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Chapter 14

PULSE MODULATION OF TRANSMITTERS

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CHAPTER 14

PULSE MODULATION OF TRANSMITTERS

1. MODULATION REQUIREMENTS

In Chap. 8 various types of oscillators are described, mainly with regard to their ability to generate continuous oscillations. In pulse radar systems, power oscillators are not employed to generate CW but RF pulses. However, since the time taken for oscillations to build up is much less than the duration of the output pulses in present-day Service equipments, this difference does not invalidate the description of the modes of oscillation given in Chap. 8.

Transmitting valves are usually held quiescent for comparatively long periods, of the order of a millisecond. They are then caused to operate, generating a RF pulse lasting for about one microsecond, after which they remain inoperative until the next pulse is required. The unit which controls the Firing or Pulsing of the oscillator is called the Modulator. This term is retained from Radio Communication, in which the RF output is modulated in a manner which conveys the intelligence. It is usually not convenient to regard the modulator from the point of view of the integrated effect of recurrent pulses but to consider it instead as a switching device, controlling the initiation, and, sometimes, the shape and the duration of the output pulses. Considered in this manner it is the function of the modulator to maintain the oscillator valve quiescent during the intervals set aside for reception, and to bring it into oscillation as and when required, maintaining the controlling potentials at suitable values throughout the duration of the pulse. Also, except in the case of self-quenching oscillators, the modulator controls the duration of the output pulses.

In some systems the modulator controls the recurrence period by its own independent timing mechanism. This may be necessary when spark gap modulators are employed, since these are prone to jitter so that they cannot readily be triggered to spark at an exact instant, and the uncertainty may lead to unnecessary timing errors. Either the modulator pulse or a rectified version of the transmitter pulse may be used to start the timing mechanism for each successive recurrence period. Where gas-filled triodes or hard valves are used as modulator valves, or when the errors due to spark-gap jitter are not important, the modulator may be externally controlled by a master timing system.

In pulse radar as commonly employed successive transmitter pulses are not phase-linked and are said to be Non-Coherent. Some types of radar system require Coherent pulses; i.e. successive pulses must conform to the phase of a continuously running oscillator. In such systems the problem of modulation is more complex. In this chapter we shall consider modulators for the production of non-coherent RF pulses only.

2. MODULATION METHODS

Pulse modulators may first be divided into two main types:-

I Grid Modulators

II Anode Modulators

Although many early radar systems employed I this is now generally abandoned in favour of II for the following reasons.

(a) It is necessary with grid modulation to maintain a high HT voltage for much longer than with Type II. This wastes power and necessitates the use of larger and more elaborate equipment.

(b) More precise control is possible with anode modulation.

(c) Most transmitter valves (magnetrons) now used for operation at centimetre wavelengths have no control grids.

Anode modulators may be classified under two general headings:-

(i) Soft Valve Modulators.

(ii) Hard Valve Modulators.

This division is more profound than the mere names suggest. Because of the high power handled by the anode modulator, which must apply to the oscillator valve the necessary HT voltage (usually several kilovolts), and which, during the pulse, carries the full current of the oscillator valve (usually several tens of amps) it is a decided advantage to use soft valves, which are capable of passing larger currents with less dissipation than hard valves. Unfortunately, soft valves, although readily triggered, are not so readily extinguished. It is necessary, when using a soft modulator valve, to employ a pulse-forming network as a supply source for the oscillator valve. The power available for each pulse is then utilised during the double transit time of the network and the oscillator is therefore unable to oscillate after that time.

Where hard valve modulators are used there is no such limitation. The shape and duration of the modulator pulse are determined by a pulse forming network at a low power level. Subsequently the pulse is amplified by hard valve amplifiers which are controllable, both at the beginning and at the end of the pulse, by the potentials of their control grids.

The divisions (i) and (ii) have, therefore, the following significance.

(i) The pulse-forming network is made to carry the whole of the pulse energy and is discharged into the oscillator in series with the soft modulator valve, which controls the start of the discharge. The problem of charging the network and causing it to discharge as and when required is the main consideration.

(ii) The pulse forming network is incorporated in a low-power stage and the pulse is subsequently amplified until the power level is sufficient for modulating the oscillator. The problem of amplification, and the choice of suitable valves, is the main consideration.

Both systems are commonly used in Service equipments.

GRID MODULATION

3. Pulse Amplifier Method

The essential circuit of this method is shown in Fig. 600.(a).

The oscillator is shown for simplicity as a single valve, valve 1, with lumped L-C tuning, although arrangements using push-pull valves and tuned lines may be employed. The RF connection from the tuned circuit back to earth is via the condenser C, but the DC path is via the resistor R, which forms the anode load of the pulse amplifier, valve 2. The cathode of valve 2 is returned to the negative supply line, $-_2V_B$. The action is as follows.

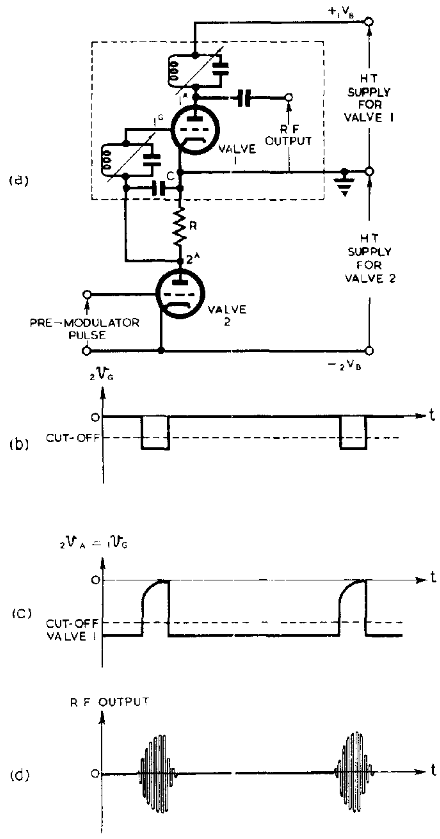


Fig. 600 - Grid modulation: pulse amplifier method.

(i) Valve 2 is normally conducting heavily, since its grid-cathode potential is approximately equal to zero. The flow of anode current through R depresses the potential at 2A, and therefore the grid bias of valve 1, below earth by an amount sufficient to cut off the anode current of valve 1, which therefore does not oscillate.

(ii) The anode current of valve 2 is cut off by means of a negative pulse applied to its grid, so that its anode voltage, and hence the grid bias of valve 1, rises towards earth potential. This allows valve 1 to conduct and generate oscillations. The waveforms are shown in Fig. 600.

Capacitance in shunt with R, including the anode-cathode capacitance of valve 2 and that of the RF bypass condenser C, delays the rise and fall of the voltage at 2A, and the grid bias of valve 1. The effect is most marked on the leading edge of the pulse, which rises with a time-constant equal to the product of the total shunt capacitance and the load R; in order to make this time-constant small, R must have a small value (e.g., $3k\Omega$). A second reason for making R small is that the grid current of the oscillator flows through it, tending to produce a negative bias. To obtain the required rise in voltage at 2A (e.g. 1500 volts) across a low resistance requires a large decrease of current (e.g. 0·5 amp). Thus valve 2 must be capable of passing such a current continuously and have an adequate permissable anode dissipation, because it is conducting for the whole of the recurrence period except for the duration of the RF pulse. For this reason, the mean power taken from the HT supply $-_2V_B$ is large. A suitable value for $_2V_B$ is 2·5 kV.

4. Cathode Follower Method

Fig. 601(a) shows the circuit for this method. The DC return path from the grid circuit of the oscillator (valve 1) is via the cathode load R of the cathode follower (valve 2). The operation is as follows:-

(i) Valve 2 normally biases itself nearly to cut-off, because its grid circuit is returned to the negative end of R, the voltage at 2K becoming positive with respect to $-2V_B$ by an amount approaching the cut-off bias of valve 2. This means that the voltage at 2K, and hence the grid bias of valve 1, are sufficiently negative with respect to earth to cut off the anode current of valve 1.

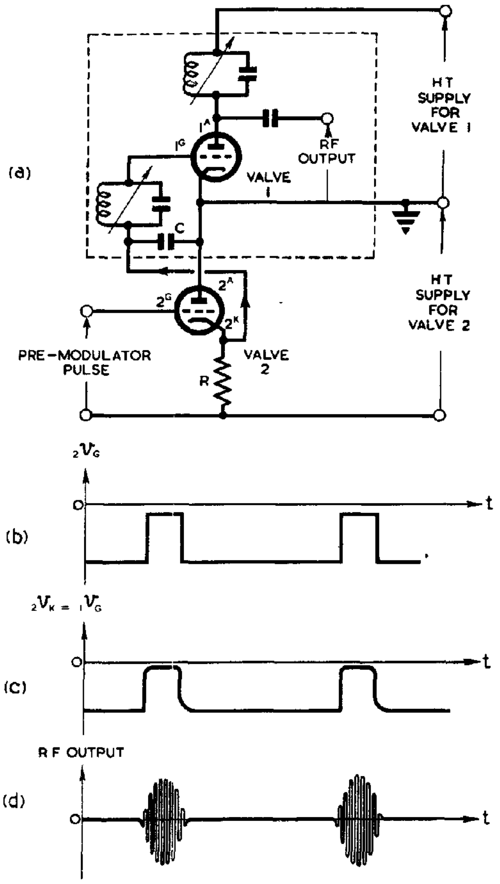


Fig. 601 - Grid modulation: cathode follower method.

(ii) A positive pulse applied to the grid of valve 2 raises the grid voltage to a value approaching earth potential, and the cathode voltage, together with the grid bias of valve 1, follows, so bringing valve 1 into conduction.

Capacitance in shunt with R produces little delay in the rise of the pulse because of the low output impedance of the cathode follower. The trailing edge, however, may be adversely affected, since for this the output resistance of the cathode follower is larger than for the leading edge. The fall in voltage at 2K tends to lag behind that at 2G so that the valve current is cut off. The output resistance for the trailing edge of the pulse is thus approximately R, and the corresponding time-constant is the product of R and the total shunt capacitance.

The waveforms for the cathode follower circuit are shown in Fig. 601.

Advantages of this method compared with the former are:-

(i) The modulator valve, valve 2, is conducting heavily only during the pulses; consequently its anode dissipation during the quiescent period is comparatively low and the valve is operated more efficiently. However, the cathode of valve 2 must still have a high emissivity because, while valve 2 is conducting, it must carry the grid current of valve 1.

(ii) The distortion of the pulse due to shunt capacitance is reduced.

The cathode-follower modulator requires to be driven by a positive-going pulse of amplitude somewhat greater than that of the output pulse required. The stage generating this pulse can be an amplifier of the type described in Sec. 3 but can have a larger output resistance than is permissable if such a stage is employed as the modulator. This is because the input impedance of the cathode follower is considerably higher than that presented by the grid circuit of the oscillator.

The design of a grid modulator is similar to that of the Pre-Modulator Stage for a hard valve anode modulator. It is therefore suggested that Secs. 9 and 10 should be read in conjunction with the above.

5. Grid Self-Quenching

If an oscillator is provided with a C-R network in its grid circuit, it may be made to generate RF pulses. The phenomenon of Grid Self-Quenching (squegging) has already been described in Chap. 8, Sec. 47. It is sufficient here to indicate the circuit to be used and the waveforms produced, illustrated in Fig. 602. In Fig. 602(a) the condenser C is shown inserted between the top end of the L-C circuit and the grid, but if more convenient it may be placed between the bottom end and earth. The resistor R is connected either to zero or to a positive potential.

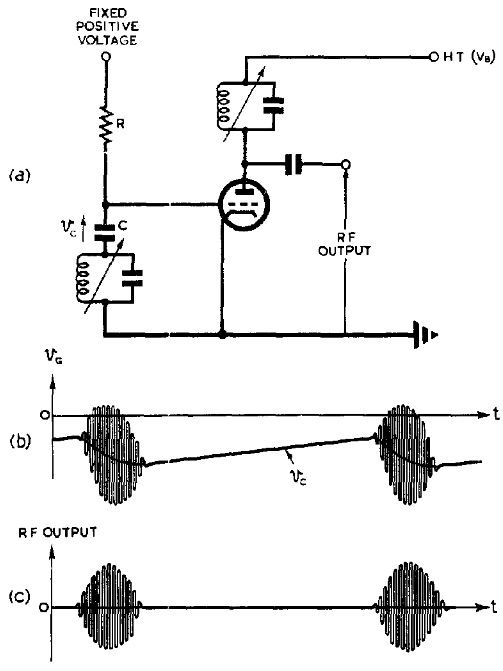


Fig. 602 - Grid self-quenching oscillator

The squegging oscillator has the merit of extreme simplicity but possesses several disadvantages:-

(i) The bias is varying throughout the duration of the pulse so that optimum conditions of oscillation are not maintained.

(ii) The pulse shape is not very good.

(iii) The repetition period is rather indeterminate because it depends on the slow exponential rise of the grid voltage to cut-off. Variation of cut-off potential with ageing of the valve or between different valve specimens can thus cause alternations of recurrence frequency. This tendency is reduced when R is returned to a positive voltage.

(iv) The RF pulse length is also rather indeterminate because it depends on the rapidity with which C charges due to grid current flow. Change of grid-current characteristics with ageing of the valve or change of valve specimen thus produces variation of pulse width.

The indeterminacy of the recurrence frequency of the freely running squegging oscillator can be eliminated by triggering the grid circuit of the oscillator by one of the methods described in Secs. 3 and 4. The triggering pulse is longer than the required duration of the RF pulse, the latter being determined as before by the rate at which C is charged by grid-current flow. The circuit may be arranged as in Fig. 603, a cathode follower triggering valve being used. From the waveforms it can be seen that the action of the oscillator is initiated by the triggering pulse and its duration is determined by the squegging action. The free-running repetition frequency is chosen to be greater than the frequency of the triggering pulses so that the valve is ready for operation by the time each triggering pulse is applied.

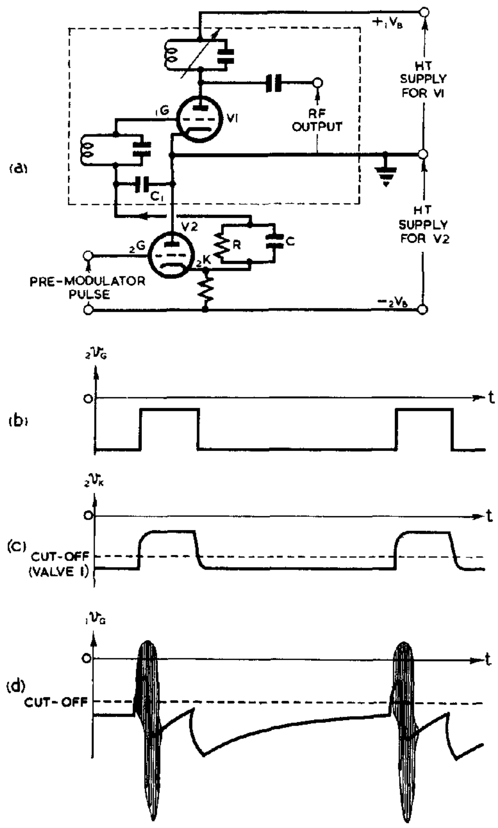


Fig. 603 - Triggered squegging oscillator.

The duration of the trigger pulse need not be shorter than is necessary to prevent double pulsing. Double pulsing occurs if the rate of rise of mean grid potential after the end of the RF pulse is such that cut-off is reached before the end of the positive portion of the triggering pulse. In this case a second RF pulse is produced by the oscillator.

The advantage of the arrangement compared with the simple cathode follower type of grid modulator is that it removes the necessity of generating triggering pulses as short as the duration of the RF pulse.

6. Anode Self-Quenching

Although this is not a method of obtaining grid modulation, it is convenient to mention it here, as it is used in conjunction with triggered grid self-quenching circuits to reduce the likelihood of double pulsing. If the anode circuit of the oscillator is fed from the HT supply via a C-R network as shown in Fig. 604, the applied HT, instead of remaining constant as with direct feed, varies in the manner shown in Fig. 604. The applied HT falls rapidly due to the discharge of C by anode current flow while the valve is oscillating during the first part of the trigger pulse; it then commences to rise as C recharges through R, but the rise during the remainder of the trigger pulse is small. Because of the low value of applied HT during this period, the cut-off value of grid potential is less negative; consequently the mean grid potential can rise further during the remainder of the triggering pulse without the oscillator coming into conduction and producing a second RF pulse. The variation of applied HT also affects the RF pulse length and shape.

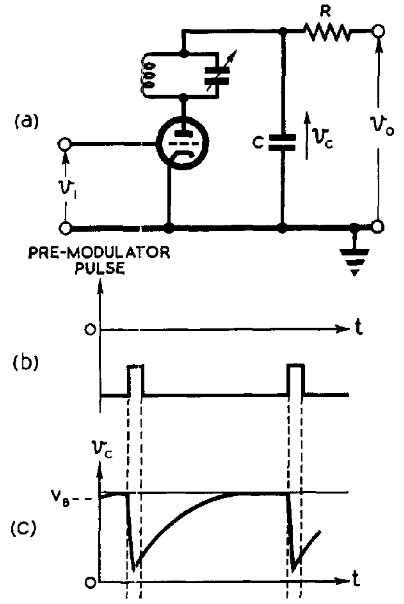


Fig. 604 - Anode self-quenching

The action of the triggered self-quenching circuits can be made more definite by substituting an open-ended delay network for the condenser C of the C-R circuit. The operation of self-quenching using delay networks will not be described as they have not been used to any great extent. Anode self-quenching using a delay network constitutes one of the important methods of anode modulation and is described in Sec. 11.

7. Disadvantages of Grid Modulation

(i) If valves with oxide-coated cathodes are used grid modulation may be seriously affected because of grid emission caused by deposition of oxide-coating on the grid. Gold plating the grid reduces this effect, the gold poisoning the sputtered coating and preventing emission. Even with thoriated tungsten filaments grid emission can occur for either of two reasons.

(a) If the grid, by reason of its proximity to the cathode, becomes very hot, primary emission from the grid may occur.

(b) If the grid is driven very positive, secondary emission may result.

In any case the Reverse Grid Current tends to cause the squegging network to charge up in the wrong sense, with possibilities of CW oscillations ensuing. In general, special precautions must be taken to prevent such reverse current from passing through the squegging network, such as the use of bypass diodes.

(ii) The HT is applied to the oscillator continuously, so increasing the tendency for the phenomenon known as flash-arcing to occur; this is the complete breakdown of the high insulation normally afforded by the high vacuum between the electrodes of the valve.

ANODE MODULATION

8. General

In this type of modulation, the mean potential of the grid (if any) of the oscillator remains constant, whilst the anode-cathode voltage is switched to the operating value for the required duration of the RF pulse. This action may be brought about by switching either anode or cathode voltage, the latter being used for oscillator valves which work with their anodes earthed.

9. Hard Valve Modulator

The basic circuit for Hard Valve Modulation is shown in Fig. 605. The oscillator and modulator valves are connected in series across the HT supply. The modulator-valve current is normally cut off by the negative grid bias and no current can flow in the oscillator valve. A positive pulse is supplied to the modulator grid so as to bring the valve into heavy conduction for the required duration of the RF pulse. During the pulse, the modulator valve should offer a low impedance so that the greater part of the HT voltage may be developed across the oscillator valve.

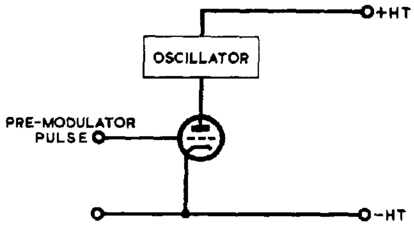


Fig. 605 - Basic circuit of hard valve modulator.

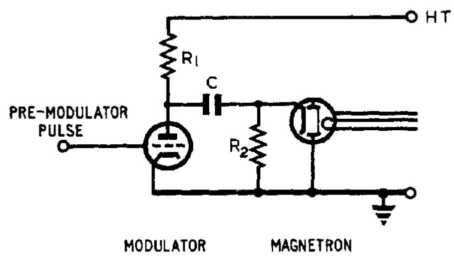


Fig. 606 - Shunt-fed modulator circuit.

If a magnetron oscillator is employed the HT positive lead must be earthed since the anode of the magnetron is at earth potential because of its mechanical construction. This may be avoided by adopting a shunt-fed arrangement, as in Fig. 606. To ensure correct operation the voltage developed across the magnetron must remain steady during the pulse. This means that a long time-constant coupling circuit should be used. The resistances involved in this coupling circuit are:-

R_1 and R_2
R_V, the resistance of the modulator valve when conducting, and
R_m, the resistance of the magnetron when oscillating.
For maximum efficiency the following results must hold:-

$$R_1 \gg R_V$$
$$R_2 \gg R_M$$
$$R_M \gg R_V$$

Hence the condition for small distortion of the rectangular pulse is that $CR_M \gg$ the duration of the pulse.

10. Modulator Valve Requirements

The Modulator valve must satisfy the following requirements:-

(i) The cathode must have a high emission, since the oscillator valve current passes through the modulator valve.

(ii) The resistance of the valve when conducting must be small compared with that of the oscillator in order that the greater part of the HT voltage may be developed across the oscillator.

(iii) The valve must be capable of withstanding the full HT.

(iv) The valve must have small inter-electrode capacitances in order to preserve pulse shape.

(v) The maximum permissible anode dissipation must be adequate.

These different requirements conflict with one another, and the design is necessarily a compromise.

Two examples of modulator valves are given below :-

(i) CV57 Beam Tetrode

Thoriated cathode. Gold-plated grid.
Max. permissible anode current = 5A
Impedance = 500 Ω
Maximum anode voltage = 11 kV
Input capacitance = 21 pF
Output capacitance = 11 pF
Maximum anode dissipation = 15W

Three of these valves in parallel are normally used for modulating a CV64 magnetron.

(ii) 715C Beam Tetrode

Oxide-coated cathode. Gold-plated grid
Max. permissible anode current = 15A
Impedance = 300 Ω
Maximum anode voltage 18kV
Input capacitance = 35pF
Output capacitance = 8 pF
Maximum anode dissipation = 60W
Rated Anode Voltage = 15kV

11. Anode Modulation using a Pulse-forming Network in Conjunction with a Spark Gap or a Gas Triode

In this method the modulating pulse is obtained by discharging a pulse-forming network into the load presented by the oscillator through a spark gap or a gas triode. The network is charged from the HT source in the intervals between pulses.

The principle of operation may be understood by first considering the action of the circuit of Fig. 607(a). This consists of the anode self-quenching circuit of Fig. 604, with the addition of a switch S in series with the oscillator. In practice, S takes the form of a spark gap or a gas triode; these possess the property of maintaining conduction at very much lower voltages than those at which they strike.

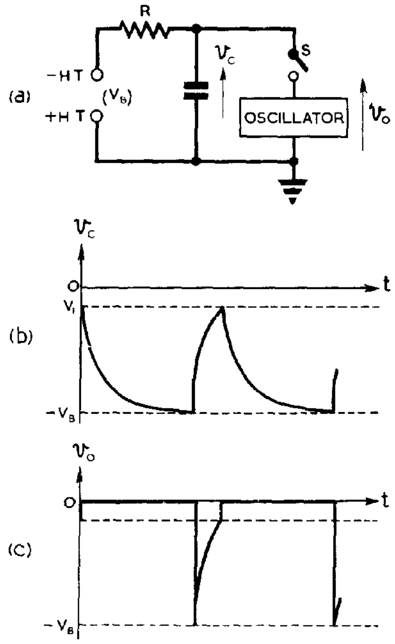


Fig. 607 - Anode modulation using C-R network.

While S is open, the condenser C charges through resistor R to the HT voltage. When S is closed, C discharges through the load presented by the oscillator, the resistance of R being very much greater than that of the load. When the voltage across C reaches some value V_1, Fig. 607(b), S opens (i.e. the spark gap or gas triode extinguishes) and C recharges through R. The voltage across the oscillator is zero between pulses and, neglecting the small loss of voltage across S, is equal to v_c while S is closed.

The circuit just described is capable of generating modulating pulses for the oscillator, but the pulse shape is far from rectangular. This disadvantage may be overcome by replacing C by an open-ended network as shown in Fig. 608. While S is open, the condensers charge through R as if they were all connected in parallel, the inductances having negligible effect during the relatively slow charge. The total charging capacitance is thus n C; (b). When S is closed the network discharges producing a more or less rectangular pulse across the oscillator; (c). The manner of this discharge depends on the following considerations:-

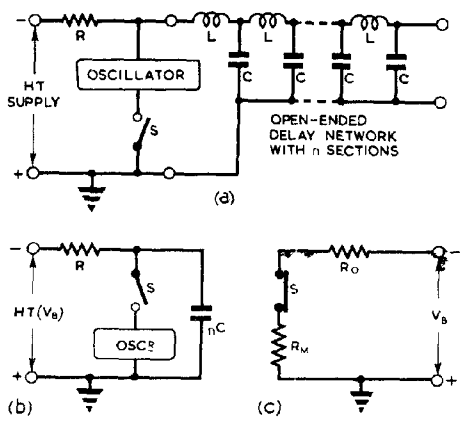


Fig. 608 - Anode modulation using network.

(i) The number of sections in the pulse-forming network. The more sections there are the more nearly rectangular the pulse becomes.

(ii) The total inductance (nL) and capacitance (nC) of the network. The duration of the pulse is $2\sqrt{nL.nC} = 2n\sqrt{L.C}$

(iii) The ratio of the characteristic impedance of the network at zero frequency to the resistance of the load. The existence of subsequent reflections, their magnitude, and polarity, depend on this ratio.

These effects are dealt with in Chap. 4, Sec. 12 and Chap. 3, Sec. 16. Various methods of charging the delay network may be employed. These are dealt with in Secs. 23 - 36.

Owing to the high voltages used, the network is usually immersed in oil to minimise leakage and the possibility of a flash-over.

PULSE TRANSFORMERS

12. General

High voltage transformers normally used in power supply circuits are designed to operate with currents which are approximately sinusoidal, and are required to produce a corresponding sinusoidal output voltage. In high-power pulse circuits the requirements are different, and ordinary transformers are liable to distort pulse-shape very considerably.

The term Pulse Transformer is applied to transformers which are designed to pass pulses without appreciable distortion of their shape.

Pulse transformers find particular application in the modulator section of radar equipments, being used for the following purposes:-

(i) Matching, e.g. matching the modulator or magnetron to the cable connecting them.

(ii) Reversal of a pulse, e.g. between the pre-modulator and modulator valves.

(iii) Coupling between two circuits at different steady voltage levels.

13. Equivalent Circuit of a Pulse Transformer

The distortion of the pulse shape produced by a pulse transformer can be determined by analysis of its equivalent circuit. An approximate equivalent is shown in Fig. 609. A mathematical justification for this equivalent circuit can be given, but it is sufficient here to indicate the significance of the various circuit elements shown.

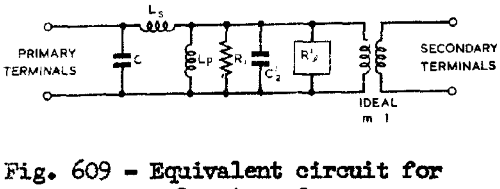


Fig. 609 - Equivalent circuit for pulse transformer.

C_1 is the primary winding self-capacitance + output capacitance of the stage feeding the primary.

L_s represents the leakage inductance between primary and secondary i.e. the inductance of the primary when the secondary is short-circuited.

L_p represents the magnetising inductance, i.e. the inductance of the primary when the secondary is open-circuited.

R_1 represents the losses due to dissipation in the iron core.

m is the step-down ratio of the transformer (the turns ratio).

$C_2' = C_2/m^2$, where C_2 is the secondary winding self-capacitance + stray shunt capacitance + input capacitance of secondary load.

$R_\ell' = m^2 R_\ell$, where

R_ℓ is the secondary load resistor.

C_2' and R_1' are the equivalent values of C_2 and R_1 reflected from the secondary to the primary circuit of the transformer.

The Ideal Transformer merely transforms the voltage in the ratio of the turns, and introduces no other circuit elements.

14. Pulse Transformer Connected Between a Modulator Valve and Magnetron

An approximate treatment of the operation of a pulse transformer connected between a modulator and magnetron (Fig. 610(a)) is now given. The modulator and magnetron valves conduct heavily during the pulse, and are non-conducting between pulses. The equivalent circuits for the duration of the pulse, and for the interval between pulses are considered separately.

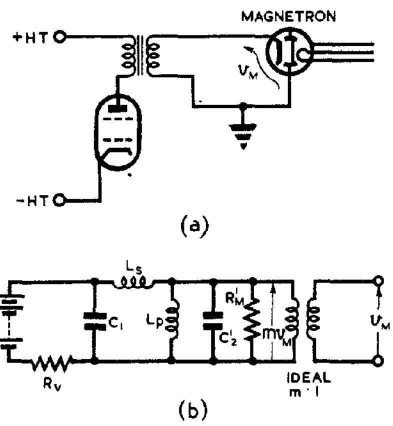


Fig. 610 - Pulse transformer connected between modulator valve and magnetron.

(i) Operation during pulse

The equivalent circuit during the pulse may be derived from Fig. 609 and is shown in Fig. 610(b). In this figure

R_v = anode-cathode resistance of the modulator valve when conducting.

$R_M' = m^2 R_M$ where

R_M = anode-cathode resistance of magnetron when conducting.

v_M = voltage across magnetron.

R_1 of Fig. 609 is ommitted because $R_1 \gg R_M'$.

This still complicated equivalent circuit can be simplified by considering separately the action of the circuit for the leading edge of the pulse (modulator just switched on) and during the steady portion of the pulse.

(a) Response on leading edge.

The equivalent circuit may be simplified by omitting C_1 , since C_1 can charge very rapidly through the low impedance R_V of the modulator. We may also omit L_p for the following reasons:-

In the first instance neglect Ls since $L_s \ll L_p$. Then the current through Lp is small provided the time-constant $\frac{L_p}{R_V}$ is much greater than the corresponding time-constant for the circuit formed by C_2', R_M' and R_V, i.e.

$$\frac{L_p}{R_V} \gg \frac{C_2' \; R_M' \; R_V}{R_M' + R_V} .$$

Normally $R_M' \gg R_V$ so that this reduces $\frac{L_p}{R_V} \gg C_2' \quad R_V$.

This condition is normally satisfied in a pulse-transformer circuit. The simplified circuit is shown in Fig. 611(a); in this the ideal transformer is omitted for clarity.

The voltage v_M across the magnetron does not instantaneously attain its steady value of $-\dfrac{V_B \; R_M'}{m(R_M' + R_V)}$, because of the presence of L_s' and C_2' which form a series resonant circuit. Oscillations occur if the damping factor ζ of the circuit is less than unity; (see Chap. 2, Sec. 10). Any reduction in damping below this critical value will

(1) introduce a ring, which is undesirable;

(2) steepen the leading edge and so accelerate the build-up, which is desirable.

A compromise must be achieved between these effects, and the value $\zeta = 0{\cdot}7$ is commonly used. This compromise is illustrated in the following typical specification:- "The pulse must attain 90% of its final voltage within 0·1 μsec, and the amplitude of the ripple must not exceed 10% of the mean pulse voltage." The pulse shape for various values of ζ is shown in Fig. 611(b).

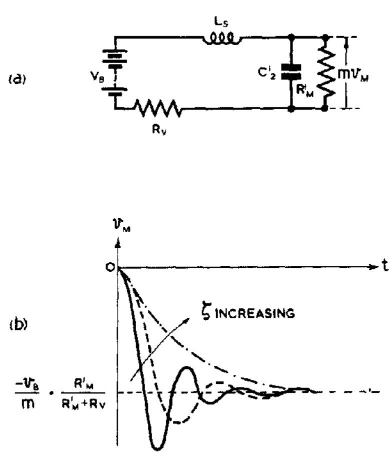


Fig. 611 - Performance of pulse transformer during leading edge of pulse.

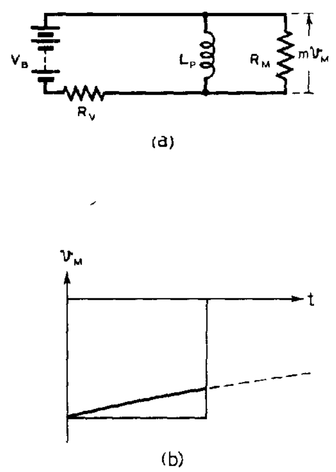


Fig. 612 - Performance of pulse transformer during steady portion of pulse.

(b) Response during steady portion of pulse.

Here the gradual build-up of magnetising current in L_p results in a decrease of current through the magnetron. Since this variation of current is relatively slow, C_1, L_s, and C'_2 may all be neglected. The resultant simplified circuit is shown in Fig. 612(a). The voltage mv_M across L_p falls from its initial value with a time constant

$$\frac{L_p\,(R_M' + R_v)}{R_v\,R_M'} \doteqdot \frac{L_p}{R_v},$$

the pulse shape being as depicted in Fig. 612(b).

Combining the effects on the leading edge and during the steady portion of the pulse we obtain Fig. 613.

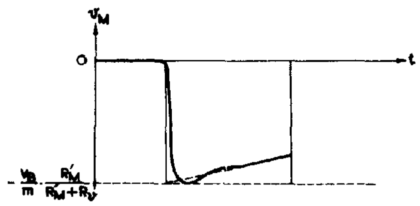


Fig. 613 - Pulse transformer: combined effects on leading edge and steady portion of pulse.

(ii) <u>Operation between Pulses</u>

We will consider the effects on the trailing edge of the pulse, when the modulator and magnetron have become non-conducting. The equivalent circuit is shown in Fig. 614(a). C_1 and C_2' are charged and must discharge before equilibrium is reached. Two types of ring can occur:-

(a) A relatively low-frequency ring due to L_p resonating with the parallel combination of C_1 and C_2', L_s having little effect for this slow oscillation. Damping occurs due to R_1.

(b) A high-frequency ring due to L_s resonating with the series combination of C_1 and C_2', L_p having little effect on this rapid oscillation.

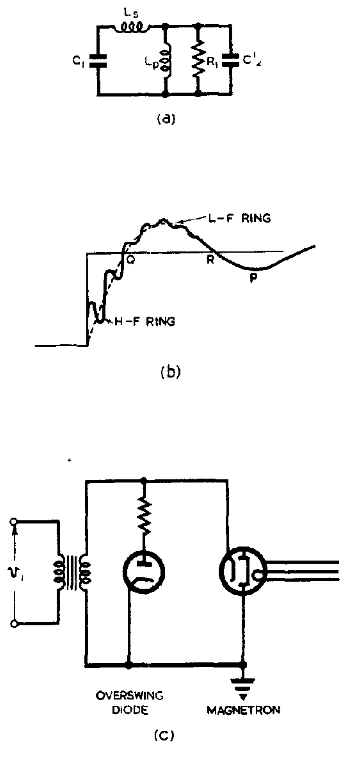


Fig. 614 - Performance of pulse transformer after the end of the pulse.

Fig. 614(b) shows the type of waveform produced, the high-frequency and low-frequency rings being superimposed.

15. Design Considerations

It follows from Sec. 14 (i) that for minimum distortion during the leading edge and steady portion of the pulse the constants must be chosen so that

$$\frac{L_p}{R_v} \gg T_p \gg C_2' R_v,$$

where T_p is the duration of the pulse.
Consequently C_2' should be as small as possible, with a large value of L_p.

In order that the trailing edge of the pulse should be sharp, the frequency of the low-frequency ring should be high; i.e.

$L_p\,(C_1 + C_2')$ should be small.

In general these two sets of conditions cannot be attained unless R_v is very small. A large value of L_p can be obtained by either using a large number of primary turns or using a high permeability core. Of these, the latter is preferable because smaller values of L_s, C_1 and C_2 are then obtained. High-permeability core materials are available in the form of certain nickel-iron alloys, of which the outstanding example is mu-metal. A considerable reduction of leakage inductance can be effected by the use of an auto-transformer, rather than one with two separate windings. This is possible only when primary and secondary circuits are at the same steady potential. **In large pulse transformers high-permeability alloys are seldom used since flux saving is very small.**

The insulation of high voltage pulse transformers is a difficult problem; the most suitable method is to enclose the windings in a sealed oil-filled container. Allowance must be made for expansion of the oil due to heat.

16. Overswing Diode

In Fig. 614(b), the negative peak at P may reach sufficient amplitude to cause the magnetron to oscillate again, leading to the production of a double RF pulse. To prevent this a diode is connected, as in Fig. 614(c) so that it conducts on the preceding positive half-cycle, i.e. between Q and R, and damps out the oscillation. A protective resistance is usually included in series with the overswing diode.

17. Pulse Transformer with Bifilar Secondary

If a pulse transformer is used to feed the HT pulse to the cathode of the magnetron (Fig. 615) the transformer supplying the heater of the magnetron must be rated to withstand the full pulse voltage. This may be avoided by the use of a pulse transformer with a bifilar secondary, connected as in Fig. 616. The two sections of the bifilar winding are numbered 3, 4 and 5, 6. The ends marked 4, 6 are effectively at earth potential so far as the pulse is concerned, 6 being connected to earth directly and 4 being connected thereto via the condenser C_1 (e.g. 0·01 μF). The heater voltage for the magnetron can therefore be applied at the points 4, 6 from a transformer of normal rating.

Equal pulse voltages appear at the ends of the winding of the pulse transformer marked 3, 5 and are applied to the two sides of the heater of the magnetron, one of which is connected to the cathode. The condenser ensures that any inequality in the pulse outputs of the bifilar winding does not damage the heater.

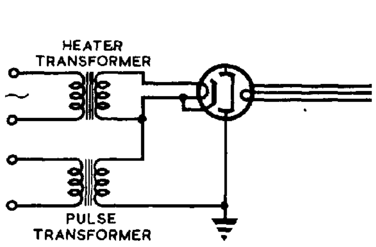


Fig. 615 - Pulse transformer feeding magnetron.

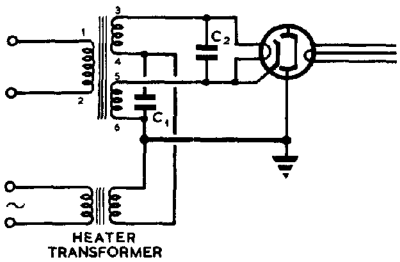


Fig. 616 - Pulse transformer with bifilar secondary.

18. Use of Pulse Transformer in Conjunction with Hard Valve Modulators.

In certain Service applications it is necessary to separate the oscillator and modulator valve circuits. To avoid distortion of the pulse shape by the connecting cable it is necessary to terminate the cable in its characteristic impedance; this may be done by using a pulse transformer to match the impedance of the magnetron into the cable. The matched cable will not in general present a suitable impedance for direct connection to the modulator, and a second pulse transformer is inserted between the modulator valve and the input of the cable. This arrangement is shown in Fig. 617; since the impedance of the cable is normally considerably less than that of the magnetron, the input pulse transformer has a step-down turns ratio, and that at the output a step-up ratio. Because of this, the voltage across the cable is considerably less than that applied to the magnetron, with the consequence that the necessary voltage rating of the cable is much reduced.

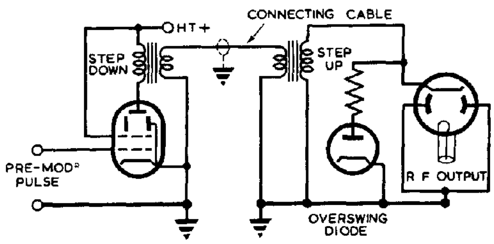


Fig. 617 - Use of connecting cable and pulse transformers between modulator and oscillator.

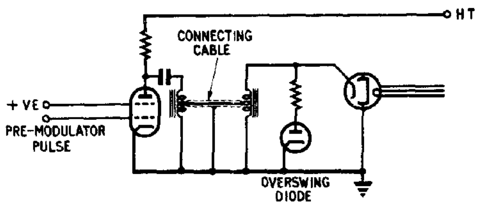


Fig. 618 - Use of auto-transformers between modulator and oscillator.

Example:-

Impedance of magnetron when conducting = $900\,\Omega$

Characteristic impedance of cable = $100\,\Omega$

Amplitude of pulse across magnetron = 12 kV.

Step-up turns ratio required between cable and magnetron

$$\sqrt{\frac{900}{100}} = 3.$$

Voltage across cable $= \frac{12}{3}$ kV = 4 kV.

The purpose of the overswing diode in Fig. 617 is described in Sec. 16.

In place of the double-wound transformer shown (Fig. 617) auto-transformers are frequently employed. An arrangement using auto-transformers is drawn in Fig. 618.

19. Use of Pulse Transformer in Conjunction with Delay Network (Soft Valve) Modulators

Pulse transformers may be used in conjunction with a delay network modulator for the following purposes:-

(i) To step up the voltage of the pulse before application to the oscillator. This allows the use of a delay network with a lower voltage rating. It increases, however, the current which must flow through the spark gap or gas triode to produce a given current in the oscillator circuit, e.g. if 12kV. 10A. are the voltage and current required in the oscillator circuit, and 4kV the output voltage from the delay network, a current of 30A. must flow through the switching device.

(ii) To enable a low impedance cable to be used between modulator and oscillator units.

HARD VALVE MODULATOR CIRCUITS

20. Pre-Modulator

A hard modulator valve requires the application of a positive-going pulse of large amplitude (e.g. 1,000 V) to bring it into conduction. The unit which generates the pulse is known as the Pre-Modulator (or Driver).

The action of the pre-modulator is initiated by a triggering pulse from the radar timing mechanism. This pulse is normally of comparatively long duration and is used to produce a short duration pulse of the desired shape by one of the methods described in Chap. 13. This pulse may go through several stages of amplification, inversion, shaping, etc. before the output stage is reached.

The output impedance of the pre-modulator unit must be low, for the following reasons:-

(i) The grid of a large modulator valve must be driven considerably positive, e.g. 100 V, in order to make the valve conduct as heavily as possible; under these conditions the input impedance is low due to the flow of grid current. Unless the output impedance of the pre-modulator is sufficiently low, limiting will occur as described in Chap. 9.

(ii) A low output impedance is necessary to preserve pulse shape.

The simplest form of pre-modulator output stage consists of a valve with resistive anode load which conducts during the intervals between pulses and is rendered non-conducting by the application of a negative pulse to its grid for the required duration of the RF pulse. In order that the stage may have a low output impedance, it is necessary to use a small anode load. To produce a voltage pulse of large amplitude across a small resistance requires a large current change. Therefore a valve capable of passing a heavy current continuously and having a large permissible anode dissipation must be used, On the other hand the stage supplying the negative pulse which cuts off the current of the output stage can be of low power since it is required to conduct only for the duration of the RF pulse.

The high power consumption of the pre-modulator output stage may be avoided by inserting a phase-reversing pulse transformer between the anode of the output stage and the grid of the modulator valve. In this case the output valve is brought into conduction for the duration of the RF pulse only, and although it must be capable of handling large peak currents, its ratings of maximum emission current and anode dissipation may be comparatively low. Such a stage may, if desired, be driven from a normally quiescent low-power stage through another phase-reversing transformer.

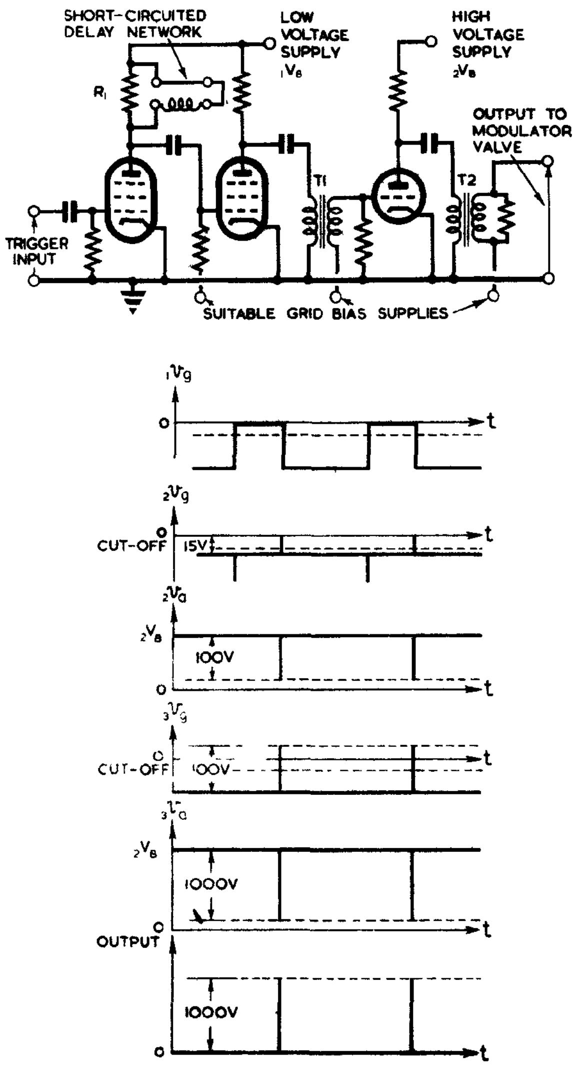


Fig. 619 - Representative pre-modulator circuit and waveforms.

A representative arrangement for a complete pre-modulator unit is shown in Fig. 619 together with the waveforms of the voltages produced at various points. The negative-going edge of the rectangular triggering voltage corresponds to the firing instant of the transmitter; this voltage is applied to the grid of valve 1 via a long time-constant circuit, which in conjunction with the grid-cathode circuit of the valve, clamps the voltage so that its positive-going level at the grid is approximately zero volts. In the anode circuit of valve 1 is a short-circuited delay network, which results in the production of positive-going and negative-going pulses of the required duration. These are applied to the grid of valve 2 which is biased so as to be brought into conduction only during the positive-going pulses. The amplified negative-going pulses at the anode of valve 2 are reversed in phase by the pulse transformer T_1 and applied to the grid of valve 3, which is also biased so as to conduct only during the (positive) applied pulses. The negative-going pulse output of valve 3 is reversed in phase by T_2 before application to the modulator valve.

21. Boot-Strap Pre-Modulator

The employment of pulse transformers, with the consequent possibilities of distortion of the pulse shape, can be avoided, whilst at the same time all valves may be kept in the quiescent state except during the pulse, by use of the Boot-Strap circuit. This circuit, together with waveforms, is shown in Fig. 620.

In the interval between pulses valve 2, a gas triode, is biased by the voltage $-V_G$ so that its grid is sufficiently negative with respect to cathode to prevent it striking. The open-ended delay network in the anode circuit of valve 2 charges to $_1V_B$ through the large resistance R_4 in series with the comparatively small resistance R_7. Valve 3 is biased to cut-off by the voltage $-V_G$.

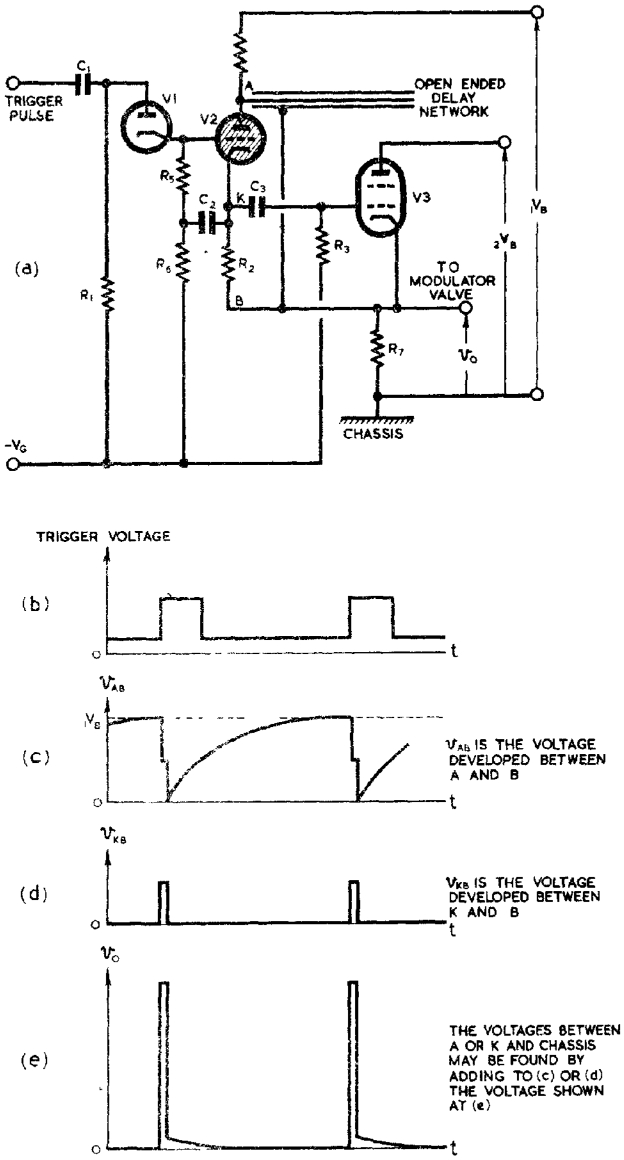


Fig.620 - Boot-strap pre-modulator.

The circuit can be triggered by any pulse with a fairly sharp positive-going edge. This pulse is applied to the short time-constant circuit $C_1 - R_1$. The positive-going portion of the output of this network is applied to the grid of valve 2 via the isolating diode, valve 1. Valve 2 strikes and the delay network discharges through the series combination of valve 2, which has negligible resistance, and R_2, the value of which is made equal to the characteristic impedance of the delay network.

At the instant of striking, the input voltage to the delay network falls to $\frac{1}{2}V_B$, and, neglecting any loss across valve 2, an equal voltage appears across R_2. The voltage across R_2 is applied via the coupling components C_3, R_3 to the grid of valve 3 which is brought into conduction. The flow of current through R_7 raises the cathode potential of valve 3, but this valve does not act as a cathode follower because the lower end of R_2 is returned to its cathode. As the cathode voltage of valve 3 rises, the voltage of R_2 as a whole rises; this raises the cathode voltage of valve 2, but because valve 2 is a gas triode, once struck it continues to conduct, irrespective of its grid-cathode voltage. The voltage across R_2, i.e. that applied between grid and cathode of valve 3, remains constant, since the anode-cathode circuit of valve 2, including the delay network, rises with the cathode of valve 3. R_4 is large and has little shunting effect on this "boot-strap" action. Provided the voltage across R_2 is of sufficient magnitude to raise the potential at the grid of valve 3 above that of its cathode, this valve can be made to conduct very heavily, the voltage across R_7 approaching the value $2V_B$.

The time for which valve 2 and valve 3 remain in conduction is determined by the transit time of the delay network. At the instant at which valve 2 is struck and the delay network starts to discharge, a negative-going rectangular wave of amplitude $\frac{1}{2}V_B$ commences to travel along the network and is reflected without change of phase at the open end. When the wave returns to the input end the line is completely discharged and the voltage across valve 2 and R_2 in series falls to zero. Valve 2 is extinguished, the driving voltage for valve 3 disappears and its current is cut off. The output pulse to the modulator valve therefore lasts for the double transit time of the delay network. The delay network subsequently recharges, valve 2 remaining non-conducting until the next triggering pulse is applied.

A condenser C_2 is connected from the cathode of valve 2 to the junction of two resistors R_5 and R_6 in order to prevent the grid-cathode voltage of valve 2 becoming highly negative when the cathode voltage of valve 2 rises during the pulse due to the "boot-strap" action. The purpose of the isolating diode is to prevent the passage of a high positive-going pulse back to the triggering circuit from the grid of the valve due to the presence of C_2.

The potentials of the supply circuits are chosen with respect to earth potential so that the output pulse to the modulator rises to a suitable value. Since this is normally not greater than earth potential the common negative supply line (shown connected to chassis) is usually several hundreds of volts negative.

22. Single Valve Pre-Modulator

A pre-modulator circuit using a single valve only is shown in Fig. 621 together with the appropriate waveforms. The circuit consists of a triggered blocking oscillator, modified by the addition of a delay network which determines the duration of the pulse produced. The anode and grid circuits of the valve are coupled together by means of two windings of a three-winding pulse transformer, and the output pulse is taken across the third winding. One end of the grid winding is connected to an open-circuited delay network, the circuit to earth being completed via the resistance R_1, across which the triggering voltage is developed. The grid is normally biased beyond cut-off through the grid leak R_2.

The circuit can be triggered by the leading edge of a positive-going pulse of sufficient amplitude. This pulse is applied to a short time-constant circuit C_1 - R_1, and the output is applied via the delay network and the grid winding of the transformer to the grid of the valve, bringing the voltage above cut-off. Regenerative action then causes the valve to conduct heavily, a large voltage of polarity as indicated in Fig. 621 appearing across the grid winding. This winding now acts as a source of voltage which charges the delay network via the resistance R_1 and the grid-cathode circuit of the valve, the latter being now of low resistance because the voltage of the grid is positive with respect to that of the cathode. As the delay network charges a negative-going rectangular wave travels towards the open end, where it is reflected without change of phase. On its return to the input end, the grid voltage falls and the valve current is cut off. The output pulse thus lasts for the double transit time of the delay network. Damping

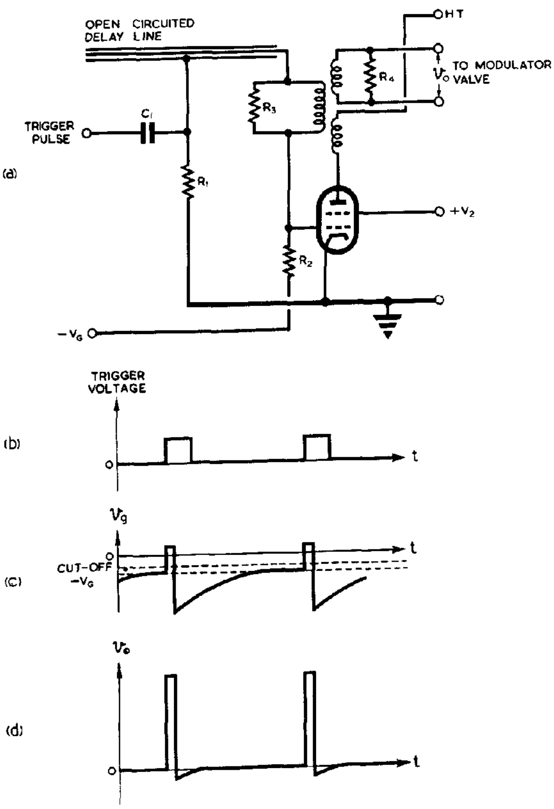


Fig. 621 - Single-valve pre-modulator.

resistors R_3 and R_4 across the windings prevent any appreciable ringing effects. While the valve current is cut off the delay network discharges relatively slowly through the circuit comprising the grid winding (in parallel with R_3), R_2, the grid bias source and R_1, in the interval before the next pulse,

The above description of the action of this circuit is an idealised one, complications occuring in practice due to the transformer characteristics.

CHARGING DELAY NETWORK FROM STEADY VOLTAGE SOURCE

23. Choke Charging

If a delay network formed by n L-C sections, is charged from a steady supply source through a resistor, the maximum network voltage is equal to that of the supply. If a choke L (Fig. 622) is substituted for the resistor, an amplitude double that of the supply may be obtained. The effective charging circuit then consists of a series resonant L-C combination. As shown in Sec. 11, the effective charging capacitance is nC. If the line is initially uncharged, application of the HT shocks the series L-C circuit into an oscillation of period $2\pi\sqrt{nCL}$. Assuming that the series resistance of the choke is negligible, the waveforms of the voltage v_C across the line, the voltage v_L across the choke, and the current i flowing in the circuit are as shown in Fig. 622. The waveforms illustrated in this figure, and in similar succeeding figures, **are** very much idealised, since in practice considerable ringing usually occurs from the oscillatory circuit formed by the choke (of the order of 100H) and its self-capacitance. This ringing is usually present during each period immediately after the line is discharged. The action is as follows:-

At time t = 0 the condensers are considered to be uncharged, i.e. $v_C = 0$. Therefore, since $v_C + v_L = -V_B$,

initially $\quad v_L = -V_B$;

but $\quad v_L = L\frac{di}{dt}$;

$\therefore$ at $t = o$, $\frac{di}{dt} = \frac{v_L}{L} = \frac{-V_B}{L}$.

v_L has its maximum negative value at t = 0; hence $\frac{di}{dt}$ also has its maximum negative value at this instant. The value of i, however, is zero because L prevents an instantaneous rise of current. Hence, at $t = o$, $\frac{dv_C}{dt} = 0$.

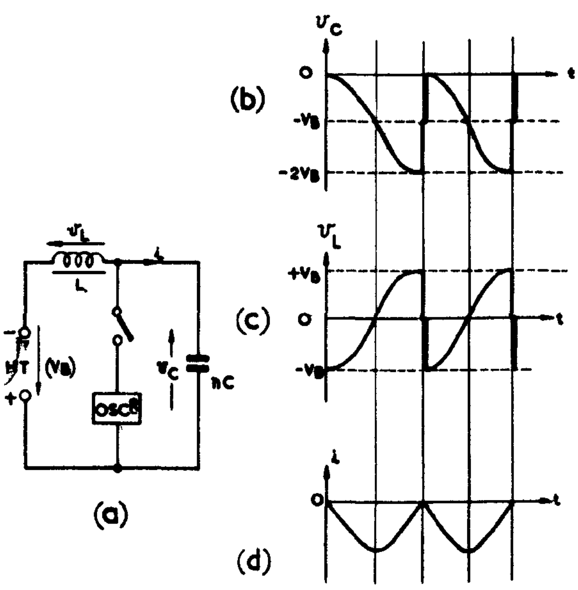


Fig. 622 - Choke charging of delay network.

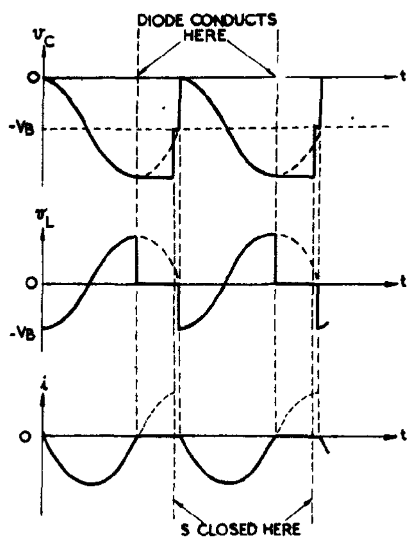


Fig. 623 - Choke charging with hold-off diode.

The circuit oscillates as shown by the waveforms of Fig. 622, until a half-period, $\pi\sqrt{nCL}$ has elapsed.

If at this instant the switch S is closed, the delay network is discharged through the matched load and the line voltage is reduced to zero after the double transit time of the line. Hence if the pulse recurrence period is made equal to $\pi\sqrt{nCL}$, output pulses of amplitude equal to the supply voltage V_B may be obtained.

When this condition is satisfied, we have what is known as Resonant Choke Charging or DC Resonant Charging.

If the recurrence period is greater than $\pi\sqrt{nCL}$ then the line voltage begins to decrease before S is closed. The oscillations during the third quarter cycle are shown dotted in Fig. 623. The reduction in line voltage may be prevented by the insertion between L and the line of a Hold-Off Diode (Fig. 624) which does not allow the current i to reverse, so that v_C is maintained at $-2V_B$ (full lines in Fig. 623).

When the recurrence period is jittery, as for example when a rotary spark gap is employed, its value should be made greater than $\pi\sqrt{nCL}$, and a hold-off diode used. Jitter does not then cause the line voltage to vary.

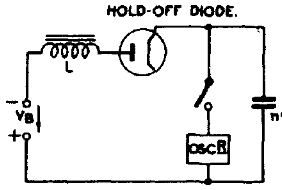


Fig. 624 - Use of hold-off diode.

24. Constant Current Charging

If the recurrence period is less than $\pi\sqrt{nCL}$ then the circuit takes several periods to reach the steady state.

Interval 1

The condensers do not charge fully to $-2V_B$ during the first period (Fig. 625). The current i is still negative when S is closed, not yet having fallen to zero.

Interval 2.

At the start, i has approximately the same value as at the end of interval 1. Therefore, the charging rate $\frac{dv_C}{dt}$ at the start of interval 2 is the same as at the end of the first interval, whereas at the start of that interval it was zero. $\frac{di}{dt}$ is, however, the same at the beginning of intervals 1 and 2, because in each case $v_C = 0$ and $v_L = L\frac{di}{dt} = -V_B$. Consequently, not only does the charging start at a higher rate than in interval 1, but, at least initially, the charging rate increases with time. The result of this is that v_C attains a much larger (negative) value by the end of interval 2. This (negative) value may be

(i) equal to,

(ii) less than

or (iii) greater than $-2V_B$,

depending on the relative values of the recurrence period and of

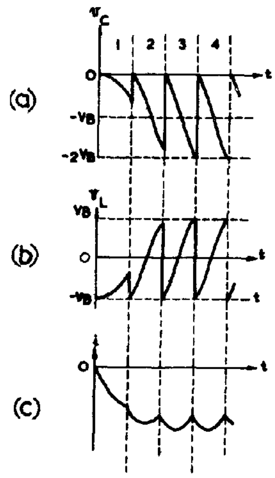


Fig. 625 - Constant current charging of delay network (discharge assumed instantaneous).

$\pi\sqrt{nCL}$. Fig. 625 shows the waveforms for case (ii).

Interval 3

(i) If v_C reaches $-2V_B$ by the end of interval 2, then the current flowing is the same at the start and finish ot interval 2, the conditions at the start of intervals 2 and 3 are the same, and the action is repeated.

(ii) If v_C does not reach $-2V_B$ in interval 2, the current flowing at the end of interval 2 is greater than at the start of that interval. Therefore, the current and charging rate at the start of interval 3 are greater than at the start of the second interval and the (negative) value of v_C reached by the end of the interval will be greater.

(iii) If the (negative) value of v_C is greater than $-2V_B$, the current flowing at the end of interval 2 is less than at the start of that interval. Consequently the (negative) value of v_C reached by the end of interval 3 will be less than at the end of the second interval.

Succeeding Intervals

The process is repeated, a steady state being reached when v_C attains the value of $-2V_B$ at the end of each interval. When this occurs the values of i at the beginning and end of each interval become equal, and the current variation is a relatively small one about a constant level. It is from this fact that the term Constant Current Charging is derived.

The actual value of the pulse recurrence period does not affect the steady state condition that $v_C = -2V_B$ at the end of each interval provided that its value is steady. Constant current charging is not suitable for use with a jittery recurrence period.

25. Symmetrical Charging

It has been shown that by charging a delay network through a choke a pulse of amplitude approximately equal to V_B may be obtained. However, with the method described above, it is necessary for the delay network to be capable of withstanding $2V_B$, since it becomes charged to this voltage. This can be avoided by arranging that the voltage across the delay network swings equal amounts above and below zero. Fig.626.(a) shows a circuit whereby this action may be produced. In this circuit the network charges rapidly during the intervals when the switch S is closed and discharges slowly through L while S is open.

Suppose that the condensers of the delay network are initially

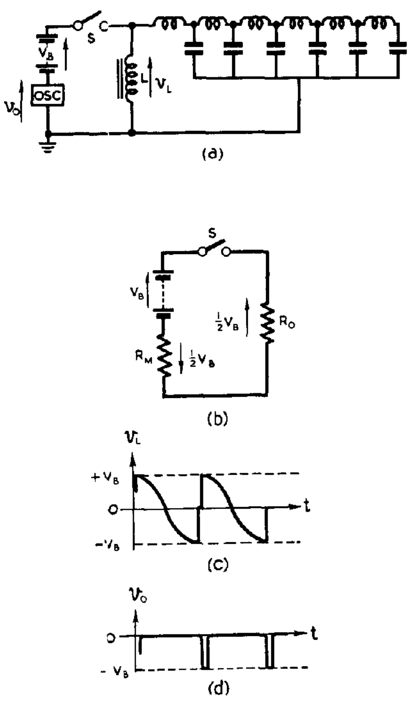


Fig. 626 - Symmetrical charging of delay network.

uncharged. When S is closed V_B is applied to the network, of characteristic impedance R_o, and the oscillator, of resistance R_M, in series (Fig. 626(b)). If $R_o = R_M$, a voltage $\frac{1}{2}V_B$ appears across R_o, and an equal negative voltage across R_M. As the network charges a rectangular wave of amplitude $\frac{1}{2}V_B$. travels along the line.

The wave is reflected without change of phase at the open end and returns to the input end, the voltage across the line increasing to V_B. If $R_o = R_M$ the wave is completely absorbed at the input end, the charging current falls to zero, and S opens.

The line now discharges through L (Fig. 627(a)). Considering first that the recurrence period is equal to $\pi\sqrt{nCL}$, the voltage across the network swings from $+V_B$ to $-V_B$ during the discharge as shown in Fig. 626(c). The second time that S is closed, the effective voltage acting in the circuit is equal to $+2V_B$ (Fig. 627(b)), and if $R_o = R_M$ the voltage appearing across the oscillator is equal to $+ V_B$. As the delay network charges a rectangular wave of amplitude $+ V_B$ travels along the network, the voltage across the line changing from $-V_B$ to zero on the forward journey and from zero to $+ V_B$ on the return. The action is then repeated.

If the recurrence period is less than $\pi\sqrt{nCL}$ a constant current form of discharge builds up in the way described in Sec. 24, and the voltage across the line again swings from $+ V_B$ to $-V_B$ as for the resonant case.

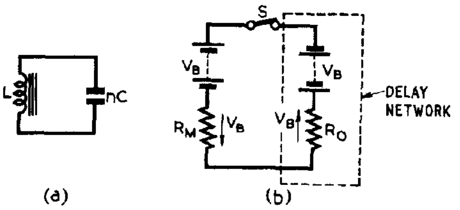


Fig. 627 - Equivalent charging (a) and discharging (b) circuits.

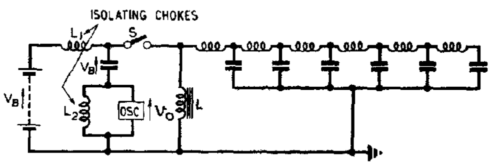


Fig. 628 - Symmetrical charging of delay network: practical circuit.

This circuit therefore generates pulses of amplitude equal to V_B without the voltage across the delay network exceeding this value, but in its present form neither side of the HT supply can be earthed. This disadvantage may be overcome by re-arranging the circuit as shown in Fig. 628. The oscillator, shunted by a coil L_2, is inserted in series with the smoothing components L_1, C_1 of the HT supply. The circuit comprising the HT supply, L_1 and C_1, is completed by L_2 during the intervals when the oscillator is not conducting. The Isolating Inductances, L_1 and L_2, are both small compared with L, but are of sufficient inductance to act during the charging process as virtual open-circuits. The action of the circuit is similar in every way to that described immediately above, the condenser C_1 taking the place of the HT supply, the time-constant C_1 $(R_M + R_o)$ being very much greater than the pulse duration.

26. TYPICAL SPARK GAP MODULATOR

The circuit of a typical spark gap modulator is shown in Fig. 629.

Valves 1 and 2 form a multivibrator, producing a 20-microsecond pulse output, which is fed to the grid of the trigger valve, valve 3. This valve is normally biased beyond cut-off, but is brought into heavy conduction by the positive-going 20 microsecond pulse. On the trailing edge of this pulse, valve 3 is cut-off, and

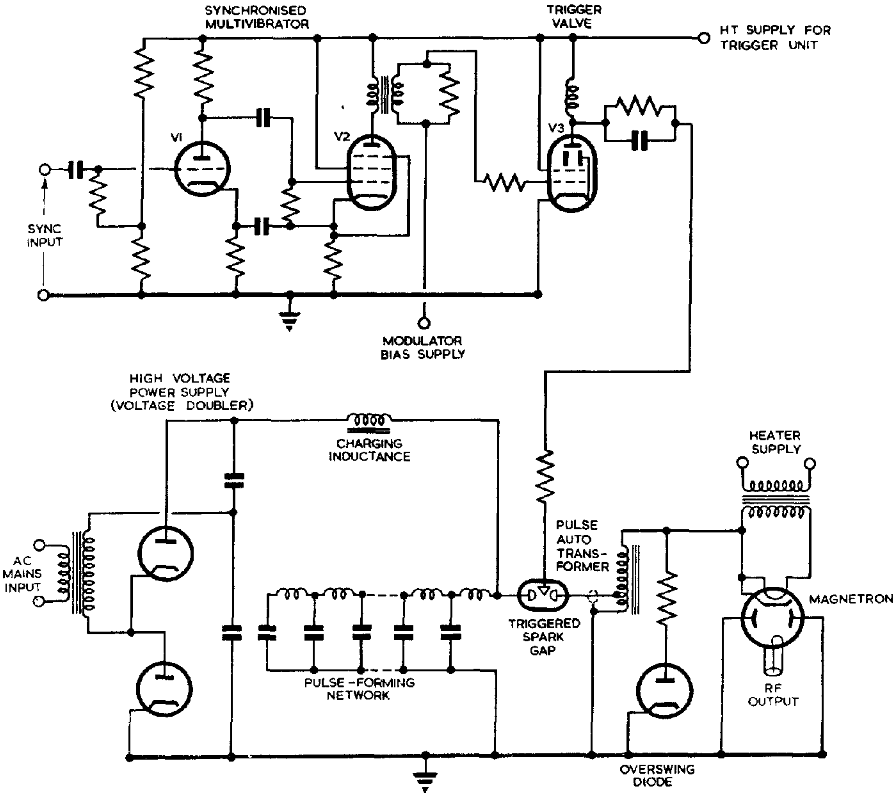


Fig. 629 - Typical spark-gap modulator.

its anode circuit rings, the first positive-going swing being of large amplitude. This is used to trigger the spark gap, which discharges the delay network. The 1-microsecond pulse so formed is fed to the magnetron via a matched connecting cable, and a step-up pulse transformer. The latter matches the cable to the load presented by the magnetron when conducting, and converts the pulse from 4kV, 40A to 16kV, 10A. The delay network is charged through a choke from a voltage doubler power pack.

CHARGING DELAY NETWORK FROM AN ALTERNATING VOLTAGE SOURCE

27. Advantages of Alternator Charging

Although it is simpler to charge a line modulator from a source of steady voltage, there are definite advantages to be gained from the use of an unrectified alternating supply. The principal advantages are:-

(i) No high voltage recitifer is required.

(ii) No high voltage smoothing circuit is needed.

(iii) The high voltage transformer need not be so large.

(iv) A hold-off diode is unnecessary (see Sec. 23).

(v) The maximum HT voltage is developed across the line for only a small fraction of the recurrence period. This helps to ensure that the transmitter does not oscillate except when required.

These advantages considerably reduce the bulk and weight of the modulator unit. However, this method of charging is usable only if the alternator frequency is constant within narrow limits. This frequency is also the recurrence frequency of the output pulses.

28. The Charging Circuit

Fig. 630(a) shows a simplified diagram of a circuit for charging a line modulator from an alternator. The line discharging circuit (b) is connected between A and B. While the discharge valve is not conducting AB is open-circuited, and, to the alternating voltage of the frequency normally used, of the order 500-1000 c/s, the impedance of the line to the right of AB is a small capacitance C. The input impedance of the line transformer T_1 may then be shown to be approximately $\frac{-j}{m^2\omega C}$ where m is the step-up ratio of the transformer and $\omega = 2\pi f_a$, f_a being the alternator frequency. This result neglects the magnetising inductance of the transformer. A more accurate result gives the input admittance of T_1 as the sum of the susceptances of the magnetising inductance and of the input capacitance m^2C.

* * * Input Impedance of Transformer when Secondary Load is a small Capacitance

The input impedance z_t of the transformer T_1 terminated in C is given by

$$z_t = j\omega L_1 + \frac{\omega^2 M^2}{j\omega L_2 + \frac{1}{j\omega C}}, \quad \text{(see Chap. 1 Sec. 20)}$$

where L_1, L_2 are primary and secondary inductances respectively and M is the mutual inductance between primary and secondary windings.

$$\text{Ideally, } \omega L_2 \gg \frac{1}{\omega C},\ M^2 \doteqdot L_1 L_2,\ \text{and } \frac{L_2}{L_1} \doteqdot m^2 \text{ where m is}$$

the transformer step-up ratio.

The equation for z_t simplifies as follows :-

$$z_t \doteqdot j\omega L_1 - \frac{j\omega^2 L_1 L_2}{\omega L_2 - \frac{1}{\omega C}}$$

$$\doteqdot j\omega L_1 - \frac{j\omega^2 L_1 L_2}{\omega L_2 \left(1 - \frac{1}{\omega^2 C L_2}\right)}$$

$$\doteqdot j\omega L_1 - j\omega L_1 \left(1 + \frac{1}{\omega^2 C L_2}\right)$$

$$= -\frac{j\omega L_1}{\omega^2 C L_2}$$

$$\doteqdot -\frac{1}{\omega C} \cdot \frac{1}{m^2}.$$

Hence the transformer, when terminated in C, "looks like" a capacitance m^2C.

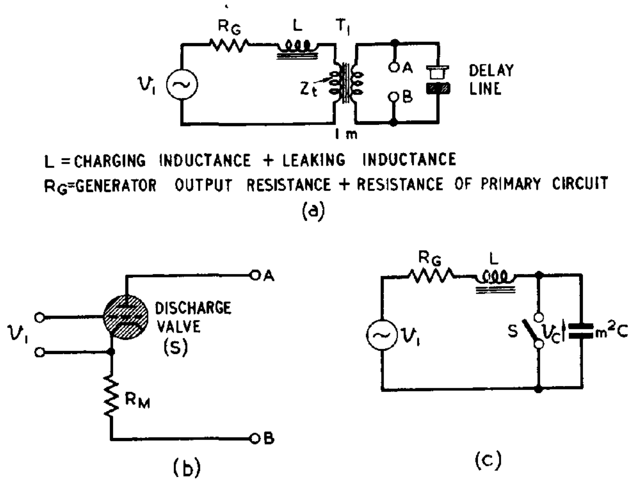


Fig. 630 - Alternator charging: basic circuits.

29. The Equivalent Charging Circuit

Neglecting the magnetising inductance of the transformer we may represent the charging circuit by the equivalent arrangement shown in Fig. 630(c).

When the switch S is closed the line discharges through the oscillator load R_M. This discharge action is described in Chap. 4, Sec. 12. For a resistive load the line may be assumed to be completely discharged in a few microseconds, even if it is not properly matched to the load. So far as the low frequency charging circuit is concerned, this discharge may be considered as instantaneous, and may be represented by the momentary short-circuiting of the terminals AB (Fig. 630(a)).

In the case of a magnetron load, and in some other instances, an overswing generally occurs. The reasons for this overswing, and its effect, are discussed later; for the present it is assumed that the line is discharged completely during the momentary closure of the switch S (Fig. 630(c)) representing the firing of the discharge valve.

30. Transients Due to Line Discharge

The closure of the switch S (Fig. 630(c)) reduces the impedance across the effective line capacitance m^2C to zero (the justification for this may be found by putting zero in place of $\frac{1}{\omega C}$ in the analysis given in Sec. 28). The corresponding input voltage v_C also becomes zero. Since the short-circuiting is only of momentary duration no appreciable change in current occurs in the primary circuit during the discharge of the line, the time constant $\frac{L}{R_G}$ being much greater than the duration of the discharge.

An interval follows during which the circuit passes through a transient state and, in the absence of further irregularities, tends gradually towards the steady state, or normal AC working condition.

In the actual circuit this steady state is never reached, but for ease of explanation it will be assumed in what follows, in the first instance, that the switch is closed once only, when the line voltage is at any portion of its steady state condition (Fig. 631). Later it will be described how the line voltage is maintained in a condition little resembling a steady sinusoidal form.

31. Graphical Derivation of Transient State

The manner in which the line voltage returns to its steady state after a single momentary discharge is illustrated in Fig. 631(a). This picture is derived as follows.

The steady state voltage across m^2C (Fig. 630(c)) is indicated by the full curve from 0 to E and from then on by the dotted curve. It is to this steady state that the transient conditions ultimately return. The actual voltage v_C is shown by the full line.

When the line is momentarily discharged (E) v_C falls to zero (F). The line current given by $i = m^2C \dfrac{dv_C}{dt}$

is unchanged, so that $\dfrac{dv_C}{dt}$ has the same value immediately after the discharge as it had immediately before. The gradients of the pre-discharge and post-discharge curves at E and F respectively are therefore the same.

The resultant line voltage after the discharge may be obtained by adding to the steady state solution (dotted curve Fig. 631(a)) the appropriate transient (Fig. 631(b)). (This corresponds to the addition of the complementary function (transient solution) to the particular integral (steady state solution) in the solution of the linear differential equation of the system).

To obtain this appropriate transient, first draw the characteristic curve of the response of the series ringing circuit, formed by R_G, L and m^2C, to a sudden change in input voltage. This is illustrated in Fig. 632(a). This curve is one of an infinite sequence of curves, some of which are shown in Fig. 632(b). From such a set of curves choose the one which satisfies the conditions, i.e. having the right magnitude and slope.

In the case illustrated in Fig. 631 the initial magnitude is equal to EF, its value is negative, and the initial slope is zero. It is this curve, starting from the point H, where it meets PQ (Figs. 632(b) and 631(b)) which is the correct transient for the case considered. When this is added to the steady state solution (Fig. 631(a)) the resultant amplitude (zero) and gradient (shown at E and F) are in accordance with the initial conditions.

The resultant curve is thus obtained (full curve, Fig. 631(a)).

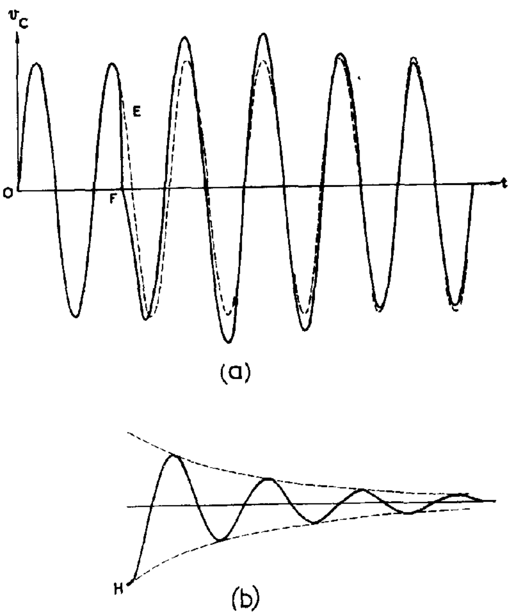


Fig. 631 - Transient effect of short-circuiting the transformer T_1.

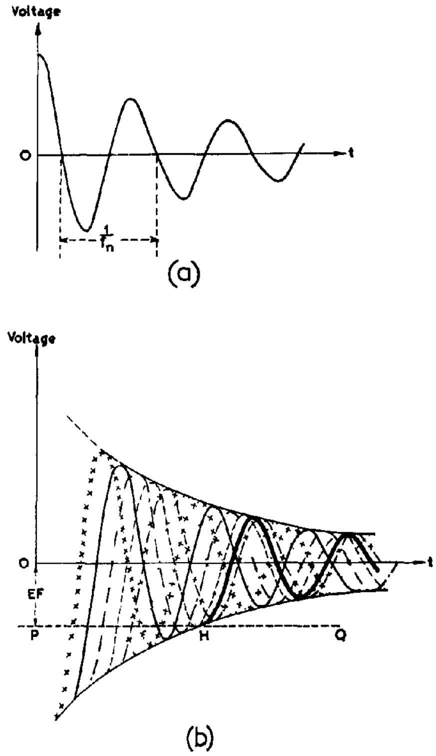


Fig. 632 - Determination of the correct transient.

32. Normal Transient Conditions

In practice the circuit is not allowed to return to the steady state. The line discharge valve is triggered each time the line voltage reaches a maximum and the voltage v_c varies as shown in Fig. 633. Since the voltage is a maximum the current $i = m^2C \frac{dv_c}{dt}$ is zero, both immediately before and immediately after the discharge. The transient curve which must be added to the steady state curve to give v_c must therefore satisfy the following conditions;-

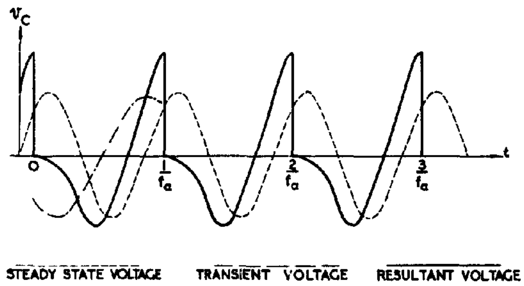


Fig. 633 - Line charging curve.

(i) It must have the same magnitude as the steady state voltage at the moment of striking (t = o) and be of opposite polarity.

(ii) It must have a gradient equal and opposite to that of the steady state curve at t = o.

(iii) When it is added to the steady state voltage the resultant must have a maximum value at $t = \frac{1}{f_a}$, where f_a is the alternator frequency.

After $t = \frac{1}{f_a}$ the line discharges again and the cycle repeats (Fig. 633).

The form of the line voltage depends chiefly on the ratio f_n/f_a, where f_n is the natural frequency of the series resonant circuit formed by R_G, L and m^2C. The optimum value for this ratio is dealt with in Sec. 34. The value $\frac{f_n}{f_a} = 0{\cdot}7$ is usually chosen, and this is the value used for the charging curve of Fig. 633.

33. Effect of Overswing

If for any reason an overswing occurs (see Sec. 35) the "initial" voltage v_C immediately following each discharge of the line is not zero, but is $-k\hat{v}_C$, where $\hat{v}_C$ is the value of v_C immediately before the discharge and k is the overswing ratio. This causes the form of v_C to be changed from that of Fig. 633, as indicated in Fig. 634.

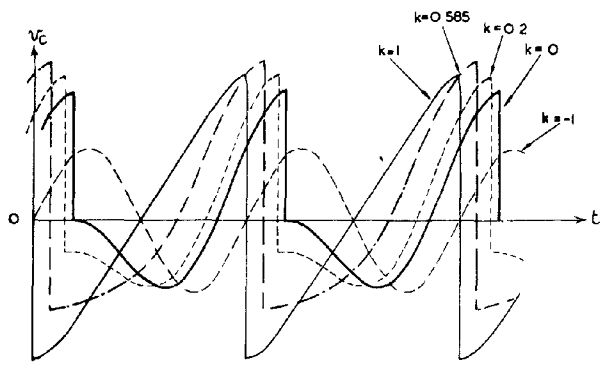


Fig. 634 - Effect of overswing, showing good regulation ($f_n/f_a = 0{\cdot}7$).

The instant at which the line discharges, relative to the alternator voltage (or to the steady state voltage at the line transformer) changes so that the new conditions are maintained for each discharge of the line. The phase angle at which the line discharges relative to the alternator, is called the Angle of Fire. A change in the overswing ratio k thus causes a change in the angle of fire, as shown in Fig. 634. A small overswing is advantageous in the case of most transmitters, as this holds the transmitter valve supply voltage at the opposite polarity to that required for oscillation, and facilitates a rapid dying away of the output pulse. It is possible, however, for overswing to cause the line voltage to acquire dangerously large values.

34. Choice of the Ratio f_n/f_a

The effect of the ratio f_n/f_a upon the charging curves is complicated and may be divided into three parts:-

(i) It determines the angle of fire and the shape of the charging curve, i.e. the form of the line voltage.

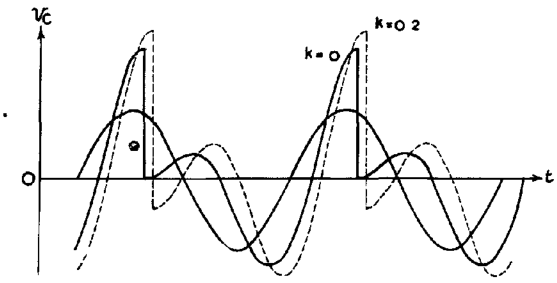
(ii) It determines the ratio of the line voltage to the alternator EMF, since this depends on the frequency of operation f_a in comparison with the resonant frequency $\frac{1}{2\pi\sqrt{LC}}$, and this is very nearly the same as f_n. In general, operation near resonance means that only a small alternator EMF is required.

(iii) It affects the reaction of the circuit to overswings. These may occur through the normal operation, in which case they are expected and can be allowed for, or they may arise through abnormal loading conditions, such as a **flash-over**, or an open-circuit due to faulty connection. The effect on the line voltage of an overswing determines the Regulation of the circuit. If the line voltage increases considerably when an overswing occurs the regulation is said to be poor.

The various values for f_n/f_a between 0 and 2 are divided into five sections as under:-

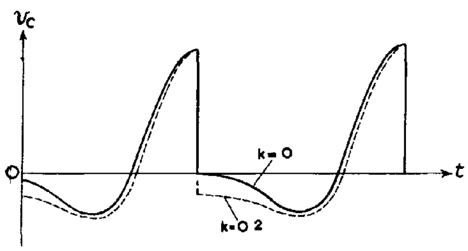
f_n/f_a	Remarks
(a) 0 - 0·5	Requires large alternator EMF
(b) 0·5 - 0·8	Optimum conditions
(c) 0·8 - 1·3	Poor regulation
(d) 1·3 - 1·6	Good regulation but charging curve less favourable than that of (b)(see Fig.635(a)).
(e) 1·6 - 2·0	Ill-shaped charging curve, poor regulation, large alternator EMF.

Fig. 635 illustrates some of these points.



(a) $f_n/f_a = 1\cdot4$ k=O (RESISTIVE LOAD) k=O 2 (MAGNETRON LOAD)

Fig. 635 - Line charging curves.



(b) $f_n/f_a = 0\cdot44$ k=O (RESISTIVE LOAD) k=O 2 (MAGNETRON LOAD)

35. Causes of Overswing

When the output from a line-modulator is used to pulse a magnetron it is unlikely that the whole of the energy of the modulating pulse is absorbed by the transmitter. This is particularly so if the pulse shape departs markedly from the ideal rectangular form, causing the magnetron impedance at the beginning and end of the pulse to change so that the line is mismatched. This may cause the line to reverse its polarity (see Chap. 4, Sec. 12), subsequent reversals being prevented by the cessation of ionisation in the discharge valve(assumed soft).

If the load becomes short-circuited, the tendency to overswing depends on whether spark gaps (unpolarised discharger valves) or gas-filled triodes (polarised discharger valves) are used. In the former case the line energy would probably be rapidly absorbed by the spark gap and other circuit elements during the oscillatory discharge, and dissipated as heat. In the latter case the valve would probably open-circuit following a complete overswing ($k = 1$) so that the charging curve would be considerably modified.

If the line is open-circuited so that the discharger valve does not operate the transient state never develops ($k = 1$). If, however, a pulse transformer is used between the line and the transmitter an open-circuit at the transmitter means that the cable discharges through the magnetising inductance of the pulse transformer, with considerable dissipation of energy in the transformer core.

36. Method of Triggering Discharge Valve

Fig. 636 shows a method of triggering the discharge valve at each instant when the line voltage is a maximum. The current i through the primary winding of the transformer T_1, represented by m^2C in Fig. 630(b), is given by the relation

$$i = m^2C \frac{dv_C}{dt} .$$

Hence when v_c is a maximum i is zero. This current is caused to flow through the primary winding of a peaking transformer T_2 (Fig. 636(a)). This is a transformer with a small iron core which saturates for small values of primary current. The variation of flux corresponding to the primary current (Fig. 636 (b)) is shown at (c) and the secondary EMF is indicated at (d). (EMF $\propto$ rate of change of primary flux). The output voltage v_p developed across the secondary winding of this transformer is thus of a form suitable for triggering the discharge valve at the instant the line voltage reaches a maximum.

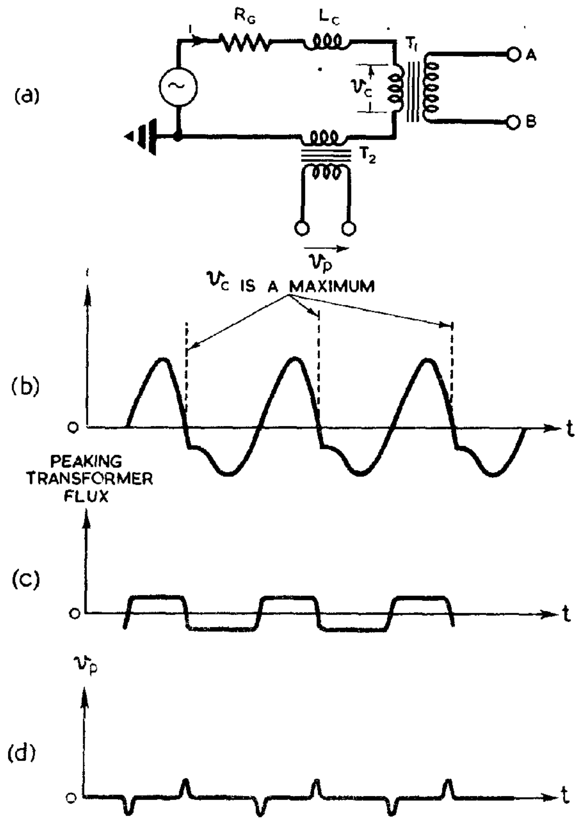


Fig. 636 - Derivation of triggering pulses for discharge valve.

VOLTAGE MULTIPLYING CIRCUITS

37. General

It is often necessary to avoid the use of high voltage supply circuits by employing voltage multiplying devices. One such device, the pulse transformer, has already been described. Other methods, involving the use of more than one pulse-forming network, are described below. They operate on the principle of being charged in parallel from the same supply and, by means of synchronous switches, discharged in series with the load. The Marx circuit is described in schematic form only. The Blumlein circuit, which is a derivative of the Marx, is dealt with more fully because its application in radar transmitting circuits is already established.

38. The Marx Circuit

The Marx Circuit is illustrated in Fig. 637.

C_1, C_2 .... C_n represent the charging capacitances of n separate networks. When the switches S_1, S_2, S_3 ..... S_n are open, C_1, C_2, C_3 ..... C_n are charged in parallel from the supply (assumed, for simplicity, to be a steady voltage source) through the charging choke L and the isolating chokes L_1, L_2, L_3 etc. The inductances L_1, L_2, L_3 etc. are all small compared with L and to a first approximation their effect on the charging process may be neglected.

If all the switches S, S_1, S_2 etc. are closed simultaneously, with the networks thus charged, the latter are instantaneously connected in series and commence to discharge through the load, R_M. Provided each inductance is $\gg C_T R_M^2$, where C_T is the total effective capacitance of C_1, C_2,.... etc. in series, it has little effect on the discharge, which takes the path of short time-constant $C_T R_M$ in preference to that of the long time-constant $\frac{L_T}{R_M}$, (L_T is the total effective inductance of L_1, L_2 etc. when the switches are closed. Thus the lines are charged in parallel and discharge in series with the load. The discharge is similar to that of a single line of characteristic impedance equal to the sum of the characteristic impedance of the individual lines i.e. n R_o, and charged to a voltage n x (the voltage on a single line). The inductances L_1, L_2 etc. are usually made equal and intermediate in value between the total series inductance of each line and the inductance L of the choke.

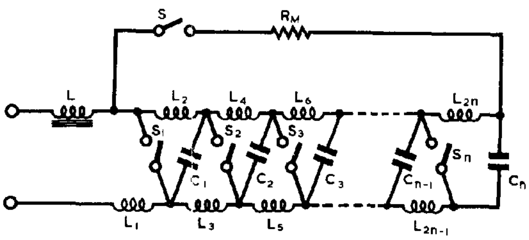


Fig. 637 - Marx circuit.

The switches S, S_1, etc. may take the form of triggered spark gaps or gas-filled triodes. The principal disadvantage of this method arises from the difficulty of synchronising the switching.

39. Blumlein Modulator

The derivative of the Marx circuit most commonly employed in Service radar is the Blumlein modulator which employs two delay lines.

To avoid the difficulty of triggering simultaneously the two switches necessitated by the Marx circuit, this variant was designed to operate with a single spark gap.

The basic Blumlein circuit is illustrated in Fig. 638. The lines are charged in parallel with the switch S open. During the discharge the generator is isolated from the network by the choke L and will be ignored in the subsequent analysis.

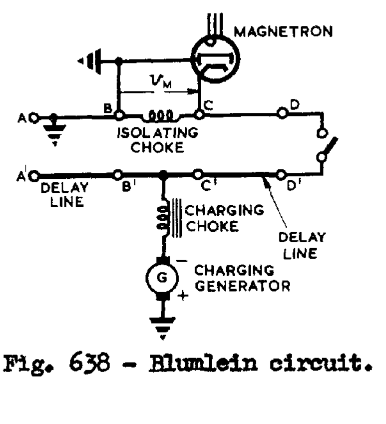


Fig. 638 - Blumlein circuit.

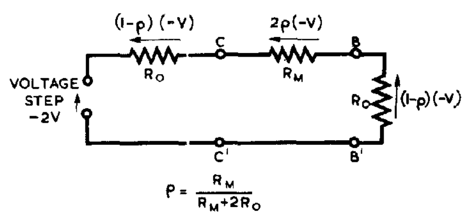


Fig. 639 - Equivalent circuit showing voltage developed across the load by a wave (-V) arriving at the termination CC'.

Suppose that the two lines are of identical construction, lossless and with characteristic impedance R_o, and transit time T. The magnetron load M has resistance R_M to a pulse of the right voltage and polarity to ensure that it oscillates correctly. Let the lines be charged to a voltage V when the switch S is closed at t = 0. This closure short-circuits the line CD so that a wave of magnitude -V travels towards C. This divides at C into three parts:-

(i) a fraction ρ is reflected towards D.

(ii) a fraction 2ρ is developed across the load.

(iii) a fraction $1-\rho$ is developed across the input to the line BA.

Thus immediately after the arrival of the wave at C, the voltages developed across the respective terminals due to the wave are as shown in Fig. 639. This result may be derived by the method indicated in Chap. 4, Sec. 9. The value of ρ is given by

$$\rho = \frac{R_M + R_o - R_o}{R_M + R_o + R_o} = \frac{R_M}{R_M + 2R_o} .$$

The sequence of waves and load voltages is illustrated in Fig. 640. It is assumed that the switch remains short-circuited throughout the action.

For a wave travelling from A to B, the division at BB' is as follows:-

(iv) a fraction ρ is reflected towards A.

(v) a fraction -2ρ is delivered to the load.

(vi) a fraction $1-\rho$ is developed across CC'.

[The reversal of sign of the voltage delivered to the load, compared with the former case of the wave travelling from D to C, is due to the fact that it is necessary to have a uniform sign convention for the voltage across the load. If this is positive when approached from D it is negative when approached from A.] Thus a pulse of magnitude $2\rho V$ is developed across the magnetron at the instant t = T. This pulse must be maintained at this magnitude until t = 3T, by which time the waves travelling along the lines CD and BA will have undergone reflections and the

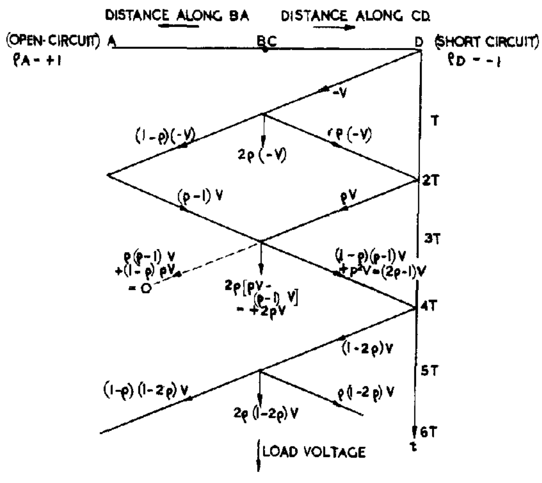


Fig. 640 - Action of Blumlein modulator.

wave fronts will have returned to the load.

The figure shows the passage of various wave fronts along the lines, suffering simultaneous reflections at the ends A and D, and returning to B and C, where fresh pulses are developed across the load. The load voltages are indicated by the vertical lines. The amplitude and polarity of the various waves are indicated by the symbols written on the appropriate lines.

Reflections at the short-circuit D involve a sign-reversal; at the open-circuit A there is no change of polarity of the reflected wave.

Further, the values shown in the diagram are incremental values. Thus at $t = T$ a pulse $-2\rho V$ is developed across the transmitter. The increment $+2\rho V$ at $t = 3T$ raises the load voltage to $-2\rho V + 2\rho V = 0$ at which value it remains for a further interval $2T$, and so on. The load voltage is illustrated in Fig. 641. This ideal form is not likely to be achieved in practice because of changes of the magnetron impedance due to the non-rectangularity of the applied pulse, unwanted reflections, etc.

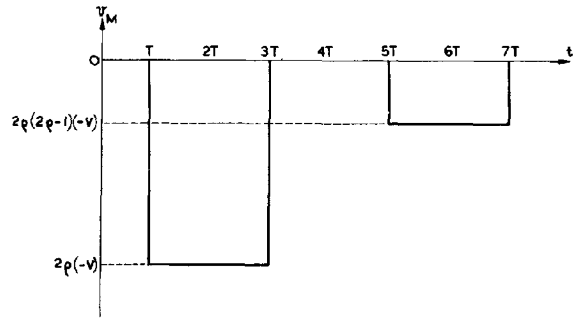


Fig. 641 - Load voltage: (the second pulse is positive if $\rho < 0{\cdot}5$; i.e. if $R_M < 2R_o$.

In the Blumlein circuit the magnetron is matched to the network when $R_M = 2R_o$. In this case, provided the pulse were perfectly rectangular and the magnetron impedance when oscillating were constant and equal to R_M the lines would be completely discharged after time $t = 3T$ from the instant S is closed.

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