1949

Line-type radar modulators

Brown, Garrison

Monterey, California. U.S. Naval Postgraduate School

LINE-TYPE RADAR MODULATORS

Garrison Brown

LINE-TYPE RADAR MODULATORS

by

Garrison Brown
Lieutenant Commander, United States Navy

Submitted in partial fulfillment
of the requirements
for the degree of
MASTER OF SCIENCE
in
ENGINEERING ELECTRONICS

United States Naval Postgraduate School
Annapolis, Maryland
1949

This work is accepted as fulfilling
the thesis requirements for the degree of

MASTER OF SCIENCE
in
ENGINEERING ELECTRONICS

from the
United States Naval Postgraduate School

[redacted]
Chairman
Department of Electronics and Physics

[illegible]

Approved: [redacted]
Academic Dean

PREFACE

A portion of the material herein presented was obtained while on duty at the Glenn L. Martin Company, Baltimore, Maryland, during the period 3 January - 18 March, 1949. Work was conducted on several line-type radar modulators under the guidance, and with the assistance of, the engineers in the Production Design Group of the Electronics Section, Special Weapons Division. The author wishes particularly to acknowledge the helpful advice and assistance rendered him by Mr. L. J. Hruska, the Group Engineer, and Messrs. D. A. Bourne, J. Markwalter, and B. Hayes, Assistant Engineers.

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

TABLE OF SYMBOLS AND ABBREVIATIONS

α---- exponential damping factor

C_N--- storage capacity of PFN

δ---- pulse width in microseconds

ΔI_m-- incremental change of magnetron current

ΔV_m-- incremental change of applied magnetron voltage

E_{bb}-- d-c power supply voltage

E_m--- peak value of a-c power supply voltage

η_t--- pulse transformer efficiency

i_{L_s}-- current in saturable-core reactor

I_L--- d-c current in load

I_m--- magnetron current

I_p--- transformer primary current

I_s--- transformer secondary current

KV--- kilovolts

KW--- kilowatts

ℓ/d-- ratio of length to diameter (of a coil)

L---- inductance

L_c--- charging reactor

L_r--- value of charging reactor for resonance charging

L_s--- saturable-core reactor

ma--- milliamperes

n---- transformer turns ratio

pp.s.- pulses per second

PFN-- pulse forming network

PRF-- pulse recurrence frequency

R---- resistance

R_c--- charging circuit resistance

R_C--- characteristic resistance

R_L--- load resistance

s---- Laplace-transform operator

t---- time

T_o--- time at which charging current is maximum

T_r--- pulse recurrence period

v_N--- voltage on the PFN

V_L--- load voltage

V_m--- applied magnetron voltage

V_p--- transformer primary voltage

V_s--- transformer secondary voltage

$Y(s)$- Laplace-transform admittance function

Z---- impedance

Z_o--- characteristic impedance

Z_p--- transformer impedance viewed from primary

Z_s--- transformer impedance viewed from secondary

$Z(s)$- Laplace-transform impedance function

CHAPTER I

INTRODUCTION

1. The Purpose of Radar Modulators

Present day radars are very nearly all of the pulsed type, and with few exceptions employ cavity magnetrons to generate electromagnetic oscillations. In order to excite the magnetron at the proper frequency and at the proper power level, provision must be made for supplying high-voltage, high-power pulses to it. In addition to proper pulse voltage and power, the shape of the pulse is of primary importance for proper magnetron operation. The unit which furnishes these pulses of voltage to the oscillator is known as the modulator; it is also called the "pulser" or "keyer".

2. Fundamental Considerations

In order to discuss modulators, certain parameters affecting the design must be considered: pulse width, peak power, pulse recurrence frequency, duty ratio, and average power. Since the output waveform is not rectangular, it is necessary to define various terms connected with pulse shape. An idealized waveform is illustrated in Figure 1. For purposes of discussion, the following terms are defined:

<u>Pulse width</u> - elapsed time in microseconds between leading and trailing edges, measured at .707 of maximum pulse amplitude (unless otherwise stated).

<u>Rise time</u> - elapsed time in microseconds between

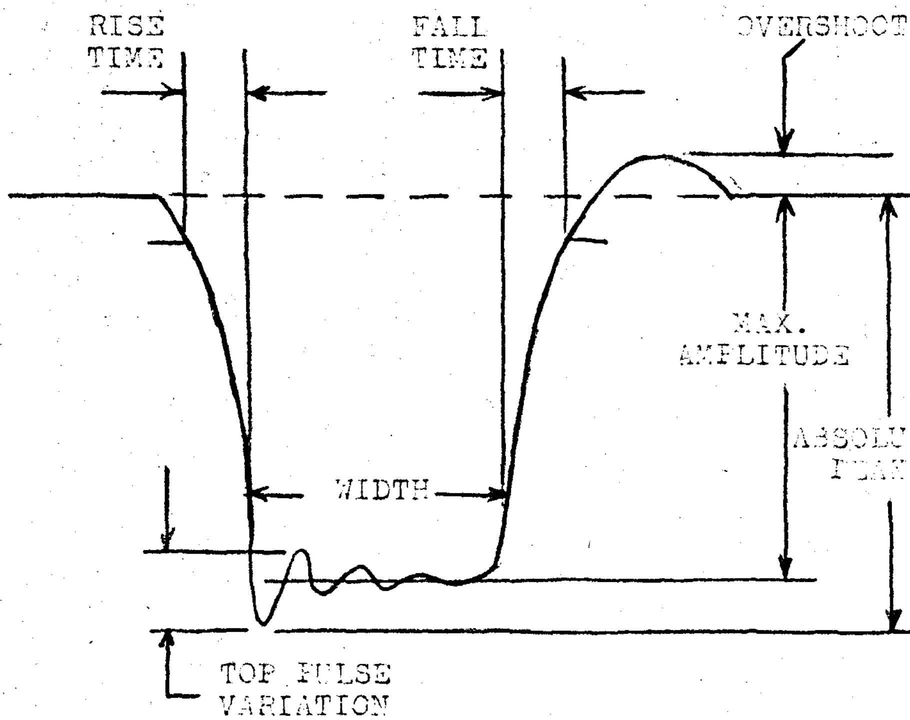

RISE TIME
FALL TIME
OVERSHOOT
MAX. AMPLITUDE
ABSOLUTE PEAK
WIDTH
TOP PULSE VARIATION

Fig. 1

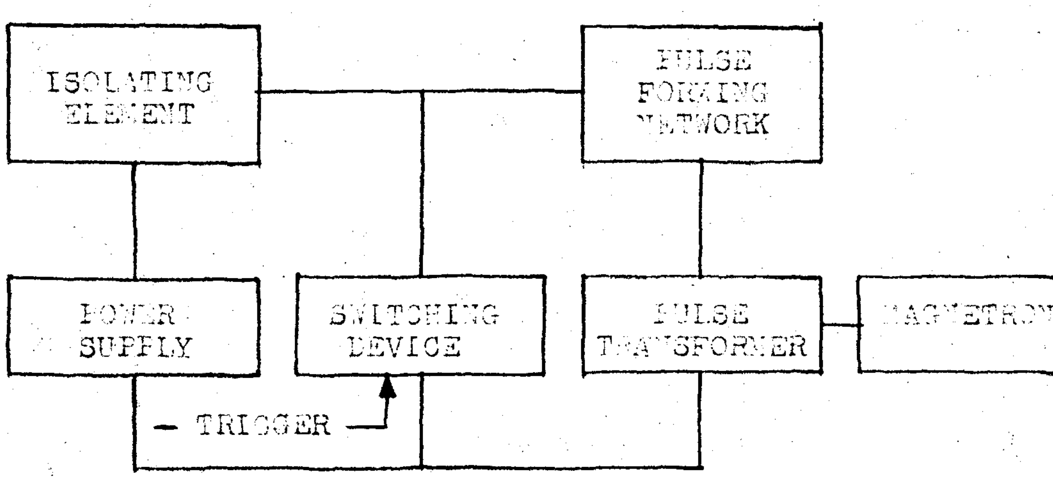

ISOLATING ELEMENT
PULSE FORMING NETWORK
POWER SUPPLY
SWITCHING DEVICE
PULSE TRANSFORMER
MAGNETRON
— TRIGGER

Fig.2

15% and 85% of maximum pulse amplitude, measured on leading edge.

Fall time - elapsed time in microseconds between 85% and 15% of maximum pulse amplitude, measured on trailing edge.

Maximum pulse amplitude - maximum value of the pulse in volts (or amperes), measured to the average of any oscillations that exist on the top of the pulse.

Peak power - the product of maximum pulse amplitude of voltage and maximum pulse amplitude of current.

Pulse recurrence frequency (PRF) - the number of pulses occurring in one second.

Duty ratio - the product of pulse width and PRF. Also called "duty cycle".

Average power - the product of peak power and duty ratio.

3. The Basic Circuit of the Line-type Modulator

Fundamentally, there are two types of pulsers:

(a) Those in which a small fraction of the stored energy is discharged into the load during the pulse,

(b) Those in which all of the stored energy is discharged during each pulse.

The former are called "hard-tube" modulators, since a high-vacuum tube is used for switching purposes. The latter type are known as "line-type", since the energy-storing device is basically an artificial transmission line. It is with the second type that this paper is concerned, and further discussion will be limited to that subject.

The basic block diagram of the line-type pulser is given in Figure 2. The operation of this circuit is as follows: An artificial transmission line (generally known as a pulse forming network, and hereinafter termed "PFN") is charged through an isolating element during the interval between pulses when the switching device is non-conductive. At the instant the pulse is desired, the switching device is made to conduct, thus effectively establishing a short circuit across the load. This causes the PFN to discharge and thereby form the pulse, which is coupled to the magnetron through an impedance matching transformer. Directly after the discharge occurs, the switch again becomes non-conductive, the PFN begins charging again, and the cycle is repeated.

As shown in Figure 2, the line-type modulator consists essentially of five sections exclusive of the transmitter, viz., power supply, isolating element, switching device, PFN, and pulse transformer. Each of these components, except the power supply, will be treated in a separate chapter of this paper. Although the power supply is a vital part of the modulator system, it is usually quite conventional, and merits little further attention. Unless otherwise specified, the power supply will be considered to be a voltage source of zero internal impedance. It is not the purpose of this treatise to discuss magnetrons per se, but as these devices present a non-linear load to the pulser and therefore influence pulser design to a great extent, it is necessary to consider them from that point of view. Consequently, it should be borne in mind that this is the type load into which the modulator will work, and that certain requirements, not otherwise found with linear resistive loads, are therefore imposed on the modulator system and its components.

CHAPTER II

CHARGING METHODS

1. General

Because charging and discharging in a line-type pulser are independent of one another, it is convenient to treat these operations separately.

Three important considerations in the design of the charging circuit are:

(a) Regularity

(b) Isolation

(c) High efficiency

By regulation it is meant that the PFN must be charged to the same energy level on each cycle. The charging element must isolate the power supply from the discharging circuit during the pulse in order to prevent short circuiting the source during the interval. Isolation is also necessary for a short period immediately following the pulse to permit the gaseous-discharge switch to deionize. A high efficiency is particularly desirable since the line-type modulator operates at high power levels.

A resistance may be used as the charging element, since it fulfills the first two requirements listed above; however, the maximum efficiency is but 50%, and hence this method is precluded in practice. More efficient methods, generally employing an inductance as the isolating element, are discussed in the following sections.

2. Charging from a D-C Source

In order to analyze the d-c charing circuit using an isolating inductance, the following assumptions are made;

(a) The PFN is represented by the capacitance, C_N, appearing across its terminals; the inductances of the PFN are short circuits at the low frequency of charging.

(b) The inductance of the charging reactor, L_c, is constant.

(c) The pulser switch is ideal.

(d) The inductance of the pulse transformer primary is negligible compared to that of the charging reactor.

Under the foregoing conditions, the equivalent charging circuit is as shown in Figure 3(a). E_{bb} is the d-c power supply voltage and R_c is the total resistance that causes damping in the circuit, in this case the ohmic resistance of the charging inductance. This circuit behaves as a damped ringing circuit which is triggered by the opening of the switch. The voltage on the PFN reaches a maximum value at time T_o equal to one-half the natural period of oscillation. If the pulse recurrence period T_r (the reciprocal of the PRF) is taken equal to T_o, the voltage on the network at the time of discharge is given by (Glasoe (2)):

$$v_N = E_{bb} + \left[E_{bb} - v_N(0)\right] e^{-\alpha T_o} \qquad (1)$$

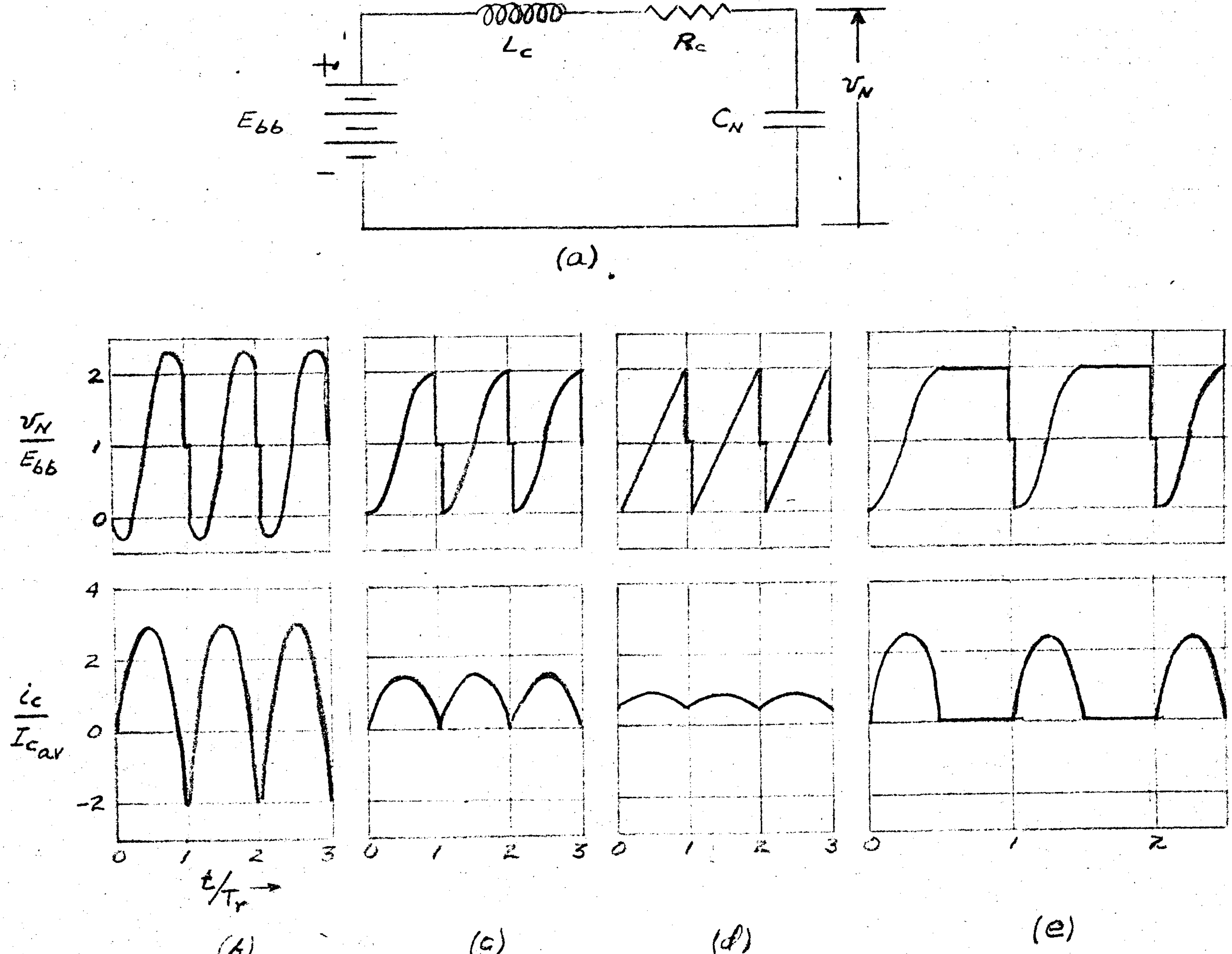


Fig. 3

where

$$\alpha = \frac{R_c}{2L_c} ,$$

and $v_N(0)$ = the initial voltage on the network. At this instant the charging current is zero, and no initial current through the inductance will exist for the next cycle. This condition is known as "resonance charging", and the following equation is then satisfied:

$$L_c = L_r = \frac{T_r^2}{\pi^2 C_N} \tag{2}$$

If T_r is different from T_o, an initial current will exist, and may be either positive or negative. For T_r less than T_o, the voltage on the network is still rising when the switch is fired, and so-called "linear charging" occurs. This is obtained in practice by using a value of inductance larger than that which would produce resonance charging. By using a large inductance and utilizing linear charging, some degree of flexibility of PRF and/or pulse width (a function of C_N) may be secured. For inductances smaller than L_r however, the initial current is negative, with the result that circuit losses are increased, and the PFN voltage is past the peak at time of discharge. For these reasons, this condition is avoided in actual circuits. Figure 3(b)(c) & (d) illustrate the current and voltage waveforms for the three cases discussed above; for simplicity in plotting, the circuit is taken as lossless, and the initial voltage on the PFN is considered to be zero.

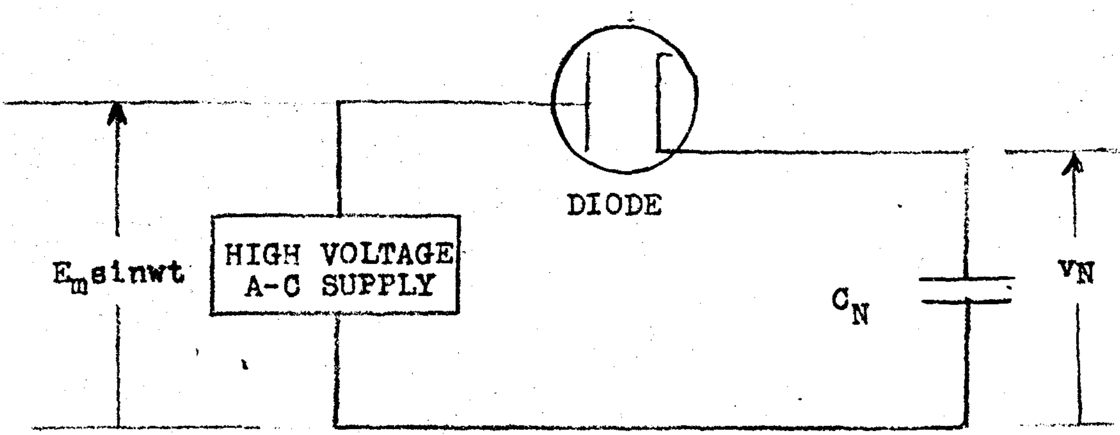
In cases where greater flexibility of PRF or pulse width, or both, is desired, or where a small value of charging inductance is indicated, a hold-off diode is employed in series with the charging reactor. This device prevents the flow of negative current and maintains the network voltage very close to the peak value until firing occurs. Equation (1) remains valid for this case, but R_c is increased by the effective resistance of the diode. This results in a slightly lower efficiency than in the case of ordinary resonance charging; typical figures being 95% for the latter and 90% for the former. Current and voltage waveforms for the case of d-c resonance charging with a diode are given in Figure 3(e).

3. Charging from an A-C Source

It is possible to use an a-c voltage source, such as an ordinary high-voltage transformer, to charge the PFN, provided the charging frequency is integrally related to the PRF. An isolating element is necessary as in the case of d-c charging; if the network were directly across the sedondary terminals, and discharged at peak transformer voltage, it would tend to recharge immediately, and short-circuit the transformer during the ensuing quarter cycle.

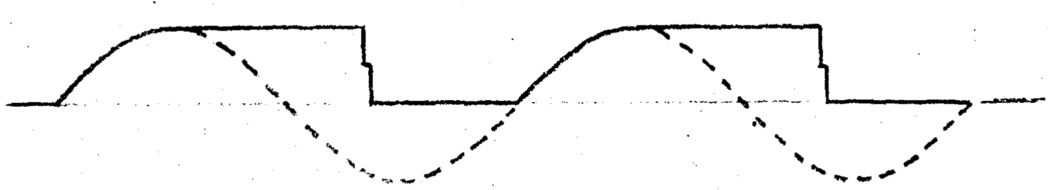
One method of isolation is to place a diode in series as shown in Figure 4(a). Neglecting the small amount of damping, the PFN will charge to the positive peak of supply voltage, E_m, and will be held at that point by the diode until the switch closes. Discharge must be made to occur during that part of the supply cycle in which the switch is non-conducting, otherwise the switch and diode would complete a short-circuit across the power supply. The maximum PRF is limited to the frequency of the supply voltage; repetition rates at any submultiples of the supply frequency are also possible. Figure 4(b) illustrates the voltage waveforms involved in this method of charging.

Another method of charging from an a-c source is to use a series inductance as the intermediate element. By maintaining the proper relation between PRF and supply frequency, and between C_N and L_c, it is possible to discharge the PFN at a voltage peak and yet preserve stable circuit conditions. The most common method is to choose L_c to resonate with C_N at a frequency equal to that of the source; this is known as "a-c resonance charging". In this case the charging circuit is a "high-Q" series resonant circuit, as shown in Figure 5(a). The voltage across C_N will build up as shown in Figure 5(b), being maximum at the zeros of the impressed voltage. Damping in the circuit limits the magnitude of oscillations after several cycles, but has little effect on the first



(a)

Equivalent charging circuit



- - - Supply voltage $E_m$sinwt

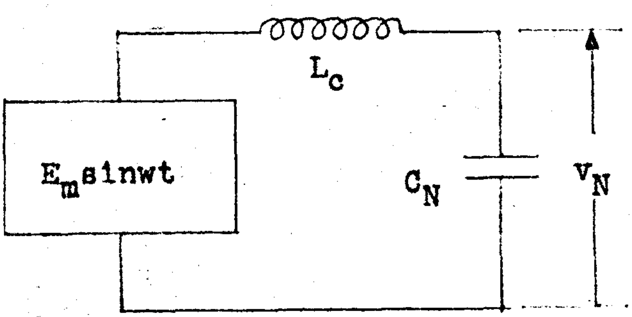
——— PFN voltage v_N

(b)

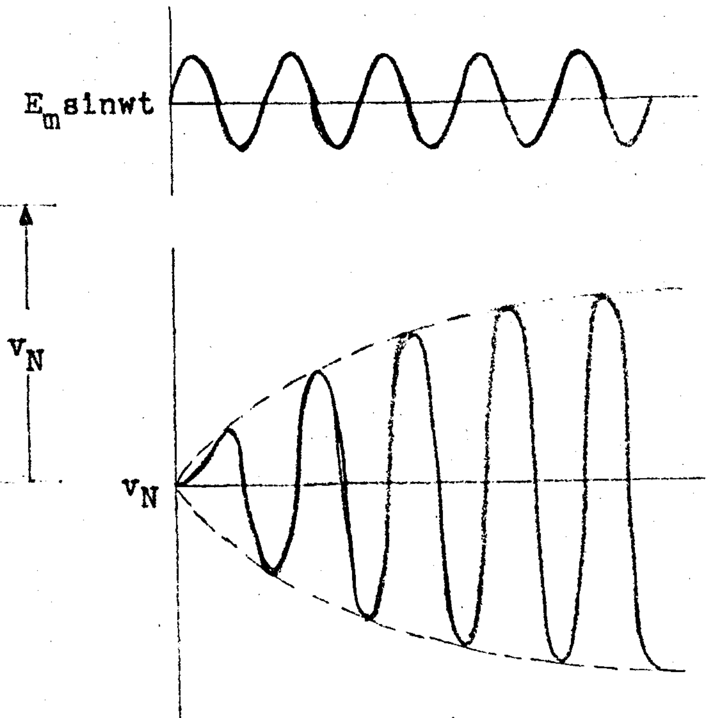
Voltage waveforms

A-C DIODE CHARGING

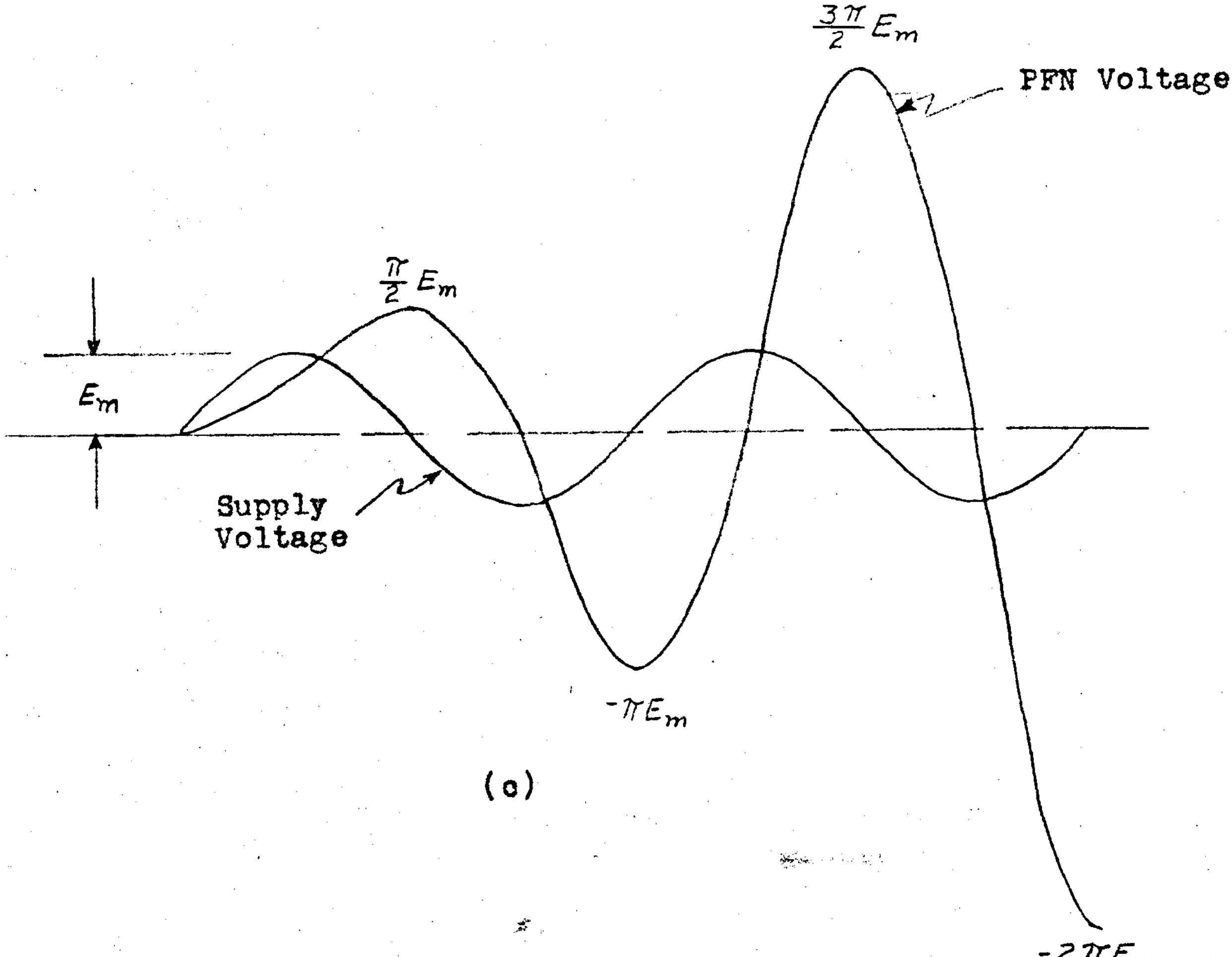
Fig. 4



(a)
Equivalent
Charging Circuit



(b)
Voltage Waveforms



(c)

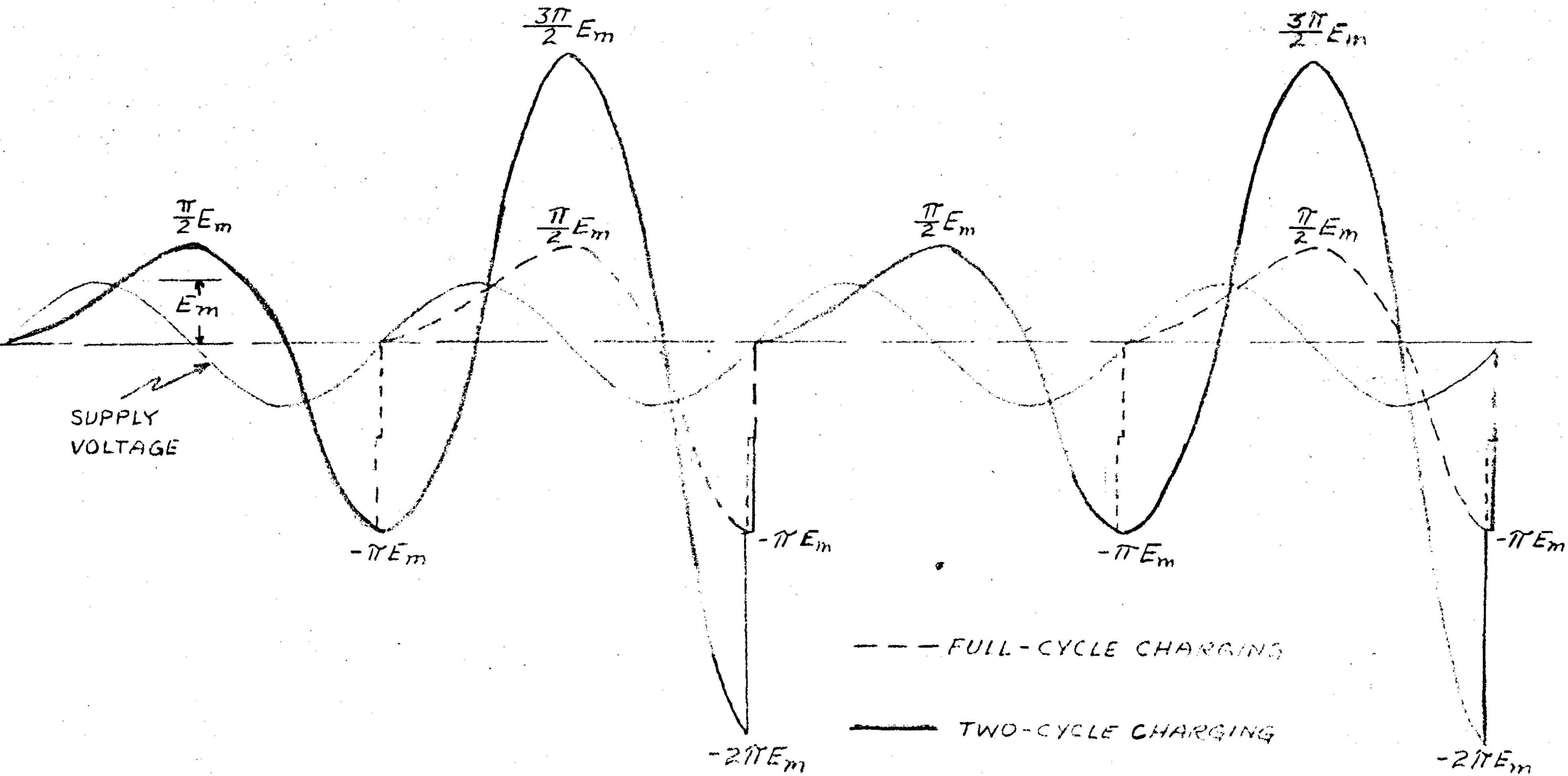
Fig. 5

few peaks. Neglecting damping, the successive voltage peaks have the values $\frac{1}{2}\pi E_m$, $-\pi E_m$, $\frac{3}{2}\pi E_m$, $-2\pi E_m$, etc., as shown in Figure 5(c). If good efficiency is to be maintained, discharge must occur on the peaks of the resonant wave; however, the discharge should occur only on those peaks which are spaced an even number of half wave lengths apart with respect to the supply voltage wave. Figure 6 illustrates this for "full-cycle" and "two-cycle" charging. If discharge occurs on the odd half wave length peaks, the direction of the pulse will alternate as shown in Figure 7. Therefore, the PRF is restricted to the frequency of the supply voltage (full-cycle charging) or integral submultiples thereof, unless some type of reversing switch is used to deliver all voltage pulses to the pulse transformer with the same polarity. In this latter case the maximum PRF is twice the supply frequency (half-cycle charging).

4. Inductive Impulse Charging

Although this method of charging utilizes a d-c power supply, the modulator circuit is somewhat different from those discussed previously, and for that reason is described separately. The simplified circuit diagram is given in Figure 8, and Figure 9 shows the most important waveforms involved. The circuit employs a saturable-core reactor switch which presents a high impedance to low currents and a low impedance to high currents. Briefly, operation is as described below:

11330



15 Fig. 6

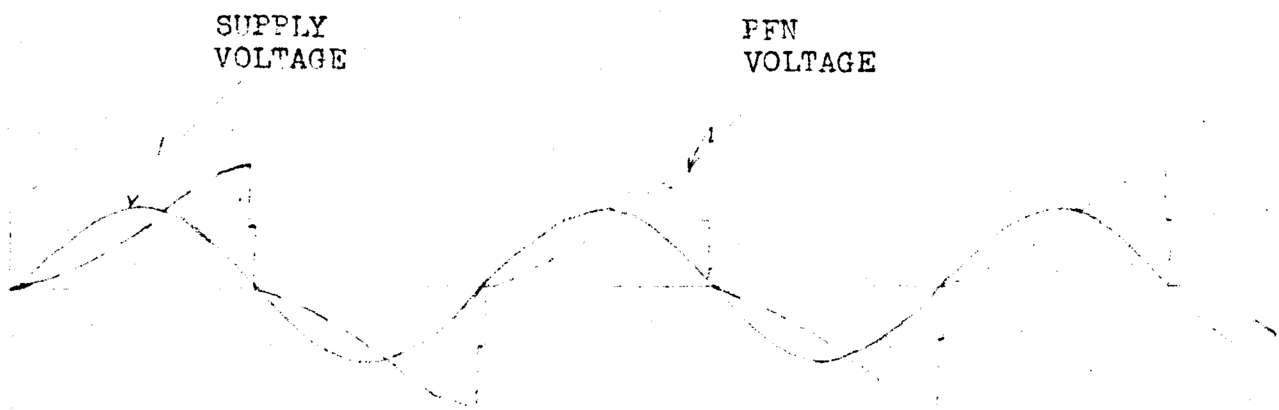

SUPPLY
VOLTAGE
PFN
VOLTAGE

Fig. 7

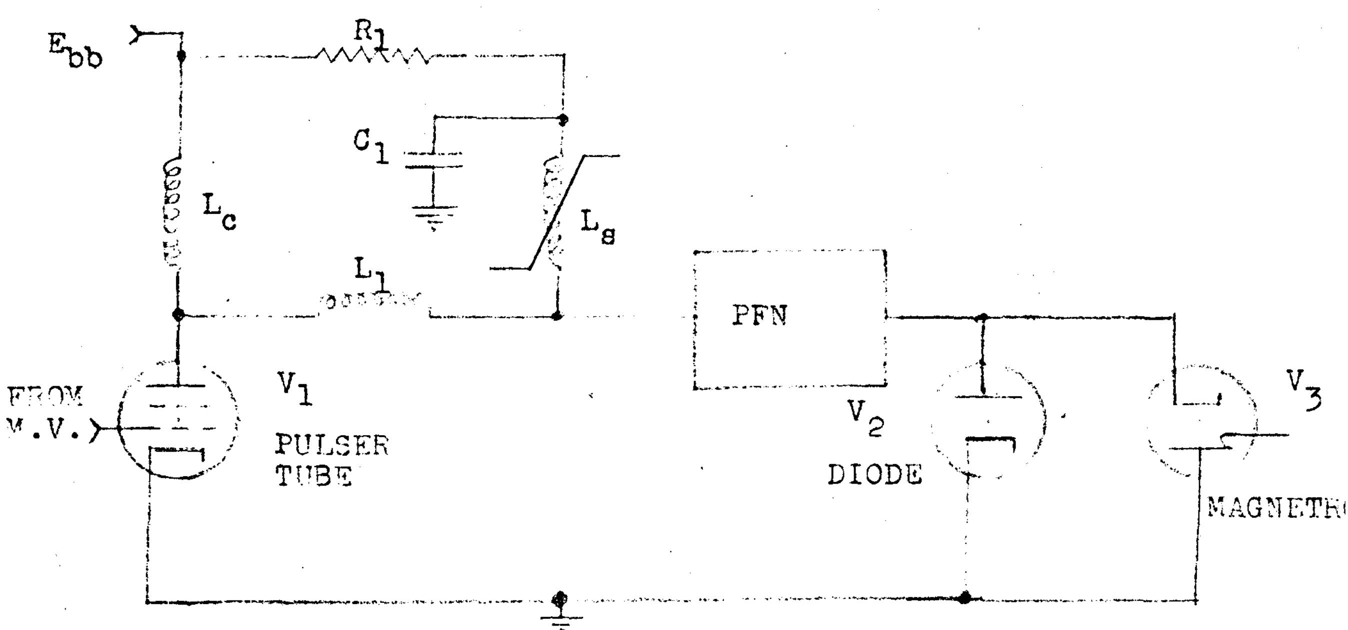

E_{bb}
R_1
C_1
L_c
L_s
L_1
PFN
FROM
M.V.
V_1
PULSER
TUBE
V_2
DIODE
V_3
MAGNETR

Fig. 8

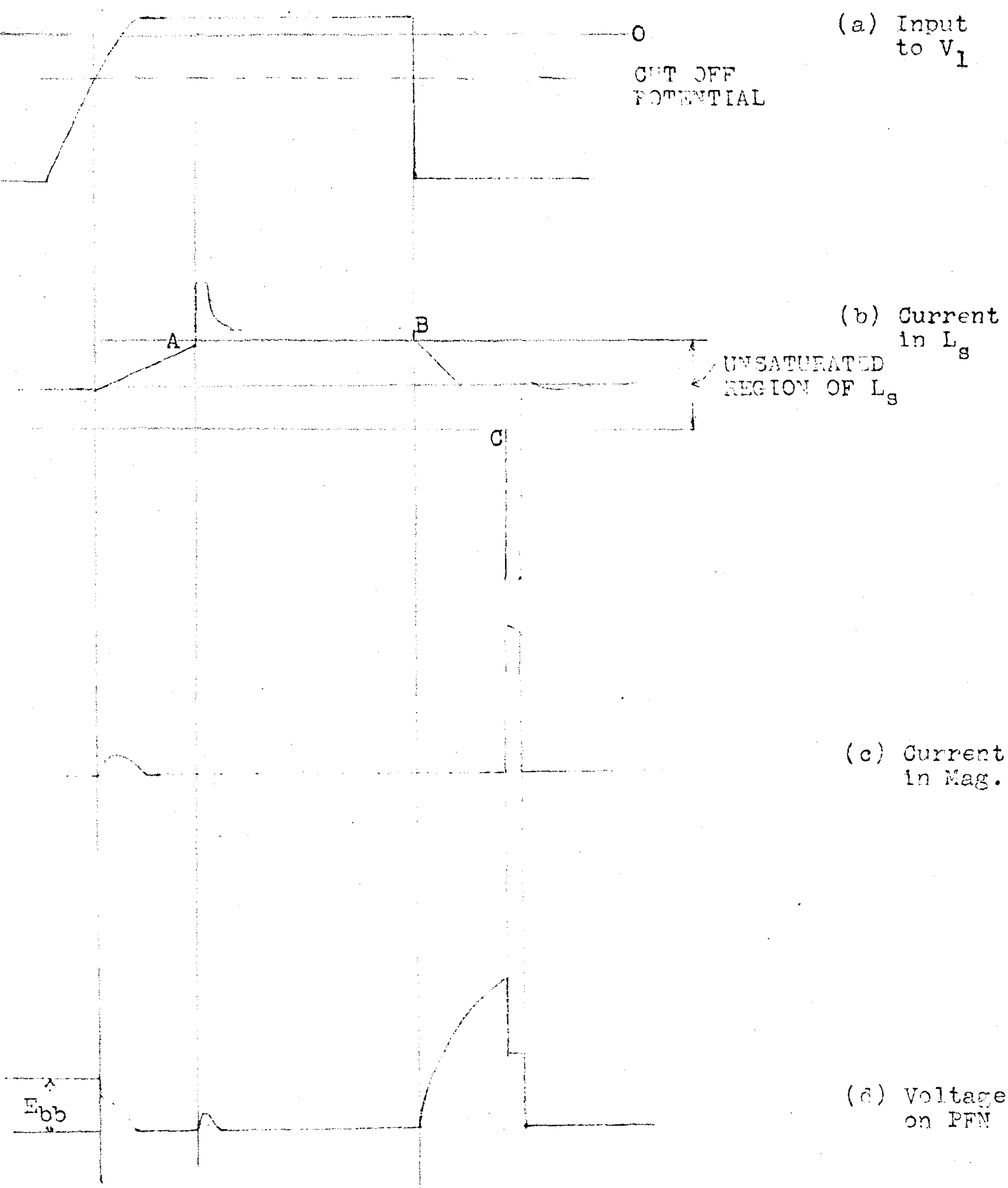

0
CUT OFF POTENTIAL
(a) Input to V_1
A
B
(b) Current in L_s
UNSATURATED REGION OF L_s
C
(c) Current in Mag.
E_{bb}
(d) Voltage on PFN

Fig. 9

A long preparation pulse (Figure 9(a) from a timing multivibrator is applied to the grid of the pulser tube, V_1, causing it to conduct. Plate current is drawn through three paths: through the charging inductor, L_c, through the saturable reactor, L_s, and through the PFN and load. When the current through L_s (Figure 9(b)) reaches the positive limit of the unsaturated region of the reactor (point A), the impedance suddenly drops, C_1 discharges rapidly, and the current i_{L_s} is then limited by the value of R_1. Concurrently, the PFN, which necessarily had been charged to E_{bb} prior to conduction of V_1, discharges and causes a small positive pip of current to flow in the load (Figure 9(c)). This could cause an undesired r-f output pulse, and therefore a small choke L_1 is included to limit the rate of discharge of the PFN. After C_1 and the PFN discharge, the circuit remains in a steady state until the input pulse to the grid of the pulser tube suddenly drops, thus commencing the actual modulator operation. V_1 is cut off, but the current in L_c cannot stop instantly, and therefore flows through the PFN and the diode to ground. The only other available path would be through L_s and C_1, but this path is denied by the high impedance of the saturable reactor, i_{L_s} having suddenly dropped to the upper limit of the unsaturated region when V_1 ceased conduction (point B). The current flowing into the PFN charges it to a voltage that is very much higher than the power supply voltage due to the inductive impulse. While the network is charging, the current through the saturable-core reactor decreases slowly; when it reaches the negative limit of the unsaturated region (point C), L_s becomes effectively a short-circuit and discharges the PFN. The discharge path includes the magnetron, and therefore an output r-f pulse is produced.

Reference to Figure 9(b) and 9(d) shows that the period for the PFN to charge to a maximum voltage must be equal to the period required for i_{L_s} to pass through the unsaturated region. The former is a function of C_N and L_c, and the latter depends upon the design of the saturable-core reactor. This critical relationship limits the usefulness of the circuit to cases where only one PRF is required.

5. A Comparison of Charging Methods

Each of the charging methods described above have certain inherent advantages and disadvantages. The most important of these are listed on the following page in outline form.

Method	Advantages	Disadvantages
D-c resonance charging	Very high efficienty	Requires heavy high-voltage power supply Inflexibility of PRF
D-c resonance charging with a diode	Continuously variable repetition rate High efficiency	Requires heavy high-voltage power supply Some loss of efficiency due to damping of diode
A-c diode charging	Light and simple	Inflexibility of PRF Requires synchronized switching
A-c resonance charging	Light and simple Good efficiency	Requires synchronized switching Flexibility of PRF severely restricted
Inductive impulse charging	Relatively low voltage supply	Critical relationship between L_c and L_s Poor pulse shape Low efficiency for pulse widths greater than 1 microsecond

CHAPTER III

SWITCHING METHODS

1. Requirements of the Switching Device

The characteristics required of a switch for a line-type modulator are clearly defined. First, the switch must be non-conducting during the interpulse interval. Second, it must be able to close very rapidly, and at a pre-determined time. Again, it must be capable of carrying the full pulse current and should have a very low resistance during the pulse. The voltage across the switch falls to zero, or very nearly so, at the end of the pulse. hence the switch is not required to be a current-interrupting device.

As discussed previously, one method of switching a line-type pulser is by the use of a non-linear inductance. Otherwise, however, the switch is usually an enclosed fixed spark gap, or a gaseous-discharge tube. Rotary spark gaps are also employed, and are quite satisfactory in some applications. The use of a grid-controlled high-vacuum tube is obviated by the fairly high resistance it presents during conduction, and also by its low cathode efficiency. Each of the three main types of switch - rotary gap, fixed gap, and hydrogen thyratron - will be treated with regard to their uses, characteristics, and special requirements.

2. The Nature of the Spark Discharge

When the voltage between two electrodes is raised to, or above, the static-breakdown point, a spark will occur. This spark is caused by the breakdown of the gas in the region between the electrodes, and is a result of the ionization of the gas molecules by the accelerated free electrons in the vicinity. This acceleration is imparted to the electrons present by the electric field existing between the terminals, and is directly proportional to the potential difference across the gap. The higher the applied voltage, or the greater the number of free electrons present, the shorter the elapsed time between application of the voltage and initiation of the discharge. For example, if the voltage be just equal to the static-breakdown potential, a "time lag" of several minutes may occur. If the voltage be raised to a value two or three times the minimum required, the time lag is reduced to the order of hundredths or even thousandths of a microsecond. After the ionization commences, a short interval, of the order of .01 microsecond, transpires before the discharge attains the properties desired for use as a switch. At the end of this very short "breakdown time", the gas between the electodes has changed from an insulating medium to one capable of carrying quite high currents. Factors determining the characteristics of discharge are the type of gas, gas pressure, gap geometry, and the shape of the applied voltage wave.

In order to utilize the above properties of the spark, the breakdown must be controlled, and this can easily be done by heavily "overvolting" the gap. In the case of a fixed gap, a high transient peak of voltage is applied to one electrode. Another method is to maintain a high difference of potential between electrodes, and vary the spacing between them at the desired rate - this is the method employed in the rotary spark gap.

3. Rotary Spark Gaps

Essentially, the rotary gap consists of an insulating disc containing a set of electrodes and rotating in the vicinity of one or more fixed electrodes. Several variations are commonly used. The pins may be set parallel to the axis of rotation, or they may extend radially outward about the periphery of the disc. They may extend on either side of the disc and discharge simultaneously to two fixed electrodes. The fixed and moving pins may overlap for a portion of their length, or pass one another with end faces opposing.

Rotary gaps have a high power handling ability and are simple and rugged in construction. Furthermore, no external triggering voltage is necessary to initiate the switching action. To partially offset these desirable features, certain disadvantages are encountered:

(a) The PRF depends directly upon the number of electrodes, both fixed and rotary, and upon the speed or rotation. The spacing between adjacent electrodes is limited by electrical considerations, such as breakdown voltage and deionization between pulses. The size of the rotor is limited by mechanical design. Hence, even with a high speed motor, the number of electrodes is limited, and so rotary gaps are not well suited for high recurrence rates. The greatest PRF used in practice is 800 p.p.s. In addition, flexibility of PRF is restricted; a change of two to one may be effected by switching a second fixed electrode in series with the one used for the lower rate, but in general each rotary spark gap is designed for one PRF only.

(b) Rotary spark gaps have an inherently large time "jitter". This refers to the uncertainty in time of the initiation of breakdown, and results from both electrical and mechanical causes. Since the spacing between electrodes at the point of closest proximity is such that a voltage of two or more times the static-breakdown potential exists, the spark will always occur before the point of closest approach is reached. The exact time of spark will vary from discharge to discharge. Increasing the relative velocity of the electrodes, i.e., increasing the speed of the rotor, tends to decrease the amount of this jitter. Mechanical tolerances in assembly of the rotating parts, in radial and peripheral spacing of the moving pins, and in rigidity must be extremely close. As an example, an error in peripheral spacing of .02 inches can cause an increase of time jitter of 15 microseconds or more.

The accumulated jitter due to mechanical and electrical causes may be as much as ±50 microseconds, but with careful design this figure has been reduced to ±20 microseconds in some instances. There are two direct consequences of this time jitter that are important. First, the entire system employing such a modulator must be synchronized in time. For example, the sweeps on the indicator tubes must be triggered by the power pulse. Second, if d-c charging is used, a variation in time of firing will result in a variation of pulse voltage amplitude, and consequently in output pulse power. While this effect is small in d-c resonance charging, it is so serious in the case of linear charging as to make the combination of linear charging and rotary gap switching unusable.

In addition to the two major limitations of jitter and low PRF, there are a number of minor difficulties encountered. Corrosive gases are generated by the breakdown of the air in the gap. These can be dispersed by forced ventilation in some installations, but where a wide variation in ambient pressure is a factor, as in an airborne pulser, the gap must be pressurized in order to maintain stability. In this case, the use of blowers is not possible, and some absorbing medium, such as activated carbon, must be enclosed in the chamber to remove the gases. Changes in gap geometry due to erosion and pitting of the electrodes will change the spark characteristics. This is combated by the use of tungsten electrodes, by making the fixed pin the high voltage terminal, and by the use of the overlapping construction.

When used with a-c resonance charging, the use of a rotary spark gap switch can result in a simple and compact pulser. The charging reactor can be designed as the leakage inductance of the transformer secondary, and the rotor of the gap mounted directly on the shaft of the a-c machine that excites the circuit. The variation of pulse amplitude with time jitter is very much less pronounced than in the case of d-c charging, and with overall system synchronization the jitter can be tolerated.

4. Enclosed Fixed Spark Gaps

The most commonly used switching circuit employing fixed gaps is obtained by connecting two or three cold-cathode gas-filled diodes in series. These enclosed gaps are of two basic designs:

(a) The cylindrical-electrode aluminum-cathode gap - such as the Western Electric types 1B22 and 1B34 and the Westinghouse type WX3240.

(b) The iron-sponge mercury-cathode gap - such as the Western Electric type 1B42.

Both types are operated in the same manner. Figure 10(a)

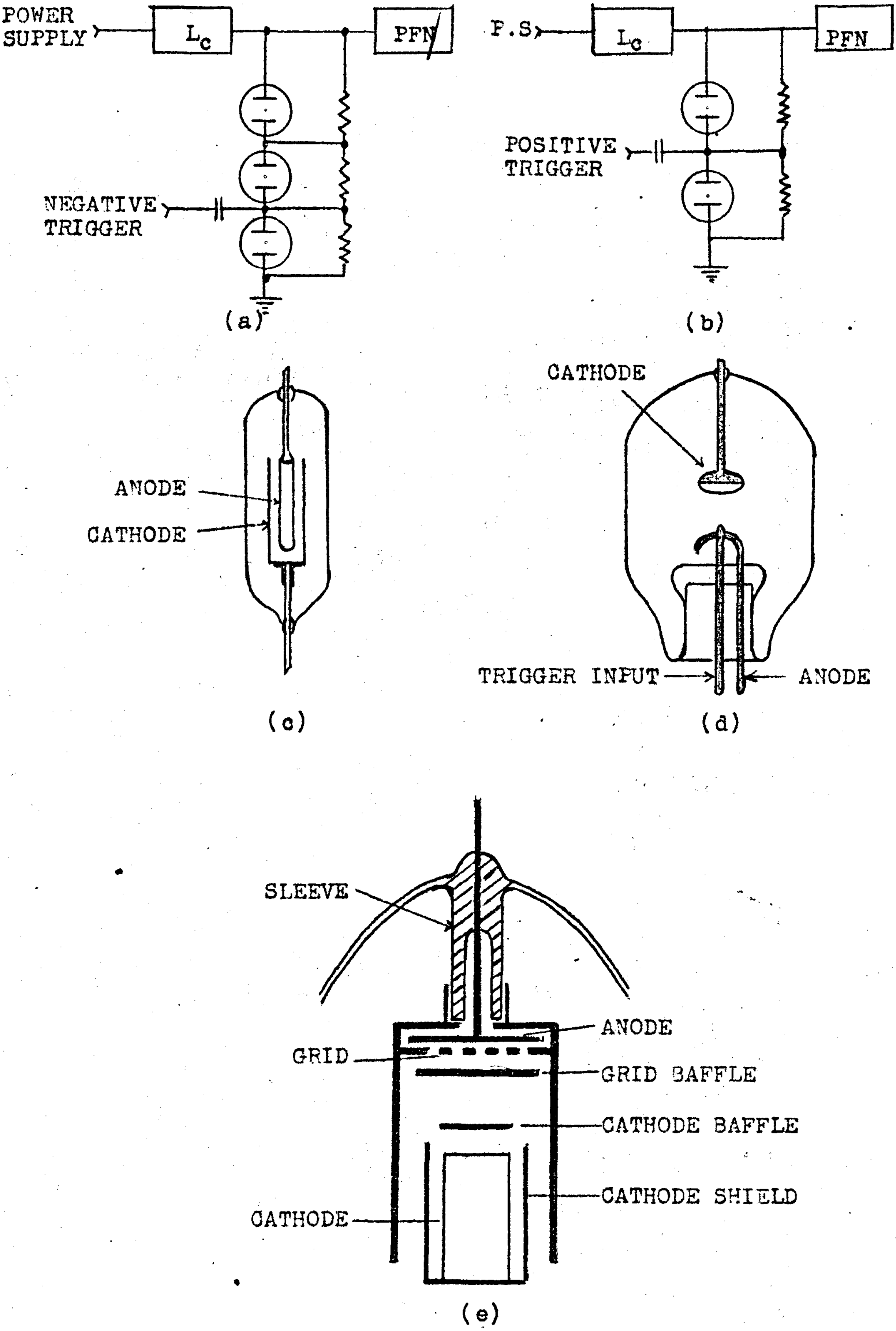

POWER SUPPLY
L_c
PFN
NEGATIVE TRIGGER
(a)
P.S
L_c
PFN
POSITIVE TRIGGER
(b)
ANODE
CATHODE
(c)
CATHODE
TRIGGER INPUT
ANODE
(d)
SLEEVE
ANODE
GRID
GRID BAFFLE
CATHODE BAFFLE
CATHODE SHIELD
CATHODE
(e)

Fig. 10

and 10(b) show two methods of employment. The voltage divider is placed across the gaps in order to divide the full PFN voltage equally across each one. By impressing a rapid-risihg, high-voltage trigger pulse, one of the gaps will break down. The full PFN voltage will then appear across the remaining gap(s), which will in turn rapidly break down, thereby completing the closing of the switch.

The cylindrical-electrode aluminum-cathode tube consists of a cylindrical aluminum cathode that almost completely encloses a long anode rod of the same material. This is illustrated in Figure 10(c). The enclosed gas is a mixture of approximately 80% hydrogen and 20% argon. By providing a large cathode area the effects of cathode erosion are minimized, and aluminum is used because of its relatively low erosion rate in the hydrogen-argon mixture compared with other metals. For a given pulse current the erosion is directly proportional to pulse duration, and this property limits tube life when wide pulses are used. The time jitter when using this type of switch is of the order of 6 microseconds.

The iron-sponge mercury-cathode gap consists of a thin molybdenum anode, and a cathode of mercury immobilized by an iron "sponge". This sponge is made of compressed iron powder containing about 60% void space. The sponge holds approximately 9 cc. of mercury, the surface tension of which maintains a film over the surface and prevents erosion of the sponge proper. The gas in the tube is pure hydrogen under relatively high pressure. The use of this gas, together with small gap spacing and a thin anode rod, makes time jitter in this type of switch very low - of the order of .02 microseconds. Low erosion makes for a longer tube life and a wider range of operating conditions than in the case of the aluminum-cathode gap.

The choice of two or three gaps in series depends upon the choice of tube and upon the PFN voltage. At low power two-gap operation of aluminum-cathode gaps is satisfactory, but in general it is desirable to use three gaps. When two only are used, a bi-directional trigger pulse should be used to insure satisfactory starting. In the case of the iron-sponge type, two-gap operation is generally satisfactory for all conditions, and a bi-directional pulse is not needed.

A third type of enclosed fixed spark gap is the three electrode type, of which the British "trigatron" is the principal example. It consists of a molybdenum cathode and anode, and a tungsten trigger pin, as shown in Figure 10(d). The gas is a mixture of 95% argon and 5% oxygen at a pressure of about 50 psi. The trigger voltage must be very high - of the order of 6 KV - and life is limited to about 200 hours by the rapid erosion of the trigger pin. Jitter is less than 0.1 microsecond.

5. The Hydrogen Thyratron

By far the most successful switch available for use in a line-type modulator is the hydrogen thyratron. The 3C45, 4C35, and 5C22 are at present on the market, the latter being rated at 16 KV peak anode voltage, and a fourth tube designed for even higher voltages is under development. A diagram of the internal structure of the tube is given in Figure 10(e). The gas in the envelope is hydrogen at a pressure of about .005 mm. of H_g, and the grid-anode spacing is very much smaller than in conventional mercury vapor thyratrons. The tube has a positive control-grid characteristic, and in order to start conduction it is necessary to drive the grid sufficiently positive to draw grid current. This current produces ions and electrons in the region external to the cathode shield, and some of these reach the area of the grid baffle. When the electron density there becomes high enough, the anode field, which is present only in the region above the grid baffle, will produce ionization in that region and breakdown will occur. The ionization time is .03 microseconds for the 3C45 and .07 microseconds for the 5C22.

Due to the chemical activity of hydrogen, the metal parts must be very pure, and great care in manufacture is necessary in order to prevent the inclusion of any contaminating substance. Should combination of the hydrogen with any impurities take place, the phenomenon known as "gas cleanup" occurs. This means that the hydrogen no longer exists in the gaseous form but has formed chemical compounds with the impurities, and the utility of the tube is lost. The cathode operating temperature is rather critical. The upper limit is about 850°C due to the reducing action of the hydrogen on the oxide cathode above that temperature; the lower limit is about 800°C due to the rapid loss of cathode emission below that temperature. For this reason indirect heating is used to minimize temperature variations over the cathode surface, and a fairly long filament warm-up period must be observed before applying plate voltage. Tube life is about 500 hours at maximum rating of pulse voltage and pulse current, but is considerably lengthened by decrease of either. Pulse rate also affects tube life, and at maximum ratings should not be in excess of 1000 p.p.s. for the 5C22 and 2000 p.p.s. for the other tubes. At reduced power output, however, recurrence rates up to 40,000 p.p.s. have been achieved. Thyratrons may be operated in series or in parallel in order to exceed the current or voltage ratings of a single tube. Special circuitry is necessary in these cases to assure balanced operation.

The hydrogen thyratron may be triggered very precisely with very small time jitter. The trigger voltage need be only about 150 volts at a rate of rise of 200 volts per microsecond. This means that a simple external trigger generator circuit can be employed. The jitter obtained when using such a trigger voltage is less than .05 microseconds. With a trigger pulse of amplitude 200 volts and a rise rate of 850 volts per microsecond the jitter has been reduced to .003 microseconds.

6. A Comparison of Switching Methods

Each of the methods for switching discussed above has its limitations. A truly ideal switch suitable to all applications has not been found. Listed on the following page in outline form are the major advantages and disadvantages of each type.

Method	Advantages	Disadvantages
Saturable-core reactor	Long life High PRF High power levels	Poor pulse shape Requires special charging circuit
Rotary spark gap	Simple and rugged High power levels	High time jitter Low PRF
Aluminum-cathode gap	High power levels	Short tube life Requires more than one gap High trigger voltage
Iron-sponge gap	Low time jitter Long tube life	Requires more than one gap High trigger voltage
Trigatron	Very low time jitter	Requires very high trigger voltage
Hydrogen thyratron	Very low time jitter Good tube life Low trigger voltage	Shorter tube life at high power or at high PRF

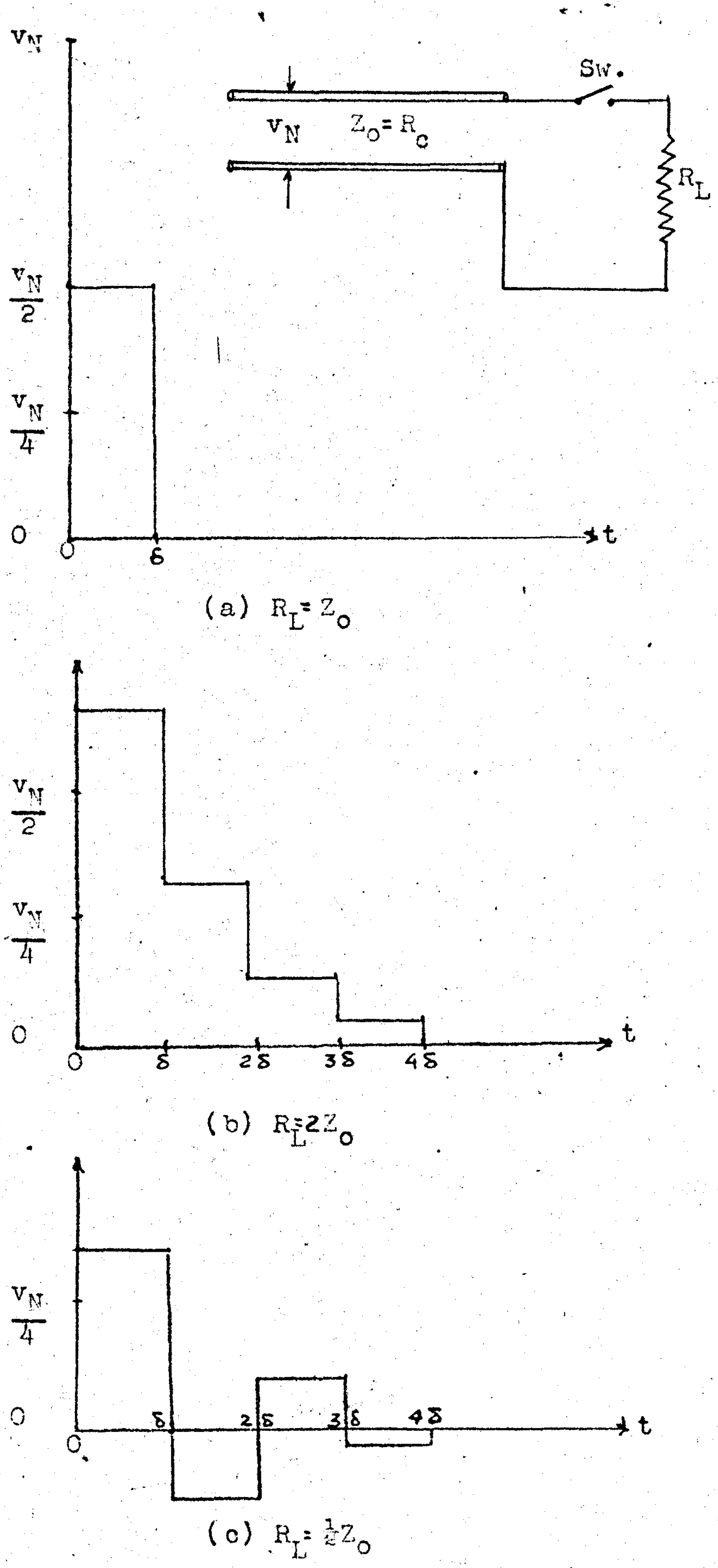
CHAPTER IV

PULSE FORMING NETWORKS

1. Elementary Theory

In the foregoing chapters it has been tacitly assumed that if the pulse forming network were charged to a voltage v_N, and then placed directly across a load equal to its characteristic impedance, a pulse of voltage of approximately $\frac{1}{2}v_N$ would appear across the load. In order to verify this assumption, first consider the PFN to be an ideal lossless transmission line, open-circuited at the far end.* Assume the load resistance, R_L, to be equal to the characteristic resistance of the line, R_C, and that the line is charged initially to a voltage v_N. (Refer to Figure 11). At the instant the switch is closed, one-half of the voltage stored on the line will immediately appear across R_L, leaving the voltage across the input terminals of the line reduced to $\frac{1}{2}v_N$. This is equivalent to introducing a negative voltage wavefront of amplitude $\frac{1}{2}v_N$ to the input terminals - this wavefront will travel down the line at the propagation velocity to the open end, leaving the line charged to $\frac{1}{2}v_N$ along its entire length. At the open end, the negative wavefront is reflected in the same phase, and returns to the input end, canceling the remaining voltage on the line as it travels. At the end of a period equal to twice the length of the line divided by

* Shorted lines are not used in practice, due to the lack of a suitable switch to interrupt the high current.



v_N
Sw.
v_N
$Z_0 = R_c$
R_L
$\frac{v_N}{2}$
$\frac{v_N}{4}$
0
0
δ
t
(a) $R_L = Z_0$
$\frac{v_N}{2}$
$\frac{v_N}{4}$
0
0
δ
2δ
3δ
4δ
t
(b) $R_L = 2Z_0$
$\frac{v_N}{4}$
0
0
δ
2δ
3δ
4δ
t
(c) $R_L = \frac{1}{2}Z_0$

Fig. 11

the velocity of propagation, the line is completely discharged. At this time the voltage across R_L drops to zero, and the energy stored in the distributed capacitance of the line has been completely transferred to the load. Considering this capacitance to be lumped, and denoting it as the "storage capacity" of the line, C_N, the following important relation is developed, (MIT (7)).

Energy stored on line = Energy transferred to load

$$\tfrac{1}{2}C_N v_N^2 = V_L I_L t$$

$$= \left(\frac{v_N}{2}\right)\left(\frac{v_N}{2R_C}\right)\delta$$

$$C_N = \frac{\delta}{2R_C} \qquad (1)$$

This relation is widely used in the design of the pulser circuit, particularly when resonance charging is to be employed.

If R_L is not equal to R_C, then the voltage across the load will accordingly be either greater or smaller than $\tfrac{1}{2}v_N$, and the effective wavefront will similarly be different. In addition, since the line is not terminated in its characteristic impedance, a series of reflections will occur, in accordance with elementary transmission line theory (assuming the switch remains closed). The voltage waveforms appearing on the load for the three conditions, $R_L = R_C$, $R_L = 2R_C$, and $R_L = \tfrac{1}{2}R_C$ are illustrated in Figure 11(a)(b)&(c), respectively. In practice, R_L is made equal to R_C in order to get a single pulse, and in order to completely discharge the PFN on each cycle.

2. Line-simulating Networks.

In order to obtain a pulse of one microsecond duration from an actual transmission line having a propagation velocity of 500 feet per microsecond (a representative value), a length of 250 feet would be required. Furthermore, this line would have to be capable of withstanding the high voltages used in radar pulsers. That such a long high-voltage cable would be much too massive and heavy for practical application is obvious. As a result, pulse forming networks that simulate actual transmission lines were developed.

The PFN is essentially a two-terminal network, since one end is open-circuited. It is a characteristic of two-terminal networks that their behavior is completely determined by the driving-point impedance function, i.e., the input impedance, expressed as a function of (angular) frequency, Glasoe (2) and Guillemin (3). The problem of finding a network which is equivalent to the ideal transmission line then becomes one of mathematical synthesis; given the impedance function, determine the network which satisfies that function. The input impedance of the ideal line is given by the equation,

$$Z = Z_0 \coth(j\omega t) \qquad (2)$$

where Z_0 is the characteristic impedance of the line, and t is the one way transmission time. The Laplace-transform impedance function is then,

$$Z(s) = Z_o \coth(st) \qquad (3)$$

where s is the Laplace-transform operator.

Figure 12(a) shows the well known "L-section" artificial line. It can be shown mathematically, Glasoe (2), that the input impedance function, Z(s), of this line will be of identical form to equation (3) in the limit as the number of L-sections approaches infinity. If a finite numbers of sections are used, there will be an upper limit to the frequencies present in the pulse, and hence the square shape will be only approximated. The greater the number of sections, the better will be the approximation, but a network of ten or more sections will generally be necessary in pulser applications.

Figure 12(b)&(c) shows two more line-simulating networks, obtained by the synthesis method mentioned above. The first of these is obtained by expanding Z(s) as a rational fraction, Glasoe (2), and the second by expanding in the same way the admittance function, $Y(s) = 1/Z(s)$,

$$Y(s)=\frac{1}{Z_o} \tanh(st) \qquad (4)$$

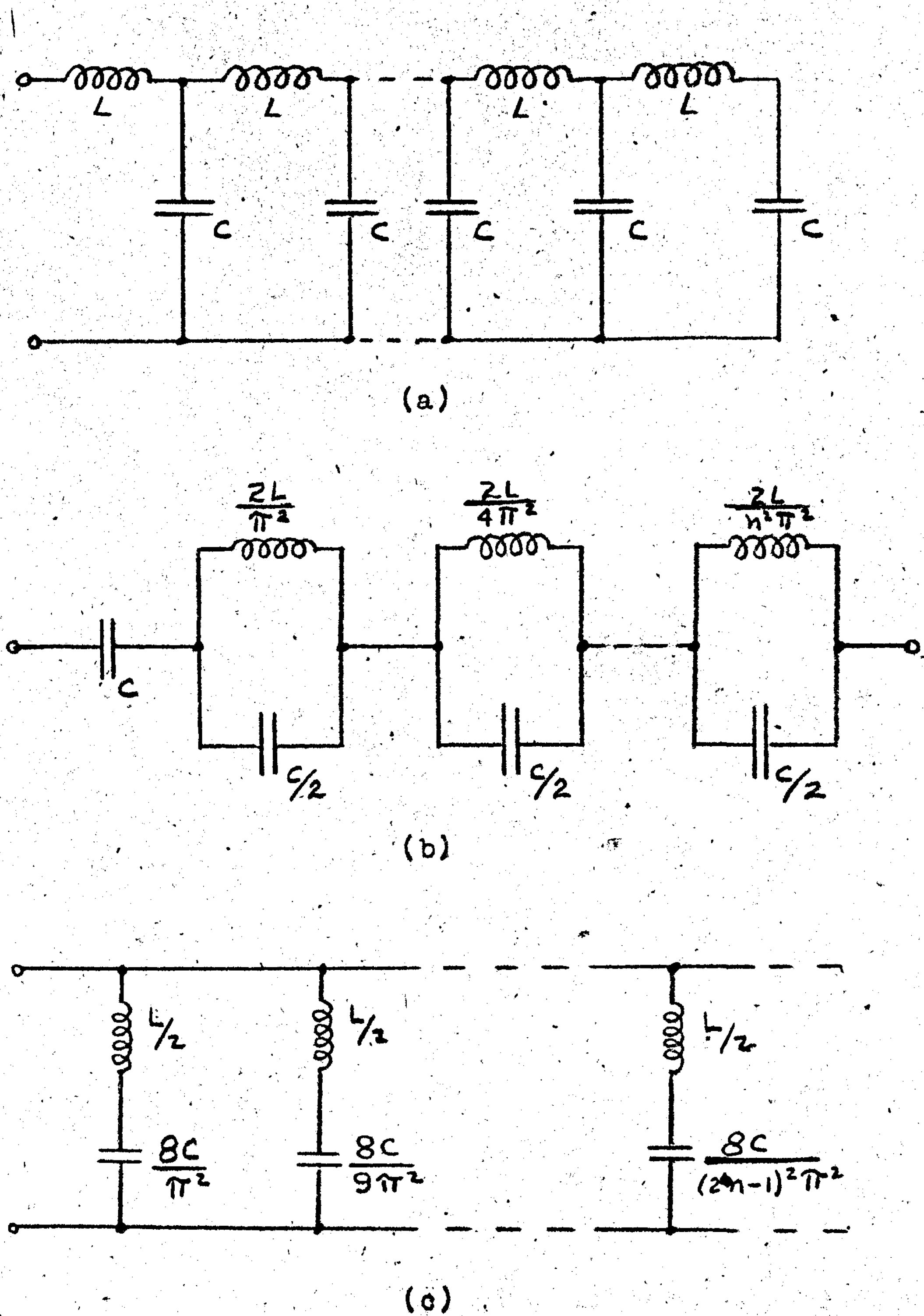

L
L
L
L
C
C
C
C
C
(a)
$\frac{2L}{\pi^2}$
$\frac{2L}{4\pi^2}$
$\frac{2L}{n^2\pi^2}$
C
C/2
C/2
C/2
(b)
L/2
L/2
L/2
$\frac{8C}{\pi^2}$
$\frac{8C}{9\pi^2}$
$\frac{8C}{(2n-1)^2\pi^2}$
(c)

Fig. 12

The rational-fraction expansion of either of these functions will result in an infinite series. Each term of the series is identified as the operational impedance function of a capacitance or inductance, or more generally a combination of the two. Comparison of the coefficients of the terms yields recursion formulae which are used to determine the values of the reactive elements in terms of Z_o and t. In these cases, again, an infinite number of sections must be employed to obtain a square pulse.

In general, it may be said of each of the three networks discussed above, and of any other that may be derived similarly, that a large number of physical elements are necessary if a reasonably square pulse is to be formed. In addition, analysis of these circuits using a finite number of sections reveals two properties that are undesirable in radar pulsers. First, overshoots will exist at the beginning of the pulse, and second, excessive oscillations occur along the top.

3. Guillemin Lines

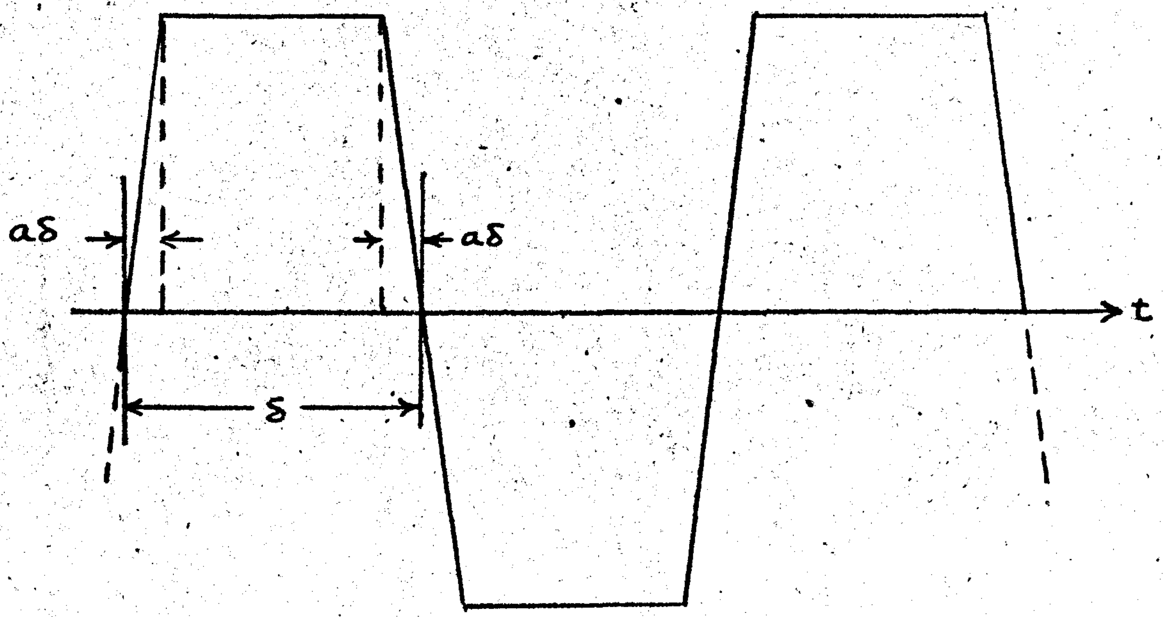
It was felt by Dr. Guillemin that the above limitations arose from attempting to generate, by means of a lumped-parameter circuit, a pulse having an infinite rate of rise and fall. That is, the pulse displayed discontinuities, and such a function cannot be produced by a finite series. Therefore, he proceeded on the premise that the pulse to be generated should have a finite rise and fall time. Mathematically, the discontinuities in the wave no longer exist, and the Fourier series is uniformly convergent. The property of uniform convergence signifies that the overshoots and oscillations may be reduced to any desired degree as more sections are added. In order to write the Fourier series of a function, the function must be periodic. Therefore, a further assuption was made: the network derived using an a-c wave will, when used in the pulser, produce a pulse similar to a section of that wave. This assumption has been justified by actual application. Two of the alternating-current waves used are shown in Figure 13(a)&(b). The Fourier series of the trapezoidal wave contains only sine terms, since it is an odd function. The series is given by

$$i(t) = I_\ell \sum_{k=1}^{\infty} b_k \sin \frac{k\pi t}{\delta} \tag{5}$$

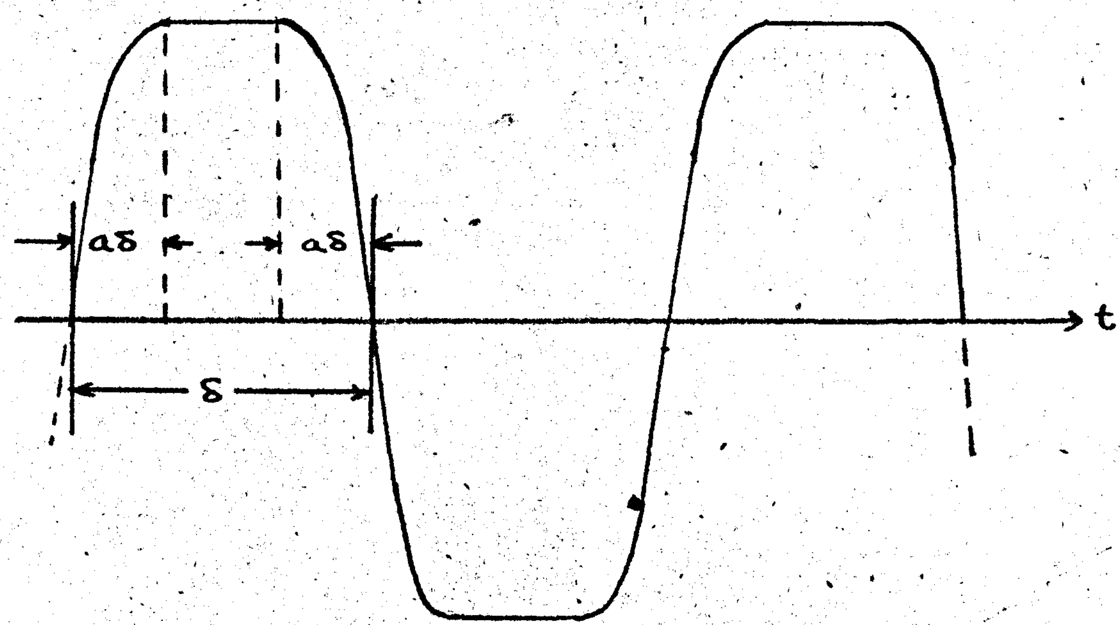
where

$$b_k = \frac{2}{\delta} \int_0^{\infty} \frac{i(t)}{I_\ell} \sin \frac{k\pi t}{\delta} dt \tag{6}$$

I_ℓ is the amplitude of the wave; i(t) is the equation of the a-c wave in terms of t, δ, and the rise time, aδ. This equation is represented by three functions, corresponding to the three regions, rise, top, and fall. i(t) is then



(a) Trapezoidal a-c wave.



(b) A-c wave with flat top and parabolic rise and fall.

Fig. 13

defined by

$$\frac{i(t)}{I_\ell} = \begin{cases} \dfrac{t}{a\delta}, & 0 \le t \le a\delta \\ 1, & a\delta \le t \le \delta - a\delta \\ \dfrac{\delta - t}{a\delta} & \delta - a\delta \le t \le \delta \end{cases} \qquad (7)$$

Substituting the above in the equation for b_k and performing the integration gives the result

$$b_k = \frac{4}{k\pi} \cdot \frac{\sin k\pi a}{k\pi a} \qquad (k = 1, 3, 5, \cdot\cdot) \qquad (8)$$

Each term in the series is a sine wave of frequency $k/2\delta$, and of amplitude b_k as defined above. Such a current, i_k, can be produced by a series L_k-C_k circuit, excited by a d-c voltage, V_o. The characteristic impedance of this circuit is

$$Z_o = \frac{V_o}{I_\ell} \qquad (9)$$

and the values of L_k and C_k are determined by comparison with the Fourier series coefficients, giving the recursion formulae

$$L_k = \frac{Z_o \delta}{k\pi b_k} \qquad (10)$$

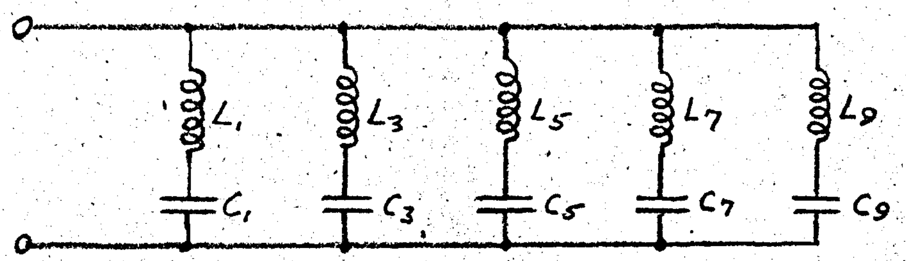
$$C_k = \frac{b_k \delta}{k\pi Z_o}$$

Since each term is generated by such a circuit described above, a pulse of n Fourier components can be obtained by adding n L_k-C_k circuits in parallel. The resulting network is of the type shown in Figure 14(a), for n = 5.

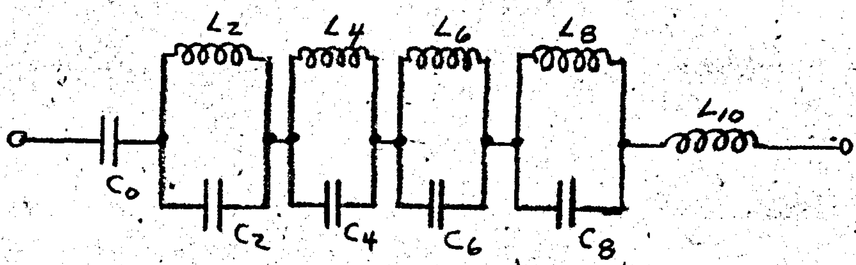
Such a network is not well suited to practical application, but is very valuable from the mathematical point of view, for it serves as a starting point in determining equivalent networks. In other words, the impedance function necessary to produce the desired pulse is now known, and it is therefore possible to synthesize other networks which will produce the same pulse shape.

The first equivalent form is found by applying Foster's reactance theorem, Guillemin (3). The mathematical procedure is tedious and involved. In brief, it consists of determining all the roots and poles of the function Z(s) of the pzevious network, and then expanding the function in partial fractions about its poles. The resulting expression is a Laurent series of n terms, involving the operator s. As before, the various terms may be identified as the Laplace-transform impedance functions of reactive circuits, and the network of Figure 14(b) results.

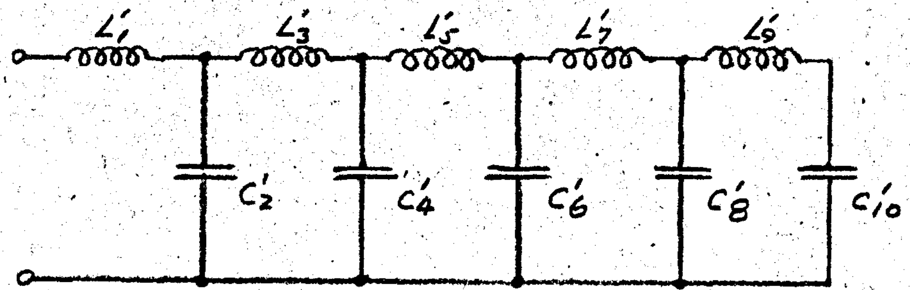
Cauer's extension of Foster's reactance theorem



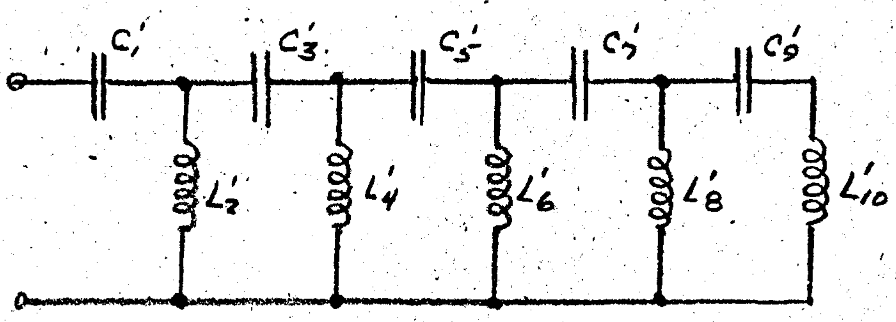
(a) 1st Foster Form



(b) 2nd Foster Form



(c) 1st Cauer Form



(d) 2nd Cauer Form

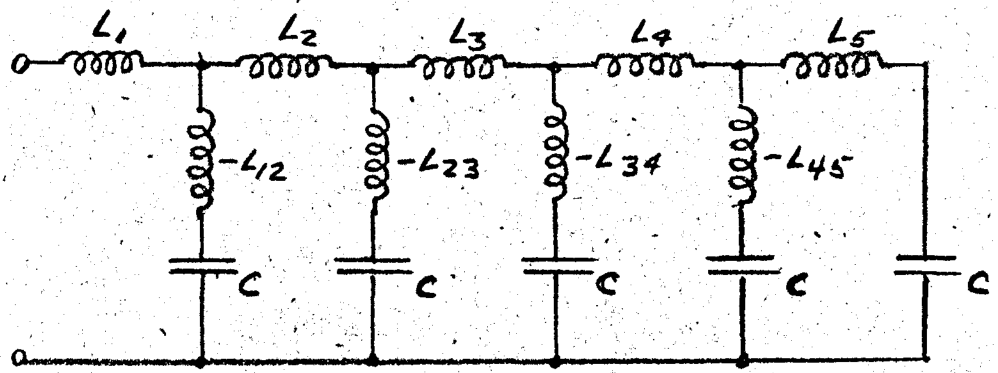
Fig. 14

yields two additional canonic forms, Guillemin (3). In this case, the procedure is to expand Z(s), or Y(s), as a continued fraction, and then identifying the terms obtained with network elements. The two Cauer networks are illustrated in Figure 14(c)&(d).

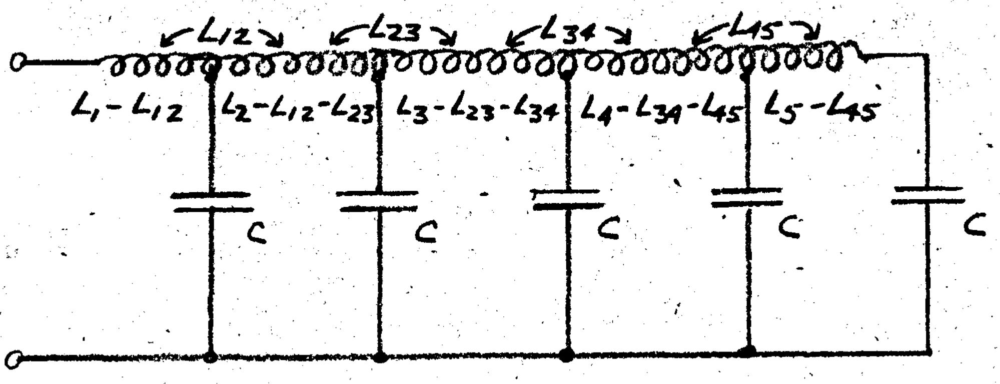
By various mathematical operations on the impedance and admittance functions, a large number of other equivalent networks may be determined. Still others may be found by combining the four canonic forms of Figure 14 in various ways. Of importance is the network having equal capacitances per section, since this results in much greater ease of manufacture. Starting with the first Cauer form, in which the capacitances are close to being equal, the network of Figure 15(a) may be found. When the values of capacitance are altered, a series inductance must be inserted in the shunt arms to compensate for the change, and this inductance is negative when the value of the capacitance is increased. Figure 15(b) is called the "type E" network, and is the equivalent of the one above, being obtained by the proper use of mutual inductance to give the effect of negative inductance. The type "E" network is the one most extensively used in practice.

4. Practical Considerations.

It should be noted that all the networks shown in Figures 14 and 15 consist of five sections. The results of experiments show that the improvement in pulse shape



(a) PFN having equal capacitances



(b) Type "E" Network

Fig. 15

as the number of sections are increased is slight for more than five sections. The network used in these experiments was derived on the basis of a trapezoidal wave shape, with a fractional rise time of 8%, i.e., $a\delta = .08\,\delta$. Use of only four sections resulted in an appreciably poorer pulse shape, and it was concluded that five sections was optimum. In applications where a steeper wave front is essential, however, a larger number of sections are necessary.

In the type "E" network the values of the inductances in the three center sections are very nearly equal. Also the values of the various mutual inductances are closely the same. For ease in manufacture it would be desirable to make them precisely equal, and this can be done without appreciably altering the pulse shape, by the following means. A continuous solenoid is wound having a total inductance of, Glasoe (2):

$$L_N = \tfrac{1}{2} Z_o \delta \qquad (11)$$

The storage capacity, C_N, is equally divided, and each condenser is tapped to the solenoid. These taps are so located that the center inductances are all equal, while the end inductances are from 20% to 30% greater. The ℓ/d ratio of the coil is such as to give a mutual inductance between adjacent sections of a value equal to 15% of the self-inductance of a center section.

Variations in design of a network, in order to change the pulse shape, would ordinarily involve going through all the mathematical labor of starting with an a-c wave, applying the Fourier series analysis, then from the network obtained getting the first Cauer form, and from that finally arriving at a type "E" circuit. Such a procedure may be very laborious. In addition, such effects as stray capacitance, lead inductance, non-linearity of the load, and dissipative losses in the network, are not accountable in the purely mathematical theory. As a result, variations in design are almost invariably achieved by experimental means.

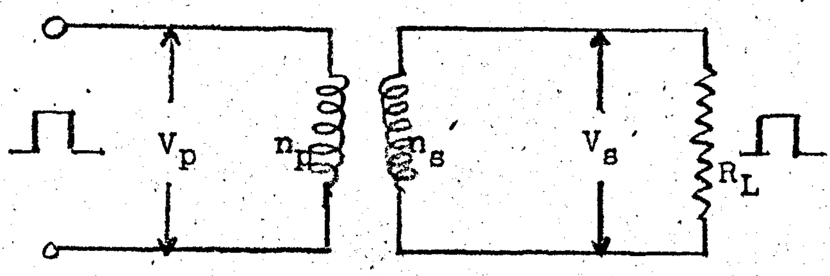
CHAPTER V

THE LOAD

1. The Pulse Transformer

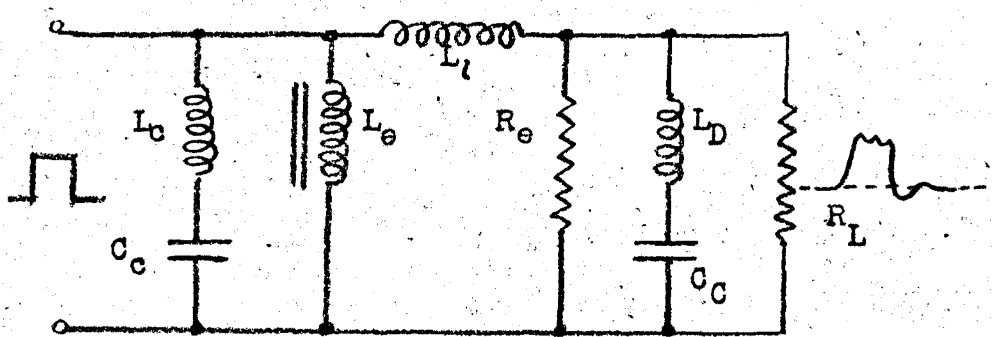
Since the pulse is generated in a PFN having, almost invariably, an impedance level of 50 ohms, and since the static impedance of the magnetron during oscillation is in the vicinity of 1000 ohms, some impedance matching device is necessary. The pulse transformer nicely performs this function, at the same time stepping up the pulse voltage and thereby making possible the use of a lower voltage network.

Although it is possible to conceive of an ideal transformer, such as shown in Figure 16(a), it is not possible in practice to construct one that will pass undistorted all the frequency components in a square pulse. In Figure 16(b) the equivalent circuit of a practical pulse transformer is given, together with an explanation of the symbols used. Since the shunt inductance, L_e, constitutes a load on the source, good design procedure dictates that it be made as large as possible. At the same time, the leakage inductance must be minimized, and as L_e is increased, this becomes increasingly difficult to accomplish. As a result, pulse transformers differ from more conventional types in the method of construction. Primary and secondary windings are wound as close together as possible on the same core. The core itself is fabricated of very thin laminations of special materials. In order to reproduce



(a) Ideal Transformer

$$\frac{V_s}{V_p}=\frac{n_s}{n_p}\ ;\quad \frac{I_s}{I_p}=\frac{n_p}{n_s}\ ;\quad \frac{R_L}{Z_o}=\left(\frac{n_s}{n_p}\right)^2$$



(b) Equivalent circuit of a pulse transformer

L_l = leakage inductance due to flux from primary current which fails to link with secondary

C_D = effect of electrostatic energy stored in the primary-secondary distributed capacity

L_D = "squirted inductance" arising from non-uniform current distribution due the charging of C_D

C_C = effect of electrostatic energy stored between primary and core.

L_C = effect of magnetic energy stored in "squirted flux" which comes from non-uniform current in primary coil arising from the charging of C_C

L_e = effective shunt or self-inductance

R_e = resistance due to eddy current in iron and to hysteresis

Fig. 16

a rapid-rising pulse, these special materials must possess a very high permeability at frequencies ranging from a few cycles up to several megacycles. The effect of the series resonant circuits, L_c-C_c and L_D-C_D, is to introduce oscillations along the top of the pulse, and possibly after the pulse itself has passed as well. Since even relatively small variations at the magnetron input can cause faulty operation, these top-of-the-pulse oscillations must be minimized by careful transformer design. Small oscillations subsequent to the main pulse might cause the magnetron to oscillate weakly. Even very low r-f output would block the sensitive receivers used in radar applications, thereby swamping out echos returning from nearby objects. For this reason a damping resistor or diode is sometimes placed across the pulse transformer secondary.

Proper impedance matching is vital, for if the PFN does not see a load equal to its own charadteristic impedance, a charge, either positive or negative, may remain on the network after the switch opens. (See Figure 11(b)&(c)). This effect can be cumulative over several charging cycles, and eventually break down the switch at an undesired time, or cause other faulty operation.

It has been experimentally determined that in a pulse transformer having an efficiency η_t, and a turns ratio n, the following relations will hold:

$$V_s = nV_p$$

$$I_s = \frac{\eta_t}{n} I_p \qquad (1)$$

$$Z_s = \frac{n^2}{\eta_t} Z_p$$

that is, the voltage transformation is conserved, and the losses appear as shunt losses affecting the current and impedance ratios. Pulse transformer efficiencies run from 75% to 90%.

In general, it may be stated that the pulse transformer must be especially designed for each application. Its function is to match source and load impedance, and at the same time preserve the pulse shape.

When used in line-type modulators, pulse transformers are commonly wound with bifilar secondaries. It is formed by laying two insulated wires side by side, so that the same secondary voltage will be induced in each one. The magnetron heater curreht is supplied from a filament transformer through the bifilar windings; in this way no high-voltage insulation is required in the filament transformer.

2. The Magnetron

Although the oscillator is not a part of the pulser proper, certain inherent characteristics of the magnetron affect the design of the pulser circuits preceding it.

The magnetron is a non-linear load; it has a fairly high static impedance and a low dynamic impedance. The static impedance determines the power drawn from the pulser, and is given by the ratio V_m/I_m at the magnetron operating point. The dynamic impedance $\Delta V_m/\Delta I_m$ is the slope of the V_m-I_m curve at this point. Since this ratio is small, a small voltage drop may decrease the current to such an extent that oscillations cease. For this reason it is important that the modulator furnish a pulse with low top variation (Figure 1).

The rise time of the supplied voltage pulse is of consiiderable importance. If the rate of rise is too fast, the oscillations in the magnetron may fail to build up at all. If the rise rate is too slow, the oscillations may build up in an undesired mode, or else the phenomenon of "mode-skipping" may occur. In this latter case the magnetron oscillations will jump erratically from one mode to another. Mode-shifting, wherein the operating mode changes during the pulse, may also occur, but little can be done in pulser design to correct for this.

Magnetron "sparking" is another aspect to consider in pulser design. All magnetrons will spark occasionally, particularly when new. This is a gaseous discharge inside the magnetron, and is usually a result of occluded gases inside the metal being released by heat. Sparking magnetrons present virtually a short circuit to the modulator, and care must be taken in the design, by the use of protective devices if necessary, to insure against failure of the circuit.

Magnetron plate current variation, in addition to distortion of the r-f output pulse and possible cessation of oscillations, causes frequency modulation of the output. A small change in anode current can quite possibly change the frequency of the r-f output to such an extent that it will be largely outside the bandwidth of the receiver, and hence a loss of received signal strength will result. This is another reason why the pulser must supply a fairly flat-topped pulse.

Finally, the allowable plate dissipation of the magnetron must be considered. This will determine the maximum duty ratio of the pulser, hence the maximum PRF for a desired pulse width. In general, duty ratios in excess of .0025 are not used in pulsed radar, and for output powers in excess of 100 KW, the maximum duty ratio is usually considered to be .001.

BIBLIOGRAPHY

1. Fundingsland, O.T., and Anna Walter. Analysis of line modulator behavior with a sparking magnetron load. Radiation laboratory report no. 765, August 10, 1945.

2. Glasoe, G.N., and J.V. Lebacqz. Pulse generators. New York, McGraw-Hill, 1948. (Massachusetts Institute of Technology. Radiation Laboratory series. No. 5).

3. Guillemin, E.A. Communication networks, vol. II. New York, Wiley, 1935.

4. Krulikoski, S.J. Technical data and operating notes for the 5C22 hydrogen thyratron. Radiation laboratory report no. 828, November 14, 1945.

5. Millman, J., and S. Seely, Electronics. New York McGraw-Hill, 1941.

6. Ridenour, Louis N. Radar system engineering. New York, McGraw-Hill, 1947. (Massachusetts Institute of Technology. Radiation laboratory series.No.1).

7. Staff, MIT Radar School. Principles of radar New York, McGraw-Hill, 1946.