 3*a* has the form $V = kt$. The differential equation for the output voltage v then becomes

$$\frac{dv_c}{dt} = \frac{kt - v_c}{RC} \quad . \quad . \quad . \quad . \quad (5)$$

or, in standard form

$$\frac{dv}{dt} + \frac{1}{RC}v = \frac{k}{RC}t.$$

Multiplying both sides by $e^{t/RC}$,

$$\frac{d}{dt}(v \,.\, e^{t/RC}) = \frac{k}{RC}te^{t/RC}.$$

Therefore,

$$(ve^{t/RC}) = k\int_0^t td(e^{t/RC}).$$

which on integration by parts yields

$$v = kt - k \,.\, RC(1 - e^{-t/RC}) \quad . \quad . \quad (6)$$

This behaviour is illustrated in Fig. 6. Initially dv/dt is zero, but after a transition period of order a few times RC, the

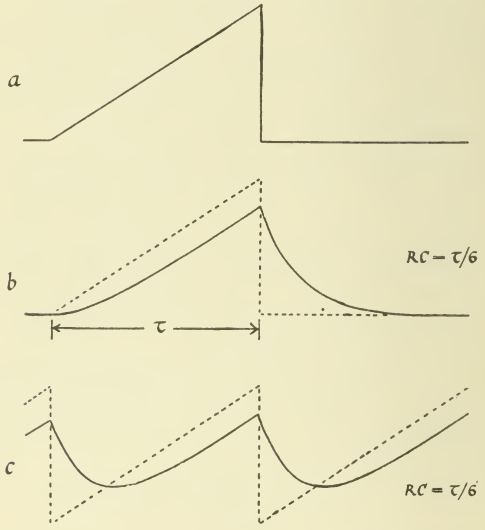


FIG. 6.—Integration of a saw-tooth.

output voltage (Fig. 6*b*) rises at the same rate as the input voltage (Fig. 6*a*), but lags by the constant amount $k \,.\, RC$; that is by the amount the input saw-tooth rises in time RC. In effect, therefore, the output voltage lags in time by the interval RC.

$$v = k(t - RC) \quad . \quad . \quad . \quad . \quad (7)$$

Physically, this happens because a steady current $C\dfrac{dv}{dt}$ is needed to charge condenser C at the constant rate, and this current is provided by the voltage drop $RC\frac{dv}{dt}$ across resistance R.

In Fig. 6*a* we show a typical input triangular waveform, and Fig. 6*b* gives the corresponding output. At the end of the sweep the input voltage returns to zero and the output decays exponentially as in the case of the square pulse. In the case of a recurrent saw-tooth the exponential decay at the end of one saw-tooth has to join smoothly to the lagging rise of the following saw-tooth with the result shown in Fig. 6*c*. Note that dv/dt is zero at the output when the input and output voltages are equal because then there is no current through R. In all cases, the integrating effect of the circuit produces departures from linearity at the beginning of the time base for the duration of several time constants.

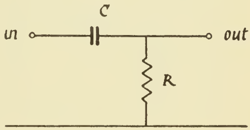


FIG. 7.—Differentiation circuit.

1.3. Differentiation. We now consider the circuit of Fig. 7 which differs from Fig. 3*a* in that condenser and resistance have been interchanged. Instead of observing the voltage across the condenser, we are now interested in the voltage across the resistance.

We can proceed as before by solving the differential equation for the output voltage but it is simpler to observe that (voltage across R) + (voltage across C) = input voltage. The voltage across C is already known from our work on the integrating circuit (see equations (2) and (3)) and we obtain for a square wave input the result

$$\left.\begin{aligned} v &= Ve^{-t/RC} \text{ at the front of the pulse} \\ v &= -\ Ve^{\ t/RC} \text{ at the tail} \end{aligned}\right\} \quad . \qquad (8)$$

The input and output waveforms are plotted in Figs. 8*a* and *b*.

Physically we can explain this behaviour by noting that the voltage across a condenser cannot in general change instantaneously. It can only change when the charge changes, and this usually happens gradually as current flows through the condenser. This means that rapid changes in voltage are transmitted by a condenser without attenuation: as far as rapid voltage changes are concerned, we can regard the condenser as a direct connection. This is an important principle which we shall use again and again in analysing pulse circuits.

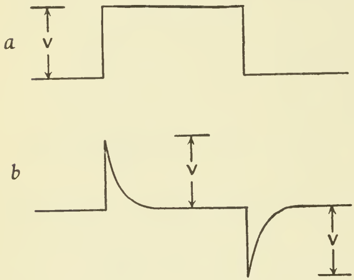


FIG. 8.—Differentiation of a square pulse.

In the present case, the steep front of the input pulse is transmitted completely by the condenser and appears undiminished at the output. The output voltage causes current to flow through R: this current must flow also through C, so that the condenser is charged and the output voltage gradually returns to zero. On the tail of the output pulse the action is similar. As before the time scale is determined by the time constant RC, and Table 1 again applies.

The output pulse obtained with a fixed input pulse and various values of the time constant RC is shown in Fig. 9. Here the pulse is modified only slightly if $RC \gg \tau$; it is the short time constants that give distortion. In the extreme case $RC \ll \tau$, the output approximates to the mathematical differential coefficient of the input waveform. For this reason the process is known as *differentiation*. Here again, we must regard the word as a technical term applying to the whole family of distortions; the pure mathematical meaning should be kept well in the background.

Let us now consider the *differentiation* of a triangular wave, $V = kt$. We can obtain the output voltage by subtraction as before, using equation (6). This yields

$$v = k \, . \, RC(1 - e^{-t/RC}) \quad . \quad . \quad . \quad (9)$$

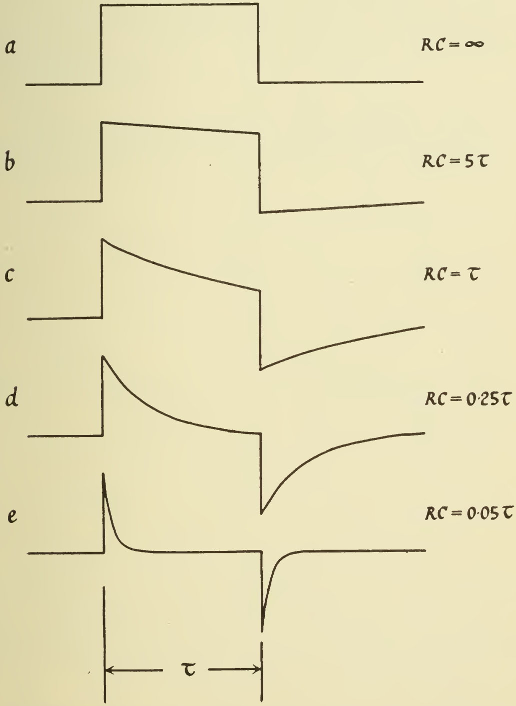


FIG. 9.—Differentiation of a square pulse by various time constants.

Figs. 10*a* and *b* show the input and output waveforms. The output voltage rises initially at the same rate as the input voltage, but after a time of order RC settles down to the steady value $k \, . \, RC$. Physically, the charging of the condenser at constant rate requires a current $C \frac{dv}{dt}$ flowing through it; this current flowing also in the resistance develops the steady voltage $RC\frac{dv}{dt}$ across it. At the end of the triangle

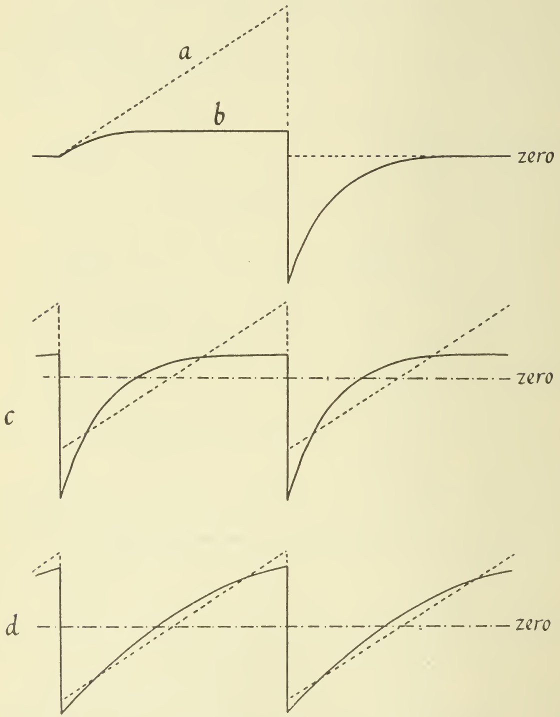


FIG. 10.—Differentiation of a saw-tooth.

the steep front is transmitted to the output in full, and is followed by a typical exponential tail. In the case of the recurrent saw-tooth this exponential tail joins smoothly to the initial transient of the subsequent cycle, as shown in

Fig. 10*c*. The input and output waveforms are parallel when the output crosses zero.

A common case is that of slight differentiation, as shown in Fig. 10*d*. The result is that the time base becomes non-linear unless the time constant RC is very much longer than the recurrence period of the saw-tooth: an important practical consideration.

1.4. Current flows *through* a condenser. Before proceeding further it is desirable to discuss the common fallacy

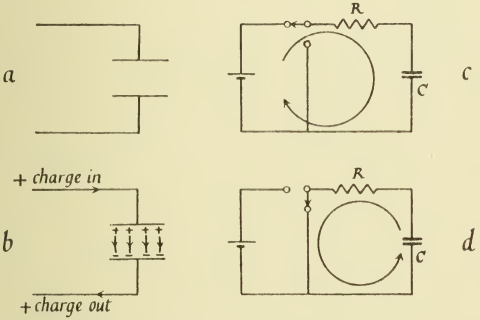


FIG. 11.—Current flows *through* a condenser.

that current flows *into* a condenser rather than *through* it. Fig. 11*a* shows an uncharged condenser. When the condenser is charged as in Fig. 11*b* positive charge flows in to the upper plate, and produces lines of force in the dielectric which must end on corresponding negative charges on the lower plate. These negative charges are those left by the positive charge flowing out of the lower plate. The net effect then is that the positive charge appears to flow *through* the condenser, although there is no direct connection. It is the charge which has flowed *through* the condenser that appears in the fundamental relation $V = q/C$.

It is important also to remember that all currents flow in closed paths. Any current flowing out of the positive pole of a generator must eventually find its way back to the negative pole of the same generator: otherwise we would have an accumulation of charge in some part of the circuit. A very simple case is illustrated in Figs. 11*c* and *d*. Here current from the battery flows through the resistance *R*, *through* the condenser *C*, and back to the battery. On throwing the switch, the current flows out of the condenser, through the resistance, and back again to the condenser. This remains true even when some point of the circuit is connected to earth. In analysing pulse circuits it will often be found helpful to follow these current paths round their complete circuit.

Note that, as far as varying signals are concerned, the behaviour of a condenser is independent of any steady voltage which may exist across it. In pulse circuits we are interested in the way the voltage across the condenser changes in response to an applied current, and the result is independent of the initial conditions. It is therefore irrelevant to inquire what is the total charge on the condenser and to decide whether the condenser is being charged (total charge increasing) or discharged (total charge decreasing). We adopt the view that a condenser is being charged whenever current flows through it, irrespective of direction, and use the term discharge only where it is convenient to indicate a reversal of the initial flow.

1.5. Thévenin's theorem. This theorem is introduced here because it allows us to make great simplifications in our thinking about pulse circuits. The idea is that any circuit which is acting as a source of signals can be replaced, as far as the subsequent circuits are concerned, by a simple signal generator in series with an impedance Z_0. This is illustrated in Fig. 12. The e.m.f. of the signal generator is equal to the voltage observed across the terminals of the generating circuit when there is no external load: it may be a steady or alternating voltage, or a pulse waveform. The impedance Z_0 is the impedance observed looking into the terminals when the internal signal generator ceases generating; it is often called the *output impedance* of the circuit. Strictly, the theorem is restricted to linear networks, but with discretion it may be applied usefully to valve circuits which are far from linear. The proof of this theorem is straightforward, and may be found in texts on radio and network theory.

The concept of output impedance is important. The output impedance is usually a pure resistance, and its value determines the current that the circuit can supply to an external load. It thus controls not only the variation of the output voltage with load, but more significantly, the rate at which stray capacities on the output can be charged. The

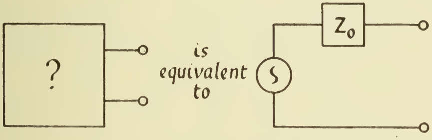


FIG. 12.—Thévenin's theorem.

stray capacity combines with the output impedance to give a standard integrating circuit, and produces the effects described in § 1.2. To estimate these effects we must know the value of the output impedance and Thévenin's theorem provides a simple rule for calculating it. It is simply the impedance we would measure looking back into the circuit. Examples of this will occur throughout the text.

As an illustrative application of Thévenin's theorem consider a simple potentiometer loaded with capacity as shown in Fig. 13*a*, and suppose that a square pulse of amplitude V is applied to the input. To calculate the output pulse, we divide the circuit along the line AA and replace the left hand side by a pulse generator giving amplitude $\frac{R_2}{R_1 + R_2}V$ and having as output impedance R_1 and R_2 in parallel, Fig. 13*b*. The circuit has an integrating action with time constant $\frac{R_1 R_2}{R_1 + R_2}C$. It follows that the pulse will be undistorted when the slider is at either end of the potentiometer, but that distortion will occur in intermediate positions. The maximum effective time constant occurs when $R_1 = R_2$, and is $\frac{1}{4}RC$, R being the total resistance. This example indicates that a simple potentiometer is not usually suitable for attenuating pulses; distortion will nearly always be introduced because of the stray capacity at the output.

The unwanted stray capacity between circuit elements, and from the circuit to earth, is an important consideration in the

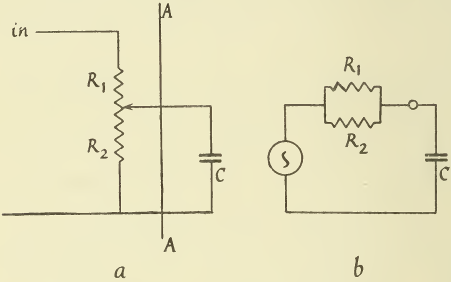


FIG. 13.—Integration by a simple potentiometer.

design of pulse circuits. The effect can always be estimated if we use Thévenin's theorem and the concept of output impedance. Replacing the circuit by a generator in series with this impedance we obtain a standard integrating circuit which will give the effects already discussed above. The presence of stray capacity implies that integration effects are inescapable so that, in practice, the corners of a pulse are always somewhat rounded. A major consideration in any design is to keep this effect within reasonable limits. This means that the output impedance of each circuit must be kept low by using small anode loads and special devices such as the cathode follower (see § 2.5).

CHAPTER II

FUNDAMENTAL VALVE CIRCUITS

In this chapter we consider some fundamental effects obtained when square pulses are applied to diodes, triodes and pentodes, and introduce some devices which are widely used in pulse circuits.

Diodes can be used in two distinct ways. If connected *in series* with the signal path, the diode transmits only a portion of the input pulse. If connected in parallel, however, the diode tends to short circuit the signal for part of the time, producing distortions which are considered in detail below.

2.1. Series diode. It is usually sufficient to regard a diode simply as a device which conducts current in one direction, and not in the other. The forward resistance is generally in the region 100 to 1,000 ohms and can often be neglected. The backward resistance in the case of a thermionic diode can be taken as infinite; but in the case of the crystal diode it has to be allowed for more carefully.

In the circuit of Fig. 14*a*, a generator of positive pulses is connected to a diode in series with a load resistance R, and the anode of the diode is negatively biased. Fig. 15*a* shows a typical input pulse, and Fig. 15*b* gives the corresponding output. Only that part of the pulse which exceeds the bias voltage is transmitted, because the diode does not conduct when its anode is negative. When the bias voltage is exceeded, however, input and output voltage are equal provided $R \gg 1{,}000$ ohms.

Fig. 14*b* shows a similar arrangement, but here instead of negative bias to the anode, we apply positive bias to the cathode by means of resistances R_2 and R_3. The action of the circuit is the same as before.

Figs. 14*c* and *d* show similar circuits, but now the input pulse is applied to the anode via the coupling network C_1R_1.

If the time constant is longer than the pulse length, the pulse will reach the anode without differentiation, but the steady voltage level of the base line will be determined by the voltage applied to the lower end of R_1. Again the output voltage will be as given in Fig. 15*b*.

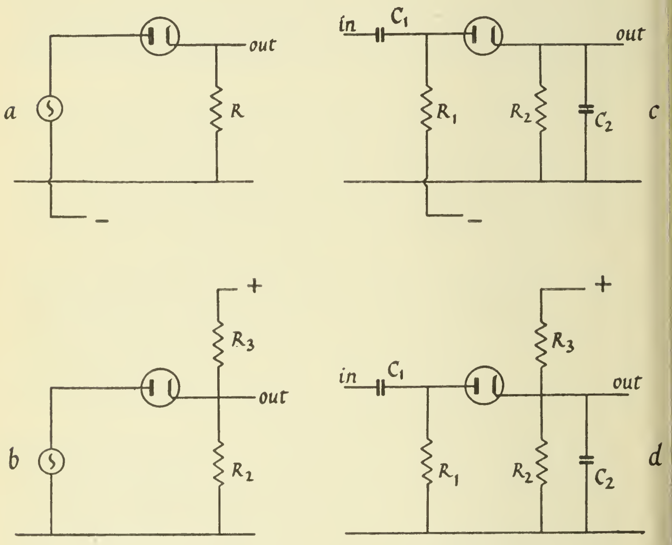


FIG. 14.—Series diode circuits.

The effect of adding capacity to the output lead of the diode, as represented by condenser C_2 in Figs. 14*c* and *d*, is shown in Fig. 15*c*. During the positive going front the condenser is charged rapidly through the low forward resistance of the diode; the output voltage follows the input pulse closely. But on the negative going tail of the pulse the condenser can discharge only through R_2, and this gives a lengthened exponential tail. Stray capacity always produces this effect to some extent; and it can be increased deliberately by adding the condenser C_2, if the circuit is to act as a *pulse lengthener*.

In circuits such as Figs. 14*c* and *d* it is desirable to have $C_1 \gg C_2$, and $R_1 \ll R_2$. The steady voltage level on the input side is then not appreciably altered by the current drawn through the diode. More generally, in fact, the pulse should be fed to the diode from a circuit of low output impedance.

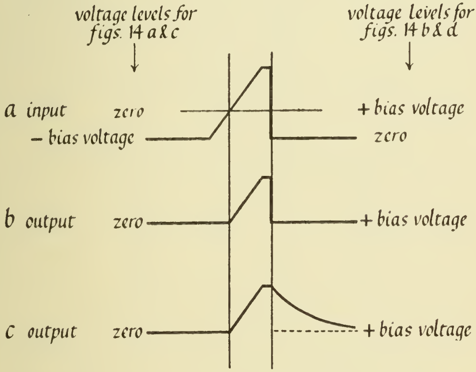


FIG. 15.—Waveforms in series diode circuit.

Exactly the same effects are produced with negative pulses, if the diode and bias polarities are reversed.

This circuit is an example of a *discriminator*, that is, a circuit which transmits only the upper portion of a pulse. In common with other discriminator circuits (§§ 4.3, 7.9) it may be used (i) to select signals of amplitude greater than a predetermined threshold value, (ii) to eliminate small spurious signals and undesirable wriggles of the base line, and (iii) (with condenser C_2 added) as a *pulse lengthener*.

2.2. Parallel diode. Fig. 16*a* shows an *RC* coupling network (for example, an intervalve coupling), with a diode connected in parallel with the resistance. This is a very common circuit, because the grid and cathode of an amplifying valve can behave as a diode: the effects to be described may therefore be produced whenever a pulse is fed through an RC coupling to the grid of a valve.

We will consider what happens when a train of positive pulses is applied to the input, assuming first that the time

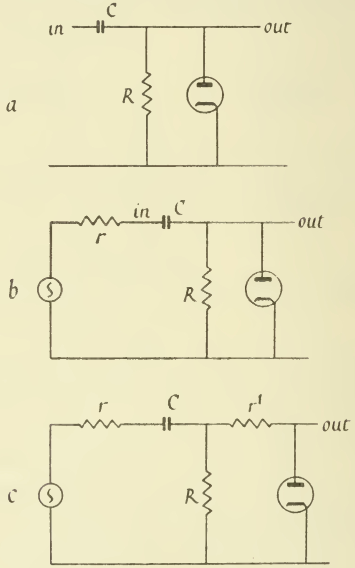


FIG. 16.—Parallel diode circuits.

constant RC is much larger than the recurrence period. The action is similar to that of the peak rectifier in radio communication. In the absence of the diode the steady voltage level at the output would adjust itself so that the mean voltage was zero, as shown in Fig. 17*a*. The base line would be depressed slightly below earth potential, and the pulses would swing well positive. When the diode is present, however, it conducts strongly whenever the anode is positive, and builds up a charge in the condenser C which leaks away only slowly through resistance R. This builds up a negative voltage on the diode side of the condenser, and a steady state is reached, with the pulses just rising to zero voltage, and the base-line correspondingly negative (Fig. 17*b*). The charge on the condenser is almost constant, the diode conducting sufficiently during the pulse to make up any charge which has leaked away in the preceding interval. Whatever the size of the pulses, or the mark-space ratio, the base-line is

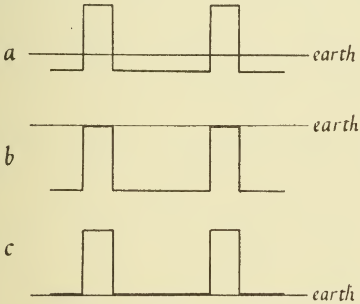


FIG. 17.—D.C. restoration by parallel diode.

automatically adjusted by the diode, so that the top of the pulse just reaches zero voltage. The diode is said to act as a *d.c.-restorer* or *diode clamp:* it restores the d.c. level which has been lost in transmission through the RC coupling network; it clamps the potential-during-the-pulse at zero voltage.

If the diode is reversed, it is the base line which is clamped at zero voltage, as shown in Fig. 17*c*, and the pulses rise positive. In this case we are d.c.-restoring on the base line. The d.c.-restorer is often used when pulses of variable mark-space ratio are encountered. In the absence of the diode the base-line would be depressed as shown in Fig. 17*a*, but by a variable amount, and this variation could upset the operation of subsequent circuits. By introducing a d.c.-restorer, however, we provide standard conditions independent of pulse width and recurrence frequency.

Consider now the situation when the time constant RC is shorter than the recurrence period. In this case, we shall have to take account of the output impedance r of the pulse

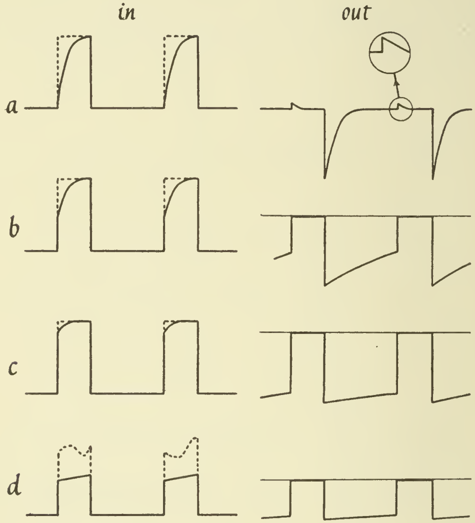


FIG. 18.—Waveforms in parallel diode circuit.

generator, as indicated in Fig. 16*b*. With a short coupling time constant, it is impossible to build up a steady bias voltage, and the diode conducts heavily on each pulse. This loads the pulse generator so that both the input and the output pulses are distorted. Fig. 18*a*, *b* and *c* show the input and output pulses for several values of RC; and for a fixed value of rC, shorter than the pulse length. The generated input pulse (i.e. the input pulse when there is no load) is shown dotted.

In Fig. 18*a*, RC is short and therefore condenser C has completely discharged by the time the pulse arrives. The diode anode is therefore driven positive by the pulse, but only slightly because the diode conducts and draws current through r (potentiometer action between r and the diode forward resistance, d). Because the voltage across C cannot change instantaneously the small positive step must appear also at the input. The input voltage cannot rise immediately to the full pulse height, but does so exponentially as C is charged by the current flowing through r, through C, and through the diode, The time constant of this charging is $(r + d)C \simeq rC$, if $r >$ diode forward resistance, d. During the pulse the output voltage remains nearly constant, though it must drop slightly as the current through the diode gradually drops to zero.

At the end of the pulse, the input voltage drops to zero, and the output is driven negative by an equal amount. Current now circulates through r, C and R, discharging the condenser and giving the exponential recovery on the output, time constant $(R + r)C \simeq RC$. Because $R \gg r$, the effect of this current on the input waveform is usually negligible.

In Fig. 18*b*, time constant RC is longer, and the condenser has not completely discharged when the next pulse arrives. There is therefore a larger positive step as the output voltage rises to zero, and a correspondingly larger positive step at the input. But when the diode conducts, the input voltage must again follow an exponential curve towards its ultimate value. In Fig. 18*c*, time constant RC is still longer, but there is a small leakage of charge through R in the interval, and this is made up through r and the diode during the pulse. As RC is increased still further the output waveform approximates to that of Fig. 17*b*.

Fig. 18*d* shows typical waveforms when rC is longer than the pulse length. The charging of the condenser during the pulse is now slower, so that the input pulse never reaches its full value; the output pulse is correspondingly reduced. When r is large the *squaring* action of the diode is particularly effective. The current through the diode is small and nearly constant, so that the top of the output pulse is accurately square and independent of the exact shape of the generated pulse (shown dotted). When good squaring action is desired a resistance r' is sometimes added in series with the diode as shown in Fig. 16*c*. The analysis is similar to the above and the results almost the same.

It should be noted that with *diode-squaring* the upper portion of the input pulse is not entirely eliminated, but is attenuated by the factor d/r' by potentiometer action between resistance r' and the diode forward resistance. After subsequent amplification, the original shape may therefore

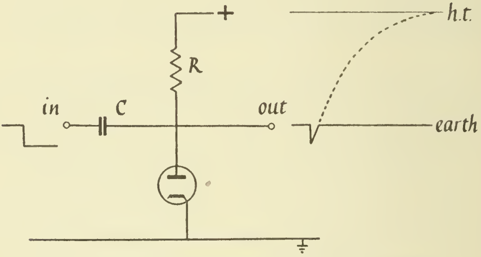


FIG. 19.—Pulse shaping by parallel diode.

reappear. In this respect diode-squaring is not so successful as *cut-off squaring* (see § 2.4), a process in which the original shape of the top of the pulse is completely lost.

To summarize, the parallel diode is used (*a*) as d.c.-restorer and (*b*) for squaring. The complex effects seen in Fig. 18 appear in many circuits.

A more specialized use of the parallel diode for pulse shaping is indicated in Fig. 19. The diode is normally conducting with current flowing through resistance R and through the diode to earth. The output voltage is clamped at earth potential, but when a negative square front is applied to the input, the output voltage falls correspondingly. Now the diode is non-conducting and the current through resistance R must flow through condenser C. As the condenser is charged the output voltage rises exponentially as if to reach the h.t. potential (dotted curve). The diode, however, conducts again as soon as the output reaches earth potential, preventing a further rise in voltage, and a triangular output pulse is obtained (solid curve). This circuit is useful for shortening an input pulse (cf. §§ 3.1, 7.8), and can be used with positive pulses by reversing the diode and the supply voltage.

2.3. Amplification. At this stage it is convenient to summarize the behaviour of amplifying valves, and to mention properties and arrangements which are particularly applicable to pulse circuits.

If a valve is biased to an approximately linear region of its characteristic, small voltage changes applied to the grid will be amplified more or less faithfully, by a factor

$$\mu \frac{R}{R + R_0} \text{ for a triode} \quad . \quad . \quad (10)$$

and

$$gR \text{ for a tetrode or pentode} \quad . \quad (11)$$

Where μ, g and R_0 are the amplification factor, mutual conductance (slope), and anode impedance of the valve; and R is the anode load. These are the standard expressions for the gain of a valve, and apply equally well in the case of pulse waveforms.

In pulse circuits it is desirable to use small anode loads so that the integrating effects due to stray capacity are reduced. This means that valves of high slope are chosen to give reasonable gains; and also that the pentode formula often suffices also for triodes, because $R \ll R_0$. To develop a reasonable pulse voltage across the small anode loads the valves are operated at high currents, consistent always with the rated mean wattage dissipation.

For amplifiers with negative feedback, the amplification is

$$A' = \frac{A}{1 + xA} \quad . \quad . \quad . \quad (12)$$

where A is the amplification without feedback and x is the fraction of the output voltage fed back to the input. If the loop-gain xA is large, $A' = 1/x$ and is almost independent of A, and therefore independent of small variations in valve characteristic, heater voltage, etc.

The output impedance of the feedback amplifier is $Z_0/(1 + xA)$, where Z_0 is the output impedance in the absence of feedback. The output impedance is therefore reduced and, here again, the loop-gain xA is the characteristic quantity, which in this case determines the amount of the reduction.

Applying (12) to the case of a pentode with anode load R, and negative feedback supplied by an unbypassed cathode resistance R_k, we find that $A = gR$, $x = \alpha R_k/R$, and $A' = gR/(1 + \alpha g R_k)$, where α is the ratio of cathode current to anode current usually 1·2. Here the amount of control exercised by the cathode degeneration is determined by the quantity gR_k. If $gR_k \gg 1$, $A' = R/\alpha R_k$, and is independent of g. Effectively, the slope of the valve has been reduced to $1/\alpha R_k$. This principle is useful in stabilizing a single stage amplifier, and as we shall see later, in generating a stable d.c. current. Note that the output impedance at the anode is not reduced in this case, because there is no feedback from the anode.

There is, in fact, a *small increase* in the output impedance because the effective anode impedance of the valve is increased (see § 2.7), and the small shunting effect of the valve on the anode load, usually negligible, becomes even more negligible. For triodes this effect may become significant.

2.4. Squaring. In many pulse circuits, a large signal is applied to the grid of a valve, driving it well outside the linear region of the characteristic. With a large negative signal, the valve is *cut off*, and the anode current is reduced to zero. Further negative excursion of the grid then produces no change in the anode current, so that any large negative pulse will appear as a flat topped positive pulse at the anode of the valve. The valve, therefore, has a squaring action when the grid is driven beyond cut-off. This is known as *cut-off squaring*.

Similarly, with a large positive pulse, the valve is driven into grid current. The grid-cathode of the valve then acts as a diode, and produces the d.c. restoring and squaring effects discussed in § 2.2. The result is that the negative pulse on the anode has a more or less square top. But this *grid current squaring* is not very effective unless the resistance in series with the grid is large (r' of § 2.2). Small departures from squareness on the grid are amplified and appear as major irregularities on the anode (see also § 3.5).

In many circuits the valve is driven suddenly to and fro between the cut-off and the grid current regions. In this case, we can regard the valve simply as a switch, and speak of it as being turned on or off according to the situation at its grid. In these cases the concept of linear amplification loses its significance and we refer to the characteristics only to see what voltage swing on the grid is needed to switch the valve on or off.

2.5. Cathode follower. The circuit is given in Fig. 20*a*, the output signal is taken from the cathode, across the unbypassed cathode resistance R_k. The whole of the output signal, appearing on the cathode, subtracts from the input signal, and the device therefore acts as a feedback amplifier with $A = gR_k$ and $x = 1$. Equation (12) yields

$$A' = gR_k/(1 + gR_k),$$

showing that the amplification is less than unity, but approaches unity when $gR_k \gg 1$. The output impedance of the circuit is $R_k/(1 + gR_k)$ and this tends to $1/g$ when $gR_k \gg 1$. For a typical high slope pentode $g = 5 - 10$ ma/v, so that the output impedance is 100-200 ohms.

The output impedance can be obtained in another way using Thévenin's theorem. We suppose that the circuit is in equilibrium with the grid at a fixed voltage, and then suppose that the cathode voltage is raised v volts above its equilibrium value by an external battery connected to the output. The grid cathode voltage is changed by v, so that the current through the valve is reduced by gv. This current, however, still flows in the resistance and must be supplied by the battery. The battery must also supply the extra current v/R_k which flows in R_k because the voltage across it has increased. The total current supplied by the battery is $v(g + 1/R_k)$. The apparent impedance of the circuit is therefore $1/(g + 1/R_k) = R_k/(1 + gR_k)$. This is the output impedance of the cathode follower.

The equivalent circuit of the cathode follower is given in Fig. 20*b*. Here the generator voltage is equal to the voltage applied to the grid.

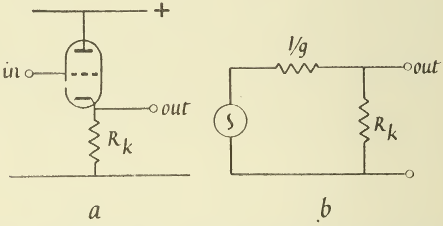


FIG. 20.—Cathode follower and equivalent circuit.

The fact that the gain is close to unity means that the cathode voltage follows the grid voltage: hence the name cathode follower. The physical reasoning is as follows: if the cathode lags behind the grid on a positive going wavefront, the valve conducts strongly, and raises the cathode. Conversely, if the cathode should become too positive the valve is cut off, and the cathode voltage must fall. In both cases the valve reacts so as to restore the cathode voltage to its correct value. It is this dependence of the valve current upon the output voltage that gives the circuit its low output impedance. The circuit reacts with a large correcting current, whenever the output voltage is in error.

The input impedance of the cathode follower is higher than that of the same valve used as an amplifier. A fraction $gR_k/(1 + gR_k)$ of the input voltage appears at the output, therefore the grid-cathode voltage is the fraction $1/(1 + gR_k)$ of the input voltage. This means that the charge required in the grid circuit to charge the grid-cathode capacity is correspondingly reduced, so that in effect the grid-cathode capacity is reduced by the factor $(1 + gR_k)$. The grid-screen or grid-anode capacity remains as before. This feature of the cathode follower is useful in wide band amplifiers.

The main use of the cathode follower is as an impedance transformer. It can receive a signal from a high impedance source and transmit it to a low impedance at the output (e.g. a concentric cable, impedance about 100 ohms). The low output impedance of the cathode follower is useful especially when a pulse is to be developed across a substantial stray capacity. But the concept must be used with caution in the case of large pulses, because of the non-linearity of the valve. On a positive going wavefront, the cathode tends to lag if it is loaded with capacity C, but the valve then conducts strongly, and the capacity is rapidly charged. The positive going front is therefore transmitted rather faithfully. On a negative going front, however, the valve will often be cut off. The condenser must then discharge at a comparatively slow rate through R_k. For positive going fronts, the time constant will be C/g: for large negative going fronts it is R_kC. Typical input and output waveforms are shown in Fig. 21. Notice the slow exponential tail on the output pulse, which ends abruptly when the valve regains control at X.

2.6. Long-tailed pair. The circuit is shown in Fig. 22. The voltages v_1 and v_2, applied to the grids are in the neighbourhood of zero, while the cathode resistance goes to a high negative voltage V (for example, $-$ 300 volts). Because of cathode follower action, the cathode voltage will be in the neighbourhood of the grid voltages, that is, near zero; the current through R_k is therefore close to V/R_k, and varies only slightly when v_1 and v_2 are changed. We shall assume that the cathode current is exactly constant, independent of v_1 and v_2. This condition can be realized as accurately as we please by making V large enough, and then choosing R_k to give a current of the desired magnitude. The constant current flowing through R_k is divided between *V1* and *V2*. If $v_1 = v_2$ and the valves are identical, half the current will flow in each valve. Starting from this condition, suppose we apply a signal voltage $+ v$ to each grid. The cathode voltage will rise by v but the cathode current will be almost unchanged, and will still be equally divided between the two valves. A symmetrical signal applied to both grids will therefore produce no effect at the anodes.

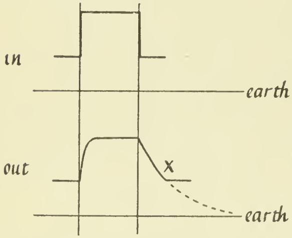


FIG. 21.—Waveforms in cathode follower circuit.

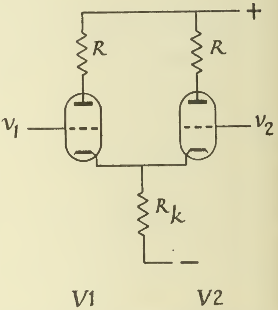


FIG. 22.—Long-tailed pair-circuit.

Now suppose a signal $+ v$ is applied to the grid of *V1*, and $- v$ to the grid of *V2*. The cathode voltage will be unaltered, the current through *V1* will increase by gv, and the current through *V2* will decrease by gv; the total current will be unaltered. Equal and opposite voltage changes will occur on the anodes, numerically equal to $gR \times v$ (pentode approximation).

For the case of a signal $+ v$ applied to the grid of *V1*, with the grid of *V2* held constant, we can obtain the result in two steps. First, apply $+ \frac{1}{2}v$ to each grid: this raises the cathode by $\frac{1}{2}v$ but gives no effect at the anodes. Then add $+ \frac{1}{2}v$ to *V1* and $- \frac{1}{2}v$ to *V2*; this gives $- \frac{1}{2}gR \times v$ at the anode of *V1* and $+ \frac{1}{2}gR \times v$ at the anode of *V2*. The overall effect therefore is that the cathode rises by $\frac{1}{2}v$, and a push-pull signal equal to $\frac{1}{2}gR \times v$ appears at each anode.

We may summarize as follows. The circuit responds only to the difference in the voltage between the two grids. The difference voltage is amplified and appears as a balanced push-pull signal at the anodes. The change on each anode is $\frac{1}{2}gR \times$ voltage difference at the grids: the gain to each anode is therefore half that of a single valve. But the voltage between the anodes is $gR \times$ voltage difference at the grids; therefore the gain *between the anodes* is the same as that of a single valve. It makes no difference whether the input is applied to one grid only, or is a symmetrical push-pull signal applied to both.

So far the discussion has been limited to small voltage changes with the valves operating in the linear region. Suppose the grid voltage of *V1* is now raised further from the symmetrical value, the grid voltage of *V2* remaining fixed. *V1* gradually draws more of the available cathode current, and *V2* less, until finally all the current is flowing in *V1*, and *V2* is cut off. After this there is no further change in the anode currents, but the cathode voltage rises, following the grid of *V1*. Fig. 23 gives the anode and cathode waveforms when a steadily rising voltage is applied to *V1*. The rule is always that the cathode follows the more positive grid. Note that when the current is not symmetrically divided, the slopes of the two valves g_1 and g_2 will be different; the gain to each anode is then $\frac{1}{2}g_1g_2R/(g_1 + g_2)$. This means that we must average the reciprocals of the slopes, and the valve with the lower slope dominates in the gain formula.

The circuit is called the *long-tailed pair*, indicating that the cathode resistance goes to a high negative voltage. In practice the operation is not greatly disturbed if this voltage is reduced, provided always that $gR_k \gg 1$. When it is vital that the cathode current remain constant, independent of grid voltages, resistance R_k can be replaced by a constant current valve (§ 2.7). Instead of using a negative supply the cathode resistance is often taken to earth, and the grids raised correspondingly positive. For brevity, this circuit is sometimes referred to simply as a *pair*.

2.7. Constant current valve. We often need to supply to, or draw from, a point in the circuit a current which is independent of the voltage at the point. An example of this has already occurred in the constant current to be drawn from the cathodes of the long-tailed pair. To avoid the use of large voltages, we can replace the cathode resistance by a pentode operating with fixed screen and grid voltages, as shown in Fig. 24. Because the anode current of a pentode is nearly independent of anode voltage over a wide range, this circuit will supply an almost constant current in spite of

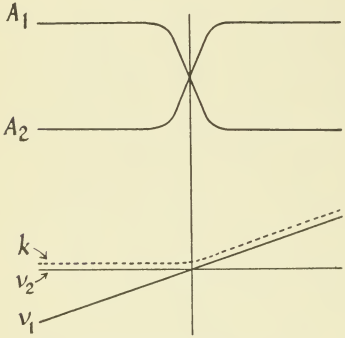


FIG. 23.—Long-tailed pair-waveforms.

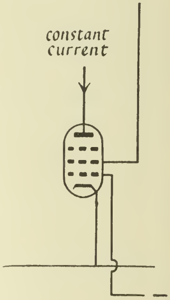


FIG. 24.—Constant current by pentode.

voltage changes. The anode voltage must, however, be maintained above the knee of the characteristic, which begins at 50-100 volts. The differential resistance of the circuit (that is, the constant which relates small anode voltage changes to the corresponding current changes) will be equal to the internal impedance R_0 of the valve. The actual value of the constant current in this circuit will depend very much on the individual valve characteristic, and will not be particularly stable.

An alternative method, which applies also to triodes, is to raise the grid to a voltage V, and use a large cathode resistance R_k, as shown in Fig. 25. A voltage differing little from V will appear at the cathode because of cathode follower action; the cathode current, and therefore the anode current, will be close to V/R_k, and nearly independent of anode voltage. It is easily shown that the differential resistance at the anode is μR_k provided $gR_k \gg 1$. As an example, we suppose a triode with $\mu = 30$ is used to supply a current of 10 ma. With $V = 100$ volts, and $R_k = 10$ kΩ, the desired current would be produced with a differential resistance of 300 kΩ. A 100 volt change in anode voltage would therefore change the current only 3 per cent.

It is clearly an advantage with this second circuit to use a valve of high μ, that is, a pentode. In this case the circuit is similar to that of Fig. 24 except for the added cathode resistance and the positive voltage applied to the grid. These modifications result in improved performance over the circuit of Fig. 24, the differential resistance now being $(1 + gR_k)R_0$. With only a modest positive voltage of, say, 20 volts applied to the grid, and $R_k = 2k\Omega$, a current of 10 ma. can now be produced with a differential resistance of 5-10 MΩ. The current will then change only 0·1 per cent. for a 100 volt change in anode voltage.

A further advantage of including the cathode resistance is that the current becomes substantially independent of the valve characteristic, especially when V is large.

2.8. Cascode amplifier, Fig. 26. In this circuit the signals are applied to the grid of the lower valve *V1* which is normal except that its anode load is replaced by the upper valve *V2*. *V2* has a fixed potential, say + 150 volts, applied to its grid and a resistance in the anode circuit across which the output signals are developed. The cathode potential of *V2* adjusts itself automatically, by cathode follower action, until *V2* is passing the same current as *V1*, and the same current must flow in the anode load resistance R. The cathode will clearly be a few volts more positive than the grid of *V2* and this will allow ample anode potential for *V1*.

The cathode of the upper valve presents an impedance $1/g$ to the lower valve (cf. § 2.5), and therefore the lower valve behaves as though it had an anode load resistance of this value. If the two valves are identical the gain to the anode of *V1* will therefore be $g \times 1/g = 1$, the pentode approximation being valid for *V1* because of the small anode load. The input signals therefore appear at the cathode of *V2* inverted but not amplified. Since the grid of *V2* is held at constant

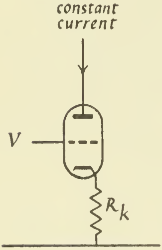


FIG. 25.—Constant current by triode.

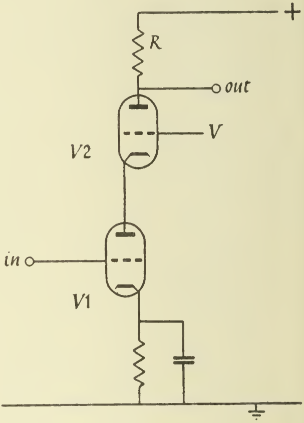


FIG. 26.—Cascode circuit.

voltage, these cathode variations give rise to current variations in *V2* and produce an inverted signal amplified by a factor gR at the anode. Here we have assumed that the voltage variations at the anode of V2 do not react back on the anode current as is usual with a triode, and we shall now show that this is so. In fact, with two identical triodes of slope g, amplification factor μ and anode impedance R_0, the circuit is equivalent to a single valve of slope g, amplification factor μ^2, and anode impedance μR_0, values which are typical of a pentode.

The high effective anode impedance can be seen in the following way. Suppose the grid voltage of *V1* is held constant, but the grid voltage of *V2* is raised 1 volt. This raises the anode potential of *V1* by 1 volt and therefore increases the current by $1/R_0$. Now suppose that the grids of *V2* and *V1* are held constant but the anode potential of *V2* is raised 1 volt, the influence of the anode is $1/\mu$ times the influence of the grid, and therefore the increase in current in *V1* is $1/\mu R_0$. This must also be the change in current in *V2*, and it follows that the effective anode impedance is μR_0. As far as the slope is concerned we have seen that the pentode approximation is valid for *V1*, and therefore the dynamic characteristic will have the full slope g. The same current must flow in the upper valve, so the amplification factor is $g \times \mu R_0 = \mu^2$. This means that a typical double triode * with $g = 5$ ma./v, $\mu = 50$, $R_0 = 10$ kΩ, can be connected to behave as a single valve of the same slope, but $\mu = 2{,}500$, $R_0 = 500$ kΩ, values which are typical of a pentode. This means that large anode loads can be used with advantage, and the pentode formula, gain $= gR$, is valid.

This circuit also reduces the capacitative feedback from the anode to the grid of a triode (Miller effect). There is now no direct capacity link from output to input, and because the gain to the anode of *V1* is unity the grid-anode capacity in *V1* is in effect increased only by a factor 2. It follows that as far as input impedance is concerned the circuit is again virtually equivalent to a pentode. This cascode circuit can be used with advantage in wide band amplifiers.

Another application of the cascode circuit is as a constant current device. Essentially the impedance presented at the anode of *V2* is $\mu \times$ the impedance in the cathode circuit. If then *V1* is connected as a constant current valve (e.g. Fig. 25), presenting a high impedance to the cathode of *V2*, the impedance at the anode of *V2* will be higher still. And a still higher impedance may be obtained by adding a third valve at the top and so on: the differential impedance is multiplied by μ for every valve added, and soon becomes very high indeed.†

* Mullard ECC 81, R.C.A. 12AT7.

† For an application to stabilized power supplies see P. Fellgett *J Sci. Instrum.* 31, 217, 1954.

2.9. Economic and Stable Design. In pulse circuits there is usually more than one way of achieving the desired result. The choice will be dictated by stability of performance and considerations of economy.

To develop pulses across the small load resistances which are required for sharp-fronted pulses (§ 1.5), large valve currents are often required. It is important, however, to minimize the mean current taken by each valve both to keep within the permitted power dissipation, and also to economize in the h.t. power supply. Economy in current can often be achieved by arranging that, as far as possible, valves are turned on during the pulse, rather than off. If the valve conducts only during the pulse, the pulse current can be large, while the mean current is small. This means that steep-sided pulses can be produced economically, at the anode of a valve if they are negative, at the cathode of a valve if they are positive. A cathode follower will be used, for example, to develop a positive pulse across a low impedance, while an anode follower (see § 6.8) will be more suitable for a negative pulse. To ensure stability, measures are taken to minimize effects of small variations in valve characteristics and supply voltages. The performance is made to depend as far as possible only on the values of resistances and condensers in the circuit. In amplifiers, this means that the gain will be controlled by negative feedback; and in trigger circuits the signal will be large compared to the grid base of the valve, which then acts primarily as an on-off device, switching from one controlled current to another.

In pulse circuits a triode or pentode can often be used interchangeably, but the pentode gives a slightly better performance because of its lower anode-grid capacity. For simplicity a triode will normally be indicated in the circuits, but we note here that a pentode can in general be substituted with advantage. The screen and suppressor grid connections of a pentode will often be omitted from the diagram: it is then to be understood that they go to h.t. and cathode respectively.

CHAPTER III

SQUARE WAVE GENERATORS

A SQUARE pulse or step waveform can be generated most simply by operating an on-off switch. Indeed, if the circuit is made and broken by liquid mercury, pulses of almost perfect square shape, rising in less than 10^{-9} sec., can be made in this way. This form of pulse generator is increasingly used for testing equipment designed to handle very rapidly rising pulses, but mechanical switching is severely limited in recurrence frequency and in flexibility, with the result that pulse waveforms are more usually generated by means of valve circuits. There are circuits which produce a single square pulse in response to an initiating signal, usually called *trigger circuits* (see Chap. IV); there are circuits which give out a train of regularly spaced pulses, the free-running square wave generators considered in this chapter; and there are circuits which produce saw-toothed waveforms, either single or recurrent (see Chap. V). These basic elements, together with the fundamental principles of Chaps. I and II, are the foundation of all pulse circuits, and enable an infinite variety of effects to be obtained.

All square wave generators include an amplifier with strong positive feedback from output to input. Usually two valves are used so that the output is in phase with the input, and the positive feedback connection can be made directly. The circuit is arranged to switch suddenly from one temporarily stable state to another, thus generating the square wave. This trigger or snap action is considered in detail in § 4.1. In the present chapter we discuss the circuits which oscillate spontaneously, generating a repetitive square wave.

3.1. Multivibrator. In the multivibrator there are two valves connected with positive feedback as shown in Fig. 27. One can check that the feedback is positive by following the consequences of a small positive signal applied for example to the grid of *V1*. In practice the circuit is not stable if both valves are conducting, but oscillates between two temporarily stable extremes: in the first *V1* is fully conducting, while *V2*

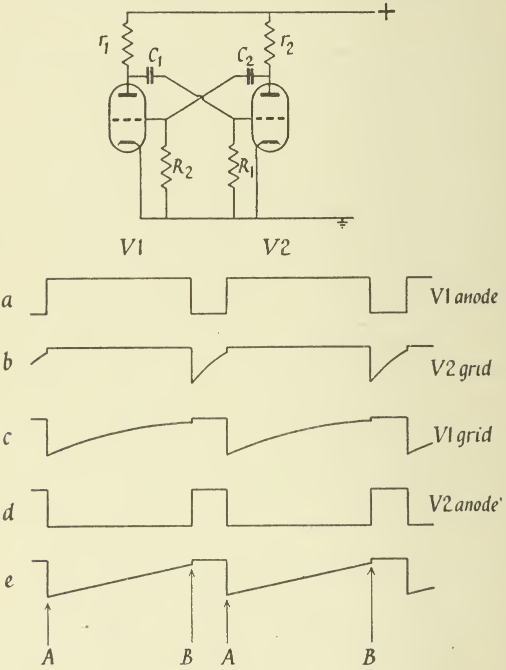


FIG. 27.—Multivibrator.

is cut off, in the second *V1* is cut off and *V2* is conducting. The transition between the two states is very rapid, and gives rise to square pulses at the anodes as the current flowing in the anode load resistances is alternately turned on and off. The time lapse in each state before the circuit makes a transition is determined by the coupling networks R_1C_1 and R_2C_2, as will now be explained with reference to the anode and grid waveforms of Fig. 27*a*, *b*, *c* and *d*.

Starting at the point *A* on the diagram, that is the moment when *V2* begins to conduct, the sequence is as follows. *V2* conducts and its anode voltage falls sharply. This negative wave front passes through the coupling condenser C_2 to the grid of *V1* and cuts off the current in *V1* : its anode potential therefore rises, driving the grid of *V2* positive and making *V2* conduct as postulated. The grid of *V2* does not rise above earth potential, however, because of the diode action between the grid and cathode when grid current flows (see § 2.4).

The net result of this process is to leave *V2* conducting with its grid at zero potential (a stable state of affairs because the grid leak is connected to cathode), and *V1* non-conducting with its grid highly negative. The situation at *V1* is, however, not permanent because current flows through resistance R_2 charging condenser C_2 with the result that the grid potential returns exponentially towards earth with time constant R_2C_2. Eventually at point *B* the grid reaches the cut off voltage of *V1* (say $-$ 5 volts), *V1* starts to draw current, and the circuit snaps over into the opposite state. The anode potential of *V1* now drops, a negative voltage is transmitted through C_1 to *V2*, and so on. The cumulative action in this case leaves *V1* conducting, and *V2* cut off with its grid highly negative.

The sequence now continues correspondingly, with the grid of *V2* recovering towards earth potential with time constant R_1C_1 until *V2* again conducts and we again reach the situation at point *A*. In drawing the waveforms it has been assumed that time constant R_1C_1 is less than R_2C_2. This produces the asymmetrical square waves at the anodes as shown. With equal coupling time constants a symmetrical square wave is obtained from either anode.

There is a notable similarity between the grid waveforms in the multivibrator and the waveform Fig. 18*a* which was produced by the diode circuit, Fig. 16*b*. The grid-cathode of a triode behaves as a diode, so that in both cases a square wave is passed through an *RC*-coupling circuit on to the anode of a diode. The discussion in § 2.2 is therefore applicable to the grid waveform of the multivibrator.

One can obtain pulses with a large mark-space ratio with a multivibrator by making one time-constant much longer than the other. But there is a limit to this process. One must remember that the charge which slowly leaks out of the condenser C_2 during the long *space* period must be restored during the short *mark* period if the cycle is to repeat. The charging path is through the anode load of *V2* and grid current in *V1*. This means that to obtain a large mark-space ratio we must make the grid leak R_2 much larger than the anode load of *V2*. In increasing the time constant R_2C_2, therefore, it pays to increase the resistance R_2 rather than the condenser C_2.

The multivibrator works best with a roughly symmetrical waveform. While mark-space ratios of say 10 : 1 can be obtained easily, attempts to reach 100 : 1 or more will usually be frustrated unless some care is taken in the design. For large mark-space ratios Scarrott's oscillator, § 3.3, is superior.

Often the grid resistors R_1 and R_2 are connected from grid to h.t. instead of from grid to earth. At first sight rather surprising, this practice does not damage the valves because grid current prevents the grids from rising positive. The advantage is that the grid waveform is modified during its negative excursion. Instead of rising exponentially towards earth potential the grid voltage now recovers towards h.t. potential. The initial rise up to the cut-off potential of the valve is nearly linear and it is this portion that determines the pulse length in the multivibrator; the resulting grid waveform is shown in Fig. 27*e*. Because the steeply rising grid waveform intersects the critical potential more definitely, the result is that the pulse lengths are more exactly determined and less dependent on small changes in the cut-off potential of the valves.

To synchronize the multivibrator to an external waveform, the reference signal is injected at one anode, whence it passes through the *RC*-coupling to the opposite grid. Superimposed upon the usual grid waveform, it tends to make the multivibrator fire in synchronism with the reference wave. Direct connection to an active grid via a coupling condenser is always undesirable because the additional components connected to this high impedance point usually upset the operation of the circuit. For this reason synchronizing or triggering signals are always injected at an anode, or at a free grid if there is one available (e.g. § 3.2). For a fuller discussion of synchronization, see § 5.2.

In practice the anode waveforms of the multivibrator are not ideally square as indicated in Fig. 27. The modifications are fundamental but the treatment is deferred until § 3.5.

3.2. Cathode coupled multivibrator, Fig. 28. Here again a two-valve amplifier is operated with overall positive feedback, but in this case the valves are arranged as a long-tailed pair (see § 2.6). A positive signal on the grid of *V2* gives a positive output on the opposite anode, and a single *RC*-coupling is used to connect this back to the grid. This circuit gives a symmetrical square wave, Fig. 28*a*, on each anode, the waveform at the grid of *V2* being the differentiated square wave, Fig. 28*b*. When the grid of *V2* swings positive *V2* conducts, raising the common cathode and cutting off the current in *V1*. This phase lasts until the grid in its exponential return to earth potential, time constant *RC*, falls sufficiently for *V1* to draw current. The grid is then driven cumulatively negative until *V1* takes all the current, and *V2* remains cut off until its grid again approaches earth potential, and the cycle repeats. No grid current flows in this circuit because the cathode potential can follow the grid potential during its positive excursion : the grid waveform is therefore not squared by grid current as in the multivibrator of § 3.1.

Advantages of this circuit are the free grid (in *V1*) and free anode (in *V2*) which do not play any part in generating the oscillations. An undistorted square wave output may be taken from the anode of *V2* without disturbing the oscillator : and synchronizing signals can be injected to the grid of *V1* (see § 5.2).

If the grid of *V1* is biased slightly positive or negative with respect to the grid of *V2* the waveform becomes asymmetric; and with too great a bias difference the oscillations stop, the circuit remaining in a stable state with one of the valves

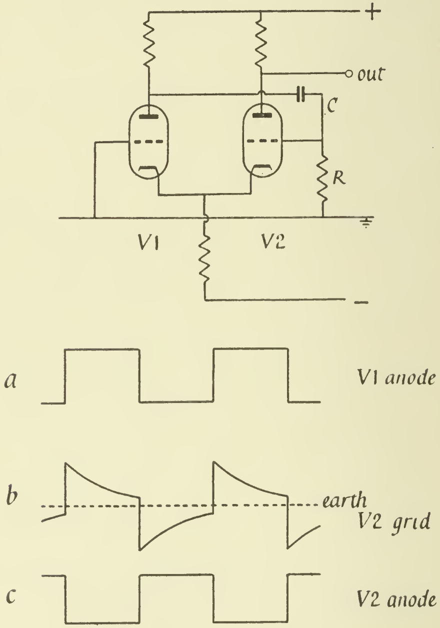


FIG. 28.—Cathode-coupled multivibrator.

permanently non-conducting. A suitable initiating signal, applied for example to the grid of *V1*, will then cause the circuit to go through a single cycle of oscillation, with the result that a single square pulse is produced at the anodes. The free-running oscillator has become a trigger circuit. This type of circuit will be treated further in Chap. IV.

3.3. Scarrott's oscillator, Fig. 29. This modification is particularly useful for generating pulses with a large mark-space ratio. The circuit is similar to that just treated with

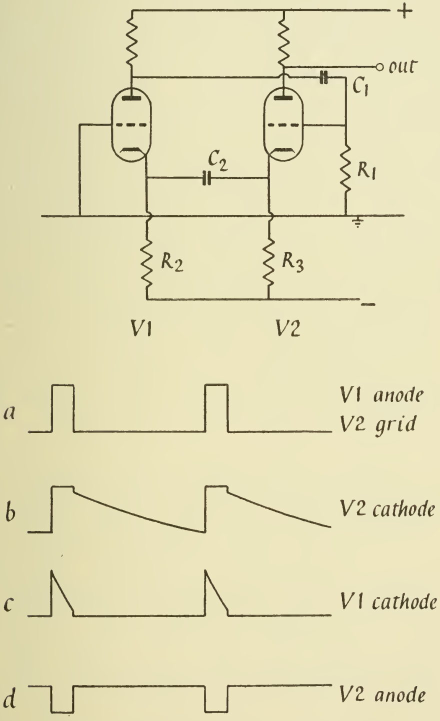


FIG. 29.—Scarrott's oscillator.

the addition of the condenser C_2 coupling the two cathodes, and the separate resistances R_2 and R_3 to the negative line. The time constant R_1C_1 is made very long to ensure that the on and off periods of the oscillator are both governed by the cathode network. The operation is as follows.

The square wave produced at the anode of *V1* passes without differentiation to the grid of *V2* and turns *V2* on and off. Suppose *V2* has just been turned on. Its cathode potential rises and drives the cathode potential of *V1* positive because of the connection through condenser C_2. This cuts off the current in *V1* giving a positive pulse at the anode of *V1* and grid of *V2* as postulated. The potential at the right-hand side (r.h.s.) of C_2 is held at a fixed potential by the cathode following action of *V2*, but the potential at the left-hand side (l.h.s.) which has been driven positive falls exponentially towards the negative line as current flows into C_2 through R_2, time constant R_2C_2. The cathode potential of *V1* therefore falls and eventually *V1* will conduct, reversing the state of the circuit. Now with *V1* conducting, but *V2* cut off, it is the current through R_3 that must flow into the condenser. The flow of current through C_2 is reversed, and with the l.h.s. clamped by the cathode follower *V1*, the cathode potential of *V2* falls towards the negative line voltage with time constant R_3C_2; eventually *V2* conducts and so the cycle repeats. Grid and cathode waveforms are given in Fig. 29*a*, *b* and *c*.

The two periods of this oscillator are controlled by time constants R_2C_2 and R_3C_2; these may be very different because R_2 may be say 10 kΩ while R_3 is say 10 MΩ giving a ratio of 1,000 : 1. The success of this circuit is due fundamentally to the low output impedance of the valve cathodes. This enables R_2, say, to be quite small without affecting the amplification of the positive feedback loop. Condenser C_2, which charges slowly through R_3, is discharged rapidly through the low impedance path formed by valve *V2* and resistance R_2.

In this circuit there is a free grid for synchronizing signals, and a free anode from which the output can conveniently be taken.

3.4. Blocking oscillator, Fig. 30. In this circuit there is only one amplifying valve and the necessary phase reversal to give positive feedback is obtained by means of the transformer coupling from anode to grid. The circuit is similar to the transformer coupled R.F. oscillator, but there is no tuning condenser and here the two windings of the transformer are tightly coupled by means of a high permeability core (iron, mumetal, ferrite, etc.). Instead of a steady

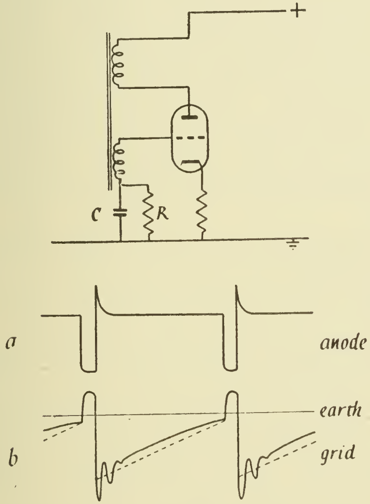


FIG. 30.—Free running blocking oscillator.

sinusoidal oscillation, the circuit gives a single pulse, negative at the anode and positive at the grid, and then becomes quiescent with the grid negative.

Consider the sequence from the moment when the valve first starts to conduct. The anode potential falls and by the action of the transformer the grid is driven positive. Cumulative action then drives the grid strongly positive, a large anode current flows, and the anode voltage falls to a low value : heavy grid current flows. This situation persists for a short time depending mainly on the characteristics of the transformer and partly on the capacity, C. A rising anode current is required in the primary of the transformer to maintain the grid positive against the influence of the grid current, but this cannot continue indefinitely. Eventually the anode current ceases to rise and the process is reversed. The valve now cuts-off, the anode voltage rises to h.t. potential or higher and the grid is driven negative. Now because the condenser C has been charged by grid current the grid is more negative than it was at the beginning of the pulse, typically 50-150 volts negative. The valve therefore remains non-conducting while the grid voltage rises exponentially towards earth potential with time constant RC, eventually reaching the point at which the valve again draws current, and the cycle repeats.

The pulse length can be as short as 0·1 microsecond with a very small transformer, and is seldom longer than 10 microsecond. The interval between pulses can, however, easily be of the order of seconds if the time constant RC is large. The blocking oscillator is therefore well adapted to generating pulses with a large mark-space ratio. As in the case of the multivibrator it is possible to connect the grid leak resistance R from the condenser to the h.t. line instead of from condenser to earth, and this makes the grid waveform more linear, as indicated in Fig. 30*b* (dotted). In this form the blocking oscillator is often used as a time base generator, the saw-tooth waveform being taken from the condenser. Square output pulses can be taken either from the anode (negative pulse), from the grid winding of the transformer (positive pulse), from a small cathode resistance (positive pulse), or from an auxiliary winding on the transformer. When the circuit is used as a time base this output pulse may be fed to the cathode ray tube to black out the spot during the fly-back period.

As a pulse generator the blocking oscillator has a very low output impedance. This is due to the absence of resistances in the circuit, and to the fact that the valve conducts very strongly during the pulse. It is not unusual, for example, to find a small receiving valve passing a current of 1 amp. during the pulse, and developing a 100-volt pulse across a 100 ohm resistance in the cathode circuit. This low output impedance is a useful property, especially when a pulse has to be developed across a low resistance or fed into a large stray capacity. The main disadvantage is that the pulse length cannot be readily varied.

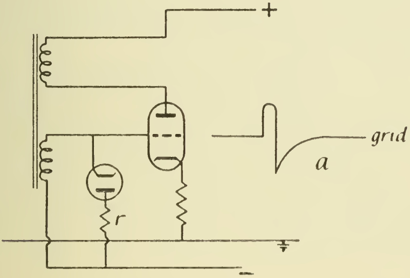


FIG. 31.—Blocking oscillator trigger circuit.

The blocking oscillator may be converted into a trigger circuit by connecting the resistance R to a negative potential so that the grid voltage does not rise sufficiently after a pulse to allow the valve to conduct again. The circuit can then be triggered to produce a single pulse by driving the grid positive, for example, with a signal in an auxiliary winding of the transformer or by applying a negative signal to the anode via an auxiliary valve. In this application it is possible to omit the resistance R and condenser C and to use a diode to give the blocking action. The circuit is Fig. 31. Here the diode conducts when the grid swings negative after the pulse and damps the circuit, preventing further cycles of oscillation. The small resistance r in series with the diode is included to speed up the recovery of the circuit after a pulse. The diode starts to conduct when the voltage across the grid winding is passing through zero, that is, when the flux in the transformer core is a maximum. At this stage the diode in effect short circuits the coil, and the flux decays to zero with time constant L/r', r' being the total resistance of coil and diode. This time constant is reduced, speeding up the recovery, if r' is increased by adding resistance r in series with the diode. Of course r must not be too great or the diode will no longer damp the circuit and the grid will again swing positive, initiating a further output pulse. Ideally the total resistance $(r + r')$ should just give critical damping. As the flux decays in the transformer core a negative voltage appears across the grid winding, giving the negative spike on the waveform as shown in Fig. 31*a*.

Other types of coupling between anode and grid circuits can be used in the blocking oscillator. A common arrangement is derived from the well-known Hartley oscillator. Here there is a single iron-cored inductance between grid and earth and the cathode is connected to a tapping on the coil. The usual blocking condenser and leak resistance are included in the grid circuit. Other arrangements will occur to the reader.*

3.5. Detail of multivibrator waveforms. As mentioned in § 3.1 the anode waveforms of the multivibrator are not perfectly square as indicated in the simplified diagrams of Fig. 27, but usually resemble those of Fig. 32. There is a negative going peak or spike at the start of the negative phase, and at the beginning of the positive phase the voltage does not rise sharply, but exponentially with a rather long time constant. This phenomenon, somewhat disconcerting at first, can be explained by a consideration of the grid current which flows when the grid swings positive. The effects are related to those obtained with the parallel diode discussed in § 2.2 and illustrated in Fig. 18.

Let us start at the moment *A* indicated in Fig. 32*a*. Valve *V2* is conducting and *V1* is off with its anode at h.t. potential, but its grid has just reached the cut-off potential and the multivibrator snaps over. Now *V2* is cut off and its anode

* For theory and further circuits see R. Benjamin, *J.I.E.E.*, part IIIA, 93, 1159, 1946.

starts to rise towards the h.t. potential taking the grid of *V1* up correspondingly. But after a rise of say 5 volts (the cut-off voltage), *V1* draws grid current and the grid is in effect clamped at earth potential. The anode of *V2* can now continue to rise only in so far as the coupling condenser C_2

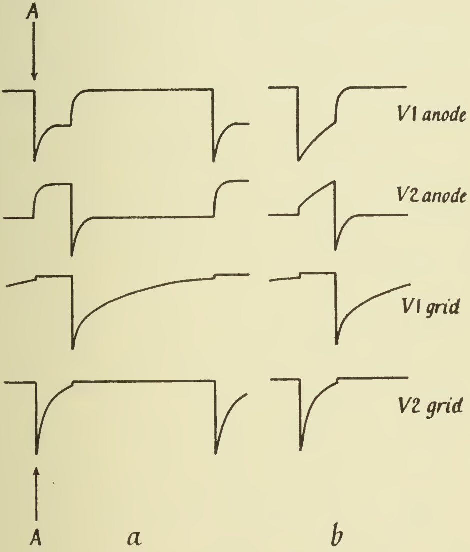


FIG. 32.—Multivibrator waveforms.

(*a*) $r_2C_2 = r_1C_1$
(*b*) $r_2C_2 > r_1C_1$

is charged through the anode load r_2 (see Fig. 27). The anode voltage of *V2* therefore rises slowly with time constant r_2C_2 as shown.

Let us now return to the grid of *V1*. It is not quite true that it is clamped at earth potential. In fact, the grid to cathode path is not of zero resistance but of order 1,000 Ω, and therefore the grid voltage rises slightly positive when grid current flows. There is, in fact, a small positive spike on the grid waveform of *V1* complementary to the exponential rise of voltage on the anode of *V2*. As the anode rises, and the current charging condenser C_2 gradually falls to zero, so the grid returns more exactly to cathode potential. This small excursion of grid voltage is usually invisible on an oscilloscope, but it is amplified by the valve *V1* and appears as a negative spike on the anode waveform. The negative spike in the anode waveform of *V1* is always complementary to the slow rise of voltage on the anode of *V2*; both are governed by the same time constant r_2C_2. A similar process occurs in the next half of the cycle and is governed by time constant r_1C_1.

In Fig. 32*a* we have assumed that $r_1 = r_2$, $C_1 = C_2$, so that the two negative spikes and the two exponentially rising wavefronts are identical. (The asymmetry in the lengths of the square pulses is due to a difference in the grid leaks, $R_2 > R_1$.)

In the alternative case of Fig. 32*b* we assume that the difference in pulse lengths is obtained by making $C_2 > C_1$. In this case the first charging time constant r_2C_2 is longer, the anode of *V2* rises more slowly and the corresponding negative spike is lengthened. If C_2 is still further increased, the anode of V_2 will never reach h.t. potential and the charge on C_2 will not be fully reset when the circuit switches to the opposite phase. With less than the normal charge to leak away through R_2, this means that the following phase will end early. Attempts to increase the mark-space ratio by increasing C_2, thus lengthening the second phase of the waveforms illustrated, partially fail because of this effect.

Tolerably square waveforms may often be obtained from the multivibrator by including large resistances in series with each grid. These limit the grid current and hence the distortion of the anode waveforms.

A superiority of the cathode coupled multivibrator circuits will now be understood. These have a free anode at which a clean positive square pulse is available, unsullied by grid current or by participation in the oscillatory mechanism.

CHAPTER IV

TRIGGER CIRCUITS

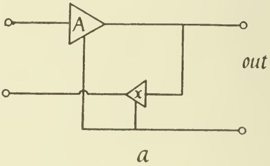
So far we have been dealing with free-running square wave generators, although we have seen that by altering the steady bias conditions the oscillations may be inhibited, and the circuit will then give out a single square pulse in response to an initiating signal. When biased in this way the circuits of §§ 3.2 and 3.4 are examples of trigger circuits.

A trigger circuit has at least one permanently stable state. It remains in this state until triggered by an initiating pulse, which causes it to snap over into a new state. What happens then depends on the circuit details: the new state may also be permanently stable until another signal arrives, or it may revert back to the original state after a lapse of time. There may also be more than two states of the system: in some counting circuits there are as many as five stable states which follow each other in succession before the initial state is reached again. The output from the trigger circuit is usually a square pulse, but it may be a saw-tooth or complex form.

As in the case of the rifle or shot-gun, the ideal trigger circuit gives a constant response, independent of the size or nature of the initiating signal applied to the trigger. Often, however, this ideal is not perfectly realized and there remains some influence of the input signal upon the output waveform. A well-known example of an electrical trigger device which approximates closely to the ideal is the thyratron. The term trigger circuit is, however, usually reserved for circuits using high vacuum (or hard) valves.

Before discussing particular trigger circuits it is appropriate at this stage to examine the trigger or snap-action itself in greater detail.

4.1. Hard valve trigger action. Consider a direct coupled amplifier of gain A with positive feedback, a fraction x of the output being fed back to the input so as to reinforce the input signal. It is assumed that x can be varied without altering the steady conditions in the amplifier when the input voltage is zero. Fig. 33 gives the block diagram of such an arrangement, and also a circuit to show how it could be realized in practice. Let us now consider the behaviour of



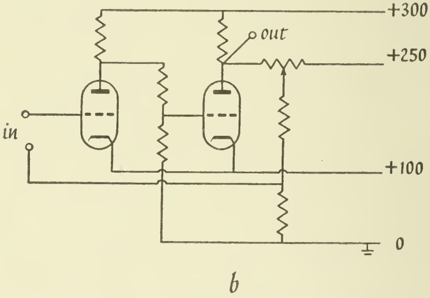


FIG. 33.—Positive feedback circuits.
(*a*) basic principle.
(*b*) a practical circuit.

such an amplifier and the effect of various amounts of positive feedback. The value of the loop-gain xA is the controlling factor.

A typical characteristic for an amplifier with no feedback is given in Fig. 34, curve *a*. There is a region of roughly linear amplification as the input voltage varies about zero; but for more positive input voltages the amplifier saturates, the output stage is cut off, and the output voltage reaches h.t. potential. Similarly, if the input voltage is too negative, the amplifier saturates with the output stage drawing grid current and with the output voltage at some steady low potential. If the input voltage is initially negative and allowed to rise, there is at first no change in the output voltage: but as the input voltage passes through zero the output voltage rises gradually to its maximum value and then remains constant.

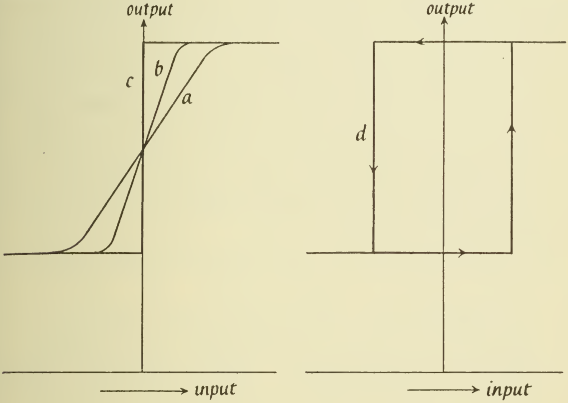


FIG. 34.—Response of positive feedback amplifier.

(a) no feedback, $xA = 0$
(b) some feedback, $xA < 1$
(c) critical feedback, $xA = 1$
(d) trigger circuit, $xA > 1$.

If some positive feedback is introduced the gain of the amplifier rises, and the characteristic becomes steeper in the region of zero grid voltage, Fig. 34 curve b. This trend continues until, with $xA = 1$, the gain of the amplifier becomes infinite. This follows from the fundamental formula for feedback amplifiers given in § 2.3 equation (12). For positive feedback the denominator is $(1 - xA)$, so that when $xA = 1$, A' becomes infinite. In this case the characteristic is given in Fig. 34 curve c; an infinitesimal change of input voltage is sufficient to change the output voltage from one saturation level to the other. Physically the change of voltage at the output is itself giving rise to sufficient signal at the input to produce the whole effect : there is no need for a true input signal. With this amount of feedback there is now a sudden, reversible, transition in the output voltage as the input signal passes through zero.

What happens if the loop gain xA is greater than 1? In this case the output voltage still makes its infinitely steep transitions between the two saturation levels, but now the change of voltage at the output is *more than* sufficient to produce the whole effect. This means that the input voltage changes *more than* necessary and swings the amplifier *well into* the region of saturation. Before the effect can be reversed it is necessary to cancel this surplus input voltage by means of an opposite input signal, with the result that the device is no longer perfectly reversible but follows a hysteresis loop as illustrated in Fig. 34*d*.

With positive feedback with loop-gain greater than 1, the amplifier has become a trigger circuit. The output voltage snaps definitely from one limit to the other in response to an input signal, but to reverse the process a finite change in the input signal is now necessary. This is called the *backlash* of the circuit. It will be clear from the above discussion that the backlash increases as the loop gain increases beyond 1. In fact, one might expect the backlash to be exactly proportional to $(xA - 1)$, but this is not universally true because it is often affected by non-linearities in the circuit.

Backlash is an inevitable feature of practical trigger circuits. First it would be impossible to maintain the loop-gain exactly equal to 1. Secondly, the presence of backlash makes the trigger action definite, because once the circuit has begun to trigger a small reversal of the input conditions will not interrupt the process.

In the discussion above we have assumed that direct coupling is employed both in the amplifier and in the feedback path. This results in a trigger circuit with two permanently stable states. If, however, an *RC*-coupling is used in place of direct coupling somewhere in the circuit both states will not be permanently stable. The charging and discharging of the condenser is then equivalent to a slowly changing input signal. When a change sufficient to overcome the backlash has taken place, the circuit will snap over into a new state. The time interval required depends both on the amount of backlash, and on the time constant of the RC-coupling.

It is now apparent that the free-running square wave generators already discussed can be regarded as trigger circuits oscillating between the two limits set by the backlash of the circuit. At any moment there is an effective input voltage which depends on the charge on the condensers. As the condensers charge this slowly changes and when it reaches the appropriate limit the circuit changes its state. The effective input voltage then returns towards the opposite limit as the condensers charge or discharge. The free-running oscillators are really examples of trigger circuits, but it is convenient in practice to confine the term to circuits with at least one permanently stable state, as we have done so far.

4.2. Scale-of-two circuit, Fig. 35.* This trigger circuit is very similar to the circuit of Fig. 33 used to illustrate the discussion on trigger action. It is, in effect, a direct coupled multivibrator (compare Fig. 27). The circuit has two stable states, and is triggered from one to the other by means of negative pulses applied to the resistance R_1 which is common to the two anode circuits. As each trigger pulse gives a change of state the output square wave taken from either anode is negative-going only once for every two input pulses : hence the term *scale-of-two* circuit. If a second circuit is triggered by the output of the first only the negative going fronts are effective (see below), with the result that the final output is negative-going once for every four input pulses, and so on. In this way scaling circuits which scale by any power of two can be built up.

* This is often called the Eccles-Jordan scale-of-two to distinguish it from thyratron scaling circuits, because hard valve trigger action was first suggested by these authors in 1919.

The circuit is completely symmetrical. The anode of a valve is connected to the opposite grid via the potentiometer R_3, R_4 which terminates at a fixed negative voltage, typically

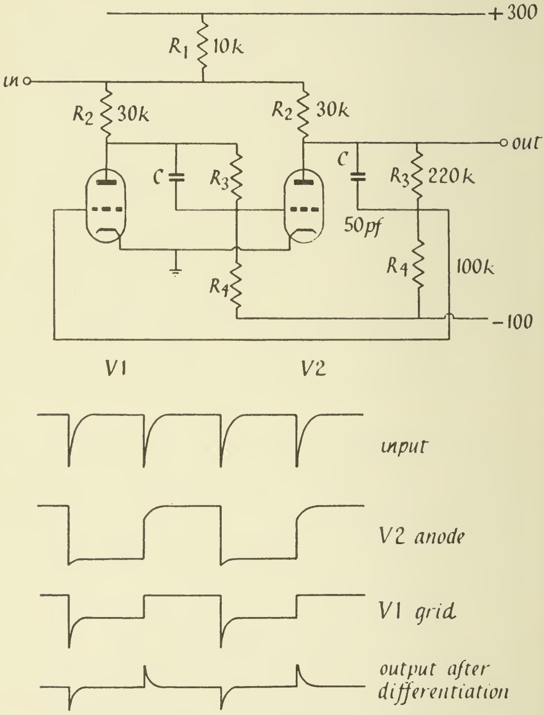


FIG. 35.—Scale-of-two circuit.

− 100 volts. In this way a fraction $R_4/(R_3 + R_4)$ of the voltage variations at the anode (usually about 1/3) are transmitted to the opposite grid. Resistances R_3 and R_4 are chosen so that the grid is well negative when the opposite valve is conducting, and would rise positive when the opposite valve is cut off but for the restraining influence of grid current.

The condensers C are added to compensate for the stray capacity from grid to earth. This capacity, coming after the high impedance potentiometer R_3, R_4, would slow down the rising and falling pulse fronts transmitted from the anode, with the result that the circuit would trigger slowly, and fail to trigger altogether on short input pulses. The condenser C, small but larger than the stray capacity, transmits the steep wavefronts directly to the grid.

As already mentioned the circuit is triggered by negative pulses applied to the common anode resistance R_1. This negative signal is transmitted through the anode loads R_2 to the anodes and through condensers C to both grids. Here the valve (say *V1*) which is already non-conducting is unaffected; but the other valve *V2*, which is conducting with its grid at zero, is cut off by the input signal. Its anode potential rises, swamping the negative input pulse at this point, and the opposite valve *V1* is turned on. Cumulative action then completes the transition in the usual way. Positive input pulses do not trigger the circuit unless they are very large. Transmitted to the grid of the conducting valve they are attenuated by grid current squaring (see § 2.4) and in any case tend to perpetuate the existing state of the circuit; transmitted to the non-conducting valve they are not usually large enough to overcome the steady negative bias (of order − 40 volts) and the valve does not conduct. The circuit is therefore far more sensitive to negative than to positive input signals. For this reason when the input signal is a square waveform from a previous scaling stage, the circuit responds only to the negative going wavefronts.

Condenser C must be kept small if the circuit is to trigger correctly on input pulses which follow each other in quick succession. The condenser transmits the steep pulse fronts to the grid without attenuation by the potentiometer R_3, R_4, and accordingly there is an excessive signal on the grid until the charge on the condenser has been readjusted to correspond to the equilibrium conditions set by R_3 and R_4. Also, when the grid swings positive, grid current charges the condenser again and slows up the rise of potential on the opposite anode. The circuit is therefore difficult to trigger again after a pulse until the equilibrium conditions are established, and this gives rise to a *dead-time* which is roughly equal to the time constant of C with R_3 and R_4 in parallel. This time constant should be kept short ; but R_3 and R_4 must be large enough to avoid drawing heavy current from the anode circuits, and there is a minimum value of C; therefore a compromise must be struck. With the typical values indicated in Fig. 35, the dead-time is about 4 μs.

4.3. Schmitt trigger circuit, Fig. 36. The scale-of-two circuit discussed above is in effect a direct coupled multivibrator. Correspondingly the Schmitt circuit is a direct coupled version of the cathode coupled multivibrator (§ 3.2). It can also be regarded as the cathode coupled modification of the circuit of Fig. 33 used in the discussion of backlash (§ 4.1).

The coupling from the anode of *V1* to the grid of *V2* is by means of the potentiometer R_3, R_4 and once again the small condenser C is added to pass the pulse fronts without attenuation by stray capacity (cf. § 4.2). The circuit has two permanently stable states, the transition from one to the other being controlled by the voltage applied to the grid of *V1*. As usual there is backlash in the circuit so that the transition to *V1*-conducting occurs at a more positive voltage than the transition to *V1*-off. These two critical potentials at which the transitions occur depend on the potentials on the grid of *V2* in the two states. To a first approximation the transitions occur when the two grid voltages are equal, for then the cathode current is equally divided between the two valves. We therefore obtain the waveforms shown in Fig. 36.

Initially, with *V1* non-conducting the grid voltage of *V2* is at its upper level, and the circuit triggers at A when the input voltage rises to this value. This drives the grid of *V2* down to its lower level so that the circuit does not reset until the input voltage falls well below the initial triggering voltage at B. The overshoot in potential at the grid of $V2$, caused by condenser C, and the positive output pulse at the anode of $V2$ also appear in the diagram.

According to this simplified picture the backlash of the

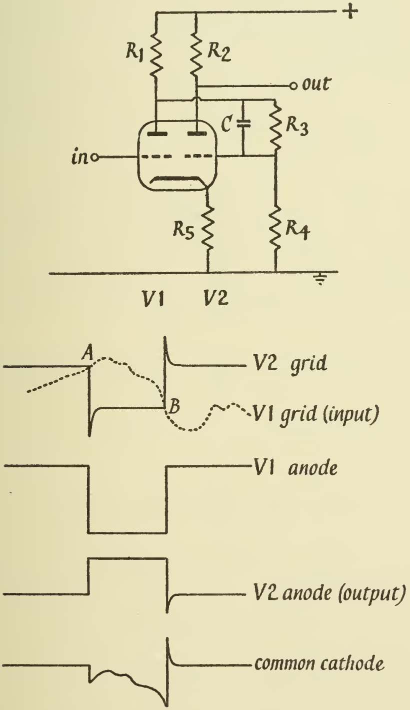


FIG. 36.—Schmitt trigger circuit

circuit is just equal to the change in potential at the grid of $V2$. Usually the backlash is somewhat less than this because the circuit triggers slightly earlier when $V1$ just begins to draw current, and also resets earlier just as $V2$ starts to conduct. Both factors reduce the backlash.

The backlash is also reduced if the grid of *V2* swings below earth potential. In this case the current in *V1* is cut off as soon as the input voltage swings much below zero, and the circuit resets. This effect occurs especially if R_5 is small so that the grid voltages are initially not far above earth potential. A small backlash can therefore be achieved by making resistance R_5 small.

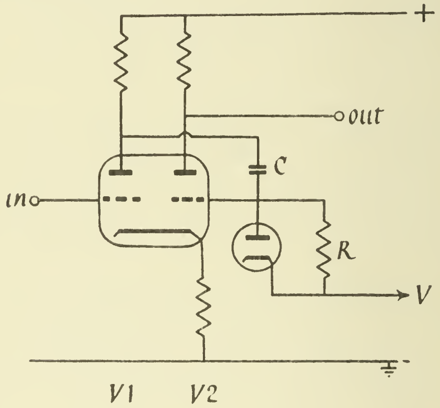


FIG. 37.—Schmitt trigger with long time constant coupling.

The Schmitt circuit is used primarily as a *discriminator*, that is, to give an output signal when an input pulse reaches a predetermined threshold. Compared with the simple diode discriminator (§ 2.1), it has the advantage of the all-or-nothing trigger action, and of giving an output pulse of standard height, although of course its length depends on the length of the input pulse. It is also useful for converting a sinusoidal or other waveform into a square wave of the same frequency.

An interesting variation of the Schmitt circuit, suitable for short input pulses, is given in Fig. 37. Here instead of the potentiometer between anode and grid a long time constant *RC*-coupling is used. For short periods the voltage across the condenser will remain constant and the coupling is therefore equivalent to the d.c. coupling used in the standard circuit. For short input pulses, therefore, the circuit behaves as a Schmitt trigger; but with long input pulses the circuit will reset spontaneously before the end of the pulse because of the charging of condenser *C* (cf. § 4.4). In practice a small leakage of charge from the condenser during the pulse is inevitable, but this is quickly restored through the diode when the circuit resets.

This principle of using a long time constant a.c.-coupling instead of a d.c.-coupling has wide applications.

4.4. Kipp relay, Fig. 38. This is a useful form of the cathode coupled multivibrator. It has one permanently stable state to which it returns, if triggered, after a lapse of time determined by the time constant R_4C. A single square positive pulse of predetermined length appears at the anode of *V2*.

In the normal quiescent state, *V2* is conducting with zero potential between grid and cathode because of the grid leak R_4. The anode current is limited by the anode load resistance R_2 and the associated drop in anode potential. This current develops sufficient voltage across the small cathode resistance R_5 to keep *V1* cut-off, because the grid leak of *V1* is taken to earth. When the circuit is triggered, either by a positive pulse applied to the grid of *V1*, or a negative pulse applied to its anode, the grid of *V2* is driven strongly negative, the cathode potential falls, and *V1* conducts drawing a reduced current through the self-bias resistance R_5. Condenser *C* is then charged in the usual way, by current flowing through R_4 and the circuit resets when the grid potential of *V2* rises sufficiently.

In this circuit the charging of condenser *C* determines the length of the output pulse, and it is important to remember in such cases that the charge must be reset at the end of the pulse before the circuit is ready to fire again. In this circuit the recharging of the condenser is rapid, because at the end of the pulse the grid of *V2* is driven positive, the cathode potential cannot follow because R_5 is small, grid current therefore flows and the charging time constant is R_1C. This is normally less than the time constant R_4C which controls the pulse length, and therefore the circuit recovers in a time short compared with the pulse length.

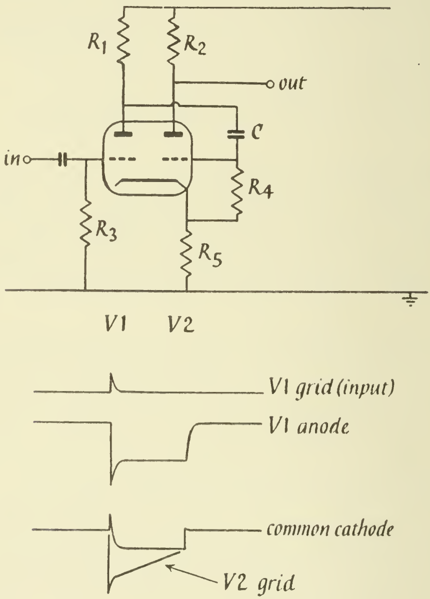


FIG. 38.—Kipp relay.

A defect of this circuit is apparent if a positive initiating pulse applied to the grid is followed by a negative overswing of the base line (see § 6.6). Quite a small negative excursion is sufficient to cut off the current which has just been established in *V1* with the result that the circuit resets prematurely giving a shortened output pulse. This behaviour of the circuit is similar to that of a Schmitt trigger with low backlash, and the analogy with Fig. 37 will be apparent.

The defect can be avoided by biasing the grids positive as in Fig. 39, and increasing the cathode resistance correspondingly. The backlash is now larger, and the current in *V1* is affected only slightly by a negative overswing at the input. A disadvantage is that now *V2* will no longer be driven into grid current when the circuit resets, so that unless a long recovery time can be tolerated a diode must be added across

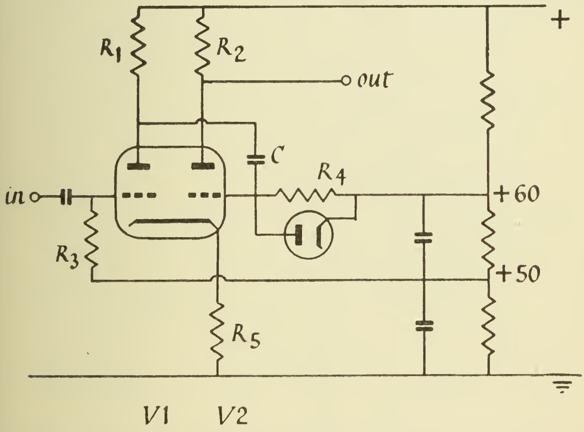


FIG. 39.—Kipp relay with grid voltages raised to avoid resetting by a negative overswing at the input.

R_4, making the charging time constant still R_1C. The circuit of Fig. 38 is to be preferred on the grounds of simplicity when it can be used, and moreover it has no tendency to reset prematurely if the triggering signal is applied to the anode of *V1* (negative pulse) instead of to the grid (positive pulse).

In another modification of Fig. 38, the leak resistance R_4 can be taken to the h.t. line, or connected across condenser *C*. Because the cathode resistance R_5 is small *V2* draws grid current and limits the rise in grid potential. This circuit is useful when short output pulses are desired, and generally gives a more stable output pulse length (cf. § 3.1 and Fig. 73).

4.5. Bootstrap trigger circuit, Fig. 40. In the circuits described so far a positive output pulse is available at the anode of the valve which is turned *off* by the trigger action. In the present circuit a positive pulse is generated at the cathode of a valve which is turned *on*: a heavy current can therefore be passed for the duration of the pulse, so that the

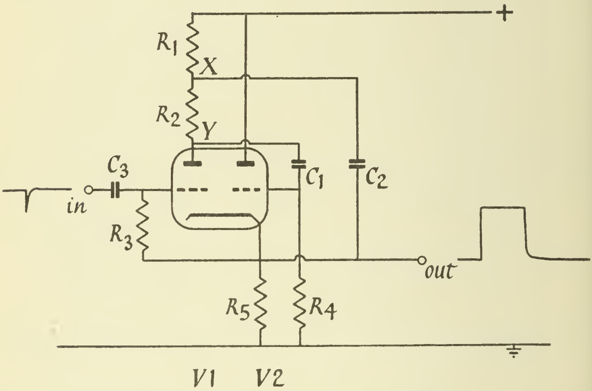


FIG. 40.—Bootstrap trigger circuit.

load resistance can be small, and, moreover, stray capacities at the output are charged rapidly by cathode follower action. This circuit is therefore well adapted to producing a large, rapidly rising positive square pulse.

The circuit is another variant of the cathode coupled multivibrator. In the quiescent state *V1* conducts drawing sufficient current through the cathode resistance R_5 to keep *V2* cut off. When a negative initiating pulse is applied to the grid of *V1*, its anode rises and turns on *V2* which conducts heavily, raising the common cathode potential with the result that *V1* is completely cut off. A positive output pulse appears at the cathodes.

A special feature of this circuit is the inclusion of the large condenser C_2 and resistance R_1 which gives rise to the following action. The output pulse is transmitted through C_2 to point X, and thence to Y, and back to the grid of *V2*. Thus the positive output pulse produces a positive signal at the grid, which in turn makes the output more positive and so on. There is positive feedback (although the loop gain is less than 1), and the result is that the final output voltage can be much greater than the voltage initially developed across resistance R_2. The circuit ' pulls itself up by its own boot-straps ' : hence the name *bootstrap* circuit.

To analyse the behaviour more exactly we note that the point X is tied to the cathode of *V2* through condenser *C2*, while point Y is tied to the grid through condenser C_1. This means that any change of voltage between X and Y must appear between grid and cathode of *V2* irrespective of any changes of these potentials relative to earth. The result is that the full voltage developed across R_2 is applied between grid and cathode of *V2* and amplified. There is no degeneration or negative feedback as in the cathode follower, and the amplified output appears across R_5. The *bootstrap* connection thus allows *V2* to function as an amplifier instead of a cathode follower. This useful principle has many applications (see §§ 5.4, 6.10).

In a typical circuit the change of voltage across R_2 is sufficient to drive *V2* into grid current, and this grid current continues to flow although the cathode potential rises. *V2* therefore conducts heavily and develops a large output pulse across the comparatively small load resistance R_5.

The resetting of the circuit is controlled by the conditions at the grid of *V1*. The grid voltage is at first held down near earth potential when the circuit is triggered by the restraining influence of C_3 (which we assume connected to some low impedance signal source), but rises gradually towards the new cathode potential as charge flows into C_3 through R_3. Eventually *V1* will conduct again and the cumulative action is reversed. During the pulse, current is flowing through condensers C_1 and C_2. These should be large to avoid voltage changes across them, but inevitably some time will be required after the pulse before the charge is readjusted and the voltages return to their quiescent values.

4.6. Transitron, Fig. 41. This is a one-valve trigger circuit, and depends on an interaction between screen and suppressor grid potentials. If the suppressor grid of many pentodes is made sufficiently negative the electron current is diverted from the anode, and flows instead to the screen.

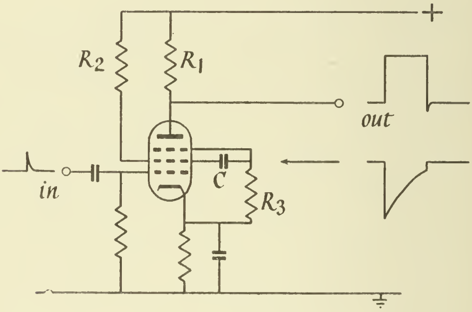


FIG. 41.—Transitron.

The screen current, flowing in its load resistance R_2, increases and therefore the screen potential falls. A negative signal on the suppressor grid therefore drives the screen negatively, and with suitable operating conditions the system will amplify. If the gain is greater than unity a trigger circuit can be formed by interconnecting the screen and suppressor grids by means of the condenser C as shown in the figure.

In the initial state the suppressor is at cathode potential and most of the valve current flows to the anode. The screen is at its most positive level. The circuit can then be triggered by means of a positive pulse at the grid, which increases the total current and thus causes the screen potential to drop. This negative front is transmitted to the suppressor through condenser C, and by cumulative action the screen and suppressor voltages are driven down, all the current flows to the screen, and a positive pulse appears at the anode. This situation is, however, not stable because the suppressor voltage returns towards cathode potential as condenser C charges through the leak resistance R_3. After a time of the order R_3C the anode again starts to draw current and the circuit resets. Now the screen potential rises and the suppressor is driven positive, drawing ' grid ' current, and the circuit returns to the initial state when the charge on condenser C has been readjusted.

This initial state is usually stable because the amplification between suppressor and screen is small when the suppressor is near cathode potential. Owing to the shape of the suppressor grid characteristic the amplification becomes greater than unity only when the suppressor is slightly negative, a situation which is produced by the triggering pulse.

The apparent economy of the transitron circuit, which requires only one valve instead of two, is off-set by the fact that this valve is a pentode. The existence of double triode valves means that only one valve need be used for the more orthodox trigger circuits and these have the advantage of greater flexibility. A difficulty with the transitron is that the suppressor characteristics of most valves are not standardized, so that considerable variation can occur between valves of the same type. Special valves with short suppressor base characteristics are however available and with these the circuit is completely reliable.*

* For a fuller discussion of the transitron circuit see O. S. Puckle, *Time Bases*, Chapman & Hall, London, 1952.

CHAPTER V

TIME BASES

THE function of the time base circuit is to produce a voltage which changes gradually and, for preference, linearly with time. The voltage excursion may be spontaneously recurrent, as in the free-running time base, or intermittent as in the triggered time base. A triggered time base is usually a combination of a time base circuit with a trigger circuit such as those already described, but it is not always easy to draw a dividing line between the two parts. The discussion in this chapter will deal with the basic time base methods, while examples of the more complex circuits appear in §§ 7.4 and 7.5.

For many applications it is desirable to have a linear time base waveform, and it will be seen how this can be achieved. Once having obtained a linear sweep, it is important to avoid distortion by subsequent coupling circuits. In particular, we recall the result of § 1.3 that slight differentiation by a RC-coupling combination will give rise to non-linearity. This effect can be minimized by using a very long coupling time constant, or avoided completely by direct coupling from the time base generator to its point of application.

5.1. Thyratron time base, Fig. 42. This is a free-running time base in which condenser C is in turn discharged slowly through resistance R_2 and then charged rapidly through the thyratron. With the output voltage initially close to the h.t. potential, and the thyratron extinguished, the condenser discharges through R_2 giving an exponential waveform at the output with time constant R_2C. When, however, the cathode potential of the thyratron has fallen sufficiently, the thyratron will strike and return the output voltage nearly to the h.t. potential. If R_2 is large there will then be insufficient current to maintain the discharge, the thyratron will go out and the cycle will repeat. The cathode voltage at which the thyratron strikes is regulated by the setting of the potentiometer R_1 which controls the grid voltage. R_1 therefore controls the amplitude of the generated saw-tooth waveform. If the amplitude is kept small, so that only a small initial portion of the exponential is used, then the time base will be approximately linear.

To linearize the time base for larger amplitudes the lower end of R_2 can be connected not to earth, but to a high negative potential which is often available from the cathode ray tube power supply. In this case the condenser C discharges exponentially towards this high negative potential and the

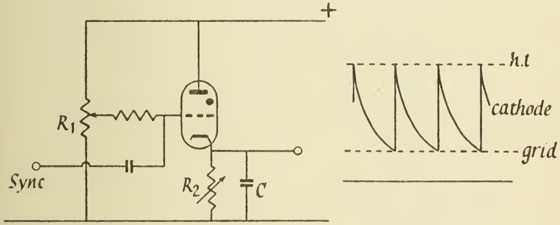


FIG. 42.—Thyratron time base.

early part of the curve will be more nearly linear, because it represents a smaller fraction of the possible voltage excursion.

Another method of obtaining a linear time base is to use a constant current device in place of the resistance R_2: if a constant current flows into the condenser the voltage across it must fall at a constant rate. Any of the devices discussed in §§ 2.7, 2.8 are suitable. With a constant current pentode we reach the circuit of Fig. 43 in which R_3 controls the charging current, and therefore the speed of the time base. Linear sweeps of several hundred volts amplitude can readily be obtained with this circuit.

The thyratron time base is limited in frequency because of the de-ionization time of the valve. The thyratron will strike prematurely if voltage is re-applied too soon after an initial discharge. Typically this makes the maximum operating frequency about 10 Kc/s : hard valve time bases must be used if more rapid time base wave forms are to be generated.

5.2. Synchronizing. It is frequently desired to synchronize a time base so that it keeps exactly in step with a repetitive waveform, and gives a stationary picture on the oscilloscope.

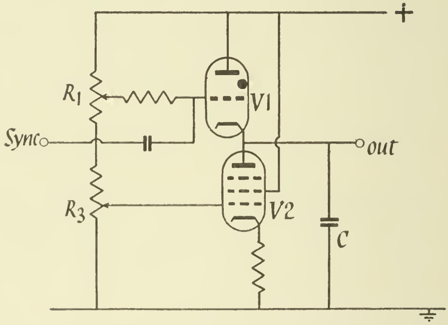


FIG. 43.—Thyratron time base with constant current valve.

More generally it may be necessary to synchronize a free-running oscillator of any type (e.g. a square wave generator), to a given reference signal. The nature of the synchronizing process is always the same, and will be discussed here using the thyratron time base as an example.

The reference waveform is applied to the oscillator as a synchronizing signal in such a way as to force the oscillator into step. The two waveforms will remain in step only if the two frequencies are identical or in simple numerical ratio. It follows that if the natural frequency of the oscillator is incorrect the synchronizing signal must alter it. In the case of a free-running time base this can be done in two ways, (i) by altering the amplitude of the sweep, or (ii) by altering the rate of sweep. In most circuits it is the amplitude that is altered by the synchronizing signal, but in some television receivers the synchronism is effected by controlling the mean rate of sweep in a somewhat complex circuit.

In the thyratron time bases of Figs. 42 and 43 the synchronizing signal is applied through a condenser to the grid of the valve, as shown. The action is indicated in Fig. 44 in which we postulate a sinusoidal reference wave and assume for simplicity that the thyratron fires when grid and cathode voltages are equal. Consider an initial firing point *A* and

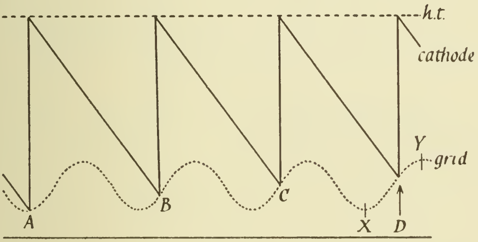


FIG. 44.—Synchronizing the thyratron time base.

uppose that the time base is running too slow, so that the uccessive firing points become later in phase as indicated t *B* and *C*. The firing point therefore slowly climbs up the ising part of the sine-wave, and as it does so the amplitude f the time base is reduced and its frequency therefore ıcreases. If the adjustment required is not too great the wo frequencies will coincide when the firing point is say at *C*, nd the time base will then fire regularly at the same part of ıe reference wave. A type of equilibrium is thus set up, nd the argument just given shows that the equilibrium is able. The time base will adjust itself to small changes in ıe reference frequency by automatically shifting the firing oint between the limits indicated by *X* and *Y*. If, however, ıe firing point comes on the falling part of the reference wave e situation is unstable and it will rapidly move over to the ext rising portion.

It appears from this discussion that the phase at which the time base locks will depend on its frequency relative to that of the synchronizing signal. If the natural frequency of the time base is correct it will be firing when the signal voltage passes through zero. But as either frequency is varied there is a gradual change in relative phase until when the limit X or *Y* is reached the time base no longer locks. This behaviour can be observed on most oscilloscopes: the stationary

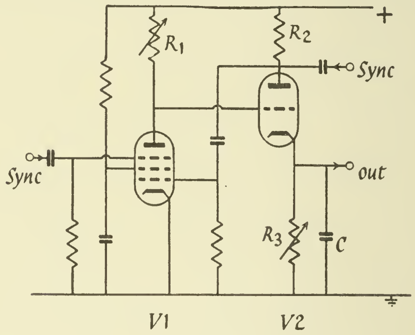


FIG. 45.—Puckle time base.

pattern shifts sidewards on the screen if the time base frequency control is varied until finally the variation is too great and the synchronism is lost.

5.3. Puckle time base, Fig. 45. This is a free-running time base in which a hard valve trigger device replaces the thyratron of Fig. 42. The circuit is based on the multivibrator (§ 3.1), but uses a direct coupling link between one anode and the opposite grid. The time base waveform appears on the cathode of *V2* and is due to the discharging of the condenser *C* through resistance R_3.

Suppose that initially this output voltage is near h.t. potential, *V1* is conducting giving a potential drop of a few hundred volts across R_1, so that the grid of *V2* is negative with respect to cathode and *V2* is cut off. This situation persists while the output voltage falls with time constant R_3C until the cathode potential of *V2* has fallen nearly to its grid potential. *V2* then starts to conduct, and by cumulative action *V1* is cut off, its anode voltage rises to h.t. potential, and the condenser *C* is charged through *V2*, which is now fully conducting. As the output potential approaches the h.t. potential, however, the current through *V2* gradually gets less and less, and eventually the potential across R_2 is insufficient to keep *V1* cut off. *V1* then conducts, the trigger circuit resets, and the initial conditions are restored. It will be seen that the *RC*-coupling time constant between *V2* and *V1* does not affect the operation provided it is longer than the fly-back time, i.e. the time taken to charge condenser *C* through *V2*.

The output voltage at which the trigger action starts is controlled by the grid potential of *V2*, and can therefore be adjusted by changing the value of R_1 or by changing the current through *V1*. Resistance R_1 therefore becomes the control of time base amplitude, and the time base is synchronized by using the reference signal to control the current in *V1*. This synchronizing signal can be connected either to the anode of *V2* (when it will reach the grid of *V1* via the *RC*-coupling), or preferably to a subsidiary grid of *V1*. Thus in *V1* the grid and suppressor can both be used, one connected to the synchronizing signal and the other to the anode of *V2*.

In the circuit given the time base waveform would be exponential. For a linear time base R_3 is replaced by a constant current valve as described in § 5.1 (see Fig. 43). The negative pulse which appears at the anode of *V2* during the fly-back can be taken to the cathode ray tube, to turn off the electron beam during this period and ensure that the flyback is invisible on the screen.

5.4. Bootstrap time base. Another method of obtaining a linear time base is shown in Fig. 46. In this circuit condenser C_1 is charged slowly through resistance R_1 generating a positive going time base sweep, and finally the condenser is discharged through valve *V1*. The action is controlled by a square wave applied to the grid of *V1* from a square wave generator (not shown). This circuit is primarily adapted to producing an intermittent time base sweep, so we shall assume an initial quiescent state in which *V1* is conducting and the potential at its anode (point *B*) is near earth potential.

When *V1* is cut-off by a negative square pulse at its grid, the current flowing through R_1 can no longer flow through the valve, but flows instead through condenser C_1, and the

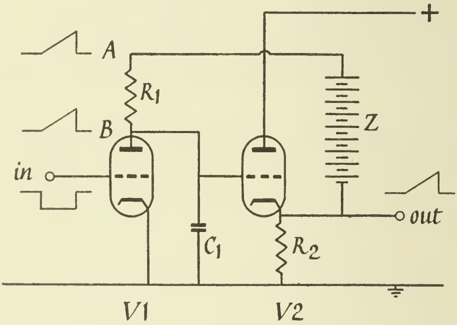


FIG. 46.—Bootstrap time base—principle.

potential at *B* slowly rises. If the upper end of resistance R_1 (point *A*) were connected to a fixed h.t. potential the condenser would charge exponentially in the usual way. To obtain a linear time base we require the current flowing through R_1 to be constant : this means that the voltage across R_1 must also be constant, that is, the waveform that is generated at *B* must appear also at *A*. This is achieved by connecting *A*, not to the h.t. line, but indirectly to the cathode of *V2* as shown in the diagram.

Valve *V2* is a cathode follower : the time base waveform applied to its grid therefore appears also at its cathode. Because point *A* is connected to this cathode via the battery *Z* there is a constant potential difference between *A* and the cathode, and the desired result is achieved. Ideally the voltage across R_1 is exactly constant, but this is not quite true because the gain of the cathode follower is not exactly 1. With a gain just below 1 the voltage at A lags slightly, the current through R_1 therefore falls slightly and the time base is slightly non-linear.

The path of the charging current is as follows. Starting at the positive h.t. line the current flows through *V2* to the cathode, through the battery Z, through resistance R_1,

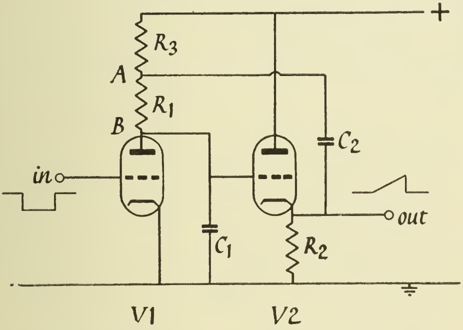


FIG. 47.—Bootstrap time base—practical circuit.

through the condenser C_1 and so back to the negative pole of the h.t. supply.

In practice it is usual to replace the battery Z by a large condenser which acts in effect as a constant voltage device for the duration of the time base. The circuit is Fig. 47. In the quiescent period current flows through R_3 and R_1 and through *V1* to earth. During the sweep the current through R_1 continues as before but the current through R_3 decreases, the difference coming from the condenser C_2. The discharge rate of this condenser is therefore governed initially by the time constant R_3C_2 (independent of R_1); this time constant must be large enough to ensure that the voltage change across C_2 during the sweep is small: otherwise the voltage across R_1 will not be constant and there will be non-linearity.

5.5. Miller time base, Fig. 48. This is an outstanding linear time base circuit which has wide application in electronics. The linearity can be of a high order, and, moreover, single valve is used both to control the charging of the condenser and also to discharge it. A square pulse applied to the suppressor grid of the valve actuates the mechanism.

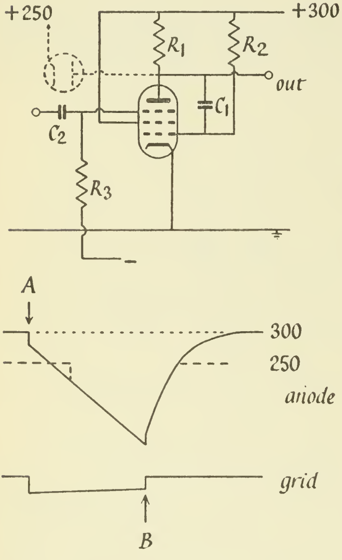


FIG. 48.—Miller time base. Anode-catching diode shown dotted.

Suppose that initially the suppressor grid is held negative so that no current flows to the anode. The anode is at h.t. potential, the grid at zero potential with grid current flowing, and the valve conducts with all the current flowing to the screen. When the suppressor is driven up to zero potential by a positive square pulse, current flows to the anode and the anode potential falls suddenly. But not much : because the fall in anode potential is transmitted through C_1 to the grid and tends to reduce the anode current to zero. Equilibrium is reached with the valve nearly cut off, but drawing sufficient current to give the necessary small voltage drop at the anode. This voltage drop is therefore nearly equal to the grid cut-off potential of the valve, say 5-10 volts. (Point A on the waveform diagram.)

This equilibrium situation is, however, not permanent because the current through R_2 (which originally flowed to earth as grid current), must now flow through C_1. The voltage across C_1 therefore gradually falls and as it does so the grid voltage rises slightly and the anode voltage falls. In fact, however, the change in voltage across C_1 appears almost wholly at the anode rather than at the grid, because the interaction between anode and grid via C_1 leads to a transient equilibrium situation at every instant. If the grid voltage should rise too much, the anode voltage would fall, driving the grid negatively back to its equilibrium voltage. The net result is that if the amplification of the valve is A, a drop in anode potential V is accompanied by a small rise in potential V/A at the grid, and the potential change across the condenser is $V\frac{(1 + A)}{A}$.

The time base waveform generated at the anode is linear because condenser C_1 is charged by a nearly constant current: constant because the voltage across R_2, through which all the current must flow, is almost constant. In fact, for a sweep amplitude V the grid voltage changes by V/A, and this is a small fraction of the total supply voltage V_0 which is applied across R. The slope of the time base therefore changes from beginning to end only by the fraction V/AV_0. For example, with a valve giving a gain $A = 100$, and $V = \frac{1}{2}V_0$, the slope of the time base changes by only $\frac{1}{2}$ per cent. from beginning to end. The linearity can in principle be improved indefinitely by increasing the amplification, if necessary using a multistage direct coupled amplifier in place of the single valve.

The slope of the quasi-linear sweep at the anode is determined by condenser C_1, resistance R_2, and the voltage V_0 applied to R_2. To a good approximation the slope is $V_0/R_2C_1 \times A/(1 + A)$ if V_0 is large compared with the cut-off potential of the valve. It is practically independent of A, and therefore of R_1 and the valve characteristic, provided $A \gg 1$.

It is interesting to note that non-linearity of the valve characteristic does not disturb the output waveform. Even for a non-linear amplifier, the argument given above still demonstrates the linearity of the time base; and this implies that the waveform at the grid must be non-linear in such a way that when amplified it compensates perfectly for the non-linearity of the valve!

To complete the picture we now suppose that the positive square pulse applied to the suppressor grid ends when the point B on the waveform is reached. As the suppressor goes negative the anode current is cut off and the anode voltage rises sharply carrying the grid voltage up also until grid current flows. Thereafter the anode voltage can rise only exponentially as condenser C_1 is charged to its original voltage through anode load R_1, time constant R_1C_1. The flyback or recovery time of the Miller time base therefore tends to be long. It may be reduced by including a diode which prevents the anode voltage from rising to the full h.t. potential (called an *anode-catching* diode). This is shown by the dotted connections on the circuit, and corresponds to the dotted anode waveform. The rather long recovery time means that the simple Miller time base can be used only in intermittent operation, for example, as a triggered time base (see, however, § 7.4).

There is, of course, a limit to the amplitude of linear sweep that can be obtained with the Miller circuit. As the anode voltage reaches the ' knee ' of the pentode characteristic (about 50-100 volts) the valve current tends to flow to the screen rather than to the anode. Eventually a point is reached at which the anode voltage remains constant and the grid voltage then rises to zero. Alternatively if R_1 is small the grid current region may be reached before the anode voltage drops to the knee of the characteristic, and the linear sweep will end suddenly. In both cases potentials in the circuit will then remain steady until the pulse on the suppressor ends and the recovery phase begins.

A useful property of the Miller circuit is its low output impedance. We have already seen that the slope of the time base waveform is practically independent of the anode load R_1, provided the amplification A is large. This means that if another external load resistance is added in parallel with R_1 there will be very little change, i.e. the output impedance of the circuit is low. To make a quantitative argument we note that any voltage variations at the output anode are fed back to the grid through the coupling condenser C_1 (apart from the steady charging of the condenser which continues relentlessly generating the time base). The circuit is that of an amplifier with the whole of the output fed back negatively to the input. To find the output impedance we apply Thévenin's theorem (§ 1.5), that is, we calculate the impedance presented by the circuit to an external voltage source applied to the anode. Postulate $+1$ volt applied to the anode: this appears also at the grid, increasing the valve current by g. This current, together with the change in current through the anode load R_1 has to be supplied by the external voltage source; the apparent impedance is therefore $1/g$ in parallel with R_1: this is the output impedance, which can usually be taken as $1/g$ to sufficient accuracy. It thus appears that the output impedance of the Miller time base circuit is as low as that of a cathode follower, typically a few hundred ohms. This means that the circuit will develop its waveform with little distortion across small resistances and large stray capacities.

Stray capacities at the output of time base circuits are usually due to the deflector plates and base connections of a cathode ray tube, and tend to be large. In the previous circuits, in which the time base waveform is developed by charging a reservoir condenser, one end of which is connected to earth, the action of the stray capacity is simply to increase the size of this condenser, affecting the speed of the time base, but not its linearity. From another point of view this behaviour on capacitative load arises because the output impedance of the circuits is capacitative, not resistive.

In the Miller time base, however, the reservoir condenser is not connected to earth and the output impedance, as we have seen, is resistive ($1/g$). Together with the stray capacity this output impedance constitutes a standard integrating circuit, and the effects produced will be those described in § 1.2. There is an initial transient the duration of which is of the order of the integrating time constant (RC), and the output has a time lag equal to RC. The low output impedance of the Miller circuit, however, keeps these effects small. Taking typical figures, $R = 200\ \Omega$, $C = 100\ pF$, we find $RC = 0{\cdot}02\ \mu s$, which shows that the effects will normally be quite negligible. Integration of the Miller waveform is, moreover, slightly different from the case discussed in § 1.2 because of the short vertical step at the beginning of the time base, of amplitude say v. This can be used, in effect, to provide the delay required by the integrating circuit so that if the integrating time constant is correctly chosen, $RC = R_2C_1 \times v/V_0$, the output waveform rises at the correct rate from the very start and the initial transient is eliminated. The practical value of this trick does not, however, appear to be great.

In conclusion we note that the starting and stopping of the sweep in the Miller time base may be achieved in a variety of ways, of which the suppressor grid control used above is only one. The circuit may also be used to obtain the time integral of a waveform, because the output voltage at any moment is almost perfectly the time integral of the voltage V_0 applied to resistance R_2. This is especially true if the single valve is replaced by a multi-stage direct coupled amplifier. A similar circuit in which R_2 and C_1 are interchanged can be used for exact differentiation. A trigger circuit which combines the basic principles of the transitron (§ 4.6) and the Miller time base has been named the *phantastron* and has given rise to a whole family of derivative variations.* These are used mainly as triggered time bases, or to generate pulses of variable duration, linearly and accurately related to a pre-set control voltage (cf. § 7.4).

5.6. Time base with movable fast region, Fig. 49. This circuit has the effect of horizontally magnifying a small part of the picture on the cathode ray tube by speeding up the time base sweep over this region. It is as though a magnifying cylindrical lens had been placed over the picture. The

* See O. S. Puckle, *Time Bases;* D. Sayre, Generation of Fast Waveforms in *Waveforms,* M.I.T. Radn. Lab. Series, No. 19.

fast region can be adjusted in position to give a detailed view of any desired region of the sweep. This is achieved by means of the long-tailed pair formed by *V2* and *V3*.

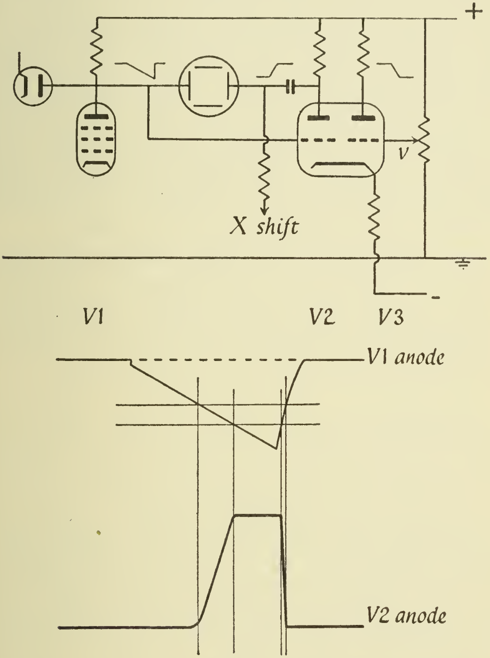


FIG. 49.—Time base with movable fast region.

V1 is assumed to be a Miller time base circuit giving a negative going linear sweep which is applied direct to the *X1* plate of the cathode ray tube and produces the main time base. The sweep waveform is also applied to the grid of *V2* which is conducting at the beginning of the sweep because the grid of *V3* is more negative. At a certain point in the sweep, however, determined by the bias voltage v, applied to the grid of *V3*, the current in the long-tailed pair is switched over and

V2 becomes non-conducting. The anode voltage of *V2* is therefore initially steady at a low potential, and rises as *V2* is cut off to a new steady level as shown in the diagram. The rising region is an amplified version of a portion of the time base waveform; and while showing some distortion due to curvature of the valve characteristics, it is tolerably linear over its main excursion. This waveform is applied to the *X2* plate of the cathode ray tube : it has no effect on the main sweep while the voltage is steady, but during the rapidly rising region it moves the spot rapidly forward, producing the speeded up section. The position of the fast region can be varied by adjusting the bias voltage v, and so a detailed examination of any part of the trace is possible.

In other applications the signal derived from the long-tailed pair can be processed by subsequent circuits and used to provide marker pulses, gating signals, separate fast time base displays, etc., all variable in position by means of the bias voltage v. One such case is treated in § 7.3.

5.7. Calibration of time bases. In the case of a free-running time base which can be synchronized by an externa signal the time calibration is easily carried out. One simpl observes a signal of known frequency on the cathode ray tube

For a triggered time base the problem is more difficult Usually the time base cannot be triggered reliably by a hig frequency repetitive waveform, because the circuit does no have time to recover fully after one sweep before the nex triggering signal arrives. The result is a hazy picture whic jitters to and fro on the screen. To obtain a clean pictur the calibrating waveform must be broken up into blocks o oscillation followed by quiescent periods. This may b achieved, for example, by applying the continuous oscillation to the grid of a valve and a square *gating* waveform to th suppressor grid (§ 7.8). Then no anode current flows whe the suppressor is negative, and a 100 per cent. square wav modulation of the oscillation results. The time base can the be triggered by the first of a block of oscillations giving a clea picture of the following calibration cycles, and the circui is allowed sufficient time to recover again during the next quiescent period, so that the traces superimpose perfectly.

An alternative method of calibration is to use a *ringing circuit*. This is a tuned circuit which is shocked into oscillations of known frequency at the beginning of the sweep. In the circuit of Fig. 50, the steady current i flowing through the tuned circuit in the anode lead is suddenly cut off when the negative square pulse is applied to the grid. The circuit then gives out a train of damped oscillations at its natural frequency, with initial amplitude $i(L/C)^{\frac{1}{2}}$. If the circuit is shocked into oscillation by a square wave derived from the

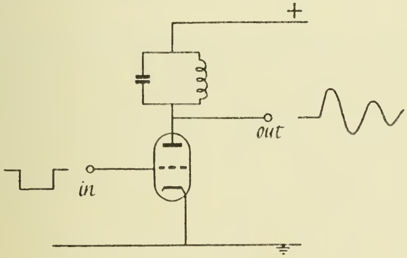


FIG. 50.—Ringing circuit.

time base circuit, the output waveform will be in perfect synchronism and can be used for calibration. The ringing circuit is adjusted to the desired frequency by tuning to resonance with a subsidiary signal generator.

Ideally the output of a ringing circuit is taken to the cathode ray tube via a cathode follower, so that the circuit is isolated from large and potentially variable stray capacities. A positive feedback coupling coil may also be included to overcome the natural damping in the circuit and in this way a very long train of shocked oscillations may be obtained (there is no danger of permanent continuous oscillation because the circuit is damped when the control valve is again allowed to conduct between the sweeps). But these are refinements: the circuit of Fig. 50 is often satisfactory, and particularly useful for rough calibration of an intermittent fast time base.

CHAPTER VI

PULSE AMPLIFIERS

THE amplification of pulse waveforms has been implicit in the previous chapters. All trigger circuits, for example, include an amplifier, and for proper operation it is essential that the amplifier responds satisfactorily to the pulse inputs. Some basic principles of pulse amplification were mentioned in § 2.3, and we now resume the subject in more detail.

There are two distinct but complementary ways of assessing the performance of a pulse amplifier or transmission system. In the first we calculate the shape of the output transients directly using the concepts of integration and differentiation discussed in §§ 1.2, 1.3. This method gives directly the shape of the output waveform, but becomes virtually impossible for a multi-stage amplifier because the input to each stage is not a simple square or triangular waveform, but is the distorted output from a previous stage. It is nevertheless fruitful to examine the response of each stage in turn to a hypothetically perfect input pulse; because this may reveal the weakest link in the chain, and will certainly lead to a fair estimate of the overall performance.

In the second method the input waveform is decomposed into sinusoidal Fourier components. The amplification and phase shift for each component is measured for the complete amplifier using a sinusoidal signal generator, or calculated using the familiar techniques of vector algebra, and the resultant output signals are then, in theory, recombined to yield the final output waveform. The Fourier analysis and synthesis is complex in the general case, but for practical amplifiers and for square input waveforms it has been possible to formulate some general working rules relating the frequency response of the system to the shape of the output pulse. The method can be applied with equal facility to an amplifier of any number of stages, provided the frequency and phase characteristics are known.

These two approaches to the problem of amplifier analysis, one based on transient response and the other on frequency response, would both lead to the same result if they were followed through exactly. In practice, however, for a multi-stage amplifier, neither method can be used rigorously, although each is sufficient to give a good estimate of the overall performance. It is as well to bear in mind both systems, and to use whichever promises to give the quicker solution to a particular problem.

6.1. Fourier spectrum of a square pulse. We will consider first the amplification of a square pulse of duration T, as shown in Fig. 51*a*. If there is only a single pulse, Fourier analysis (more exactly, Fourier transform) yields a continuous spectrum of component frequencies as shown in Fig. 51*b*. There is a d.c., or zero frequency, component because the mean potential of the waveform is not zero, and as the frequency rises the amplitude gradually falls reaching zero at a frequency $1/T$. Beyond this there are small contributions from higher frequency components as indicated.

This diagram is identical to that obtained in physical optics for the Fraunhofer diffraction pattern of a slit aperture. In fact, it can be readily proved that the Fraunhofer diffraction pattern of any aperture is always the Fourier transform of its transmission characteristic. In this case, the transmission characteristic of a slit of width d would be as indicated in Fig. 51*a* and the diffraction pattern, Fig. 51*b*, has its first zero at an angle λ/d. This correspondence between diffraction and Fourier analysis may help to establish the pattern in one field if that in the other is already known.

For a regularly recurring pulse of duration T, recurrence frequency f, the Fourier series analysis is appropriate. Only frequencies f, $2f$, $3f$, . . . etc., are now present in the spectrum, which splits up into a series of lines as shown in Fig. 51*c*. The envelope of these lines is the same as Fig. 51*b*, and corresponds to the shape of the individual pulse, while the fine line structure reflects the recurrent nature. If the mark-space ratio is small the frequency $1/T$ for the first zero is much larger than f and the line structure is comparatively fine, merging in the limit into the continuum of Fig. 51*b*.

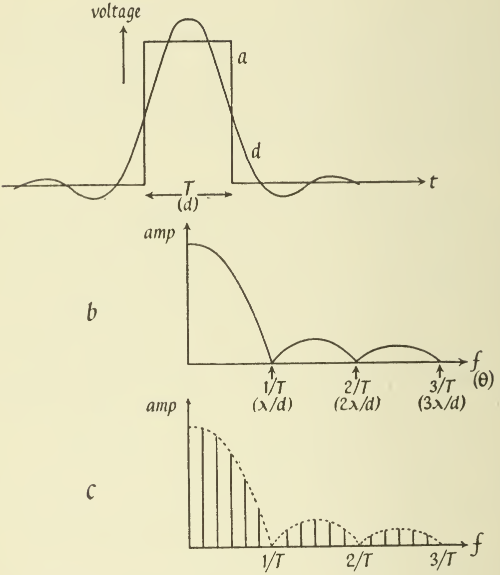


FIG. 51.—(*a*) square pulse.
(*b*) Fourier spectrum of (*a*).
(*c*) Fourier spectrum of regular train of pulses.
(*d*) theoretical output pulse if frequencies up to $1/T$ only are transmitted and there is no phase shift.

Ideally the whole spectrum extending up to infinite frequency must be transmitted if the pulse is to be reproduced exactly. It is found in practice, however, that the pulse is reproduced tolerably well if only frequencies up to the first zero at $1/T$ are transmitted. If the frequencies higher than this are excluded the output pulse is as shown in Fig. 51*d*.

The pulse is lengthened slightly and the sharp corners are rounded because of the absence of the higher frequency components. But it is still recognizably a pulse, recognizably like the original. We therefore have the practical criterion for the reproduction of a square pulse of duration T that the amplifier must transmit frequencies up to $1/T$ at least. If the pass band is smaller than this the output pulse will be further rounded and lengthened and reduced in amplitude: if, however, even higher frequencies are transmitted the pulse will be more nearly square, its sides steeper, and will generally approximate more closely to the original. The exact criterion to be adopted in a particular problem will depend on how faithfully the input pulse must be reproduced. If the presence of a pulse at the output is all that is required it will be sufficient to pass frequencies up to $1/T$. But if steep sides and a square topped pulse are required then the pass band must extend to many times this value.

It follows that an amplifier with upper cut-off frequency f_2 will reproduce pulses of duration $1/f_2$ tolerably well: longer pulses will be reproduced more faithfully with steep sides and square tops, while shorter pulses will be seriously distorted and deficient in peak amplitude. This is indicated in Fig. 52.

The steepness of the pulse fronts at the output is independent of the pulse length and is governed by the highest frequency f_2 which the system transmits. The steepness is specified conveniently by means of the *rise time*, t_r, which is commonly defined as the time for the pulse to rise from 10 per cent. to 90 per cent. of its final amplitude. An approximate working rule relating the rise-time to the upper cut-off frequency f_2 (at which the amplitude response is $1/\sqrt{2}$ times that at mid-band) is

$$t_r = \frac{1}{3f_2} \qquad . \quad . \quad . \quad (13)$$

This relation is good to about 10 per cent. and applies to all types of amplifier. t_r is often referred to as the *rise time of the amplifier*. Relating this to our criterion $T = 1/f_2$ for the shortest pulses which the amplifier can handle we find that in this case the rise time at the output is $\frac{1}{3}$ the original pulse length : a result which is clearly right as to order of magnitude. The tail of the pulse will, of course, be exactly similar in shape (but inverted), provided the transmission system is linear.

It now appears that to reproduce a pulse of given length, or a pulse front of given steepness, frequencies up to some minimum value f_2 must be transmitted. The lower frequency limit f_1 is near zero (this will be discussed later, § 6.6), and it is therefore customary to refer to f_2 as the *bandwidth* of the system. For faithful reproduction of pulses and wavefronts

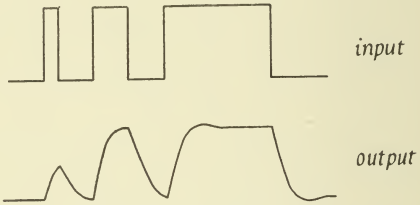


FIG. 52.—Response of an amplifier to input pulses of various widths, showing that the shorter pulses are more distorted.

the bandwidth must be large, and we therefore require *wide band* amplifiers, and must take appropriate steps to increase the bandwidth as much as possible. In the following sections we shall treat the design of wide band amplifiers.

6.2. Gain-bandwidth product. Because amplification over a range of frequencies is required the resistance coupled amplifier is the basis of all wide band design, which is in effect simply an extension of audio amplifier technique to higher frequency regions. For a pentode with anode load resistance R, slope g, the gain G at moderate frequencies is gR. At high frequencies, however, the anode load impedance is reduced by the shunting effect of the stray capacity C: this introduces a phase shift at high frequencies and causes the gain to fall off. The response is reduced to $1/\sqrt{2}$ when the impedances of R and C are equal, so we obtain for the bandwidth

$$B = f_2 = \frac{1}{2\pi RC} \quad . \quad . \quad . \quad (14)$$

Forming the product of bandwidth and gain we find

$$GB = \frac{g}{2\pi C} \quad . \quad . \quad . \quad (15)$$

independent of the value of the anode load resistance R. This relation indicates that the stage gain of an amplifier is, *ceteris paribus*, inversely proportional to the bandwidth. For example, to increase the bandwidth one must reduce the value of R, and the gain is inevitably reduced.

The stray capacity C consists of the anode to earth capacity of the valve, plus the input capacity of the following stage, together with the capacities in the valve bases and wiring. These last items add substantially to the capacities of the bare valve quoted by manufacturers and should not be overlooked. The total interstage stray capacity for an amplifier using high slope voltage amplifying pentodes is usually about 15-20 pF, but rises considerably if power output valves are used.

Equation (15) shows that the ratio g/C gives a measure of the potentialities of a particular valve as a wide-band amplifier. Because the other capacities are uncertain, only the input and output capacities of the bare valve are taken in this expression and in this form the ratio represents a figure of merit for the valve. In the main the slope of a valve can be increased only by increasing the size or number of grid wires, or by reducing the grid-cathode spacing: these changes all have the effect of increasing the grid to cathode capacity, with the result that the ratio g/C is not increased. It is a difficult problem to produce a valve with a better figure of merit, but nevertheless there is a slow but steady improvement as new types of high slope television pentode are evolved. For a typical advanced type at present * $g = 7{\cdot}4$ ma/v and

* Mullard EF80.

$C = 10{\cdot}8$ pF making g/C equal to $0{\cdot}69$ ma/v per pF so that GB is of order 100 Mc/s. This implies for example a gain of 2 per stage with a bandwidth of 50 Mc/s; but when wiring capacities are allowed for the actual performance will be worse by a factor $1{\cdot}5$—2. It should be noted that the slope of a valve increases with anode current; in a wide band amplifier, therefore, the valves are operated near maximum current, in contrast to the usual audio practice.

In this connection the cascode circuit of § 2.8 is important. We have already seen that two triodes connected in cascode, each of slope g and amplification factor μ, are equivalent to a valve of the same slope and amplification factor μ^2, and can in many respects replace a pentode. For wide band amplifiers this substitution is an advantage because the input and output capacities of the cascode circuit can be very low although the slope is comparable to that of a pentode. This arises because double triodes are available with much lower capacities than pentodes of the same slope and in the cascode circuit the Miller effect between anode and grid is almost eliminated. A typical double triode,* designed for V.H.F. oscillator application, has grid-cathode capacity $2{\cdot}2$ pF, grid-anode capacity $1{\cdot}5$ pF and anode-cathode capacity $0{\cdot}5$ pF. For the input side of the cascode circuit the grid-anode capacity is doubled by the small Miller effect (amplification 1), so that the input capacity is $5{\cdot}2$ pF giving a total $C = 7{\cdot}2$ pF. With a slope of $7{\cdot}2$ ma/v † this yields a g/C ratio $1{\cdot}00$ which is better than that of our typical pentode. The advantage does not appear so great when the wiring capacities are included, but nevertheless the cascode circuit using double triodes in place of pentodes can be recommended for all wide band amplifiers.

The possibility of using two valves in parallel to increase the value of g is sometimes envisaged. This does not increase the ratio g/C for the valves themselves because the capacities are also doubled, and therefore the gain per stage remains substantially the same as before. It is far better to employ

* R.C.A. 12AT7; Mullard ECC81.

† Triode and pentode are compared at the same cathode current.

the extra valve as a second stage in cascode if additional gain is required. But parallel operation may be desirable in an output stage if the fixed load capacity is larger than the inter-electrode capacities in the valve. In this case with two valves the slope is doubled, but the value of C increases only slightly: the result is that a larger stage gain, and larger output voltage swing can be achieved.

These considerations govern the design of the amplifier, enable its bandwidth to be calculated, and using equation (13) lead to an estimate of the rise time at the output. The alternative approach based on transient analysis leads to substantially the same conclusions. Here in a single resistance coupled stage, anode load R, stray capacity C, the voltage rises exponentially to its final value with time constant RC. The rise time t_r from 10 to 90 per cent. is therefore 2·2 RC (cf. § 1.2) which agrees reasonably well with the value 2·1 RC deduced by applying equations (13) and (14). For a multistage amplifier the overall rise time T_r is obtained from the sum of the squares of the rise time t_r of the individual stages.

$$T_r^2 = \sum t_r^2 \qquad . \quad . \quad . \quad (16)$$

These considerations will inter-relate the gain per stage and the rise time (or bandwidth) in the same way as before.

A useful application of equation (16) is to find the rise time at the output of an amplifier, rise time t_r, when the input pulse is not perfect, as assumed so far, but itself has rise time t_r'. Applying the equation we find immediately that the rise time at the output is $\{t_r^2 + t_r'^2\}^{\frac{1}{2}}$.

6.3. Inductive compensation. So far we have assumed that the coupling between valves is by means of a simple anode load resistance. The gain-bandwidth product can be improved, however, if inductances are included in the coupling circuit, the simplest of these modifications being shown in Fig. 53. The added inductance L is chosen to resonate with the stray capacity C near the upper frequency limit, thus increasing the circuit impedance and maintaining the gain over a wider range of frequency. The system forms a damped tuned circuit which responds to a transient current front as indicated in Fig. 53*a*. With no inductance the voltage rises exponentially, curve 1; if $L = \frac{1}{4}R^2C$ the circuit is critically damped and the approach to the final voltage is more rapid because the inductance tends to maintain a constant current as the condenser C charges, curve 2; for larger values of L the circuit is underdamped and there is a

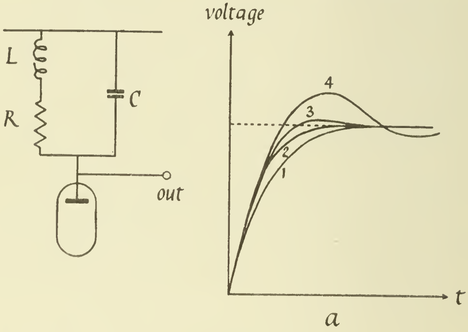


FIG. 53.—Inductive compensation.
(*a*) response to a square input front.
1. no inductance.
2. $L = \frac{1}{4}R^2C$, critical damping.
3. $L = 0{\cdot}414R^2C$.
4. $L = R^2C$.

tendency to oscillation. This at first causes a slight overshoot and for still larger inductances several cycles of damped oscillation will become apparent, curves 3 and 4. The initial slope of the transient is the same in all cases because at this point all the available current is flowing into the condenser C.

The condition $L = \frac{1}{4}R^2C$ is known as critical compensation and gives the maximum value of inductance for no overshoot. In practice a small amount of overshoot is usually tolerated and values of L range from $\frac{1}{4}$ to $\frac{1}{2} \times R^2C$. A value of $0{\cdot}414 \times R^2C$, curve 3, gives a flat frequency response of the greatest bandwidth, while for larger values there is a peak at the end of the response curve before the final drop at higher frequencies. This gives a convenient criterion for adjusting the compensating inductance, for if there is just no peak at the end of the frequency response curve there will be no more than a slight overshoot of the pulse fronts and the compensation will be near the optimum.

This type of inductive compensation is called *shunt compensation.* The gain bandwidth product of an amplifier stage is improved by a factor 1·41 for critical compensation

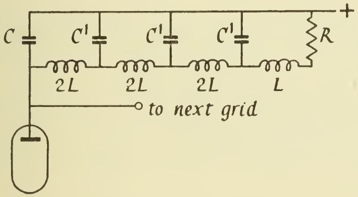


FIG. 54.—Inductive compensation with low pass filter as anode load.

($L = \frac{1}{4}R^2C$) and 1·72 when $L = 0{\cdot}414R^2C$. Formula (13) for the rise time still applies.

A further improvement can be effected by including more inductances in the circuit. The principle is illustrated by Fig. 54 in which a low pass wave filter is used as the anode load impedance. This is developed from Fig. 53 by adding a number of inductances $2L$ and the extra condensers C', each equal to the stray capacity C. Fig. 53 can then be seen as a rudimentary wave filter consisting of little more than one half-section. The point is that an infinite, or correctly terminated, wave filter of the type shown has an input impedance Z_0 constant in magnitude, though not in phase. $Z_0 = \sqrt{2L/C}$, over the whole pass band of the filter which extends up to a frequency $f_2 = 1/\pi\sqrt{2LC}$. The gain of the amplifier stage, gZ_0 is therefore constant over this band and we find that $GB = g/\pi C$, giving a two-fold improvement over the simple resistance coupled stage of equation (15). A fair approximation to this situation is obtained with a two section filter including only one added inductance $2L$ and one added condenser C', but even so it is doubtful whether the attempt to obtain the theoretical two-fold improvement in gain-bandwidth product is worth the extra complication when simple inductive compensation can itself yield a factor 1·72.

A worthwhile improvement is, however, obtained by separating the anode stray capacity C_a from the grid stray

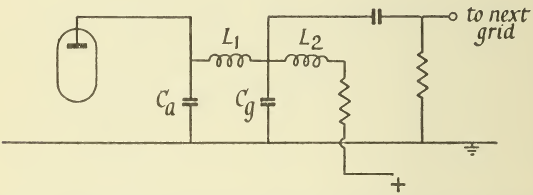


FIG. 55.—Inductive compensation with anode and grid capacities separated by an inductance: shunt-series compensation.

capacity C_g of the next valve using an inductance as shown in Fig. 55. In this case the added capacity C' of Fig. 54 is replaced by C_g and the initial stray capacity C (originally equal to $C_a + C_g$), is correspondingly reduced. In the simple case of equal anode and grid capacities this results in a further two-fold increase in gain-bandwidth product because condenser C has half its original value. In practice this benefit is not quite fully realized because the termination is not perfect and because the phase shift distortion introduced during transmission from anode to grid along the first section of the filter will increase the rise time. Unequal anode and grid stray capacities can be accommodated in this circuit by the appropriate choice of values for L_1 and L_2, and an m-derived terminating section is an advantage. The circuit of Fig. 55 is known as *shunt-series compensation*, and if inductance L_2 is omitted we have *series compensation.*

A further development of this idea, indicated in Fig. 56, allows many valves to be used in parallel and is known as *distributed amplification*. The input signal is applied to a terminated low pass filter and the grids, *G1*, *G2*, etc., of the valves are connected to successive elements as shown. The anodes *A1*, *A2*, etc., are connected to another low pass filter and the output is taken from one of the terminating resistances *R*. The capacitative elements of the filters are just the grid

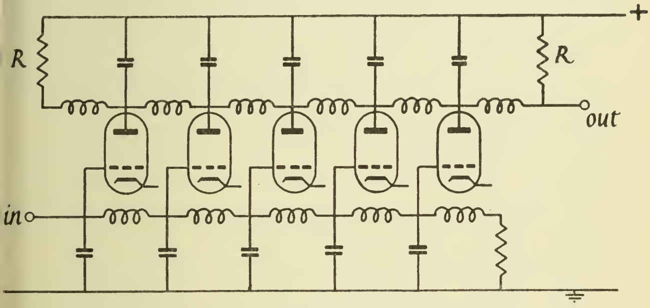


FIG. 56.—Distributed amplifier.

and anode stray capacities of the valves. The delays in the filters are carefully matched so that the signals passing via the various valves arrive simultaneously at the output and reinforce. In this way the effective slope of the combination of n valves is increased $n/2$ times (not n because half the current is wasted in the other terminating resistance), but the stray capacities do not add and the bandwidth remains the same as for a single valve. (If the stray capacities of the added valves were not present condensers would have to be introduced in making up the filters, cf. Fig. 54.) The result is that the gain-bandwidth product is improved $n/2$ times. Distributed amplification is useful when the bandwidth requirement is so large that a normal stage would have gain less than unity. A typical distributed amplifier using six valves per stage gives a stage gain of 3 with bandwidth 200 Mc/s. Rise times of order 2mμs (millimicrosecond) can be realized in this way.* The principle is not used at moderate bandwidths except for the output stage, because the gain of the valves multiplies in cascade connection but merely adds in a distributed amplifier.

6.4. Negative feedback. It is well known that negative feedback (see § 2.3) tends to flatten the response curve of an amplifier and so gives rise to an increased bandwidth. But this does not help to solve the main problem of the wide band amplifier because the increase in bandwidth is obtained at the expense of a reduction in gain. Negative feedback cannot increase the gain-bandwidth product per stage, and the results obtained are virtually equivalent to the appropriate reduction of anode load resistances.

But in the output stages it is often impossible to reduce the anode load resistance, because it is determined by the output voltage swing required and the current swing available from the valves. In fact, while the early stages of a wide band amplifier present some problems, the output stage often becomes almost impossible, requiring as it may very large current swings or resistances so large that the bandwidth cannot be achieved. It is here that negative feedback can provide a solution by increasing the bandwidth, although the load resistance and therefore the maximum output voltage remain unaltered.

Negative feedback is also useful throughout the amplifier when linearity or stability of gain is important, because with sufficient loop gain the characteristics of a feedback amplifier are largely independent of small curvatures and changes in the valve characteristics (see § 2.3).

6.5. Maximum rates of change at the output. So far we have been dealing with amplifiers working in their linear regions, but in pulse circuitry it often happens that a valve is turned hard on, or is completely cut off. In the output stages of an amplifier these situations represent the limiting

* See I. A. D. Lewis and F. H. Wells, *Millimicrosecond Pulse Techniques*, Pergaman Press, London, 1954.

conditions in which the output voltage rises or falls at its maximum rate. It is useful to compute and bear in mind these maximum rates because they clearly cannot be exceeded, and may be the limiting factor especially where large output pulses are expected. For example, if an output cathode follower can deliver a maximum of 10 ma into a load capacity of 100 pF, the maximum rate of rise of output voltage is $dV/dt = i/C = 100$ v/μs; it is clear that there is no hope of obtaining an undistorted 1 μs pulse of 100 volt amplitude at the output, although the rise time of the amplifier by a bandwidth measurement may be 0·1 μs and this may be realized for smaller output signals.

These considerations are particularly important in the case of negative feedback amplifiers for which the linear analysis can be seriously in error for large pulses. An elementary example of this has appeared in § 2.5, Fig. 21. In negative feedback systems, however, the ultimate possibilities of the output stage are usually well exploited; because if the output voltage does not accord with the input signal as related by the feedback network, the resulting error signal turns the output stage fully on or fully off as the case may be, and the output voltage is therefore corrected at the maximum possible rate. As a result, a negative feedback system will often amplify a pulse with little distortion, even when the rate of voltage change at the output approaches close to the theoretical maximum.

6.6. Low frequency response, paralysis. It was seen in § 6.1 that a single pulse contains Fourier components ranging down to zero frequency. If the lower frequency components are not fully transmitted, a distortion is produced, and this is analogous to differentiation (§ 1.3, Fig. 9). If the low frequency response is limited by only one capacity-resistance coupling the effect must of course be exactly the same as differentiation : the top of a square positive pulse then slopes negatively, and at the end of the pulse the output voltage swings negative and finally recovers slowly towards its quiescent level, Fig. 57*a*. If there is a second *RC*-coupling in the system this waveform is again differentiated, with the result that the negative overshoot of the base line is followed by a smaller positive overshoot before the quiescent stage is reached, Fig. 57*b*. With three *RC*-couplings there is a further small negative overshoot, and so on. There is a further traversal across zero voltage for each added *RC*-coupling stage, although it often happens that some of the oscillations are negligibly small in amplitude. To avoid these negative and positive excursions of the base line after a

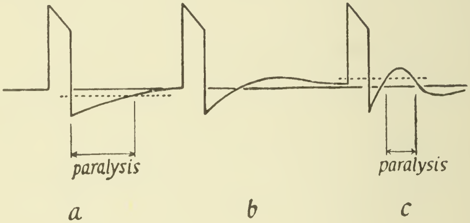


FIG. 57.—Paralysis due to differentiation of a large input pulse.
(*a*) with one differentiating circuit.
(*b*) with two differentiating circuits.
(*c*) with three differentiating circuits.

pulse the coupling time constants must be very much longer than the pulse length. In the case of a regularly recurrent pulse, the recurrence frequency is the lowest frequency present in the Fourier series analysis. Distortion will be completely avoided if this frequency is transmitted in full, so for a completely flat base line the coupling time constants must be longer than the recurrence period.

These distortions of the base line are particularly serious early in the amplifier if large overload pulses are present. The negative excursion following a large pulse may look small in comparison, but after amplification by the early stages it can be sufficient to cut off the succeeding valves so that the amplifier becomes temporarily inoperative. This phenomenon, known as *paralysis* or *blocking*, interferes with the reception of subsequent signals. In an extreme case the output voltage will swing violently to and fro between the upper and lower saturation limits, shown dotted in Fig. 57, long after the initiating overload signal has vanished.

So far we have discussed paralysis due to inadequate low frequency response. Another common cause of paralysis in *RC*-coupled amplifiers is grid current. If, during a large positive pulse, a grid is driven positive with respect to a cathode, grid current will flow charging the coupling condenser. After the pulse the grid is left negative and the final effect is the same as above : even if this particular stage is not cut-off, paralysis is likely to occur later in the amplifier. Grid current charging of coupling condensers must therefore be avoided in pulse amplifiers, either by using long-tailed pair amplifiers (no grid current), or by using direct coupling to the positive going grids (no condenser). This difficulty is not experienced in stages which have negative input pulses.

Paralysis is also caused by inadequate screen and cathode bypass condensers. This can be interpreted as a lack of low frequency response, or we can see directly that the voltage across the coupling condenser will change during the pulse and this change will produce an overshoot of the base line when the pulse has ended.

With a decoupling condenser in the anode circuit, however, the opposite effect occurs. Consider, for example, the circuit of Fig. 58 in which the decoupling network *RC* is followed by an anode load resistance *R*, and suppose that a negative pulse is applied to the grid of the valve reducing the current by *i*. A positive front *Ri* appears at the anode, but this is followed by an exponential rise, time constant *RC*, as the condenser charges and the anode waveform approaches its final amplitude 2*Ri*. If the pulse ends after a short period the output pulse is as shown in Fig. 58*a*. The top slopes positively and after the pulse there is undershoot, that is the base line does not quite return to zero. This effect, the opposite of differentiation, is associated with a frequency response which increases at low frequency, the effective anode load tending to 2*R* instead of *R*. We find then that a deficiency in the low frequency response gives overshoot of the base line, and conversely an excessive low frequency response causes undershoot.

The circuit of Fig. 58 has a practical application when it is desired to pass a square topped pulse through a coupling condenser which is rather too small. This often occurs in the beam brightening circuit of a cathode ray tube, because the square brightening pulse is passed to the grid via a high voltage blocking condenser which is of only moderate capacity to economize in cost and space. The coupling circuit is

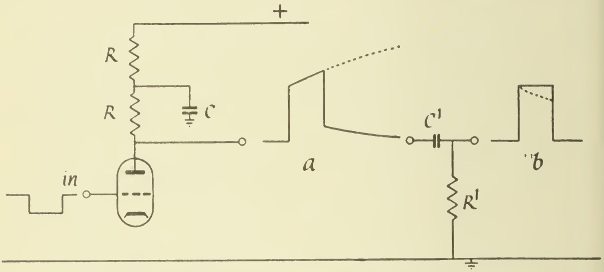


FIG. 58.—Effect of inadequate decoupling condenser C.
(a) output pulse.
(b) output pulse after differentiation, $R'C' = RC$.

represented by $R'C'$ in Fig. 58 and we assume that by itself this would cause slight differentiation of the pulse leading to a negatively sloping top, Fig. 58b, dotted. With the input of Fig. 58a, however, this differentiation is compensated by the positive slope on the input pulse, and the resulting output is perfectly square-topped. The condition for this, in the circuit given, is $R'C' = RC$. The compensating effect does not of course continue indefinitely and some fall in voltage towards the end of a long pulse will be inevitable.

With unequal resistances in the anode circuit better results can be obtained, the decoupling resistance R_d being made larger than the anode load R. The condition for compensation then remains as given, and the compensation continues to be effective for a time of order R_dC.

6.7. Practical feedback amplifiers. It has already been mentioned, § 6.4, that negative feedback is used mainly to give linearity and stability of gain rather than to increase the bandwidth of an amplifier. We give here two feedback systems which are suitable for pulse amplifiers.

A fundamental principle is that a minimum number of amplifying stages are included in each feedback loop. If more gain is required further amplifiers are added with

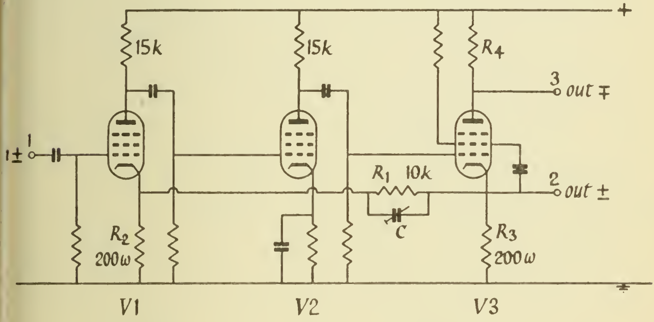


FIG. 59.—Ring of three feedback amplifier, gain 50, loop gain 20. Valves type 6AK5 (EF95).

separate individual feedback loops. The reason is that phase shifts of up to 90° per stage occur in the amplifier at high frequencies, and if a total phase shift of 180° occurs at a frequency for which the gain is still high the feedback will be positive instead of negative : the amplifier then becomes unstable and starts to oscillate. In practice, feedback is usually taken over one, two, or at most three, amplifying stages, and with careful design, instability can then be avoided.

A well tried system is that shown in Fig. 59. Here the positive input signal applied at 1 is amplified by valves *V1* and *V2* and fed to the output at 2 via cathode follower *V3*. The output signal is also positive and therefore subtracts in part from the input signal when fed back to the cathode of *V1* by means of the potentiometer network R_1, R_2. The feedback factor x (see § 2.3) is very nearly R_2/R_1, and accordingly the gain $1/x$ is R_1/R_2. With the values given in the circuit this amplifier has gain = 50, with loop gain xA = 20. The output impedance is about 10 Ω, and bandwidth, 3 Mc/s. The small trimmer condenser C is added to compensate for the stray capacity across R_2 which would otherwise reduce the

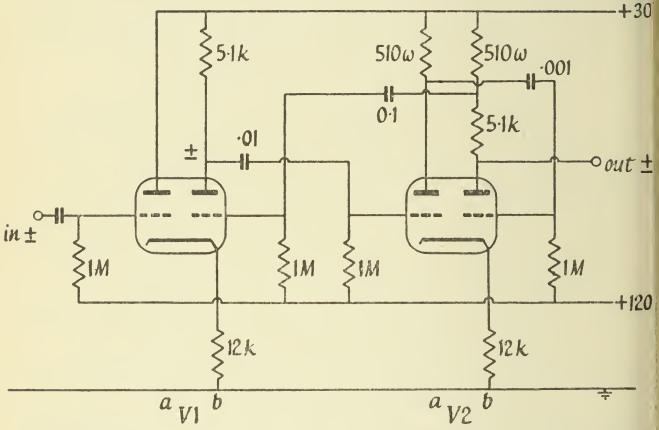


Fig. 60.—Non-paralysis feedback amplifier, gain 11, loop gain 10. Valves type 6J6 (ECC91).

feedback at high frequencies. Condenser C is usually adjusted empirically for optimum transient response. An inverted output pulse, negative if the input is positive, can be taken at 3 from the anode of $V3$. The gain will be increased in the ratio R_4/R_3 (if screen current is neglected), and the output impedance will be equal to R_4, unmodified because there is no feedback from the anode of $V3$. Note that the bypass condenser on the screen of $V3$ is connected to cathode instead of to earth. This arrangement applies the output signal to the screen with the result that the effective screen-cathode voltage is constant during the pulse, which means that a larger output signal can be obtained and also that the input capacity at the grid of *V3* is reduced (cf. § 2.5).

Another useful type of feedback amplifier unit,* given in Fig. 60, uses two long-tailed pair circuits (see § 2.6). In each case a signal is applied to the left hand grid of the pair and the amplified output taken from the opposite anode is in the same phase. Thus a positive input signal gives a positive output at the anode of *V2b*. A fraction 1/11 of this output signal is fed back to the grid of *V1b*: the feedback is negative because the signals on the grids of *V1* are both positive and have opposite effects on the anode current of *V1b*: as $x = 1/11$, the amplifier has gain 11. An additional feature is the small amount of positive feedback introduced by interconnecting the anode of *V2a* and grid of *V2b* as shown: this is insufficient to convert *V2* into a trigger circuit (cf. § 4.4) but increases the loop gain of the overall feedback loop from 5 to 10. With the components indicated this amplifier has a bandwidth of 5 Mc/s and can deliver positive or negative output signals of maximum amplitude 15 volts. The output impedance is 500 Ω. An advantage of the long-tailed pair circuits is that the valves cannot be driven into grid current by the maximum overload pulse which can be transmitted to the input from a previous stage, because, with the large cathode resistance, the cathode voltage can follow the grid in its positive excursion. The amplifier is therefore free from paralysis.

6.8. Anode follower. There are several special circuits which are useful in pulse amplification. The cathode follower has already been mentioned (§ 2.5). The anode follower, Fig. 61, is a feedback system with rather similar characteristics, but with phase reversal. It has an affinity with the Miller time base, § 5.5.

Neglecting for the moment the small condensers C_1 and C_2, we consider the general case of unequal resistance values R_1 and R_2. The action of the valve is to keep the junction of R_1 and R_2, point B in the diagram, at nearly constant potential.

* R. L. Chase and W. A. Higinbotham, *Rev. Sci. Inst.*, 23, 34, 1952.

This means that if the input potential rises the output voltage must fall correspondingly, and potentiometer action insures that if the input signal is $+v$ the output must be $-v \times R_2/R_1$: this is the condition for no signal at the point B. If, however, the output voltage differs from this value there is net signal at B, and after amplification by the valve it gives rise to a strong correcting signal at the anode which restores the output voltage to the value given.

In practice it is not quite true that there is no signal at B; in fact there is a small signal, which when amplified by the

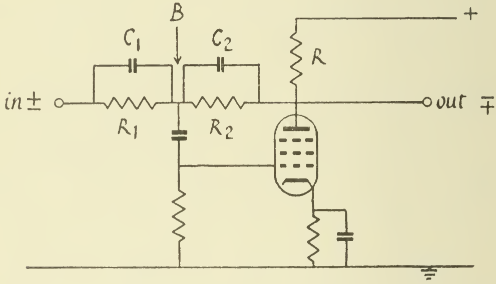


FIG. 61.—Anode follower, $R_2C_2 = R_1C_1$, gain $= R_2/R_1$.

valve, gives the correct output voltage. But if the amplification of the valve is large, the error introduced is negligible.

The small condensers C_1 and C_2 are added to swamp the effects of stray capacity from grid to earth which would otherwise reduce the loop gain of the feedback system at high frequencies. The ratio of the condenser impedances should correspond with the ratio of resistances ($R_1C_1 = R_2C_2$), so that the ratio of the potentiometer, and therefore the gain, is the same at all frequencies. In the simplest case $R_1 = R_2$, $C_1 = C_2$, and the system gives gain unity with phase reversal: this is a circuit which is often used for generating push-pull deflection signals for a cathode ray tube.

The output impedance of the circuit is low, as may be shown by applying Thévenin's theorem (§§ 1.5, 2.5); in the case of equal resistances, gain unity, the output impedance is equal to $2/g$, and is therefore twice that of a cathode follower.

A disadvantage of the circuit is that the input impedance tends to be low. Recalling that point B remains at almost constant potential we see that the impedance presented at the input terminal is simply R_1 and C_1 in parallel. This means that the previous circuit is loaded by the extra capacity C_1. It should also be observed that the capacity C_2 loads the output side of the circuit as though it was connected from anode to earth, but this may not be so serious in view of the low output impedance: (it will, however, reduce the maximum rates at the output, see § 6.5).

The principle that the valve should be turned on rather than off by the signal voltage (see § 2.9) prescribes the anode follower when large negative output pulses are required, but for positive outputs the cathode follower, or the bootstrap amplifier (§ 6.10), is more appropriate.

6.9. White cathode follower, Fig. 62. This output circuit is suitable for both positive and negative pulses. With a gain of unity, it will develop almost undistorted signals across a substantial stray capacity. On a positive going wavefront the upper valve *V1* can conduct strongly to charge the load capacity, while on negative going wavefronts the capacity is discharged by a heavy current in *V2*.

As in the simple cathode follower (§ 2.5), the action of the circuit is to maintain a constant potential difference between the grid and cathode of *V1*. If, for example, the cathode or output voltage should be too negative, *V1* will pass more current, developing a negative signal at its anode which causes *V2* to pass less current. The changes in both valves tend to restore the cathode potential to its correct value. In the quiescent state with no input signal the cathode voltage of *V1* adjusts itself so that the current in *V1* is the same as the current specified by the bias conditions in *V2*.

The circuit can be regarded as a two valve feedback amplifier with feedback factor $x = 1$ (see § 2.3). To find the gain A without feedback, we must split the feedback loop. This may be done by supposing the cathode of *V1* to be earthed, and giving *V2* an anode load $1/g$, equal to the input impedance of the cathode of *V1* which we have disconnected. The gain of *V1* (pentode approximation) is g_1R, and the gain of *V2* is $g_2 \times 1/g_1$: the overall loop-gain xA, with $x = 1$,

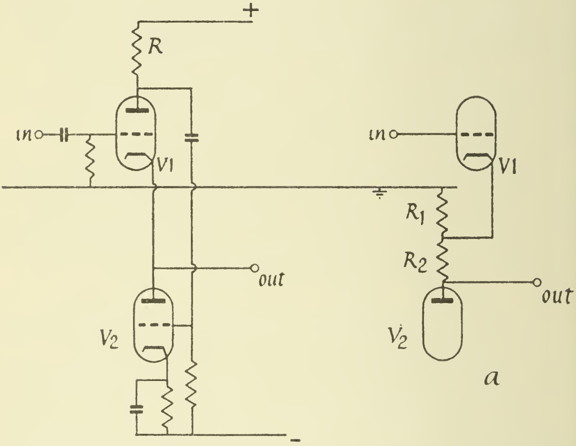


FIG. 62.—White cathode follower.
(a) adapted as an amplifier, gain $= (R_1 + R_2)/R_1$.

is therefore g_2R. The same result can be obtained by splitting the loop at the grid of *V2*, and treating the circuit as a cascode amplifier (§ 2.8). It follows that the gain of the system with feedback is $A' = g_2R/(1 + g_2R)$, the same formula as for the simple cathode follower. But in this case R can be larger without spoiling the negative going wavefronts, so that in fact a gain closer to unity can be realized.

The output impedance can be obtained using Thévenin's theorem by postulating a 1 volt positive signal applied to the output terminal from an external source. There is a current change g_1 in *V1* and $g_1R \times g_2$ in *V2*, so the total current to be supplied is $g_1(1 + g_2R)$, which means that the output impedance is $1/g_1(1 + g_2R)$. Typically, with $R = 2k\Omega$ and $g_2 = 5$ ma/v, the output impedance is 11 times less than that of *V1* used as a simple cathode follower. The White cathode follower therefore gives considerably better reproduction of small amplitude pulse fronts.

For large amplitude fronts the rates of voltage change at the output will often be limited by the current carrying capacity of the valves. In this case it may pay to abandon this circuit and to use *V1* and *V2* in parallel, either as a simple cathode follower for positive outputs, or as an anode follower for negative outputs. If both positive and negative wavefronts are to be reproduced faithfully then the White cathode follower is superior.

A useful modification, indicated in Fig. 62*a*, allows the White cathode follower to be used as an amplifier. Here only a fraction $R_1/(R_1 + R_2)$ of the output signal is fed back to the cathode of *V1*, with the result that the gain of the system is $(R_1 + R_2)/R_1$. As a result of the added resistances there is some loss of performance on large positive going fronts, but as an amplifier of negative pulses the circuit has some merit.

6.10. Bootstrap amplifier, Fig. 63. We can regard this circuit as an amplifier valve *V1* with anode load *R*, followed by a cathode follower *V2*. The signal developed at the anode of *V1* is applied to the grid of *V2*, and the output is taken from *V2* cathode. In operation, however, *V2* behaves, not as a cathode follower, but as an amplifier, because there is no degeneration, the whole signal developed across resistance *R* being applied between the grid and cathode of *V2* (cf. § 4.5). The effect of a small positive signal at the input is to produce an amplified negative signal at the grid of *V2*: the current through *V2* is therefore considerably reduced and the output voltage falls until the falling anode voltage of *V1* reduces its current to match the new current in the upper valve. Detailed analysis shows that if the two valves are similar, the overall gain rises from $\frac{1}{2}\mu$ when $R = 0$, to nearly μ when R is much larger than $1/g$. With dissimilar valves it is the μ of the lower valve which determines the upper limit to the gain. The output impedance varies from $\frac{1}{2}R_0$ when $R = 0$, to $1/g$ when R is much greater than R_0, R_0 being the anode resistance of the valves. It thus appears that while the gain is higher than that of an amplifier followed by

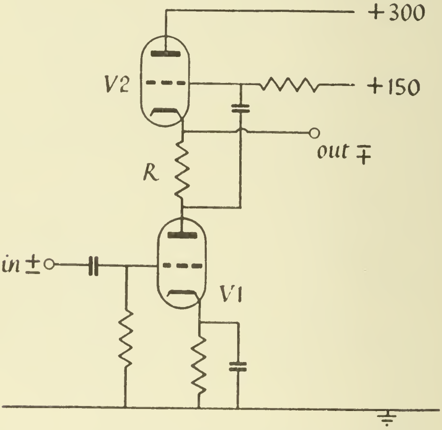


FIG. 63.—Bootstrap amplifier.

a cathode follower, there is the disadvantage that the output impedance is usually somewhat higher also.

The disadvantage of higher output impedance is nullified if the bootstrap amplifier is used as the output stage of an amplifier with overall negative feedback. The overall loop gain then serves to reduce the output impedance (§ 2.3), and this is enhanced by the high gain of the bootstrap component. In this situation the main advantage of the bootstrap circuit is its readiness to turn on a large current through $V2$ to charge stray capacities at the output. Because there is no degeneration $V2$ can be turned hard on by a fairly small signal developed across resistance R by valve $V1$, and the limiting rate of positive excursion is easily reached. The output voltage therefore approximates rapidly to the value specified by the feedback network.

The output waveform can be squared by adding a diode to limit the positive voltage excursion : it is best connected to the anode of *V1*, rather than to the output where it would have to handle the heavy current passed by *V2* (cf. § 7.2). In this form the bootstrap amplifier can generate square positive pulses of very short rise time.

In concluding this chapter on pulse amplifier design we observe that the principles and methods reviewed here are applicable also to the design of trigger circuits and pulse generators. For example, the rise time of the pulses produced by a square wave oscillator will be determined by the rise time of the amplifier in its basic positive feedback loop. Similarly, if a trigger circuit is to snap rapidly from one state to another, it must have high loop gain up to high frequencies. It follows that for steeply rising pulses and fast trigger action, small anode loads and high slope valves operated at high currents must be used ; but if a slower response can be tolerated economy of current and larger anode loads are permissible. In the practical balance between performance and economy lies the art of the designer.

CHAPTER VII

APPLICATIONS

In this final chapter we attempt to inter-relate the various circuits and principles which have been described by showing how they can be applied to particular problems. The primary interest, however, is not in the problems themselves but in the interconnection of basic circuits, in the gradual transformations of waveform, and in the variety of effects that can be produced.

7.1. Standard pulse generator, Fig. 64. The object in this case is to generate pulses of accurately known amplitude, which may be used, for example, to calibrate pulse amplifiers or measuring instruments.

Valve *V1* is a free-running square wave generator, which generates either square waves or narrow positive pulses. With the switches in position 1 the circuit is that of the cathode coupled multivibrator (§ 3.2), which generates a symmetrical square wave with frequency controlled by R_1. With the switches in position 2 the circuit becomes Scarrott's oscillator (§ 3.3), and gives a 100 μs positive pulse at the right-hand anode, with recurrence frequency controlled by R_2. With the values given both repetition frequencies are close to 50 c/s, and the oscillator in each case is locked to the mains frequency by injecting the heater supply voltage to the right-hand grid.

To generate a positive pulse of known amplitude, a known steady current flowing in the known resistance R_4 is switched off. The circuit is arranged with the positive pole of the h.t. supply at earth potential as shown, so that the voltage developed across R_4 is in fact a voltage relative to earth. It can therefore be applied to subsequent circuits without fear of modification by coupling components, or by ripple on the supply line which would normally appear between anode and earth. The long-tailed pair, valve *V2* (§ 2.7) has a constant current valve V_3 (§ 2.8) in its cathode circuit. The current flowing through *V3*, adjusted by means of R_3 and measured by the milliammeter *M*, normally flows through the right-hand half of *V2* and through R_4 to earth. But when a

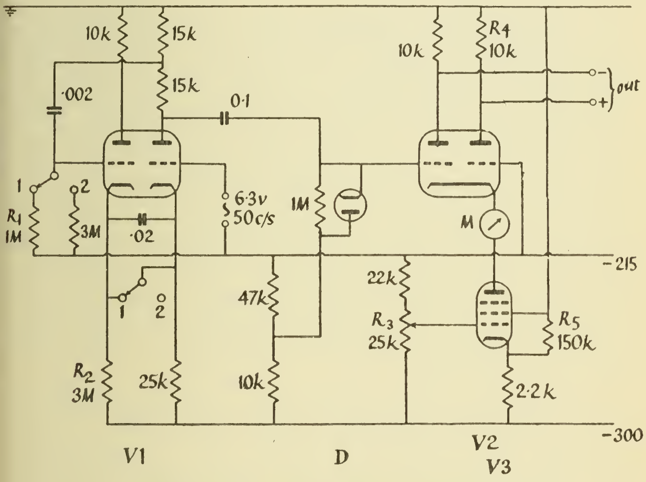


FIG. 64.—Standard pulse generator.

Double triodes type ECC81 (12AT7)
Pentode ,, EF80 (6AH6)
Diode ,, EB91 (6AL5)

positive pulse from the generator *V1* is applied to the left-hand grid of *V2* the current is diverted to this side, *V2b* is cut off and the positive pulse of known amplitude appears at the output. The constant current valve *V3* ensures that the current through the meter does not alter when it is switched from *V2b* to *V2a*, and therefore the meter correctly indicates the peak current in R_4 even in symmetrical square wave operation. It follows also that a negative pulse of known amplitude can be obtained by inserting a known resistance in the anode circuit of *V2a*. The diode *D* acts as a d.c.-restorer and maintains the correct bias conditions for valve *V2* even when a symmetrical square wave is applied. Resistance R_5 is added to raise the cathode potential of *V3* so that it is possible to reduce the current to zero by operating potentiometer R_3.*

7.2. Discriminator amplifier, Fig. 65. This amplifier is designed to ignore positive signals of amplitude less than a predetermined threshold, set by potentiometer R_1, and to amplify by a factor 5 those parts of the input signals which exceed the threshold. In effect, the output gives a magnified view of a certain range of input amplitudes. The input circuit is a long-tailed pair, *V1* and *V2*, the negative amplified signals at the anode of *V1* being passed through diode *D1* to the grid of *V3*. Valves *V3* and *V4* comprise a bootstrap amplifier (§ 6.10) and there is overall negative feedback via the potentiometer R_2, R_3 which feeds a fraction of the positive output pulse back to the grid of *V2*. With an attenuation of 1/5 in the feedback path, the overall gain is 5, with a loop gain xA of about 20.

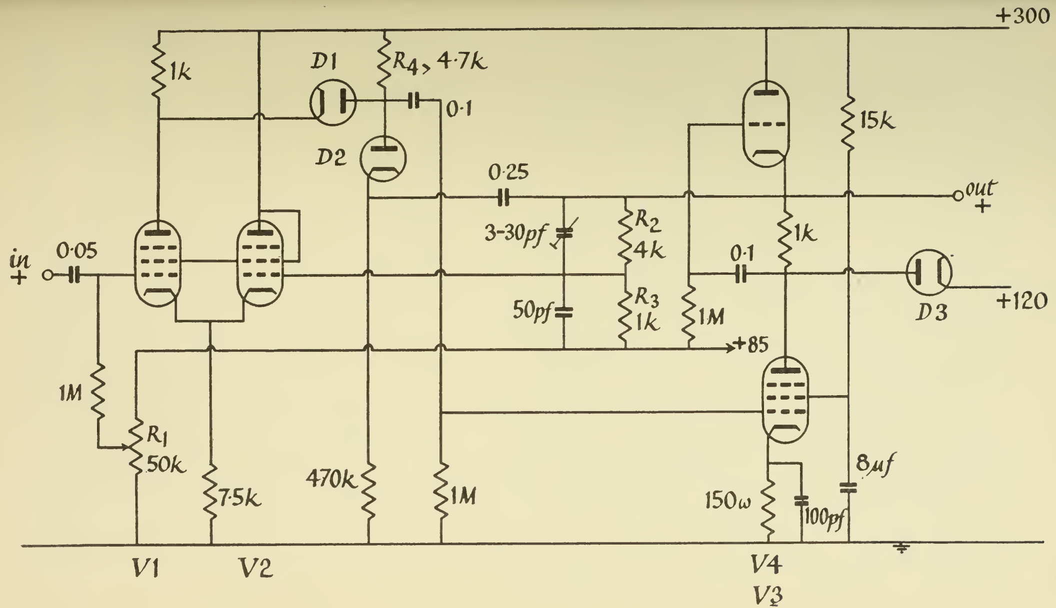
The diodes *D1* and *D2* introduce refinements which are necessary for the amplification of very short pulses. In the quiescent state a current of 0·6 ma flows through *D2* to earth, developing a potential of 3 volts across the 4·7 kΩ resistance R_4, i.e. a 3-volt delay bias voltage across the diode *D1*. This means that the anode voltage of *V1* must drop 3 volts before a signal is transmitted to *V3* ; in the operating range of the amplifier, therefore, the currents in *V1* and *V2* are about equal, and the pair is working at maximum gain.

Diode *D2* is non-conducting during the signal, its cathode being driven positive by the output pulse and its anode

* For further detail see G. W. Hutchinson and G. G. Scarrott, *Phil. Mag.*, 42, 792, 1951.

FIG. 65.—Discriminator amplifier.

Pentode	type EF80 (6AH6)
Triode	,, ECC91 strapped (6J6)
Crystal diodes	,, CG4C (IN63)


+300
1k
D1
R4, 4·7k
0·1
D2
0·25
15k
out
+
in
+
0·05
3-30pf
R2
4k
1k
0·1
R3
1k
50pf
1M
+85
D3
+120
1M
R1
50k
7·5k
470k
1M
150ω
100pf
8μf
V1
V2
V4
V3

negative, but it has an important function during the tail of the pulse. When the input signal disappears the anode voltage of *D2* rises exponentially towards the h.t. potential until *D2* conducts. Because the cathode potential of *D2* has been raised by the output voltage, the grid of *V3* is then more positive than before, and *V3* is turned hard on until the output voltage returns to its quiescent value. There is, in effect, a second feedback loop which comes into operation during the tail of the pulse, comprising *V3*, *V4* and *D*2 only. Its effect is to restore the circuit to the quiescent condition with a delay of only a few tenths of a microsecond.

The maximum excursion of output voltage is set by the diode *D3*. This means that the amplifier saturates for input signals exceeding the threshold by about 8 volts, but there is no grid current or paralysis because of the long-tailed pair input circuit (see § 6.7, p. 103). Because of the small anode loads, and the bootstrap output stage, the circuit will give faithful amplification for pulses as short as 0·2 μs.*

* For further detail see F. J. M. Farley, *J. Sci. Instrum.*, **31**, 241, 1954.

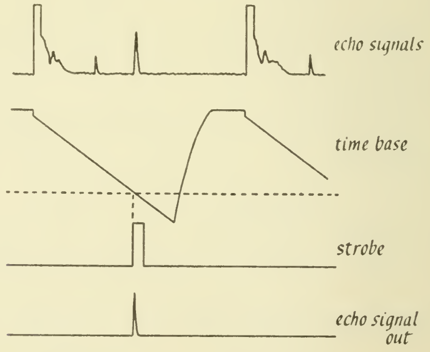
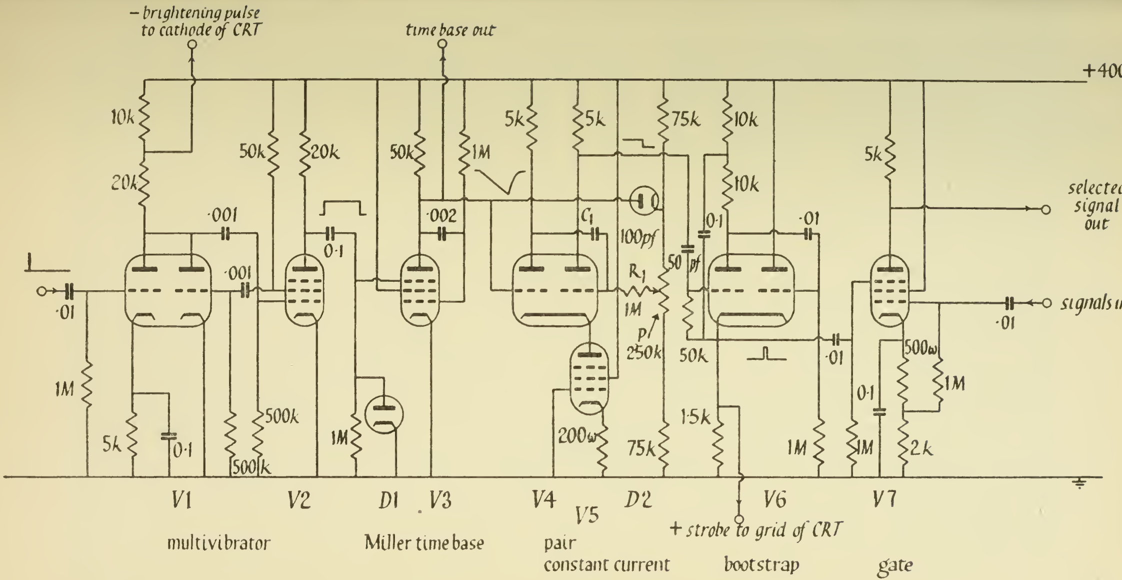


FIG. 66.—Radar time base and strobe.

Double triodes type ECC81 (12AT7)
V2 and *V5* ,, EF80 (6AH6)
V3 and *V7* ,, 6F33 (short suppressor base) (6AS6)



– brightening pulse to cathode of CRT
time base out
+400
selected signal out
signals in
10k
20k
50k
20k
50k
1M
5k
5k
75k
10k
10k
5k
·001
·001
0·1
·002
C1
100pf
R1
1M
50 pf
0·1
·01
p
250k
50k
·01
·01
1M
5k
0·1
500k
500k
1M
200ω
75k
1·5k
1M
1M
0·1
500ω
1M
2k
V1
V2
D1
V3
V4
V5
D2
V6
V7
+ strobe to grid of CRT
multivibrator
Miller time base
pair
constant current
bootstrap
gate

7.3. Radar time base and strobe circuit, Fig. 66. In radar a powerful radio pulse is transmitted at regular intervals, of for example 3 ms, and the echo pulses which arrive back at the equipment after a short time delay are studied. A typical pattern of received echoes is shown in Fig. 66*a* ; there is a strong signal from nearby objects immediately following the transmitted pulse, and two echoes from more distant objects are shown. The echoes are displayed on a cathode ray tube using a time base which starts in synchronism with the transmitted pulse, Fig. 66*b*. In some applications it is desirable to isolate a particular echo signal for special attention in subsequent circuits. This is achieved by generating a short *strobe* pulse, Fig. 66*c*, which coincides with the desired echo and using a gate circuit (§ 7.8) to transmit signals only when the strobe pulse is present. The final output signal is indicated in Fig. 66*d*.

In the typical circuit of Fig. 66, *V1b* and *V2* form a free-running multivibrator which is locked to the transmitter frequency by means of a positive locking pulse introduced via *V1a*. The screen grid of *V2* is used in the multivibrator circuit, so that a clean positive square pulse is obtained at the anode of *V2*, undistorted by the effects of grid current in the opposite valve (see § 3.5). This pulse is applied via an RC coupling, with diode *D1* as d.c.-restorer, to the Miller time base valve *V3* (§ 5.5). The negative going time base waveform at the anode is applied to the X-plate of the cathode ray tube and also to the long tailed pair *V4*. This circuit is similar to that of § 5.6, but in this case the addition of the coupling C_1R_1 from the anode of *V4a* to the grid of *V4b* turns *V4* into a trigger circuit. The result is that a sharp negative step appears at the anode of *V4b*, and the position of this step relative to the time base waveform is variable by means of the potentiometer P. The negative step triggers the bootstrap trigger circuit, *V6* (§ 4.5), which produces a short positive strobe pulse at the cathode. This pulse is applied to the suppressor of the gate valve *V7* (§ 7.8), the echo signals being applied to the grid. Because no anode current flows in the absence of a strobe pulse, only those signals which coincide with the strobe are transmitted. The strobe pulse is also applied to the grid of the cathode ray tube in order to brighten the trace in this region and so to indicate the particular echo selected by the strobe. As potentiometer *P* is adjusted the bright spot moves along the time base, and may be set to coincide with any particular desired signal.

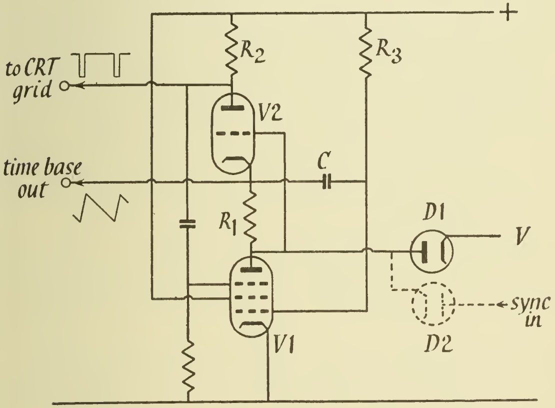


FIG. 67.—Free-running Miller time base.
Triode ECC91 strapped (6J6)
Pentode 6F33 (6AS6)

7.4. Free-running Miller time base, Fig. 67. This circuit is one of the family of time base circuits derived from the basic Miller time base of § 5.5. To the Miller time base valve *V1* is added the amplifier valve *V2*, the output of which is coupled back to the suppressor grid of *V1* in such a way that the whole system has trigger properties.

Let us first suppose that the suppressor of *V1* is at cathode potential so that current is flowing to the anode of *V1* and the Miller time base is being generated in the usual way. The current flowing through resistance R_3 flows through *C*, discharging the condenser, through R_1 and through valve *V1* to earth. This current develops a voltage across R_1 sufficient to hold valve *V2* completely cut-off. Note also that the current through R_1 and the valve is constant during the time base sweep (it is equal to the current through R_3), and therefore the grid voltage of *V1* does not change at all, and the time base is perfectly linear; * an improvement on the simple Miller time base of § 5.5. But eventually the anode voltage falls beyond the knee of the valve characteristic and the flow of anode current cannot be maintained. The voltage across R_1 now falls, allowing *V2* to conduct, with the result that a negative signal is developed at the anode of *V2* and transmitted to the suppressor of *V1*. This interrupts the anode current in *V1*, with the result that the action is cumulative, and in practice the circuit triggers to a new state in which *V2* is fully conducting and the anode current of *V1* is cut off by the suppressor.

Condenser *C* is now recharged rapidly by the current flowing through *V2*, through *C*, and via grid current in *V1* to earth. The voltage at the anode of *V1* therefore rises until the diode *D1* conducts, and draws current through R_1, thus again cutting off valve *V2* so that the cycle repeats. The voltage *V* applied to the cathode of the diode therefore controls the potential at which the negative going time base phase begins, that is it controls the time base amplitude.

This circuit produces an accurate linear time base with rapid flyback and is useful for general purpose oscilloscopes. The circuit may best be synchronized by adding the diode *D2* (shown dotted), which then controls the potential at which the flyback begins. Positive signals applied to the anode of *D2* will then tend to initiate the flyback and synchronize the time base in the usual way.

7.5. High speed triggered time base. A triggered time base is often required for observing a transient phenomenon which occurs either once only or at irregular intervals. The

* This is not completely true, because as the anode voltage of *V1* falls the grid voltage must rise very slightly to maintain the constant current.

time base is initiated by the event to be observed, and it is highly desirable that the time base should start with the minimum of delay, so that practically the whole of the triggering signal is visible on the screen. It is usual to have the cathode ray beam cut off by a negative potential applied to the grid of the cathode ray tube until the time base is triggered, to avoid burning the screen. A positive brightening pulse is then applied to the grid, to turn on the beam when the sweep is started. In practice the delay in turning on the beam is usually greater than the delay in starting the time base, because the former requires a definite voltage excursion, while for the latter we have only to establish a rate of change of voltage. For minimum delay attention should therefore be given to the rise time of the brightening pulse, and this suggests the use of a bootstrap output circuit.

A simple way to generate a fast linear time base is to charge a capacity, perhaps only the stray capacity, in the anode circuit of a pentode. The pentode is turned on to generate a negative going sweep and because of its constant current characteristic (§ 2.7) the time base will be almost linear.

A circuit incorporating these ideas is given in Fig. 68. Here valve *V1* is normally biased beyond cut-off, but is turned on to generate the time base, charging condenser C (and the stray capacity), with constant current. The charging current also flows through the small resistance R_2, across which a square pulse sufficient to drive *V3* is obtained. *V3* and *V4* form a bootstrap amplifier (§ 6.10) which generates a positive square pulse, for brightening the beam and also for driving *V1*. There is thus a complete positive feedback path, and the circuit has trigger properties. Valve *V2* conducts during the flyback to recharge the condenser, and plays a rôle similar to that of *V2* in Fig. 67 just discussed.

In the quiescent state *V1* is non-conducting with the cathode of *V2* at h.t. potential, but *V3* and *V4* are conducting. When the circuit is triggered, by a negative pulse applied to the suppressor of *V3*, a 20-volt positive pulse appears at the cathode of *V4* and turns on a current of 15 ma in *V1*. This

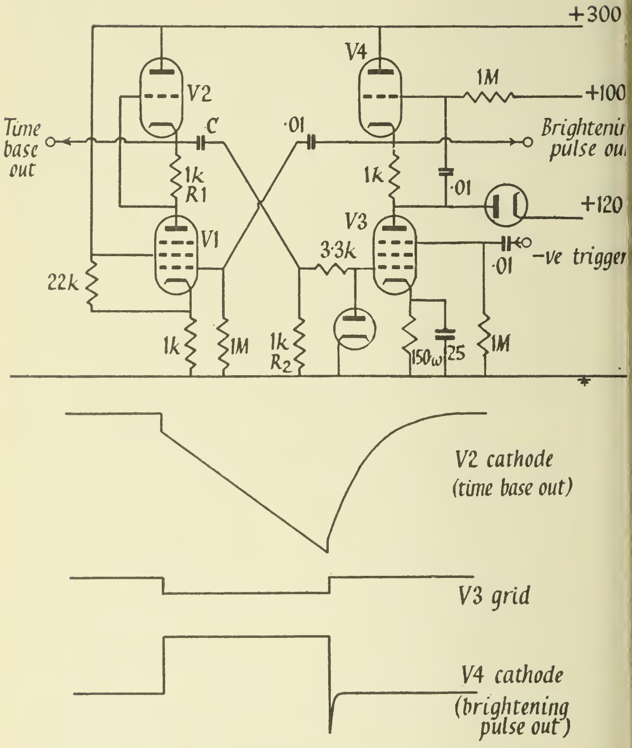


FIG. 68.—High speed triggered time base.

V1	EF80 (6AH6)
V2, V4	ECC91 strapped (6J6)
V3	6F33 (6AS6)
Crystal diodes	CG4C (IN63)

current develops sufficient voltage across R_1 to hold *V* cut-off and therefore flows through condenser C and R_2 t earth, developing across R_2 a negative voltage sufficient t keep *V3* cut off and thus complete the trigger action.

The anode potential of *V1* now swings negative generating the linear sweep, until as it approaches earth potential the current must decrease. The negative potential across R_2 must decrease correspondingly and eventually *V3* starts to conduct, reversing the trigger action and resetting the time base. At this stage *V2* conducts, rapidly recharging the condenser C and driving the grid of *V3* positive until the diode conducts. Finally, at the end of the flyback we again reach the initial quiescent state.

It will be noted that with this circuit the triggering signal gives rise directly to the brightening pulse, which is thus generated with minimum delay. The time base is then turned on and the trigger process completed. Typically the full trace brightness and time base velocity can be established in about 0·1 μs. And the time base velocity, for example, with $C = 100$ pF and an additional stray capacity of 100 pF, with charging current 15 ma, would be $dV/dt = i/C = 75$ v/μs.

7.6. Television time bases. In television receivers the beam of the cathode ray tube is deflected, not by electric fields as in the cathode ray oscilloscope but by magnetic fields produced by coils disposed round the neck of the tube. The time base circuit therefore must produce linearly increasing current waveforms instead of the linear voltage waveforms which have been treated so far. Of course if the impedance of the deflection coils is purely resistive there is no difference between the two cases, because a linear voltage waveform will give rise to a linear current sweep. This is in effect true for the frame time base which has a recurrence frequency of 50 c/s, low enough for the inductance of the coils to be negligible. But for the line time base, recurrence frequency say 16 kc/s, the inductive effects are dominant. Even for the frame time base, however, inductive effects will have to be considered if a transformer is used to couple from the output stage to low impedance deflection coils, and this is the usual practice. We now consider typical circuits for the frame and line time bases.

Fig. 69 shows a typical frame time base generator.* Positive locking pulses applied to the lock amplifier *V1*

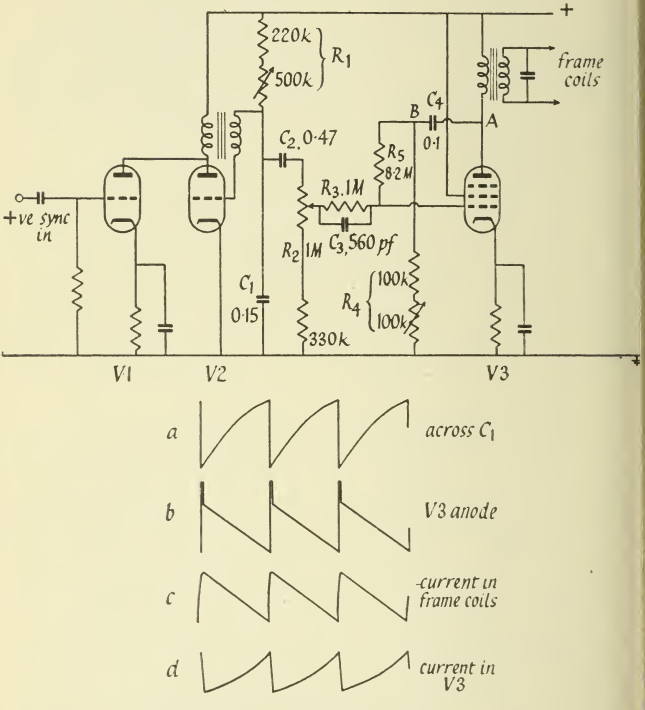


FIG. 69.—Television frame time base.

Double triode ECC81 (12AT7)
Pentode EF80 (6AH6)

synchronize the free-running blocking oscillator *V2* (§ 3.4), which generates a saw-tooth waveform of period 20 ms, Fig. 69*a*, across the condenser C_1. This signal passes with little

* From Television Receiving Tubes, Philips Electronic Tube Division.

distortion through the long time constant coupling R_2C_2 (600 ms), to the output stage *V3*, R_2 serving as amplitude control. The output valve has a feedback network designed to achieve a linear current sweep in the frame deflection coils. There are two sources of non-linearity which have to be corrected, first, the non-linearity of the input waveform itself, and secondly, that introduced by the frame output transformer : these will be discussed separately, although the same device proves sufficient to correct for both effects.

Valve *V3* is essentially an anode follower (§ 6.8) which inverts the time base waveform and amplifies it by the ratio R_5/R_3. Condenser C_3 does not affect the main part of the sweep waveform because time constant R_3C_3 is only 0·5 ms, and will be neglected for the present. The coupling R_4C_4 is, however, important because the time constant of 10-20 ms is comparable with the period of the time base. The feedback action of the anode follower gives rise to a waveform at point *B* in the diagram which is ideally a perfect inverted and amplified replica of the input waveform. The waveform produced at *A* must therefore be such that when differentiated by R_4C_4 it becomes a perfect replica of the input waveform. In fact, if the circuit is correctly adjusted the waveform at *A* becomes perfectly linear although the input waveform and the matching waveform at *B* are both exponential, and we shall now see how this comes about.

We suppose for the moment that the anode load of *V3* is purely resistive, and see how the shape of the input waveform can be corrected. Postulate a perfectly linear sweep $V = kt$ at *A*. Then the differentiated waveform at *B* is given by equation (9), § 1.3, $V = k \,.\, R_4C_4(1 - e^{-t/R_4C_4})$. This is, however, to be a replica of the input waveform which follows the law $V = V'(1 - e^{-t/R_1C_1})$ which regulates the charging of the reservoir condenser C_1 in the blocking oscillator. Clearly the two waveforms have the same shape if $R_4C_4 = R_1C_1$, so in this case our postulate of a linear waveform at *A* would be fulfilled. Under these conditions the valve generates a linear sweep at the anode, with slope R_5/R_3 times the initial slope of the input waveform. There is no net signal at the grid in the ideal case of infinitely high loop gain, because the input waveform and its replica at B exactly cancel. It is interesting to note that the compensation for non-linearity is perfect for all time: even when the input waveform is steady at its maximum value the output continues to rise linearly: it must do so in order to maintain the steady voltage at B which balances the steady input signal.

In the circuit given R_4C_4 (10-20 ms) is less than R_1C_1 ($\sim$ 75 ms), so that the compensation is excessive and would apparently lead to non-linearity in the opposite sense, if it were not for the rather low loop gain of the actual circuit and the effect of the output transformer. We shall now consider this effect assuming that the input waveform is perfectly linear.

The transformer with its resistive load is equivalent to a pure resistance in parallel with an inductance equal to the inductance of the primary when the secondary is on open circuit. The current flowing through the resistance is magnified by the turns ratio and transferred to the deflection coils; therefore we still require a linear voltage waveform at the anode of the valve to generate the linear current sweep in the resistance. The current through the valve does not rise linearly, however, but must increase more rapidly at the end of the sweep to provide for the growth of current in the inductance. Writing $di/dt \propto v \propto kt$ we find that this current rises according to a square law, $i \propto t^2$. The grid voltage of valve $V3$ must therefore have a square law component. This is provided if the input voltage rises linearly, and the voltage at B falls exponentially according to equation (9) as mentioned above. The difference between these two waveforms appears at the grid and it is readily verified by expanding the exponential as far as terms in t^2 that this difference is proportional to t^2/R_4C_4 plus higher terms. Therefore, if t is not too great, the circuit provides just the waveform required at the grid to give the linear time base sweep. The amplitude of the square law term is controlled by adjusting the time constant R_4C_4 until the sweep is linear.

To recapitulate, the partial differentiation by R_4C_4 in the feed back path corrects perfectly for the non-linearity of the blocking oscillator waveform; and also provides the square law component of current which is required in the primary of the output transformer, in addition to the main linear sweep.

It remains to explain the function of condenser C_3. As already mentioned it may be neglected during the time base sweep, but it plays a rôle during the flyback. The steep flyback waveform from the blocking oscillator is transmitted in full through C_3 to the grid of the output valve *V3*, thus interrupting the current in the frame coils. The inductance in the anode circuit, due to the frame coils and the output transformer, gives rise to a high positive voltage at *A* during the flyback which passes through the feedback network tending to make the grid of *V3* positive. It is to swamp this effect, and to ensure that the grid goes well negative, that condenser C_3 is included.

We now proceed to a circuit for generating the line time base, of recurrence frequency say 16 kc/s. In this case it is convenient to design the deflection coils so that the resistance is negligible, and to obtain a linear current sweep ($di/dt =$ const.) the voltage (Ldi/dt) applied to the coils must then be constant. If the resistance of the coils is not negligible then the applied voltage must increase slightly as the current builds up. This means that to generate the linear sweep we could apply to the coils, perhaps via a cathode follower, a positive square wave with a positively sloping top (cf. Fig. 58*a*). This method is excellent for an intermittent time base, but does not allow the rapid flyback essential for television. In the recovery phase a large reverse voltage across the coils is inevitable and this would not be allowed by the cathode follower.

A practical circuit designed to run from an h.t. supply of only 180 volts is given in Fig. 70.* This circuit allows current to flow in both directions through the line deflection coils, Fig. 70*c*, and also provides an extra h.t. voltage for the output stage by means of the diode *V2* which rectifies the time base waveform. When the circuit is operating a steady 200

* Philips Electronic Tube Division—*loc. cit.*

volts appears across the large condenser C_1 increasing the anode supply voltage to 380 volts. The grid of *V1* is driven

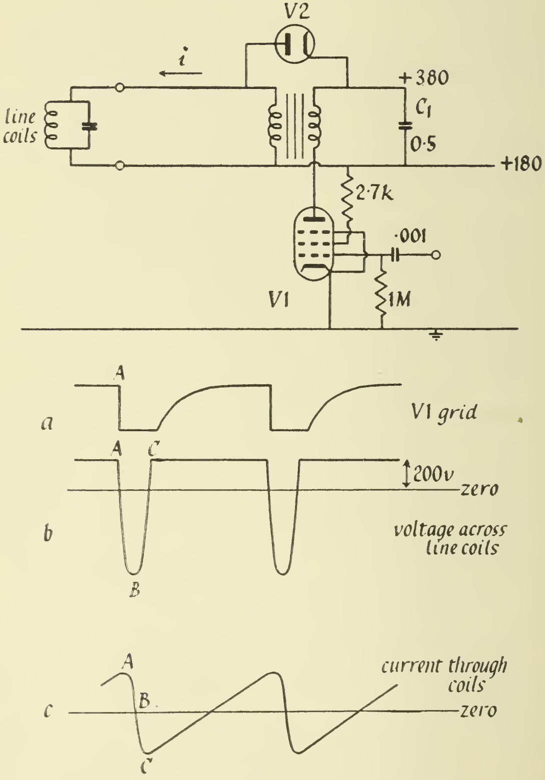


FIG. 70.—Television line time base.
Pentode PL81 (6BQ6)
Diode EB91 (6AL5)

by an input square wave, Fig. 70*a*. Consider the action at the end of the sweep when a current *i* is following through the coils in the direction shown by the arrow, point A on the waveforms. The valve is then suddenly cut off with the result that the coils become in effect isolated (we assume an ideal transformer), and a half cycle of oscillation at the natural resonant frequency of the coils follows (A to C in the figure) : the current drops to zero as charge flows into the stray capacity yielding a high negative voltage across the coils (point B), and then builds up to a nearly equal negative current in the reverse direction (point C). At this stage diode $V2$ conducts clamping the coil voltage at 200 volts, and the next sweep begins. Because the coil voltage is clamped at a constant value and the coils are purely inductive the sweep must be linear. During the first half of the sweep a gradually decreasing current is flowing out of the coils through the diode $V2$ and condenser C_1 and back again to the coils, being driven by the magnetic energy in the coils which is gradually transferred to C_1. During this phase valve $V1$ plays no part. By the time the current has dropped to zero, however, $V1$ is conducting, and through the step down transformer maintains the positive voltage of 200 volts across the coils and continues to drive current through the diode. From now until the end of the sweep the gradually increasing current i through the coils is again in the direction of the arrow and is supplied by the transformer secondary. The step down ratio means that the primary current drawn from C_1, and passed by the valve, is smaller than i and a balance of charge in C_1 over the cycle can be maintained.

In this circuit the diode $V2$ achieves three things. First it allows current to flow through the coils in both directions, with the result that the magnetic energy stored at the end of the sweep is used to provide a reverse current for the start of the sweep, instead of being dissipated in an added resistance which would be necessary to prevent oscillations if the current was to drop only to zero in the flyback. Secondly, it maintains a constant voltage across the coils during the sweep, thus ensuring a linear current waveform largely independent of the shape of the input signal. Thirdly, it provides the extra h.t. voltage for $V3$ and thus leads to higher efficiency because a smaller fraction of the total power is dissipated in the valve.

During the flyback the large negative voltage peak across the coils is transformed to a larger positive peak at the anode of the valve. By adding a further diode (not shown) this signal may be rectified and smoothed, and used to provide the positive high voltage (say 4-10 kv) for the cathode ray tube.

7.7. Pulses from nuclear particle detectors. In many of the particle detectors used in nuclear physics the signal obtained is due to the sudden arrival of electric charge, say q, on a collecting electrode. If the stray capacity from this electrode to earth is C, then a sudden change in voltage equal to q/C is produced, Fig. 71*a*. The rise time of this square step is the time during which the charge q is arriving at the electrode: if all the charge arrived simultaneously the rise time would be zero, and the pulse infinitely steep; in practice rise times less than $10^{-8}s$ have been realized in some systems. The rise may be exponential, or nearly linear as suggested in Fig. 71*a*.

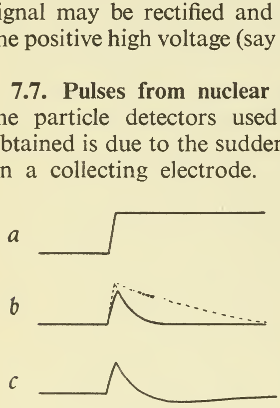


FIG. 71.—Differentiation of pulse from nuclear particle detector.

It is the subsequent treatment of this waveform that concerns us here. The object is usually to count the number of square steps produced, and often to measure their height. To avoid paralysis of the system by a progressive rise of input voltage it is essential to differentiate the signal, so that the square steps are converted into a series of pulses with exponential tails (time constant T_1), as indicated in Fig. 71*b*. The differentiation should be sufficiently sharp to ensure that successive pulses do not overlap appreciably (this depends on the mean counting rate), but if it is too sharp the pulse amplitude will be reduced because of its finite rise time. This is tolerable if the rise time is the same for all pulses

(Geiger and proportional counters), but not if the rise time is variable, or the pulses already very small (ionization chamber, scintillation counter).

The signal of Fig. 71*b* can now be handled in the usual way, amplified, measured, or displayed on a cathode ray tube. It is important, however, to avoid any further significant differentiation by coupling time constants in the system, because this would give rise to overshoot of the base line as described in § 6.6 and would interfere with the measurement of subsequent pulses. This effect is minimized by keeping all such time constants much longer than the pulse length. Significant differentiation by a second time constant $T_2 (> T_1)$ would produce a signal such as Fig. 71*c*. There is an overshoot of the base line roughly equal to a fraction T_1/T_2 of the initial pulse and this lasts for a time of the order T_2, the longer time constant.

It is sometimes supposed that by including the shortest time constant T_1 at the end of the system all long term excursions of the base line can be eliminated. This is not so, however, and, in fact, if the time constants T_1 and T_2 were interchanged in the above example the effect produced would be exactly the same. The explanation is that the first differentiation would now produce a slowly falling tail of time constant T_2, shown dotted in Fig. 71*b*, and after further differentiation by the shorter time constant T_1 this would give a negative voltage proportional to its negative slope; that is an overshoot which would die away with time constant T_2 as before.

In practice then the rule is that the signal from the particle detector must be differentiated only once; all other coupling time constants must be much longer. But it does not matter at what stage the decisive shortest time constant occurs; it may be at the beginning, middle, or end of the transmission system. A common practice is to place it in the middle of an amplifier, late enough to eliminate the low frequency noise components from earlier stages, but early enough to avoid paralysis of later stages by overlapping long pulses.

We note finally that some differentiation of the signal of Fig. 71*a* occurs at the particle detector itself. As already explained the square voltage step is produced by the arrival of charge on the collector electrode, capacity C, and we have in effect assumed that there is no leakage path so that the charge remains there indefinitely. In fact, however, there is inevitably some leakage resistance, say R, and the signal at the detector returns to zero with time constant RC. It is in effect differentiated by this time constant. There are now two main possibilities. In the first the resistance R is made as large as possible, say 100 MΩ, and the differentiation is controlled later in the system. In the second RC is made short, and becomes the master time constant. Note that the pulse size at the detector is q/C, independent of R, and remains constant as R is decreased until RC becomes shorter than the rise time of the pulse.

7.8. Coincidence and gate circuits. A common demand is for a circuit which gives an output signal when two or more independent input signals coincide in time. This is achieved by means of a *coincidence circuit*, and here it is not necessary or even usual for the output pulse to resemble the input pulses in size or shape. A similar but more stringent requirement is placed on the *gate circuit:* this has two states, in one of which (gate open) input signals are faithfully transmitted to the output, and in the other (gate shut) there is no output. The transition from one state to the other is controlled by a square pulse applied to an auxiliary input. Generally speaking a gate circuit can be used as a coincidence circuit, but not vice-versa.

Some useful coincidence circuits are given in Fig. 72. In the Rossi circuit, Fig. 72*a*, both valves are normally conducting and the common anode voltage is low. If one valve is turned off by a negative input signal the anode voltage rises

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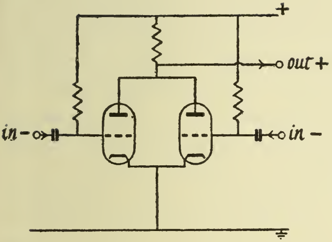
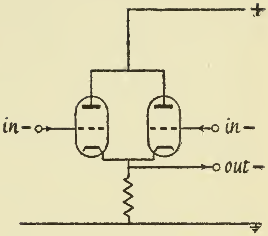
FIG. 72.—Coincidence and gate circuits.

(*a*) Rossi coincidence circuit.
(*b*) Cathode follower coincidence circuit.
(*c*) Gate using suppressor grid.
(*d*) Double diode gate.
(*e*) Double triode clamping circuit.



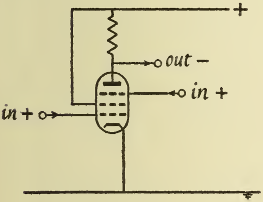
+
out +
in −
in −

a



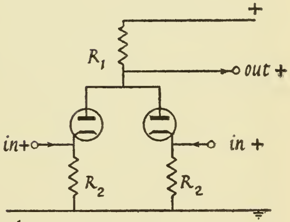
+
in −
in −
out −

b



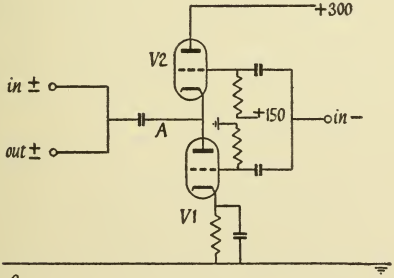
+
out −
in +
in +

c



+
R1
out +
in+
in +
R2
R2

d



+300
V2
in ±
out ±
A
+150
in −
V1

e

slightly, but if both valves are turned off by coincident input signals it rises to the full h.t. potential. The output pulse for a coincidence is therefore much greater than for non-coincident inputs, and may be distinguished by a simple discriminator circuit. The circuit works best with pentodes with the anode load chosen so as to bring the quiescent anode voltage below the knee of the valve characteristic. Further valves may be added in parallel for additional input signals if a many-fold coincidence is to be recorded, a large output pulse then being obtained only when all the valves are turned off simultaneously. Final pulse shaping can be carried out in the grid circuits as shown in the figure (cf. p. 24).

In the cathode-follower coincidence circuit, Fig. 72*b*, both valves normally conduct and negative signals are applied to the grids. The common cathode can swing negative only if both signals are present together. Here the amplitude ratio at the output for coincident and non-coincident inputs can be large, and once again the circuit can be extended to a multi-channel system by adding further valves. This circuit makes a satisfactory linear gate circuit if a gating pulse is applied to one of the grids, provided that the small transmission of the gating pulse to the output can be tolerated.

In Fig. 72*c*, the input signals are applied to the grid and suppressor of a pentode. The electrodes are biased so that anode current does not flow unless a positive signal is present at both. An advantage of this circuit is that there is no signal at all in the absence of a coincidence, but on the other hand it cannot be extended readily to record manyfold coincidences: valves with three control grids are indeed available but this is the limit at the present time. This circuit makes a very satisfactory gate circuit if the gating signal is applied to the suppressor. If the grid is biased just below cut-off we can ensure that the gating pulse itself gives no output, at the expense of some loss of the smallest input signals.

The diode coincidence circuit, Fig. 72*d*, is economical in valves, especially if crystal diodes are used, but on the other hand the input impedance is low and an extra valve may therefore be required in each driving amplifier. In the quiescent state current flows through the high resistance R_1, and through the diodes and the two low resistances R_2 to earth. A positive pulse applied to the cathode of one diode makes this valve non-conducting, and doubles the current through the other: this gives only a small output pulse. But with coincident inputs the smaller signal is transmitted in full to the output. If the steady current flowing normally in each of the resistances R_2 is i_1, and if the signal sources provide current i_2 through the same resistances, then the ratio of output amplitudes for coincident and non-coincident input signals is i_2/i_1. As i_1 is usually set by rise-time considerations at the output (see below), this relation determines a minimum value for i_2, the current to be supplied by the signal sources. The circuit can be used with negative input signals by reversing the diodes and the supply voltage. It makes a satisfactory linear gate circuit, but as in the case of Fig. 72*b* some transmission of the gating signal to the output is inevitable.

Fig. 72*e* is primarily a gate circuit. The action of the valves is to clamp the common input/output line *A* at a potential governed by the bias applied to *V2*, 150 volts in this case. With both valves conducting the impedance presented at *A* is $1/g$ and this effectively short circuits the signals if they come from a high impedance source. But when both valves are cut off by a negative pulse applied to the grids, point *A* is left floating and can follow the signals without hindrance. This circuit has the advantage that there is no disturbance of the signal line due to the gating pulse.

If a more rigid clamping action is required at *A*, *V1* may be replaced by a pentode with the switching pulse applied to the suppressor. A feedback signal can then be taken from a resistance in the anode circuit of *V2* to the grid of *V1* as in the White cathode follower (§ 6.9). This considerably reduces the input impedance to the cathode at *A* and therefore increases the clamping effect.

For all coincidence systems input signals which are not quite simultaneous will be recorded as a coincidence, but some systems are more discriminating than others. The criterion is the *resolving time* of the system. The interval between signals must be greater than the resolving time if they are not to be recorded as in coincidence.

The resolving time is determined primarily by the lengths of the pulses applied to the two inputs of the coincidence circuit, because if the two pulses partially overlap in time a coincidence will be registered. Another factor, however, is the rise time of the output signal. The basic action of the coincidence circuit is to give a larger output pulse for coincident than for non-coincident inputs. But if the input pulses are made much shorter than the rise time in the output circuit, the output pulses cannot reach their full size. The discrimination between coincident and non-coincident signals is then lost because all pulses are of a standard size determined by the rise time of the circuit and the time available. The nature of the coincidence circuit therefore sets a limit to the resolving time that can be attained by shortening the input pulses. In this respect the circuit of Fig. 72*b* is superior because the rise time of the negative output pulse is short and the output signal for non-coincident inputs is small. But Fig. 72*c* is the best circuit; here there is no output at all unless there is a coincidence and the rise time considerations do not apply; any signal at the output, however small, indicates a coincidence. With this type of circuit resolving times of order 10^{-9} second can be obtained.

In a useful modification of the Rossi circuit, shown in Fig. 73, a crystal diode is used in the anode network. The diode is initially conducting, and remains conducting for non-coincident input signals: the output signal developed across the diode is therefore very small. For a complete coincidence, however, the current through the diode drops to zero and a large output signal is obtained.

With the circuit given any number of valves may be added to form a multiple coincidence system. In the quiescent state the common anode voltage is clamped at 70 v by the diode; the current through the 22 kΩ and 1 kΩ resistances is therefore fixed at about 4 ma, and the rest of the anode current required by the valves is provided from the 70 v line via the diode. If all but one of the valves are cut-off it is only the current through the diode that changes: small output pulse. But when all the valves are cut-off the current through the 1 kΩ load resistance is also interrupted, and we obtain a 4 v output pulse.

This circuit is particularly suitable for multiple coincidence systems of high resolution, resolving times as low as 5mμs being readily achieved.*

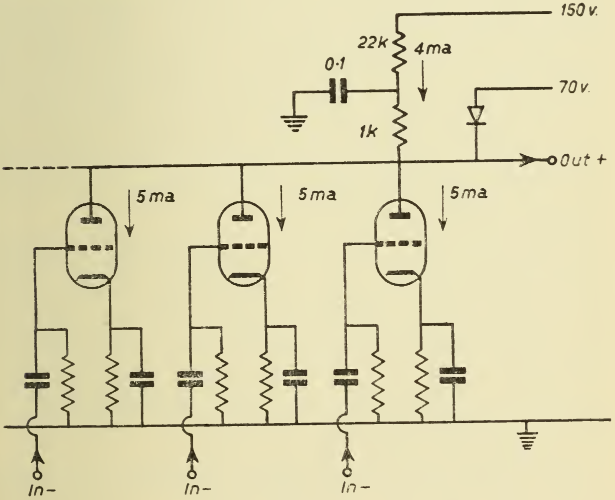


FIG. 73.

7.9. Pulse amplitude selector. Pulses which have amplitude greater than a pre-set threshold value may be selected by means of a simple discriminatior, for example by means of a series diode (§ 2.1) or a Schmitt trigger circuit (§ 4.3). A pulse amplitude selector is more elaborate; it has two threshold voltages and responds only to pulses which have amplitude lying in between the two threshold values. If the interval between the thresholds is small the device therefore

* R. L. Garwin, *Rev. Sci. Instrum.*, 24,618, 1953.

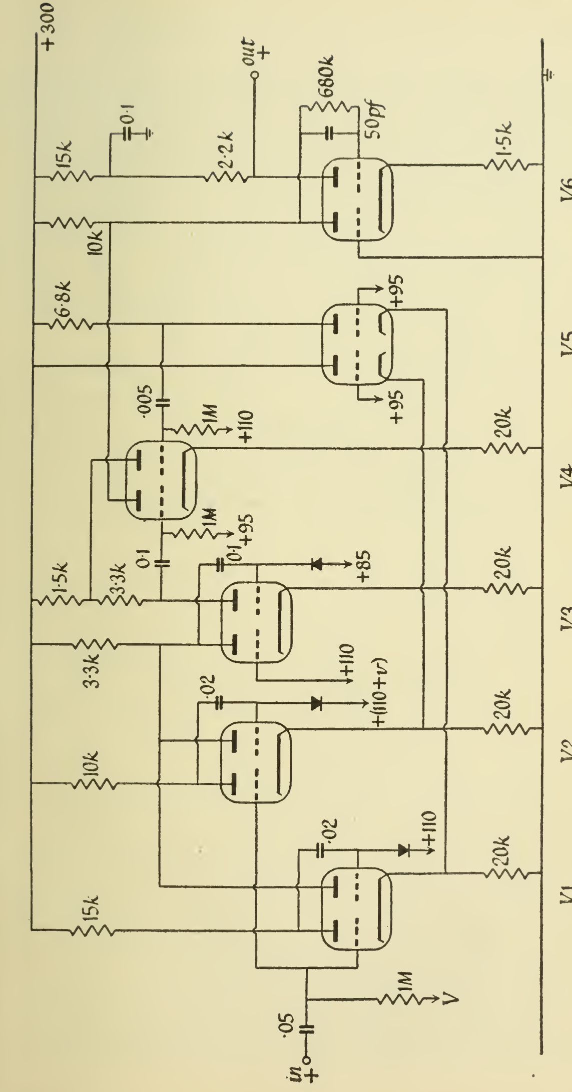
transmits only pulses of a particular height with a small margin of tolerance, and may be used to select such pulses out of a miscellaneous congregation.

In practice a pulse amplitude selector includes two simple discriminators, which define the two thresholds, and an arrangement for generating a standard output pulse when the input exceeds the lower, but not the upper threshold. The output is initiated by the lower discriminator, but blocked when the upper discriminator is also triggered. Inevitably the lower discriminator is actuated first, but the output pulse must be delayed so that the blocking action can take place if required. This is achieved by generating the output pulse not when the lower discriminator triggers, but when it resets. This occurs during the tail of the input pulse, and if at this stage the upper threshold has not been crossed it is clear that it never will be.

In Fig. 74, valves *V1* and *V2* are each Schmitt trigger circuits (§ 4.3) with a long time constant coupling between anode and grid (cf. Fig. 37) : (no resistance is needed across the diodes in this case, because crystal diodes are used and these have a back-resistance of about 1 MΩ). With the bias voltages shown the Schmitt circuits act as discriminators and define the two thresholds. The voltage interval between the thresholds is v, and they are adjusted together by altering V, which ranges from 0-90 volts.

Each Schmitt circuit is modified by connection to one half of *V5*, which acts as a backlash control. *V5* is normally non-conducting, and plays no part in the trigger action when a positive pulse is applied to the grid of say *V1a*. On the tail of the pulse, however, *V5b* conducts and prevents the common cathode potential from falling more than 15 volts below its original steady level. As the input voltage continues to fall *V1a* is cut off, and the trigger circuit resets;

FIG. 74.—Pulse amplitude selector.
V1, 2, 3, 4, 6 ECC91 (6J6)
V5 ECC81 (12AT7)
Crystal diodes CG4C (IN63)



+300
out
+
680k
50pf
0·1
15k
2·2k
10k
6·8k
1·5k
V6
+95
+95
V5
·005
1M
+110
20k
V4
1M
+95
0·1
0·1
+85
1·5k
3·3k
20k
V3
3·3k
+110
·02
+(110+v)
20k
V2
10k
·02
+110
20k
V1
15k
in
+
·05
1M
V

V1b now conducts raising the common cathode voltage to its original level and *V5b* is once more non-conducting. The backlash is therefore controlled by the difference between the steady bias voltages applied to *V1b* and *V5b*, and can be set at any desired value. It will be noted that *V5b* conducts only during the tail of the input pulse when the discriminator is resetting. A negative signal coincident with the resetting of the discriminator can therefore be obtained from the anode of *V5b*. It is this signal that is used to generate the output pulse, provided always that the upper discriminator *V2* has not been triggered by the same input pulse.

The output signal, Fig. 75*e*, is passed from *V5b* to the long-tailed pair gate valve *V4*. Normally *V4b* is conducting, but it is turned off by the signal so that *V4a* conducts and triggers the output trigger circuit *V6* (§ 4.4). If, however, the upper discriminator *V2* has been triggered a negative pulse is applied to the grid of *V4a*, with the result that this valve never conducts, the signal path is therefore blocked and there is no output.

The negative blocking pulse is generated by the trigger circuit *V3*. The left-hand anode is connected directly to the free anodes of the two discriminators, and therefore when the discriminators are triggered by an input pulse such as Fig. 75*a* a composite signal Fig. 75*b* appears at the common anode. This signal passes through the coupling condenser to the grid of *V3b* ; on the second step the bias voltage is overcome and the circuit triggers. *V3a* now becomes non-conducting and a third positive voltage step appears at its anode, the actual waveform being as shown in Fig. 75*c*. The result is that when the discriminators reset the first negative step is now insufficient to restore *V3* to its initial state, and *V3b* remains conducting until the lower discriminator *V1* resets. The overall waveform at the anode of *V3a* is Fig. 75*c*, and the blocking pulse generated at the anode of *V3b* is given in Fig. 75*d*. Note that the blocking pulse is not generated unless the upper discriminator is actuated, and that when it is generated it lasts until the lower discriminator resets, that is just long enough to block the signal from *V5b*,

Fig. 75*e*. The circuit thus automatically adjusts the length of the blocking pulse to correspond to the duration of the input signal, so that the blocking action is secure but not excessive whatever the length of the input pulse. (Valve *V3* can be regarded as a Schmitt trigger circuit with the input

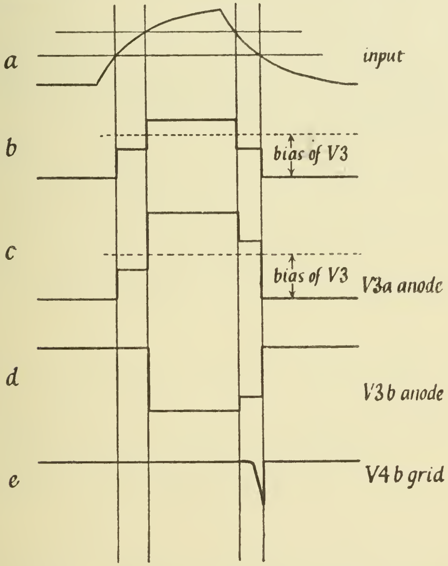


FIG. 75.—Waveforms in pulse amplitude selector.

of Fig. 75*b* applied to the anode instead of the grid; the backlash is chosen to ensure that the circuit triggers on the second positive step, but resets, not on the first, but on the second negative step.)

One further feature remains to be mentioned. When the input pulse has a slowly falling tail the signal Fig. 75*e* rises too slowly to trigger the output circuit *V6* which is set to give very short output pulses. This difficulty is overcome by connecting the free anode of *V4b* on to the anode load of *V3b* as shown in the diagram. *V4* now becomes a trigger circuit with a backlash of about 7 volts, and is triggered by the signal Fig. 75*e* from *V5*. The waveform transmitted to the output circuit *V6* is therefore steep-sided and actuates it without difficulty. If, however, a negative blocking pulse is present at the grid of *V4a* then the circuit never triggers and there is no output. It will be seen that *V4* combines the functions of gate circuit and trigger circuit in one double triode valve.

This example shows how long-tailed pair trigger circuits may be interconnected, and illustrates the use of a free anode as part of another circuit.*

* For a combination of this pulse amplitude selector with the amplifier of § 7.2, see F. J. M. Farley, *loc. cit.*

FOR FURTHER READING

A SHORT list of useful books and references is given below for readers who wish to pursue the subject further, but no attempt has been made to provide a complete bibliography.

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INDEX