THE DESIGN AND DEVELOPMENT OF AN OSCILLOGRAPH FOR HIGH VOLTAGE IMPULSE TESTING

A Thesis

Presented to

the Faculty of Graduate Studies and Research

of the University of Manitoba

In Partial Fulfillment

of the Requirements for the Degree

Master of Science in

Electrical Engineering

by

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May 1961

ABSTRACT

This thesis describes the design and development of an oscillograph for use in a high voltage impulse testing system. The oscillograph differs from the conventional low voltage type in that no amplification of the signal is required. Wherever possible, the pros and cons of several designs are discussed and the selected design is presented in detail. The final overall design consists of three five inch cathode-ray tubes. A separate time-base generator which has a sweep speed from 1 to 250 microseconds is provided for each tube. These three channels are supplied by common power supplies. The vertical and horizontal scale calibrations are obtained from a separate calibration power supply and a marker generator respectively. A short introduction is included in which are discussed the fundamentals of impulse testing.

PREFACE

The subject of this thesis is the design of an oscillograph to be used in a high voltage impulse testing system. Chapter I, the introduction, is included to provide a background on impulse testing. Chapter II describes a proposed system to trigger the oscillograph time-base in synchronism with the tripping of the impulse generator. In the third chapter, the design of the oscillograph is discussed in detail. The marker generator and power supplies are presented in the fourth and fifth chapters respectively. The power supplies are generally much heavier than would be used in commercial equipment; this was merely a matter of economy and convenience in using existing available components. Since the final installation will not be portable, weight is of little consequence. In chapter VI various oscillograms are presented which simulate the waveforms generated by an impulse generator. The manufacturer's data on the cathode-ray tube used in the oscillograph is given in appendix A. Appendix B describes the postdeflection acceleration principle and appendix C describes the use of quartz crystals in oscillator circuits. The final appendix, D, gives a partial analysis of the sweep circuit and also includes oscillograms of the sweep and unblanking pulse voltages.

The author wishes to acknowledge the assistance of Professor J. P. C. McMath in the design of this oscillograph. Appreciation is also extended to the University of Manitoba for its financial assistance in this project.

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* - Indicates an Oscillogram

* Indicates an Oscillogram

CHAPTER I

INTRODUCTION

1.1 HISTORY

This oscillograph was designed for use in conjunction with a high voltage impulse generator in the Fetherstonhaugh High Voltage Laboratory at the University of Manitoba. Construction of the impulse generator commenced in 1959 and was still in progress at the time of this writing. The designed generator consisted of twenty-four stages with each stage being charged to 100 kv thus theoretically generating an impulse of 2,400 kv. The initial objective however, was the completion of twelve stages. The impulse generated by these twelve stages is estimated to be 1,000 kv allowing for the loss in amplitude caused by the inherent wave shaping resistors.

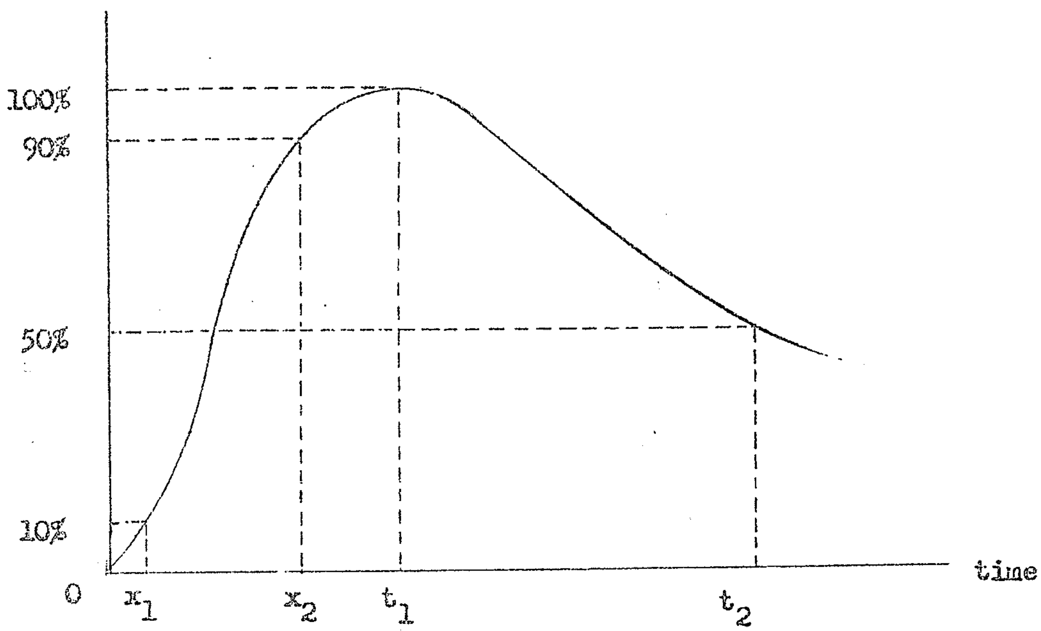
Impulse voltage tests are very important in the investigation of the electrical strength of electrical apparatus, especially power equipment such as transformers. The tests are designed to simulate lightning surges which may arise on transmission lines or associated equipment.

1.2 IMPULSE VOLTAGE TESTING SYSTEM

Very briefly, the high voltage impulse testing system consists of an impulse generator, a voltage divider, a high-speed oscillograph, a test object and triggering circuits. The voltage divider is required to produce a low voltage replica of the impulse waveform to be applied to the vertical deflection system of the oscillograph. This voltage divider must have a high frequency response so as to reproduce the impulse waveform without an undue amount of distortion. The oscillograph, a definite requirement in any impulse testing system, makes possible the permanent accurate recording of the transient response of the test object to the impulse waveform. The low voltage waveform obtained from the voltage divider (usually of the order of a few hundred or thousand volts) is delivered to the oscillograph by means of a length of co-axial cable. Unless this cable is essentially lossless and is properly terminated, errors will occur in the recording of the waveform due to the attenuation of the waveform and reflections from the unmatched termination. Usually a length of radio-frequency cable is used and this limits the errors introduced to a very small value at frequencies below 50 mc which is sufficient for all practical purposes. The length of the cable should be no longer than necessary. The triggering circuits which are required to trip the impulse generator and start the time-base of the oscillograph will be discussed in chapter II. A block diagram which shows the general layout of the system is shown in figure 1.1.

1.3 STANDARD IMPULSE WAVEFORMS

The standard impulse waveform used for all testing in North America is a 1.5x40 microsecond full wave. The general shape of the wave is shown in figure 1.2. The actual polarity of the waveform generated by the impulse generator is usually negative, however, some authors present the oscillogram in such a manner as to have a positive waveform. This may be accomplished by connecting the signal-carrying-lead to the oscillograph vertical deflection plates so that a positive signal deflects the beam downward and vice versa. The waveform rises to peak value in 1.5 usec and decays to half-peak-value in 40 usec. The other standard impulse waveform commonly used in North America is the previous 1.5x40 usec waveform which has been "chopped" to zero value near the crest of the waveform. This second waveform is commonly called a "chopped wave".



$t_1 = 1.5(x_2 - x_1) = 1.5$ usec

$t_2 = 40$ usec

FIGURE 1.2

STANDARD 1.5x40 WAVEFORM

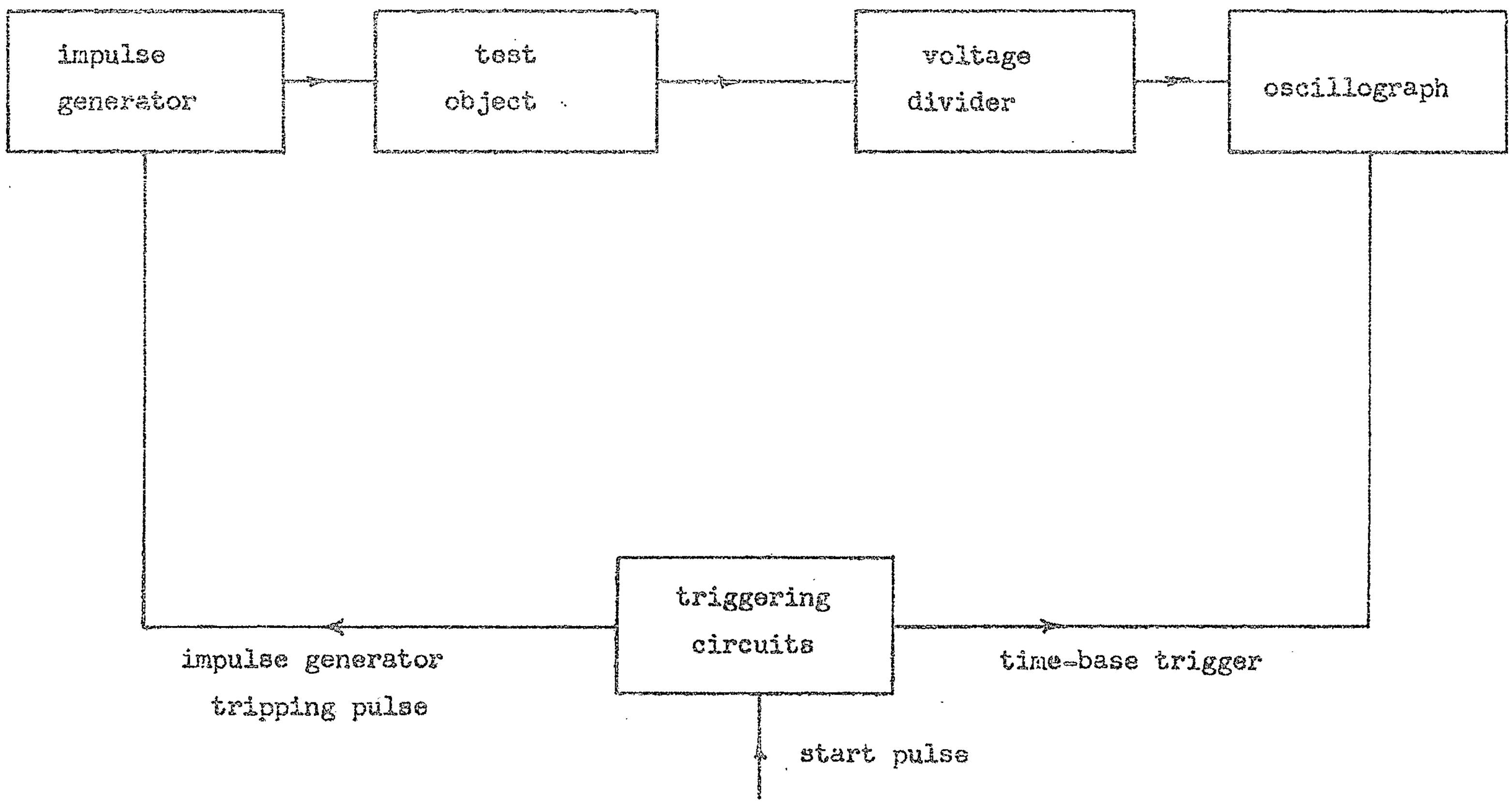


FIGURE 1.1

BLOCK DIAGRAM OF AN IMPULSE TESTING SYSTEM

1.4 TRANSFORMER IMPULSE TESTING

The procedure for the testing of power transformers is presented briefly so as to indicate the general range of sweep speeds required in the design of the oscillograph.

The following describes the basic impulse test for transformers; a series of impulses consisting of a reduced full wave, two "chopped" waves, and a full wave are applied in succession to each accessible terminal of each winding with the other terminal usually grounded. The response of a transformer whose windings are known to be sound is obtained with the reduced full wave. The "chopped" waves are meant to stress the windings with a rapidly changing voltage and the full wave is intended to detect any failures caused by the previous "chopped" waves.

Oscillograms of the voltage across the transformer winding form one method of failure detection. This method, however, fails when only a small percentage of the winding is damaged by the impulse. When a small percentage of the transformer coil fails, a current oscillogram detects the failure much better than a voltage oscillogram. There are many methods of obtaining current oscillograms; however, only the neutral current method will be described. An explanation of the neutral current method will serve to illustrate the requirements of the sweep generator of the oscillograph.

The arrangement for obtaining current oscillograms by the neutral current method is shown in figure 1.3. When an impulse is applied to one terminal of a transformer, it travels through the coil and the shape of the impulse is altered as it travels along the coil. Small oscillatory circuits composed of a few turns of wire and their stray capacitances are excited near the terminal where the impulse wavefront is steepest. These oscillations, plus any "noise" on the wavefront, travel through the series capacitance to appear at the shunt resistor as a burst of "hash", a few microseconds in duration, at the beginning of the current oscillogram. The series capacitance between windings is shown in the equivalent circuit of a single phase transformer of figure 1.4. The burst of "hash" is followed by a period of low frequency disturbances lasting about 60 microseconds which are caused by the mutual inductance between the last few turns and the rest of the winding. The slowest disturbance is caused by the build-up and decay of flux in the magnetic circuit and it may last several hundred microseconds. The last component of neutral current is caused by the rise in current occuring each time that the voltage wave is reflected from the grounded end. The four components of the neutral current are; high-frequency oscillation, low-frequency disturbances, inductive current, and reflections. It is obvious that different sweep speeds are necessary for the neutral current test; a fast sweep for the high-frequency indication and a slower sweep for the inductive-current indication.

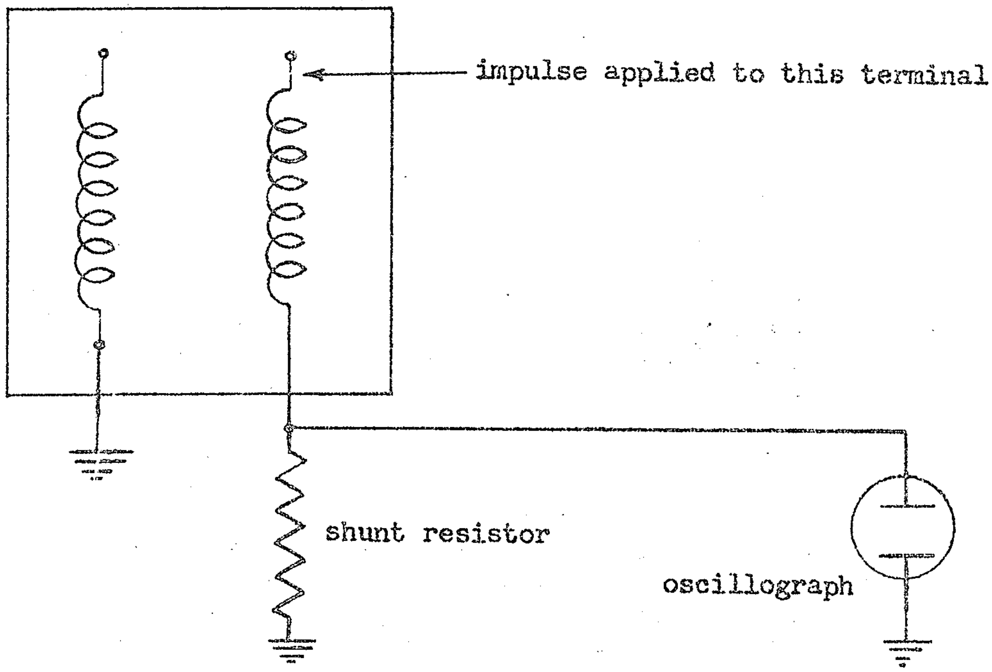


FIGURE 1.3

NEUTRAL CURRENT METHOD OF FAILURE DETECTION

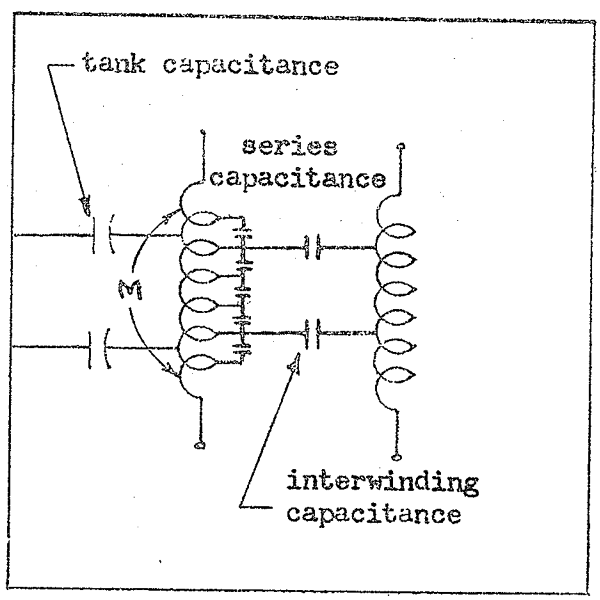


FIGURE 1.4

EQUIVALENT SCHEMATIC OF A SINGLE PHASE TRANSFORMER

CHAPTER II

PROPOSED TRIGGERING SYSTEM

2.1 INTRODUCTION

This chapter describes the tripping of the impulse generator in synchronism with the triggering of the time-base of the oscillograph. The triggering system was not designed in detail however, a description of a possible method is presented.

2.2 TRIGGERING METHODS

It is always advantageous to have the time-base of the oscillograph start just previous to the arrival of the impulse to the vertical deflection plates. The two recognized methods of achieving this control are designated as "controlled" and "uncontrolled" generator tripping. In the former method, the oscillograph time-base is triggered first and this causes a pulse to be applied to the first gap of the generator. A short delay is introduced between the initiation of the time-base and the application of the pulse to the generator gap. This enables the initial changes in the impulse wave to be clearly recorded since the sweep has traversed a small part of the screen before the arrival of the impulse. In the latter method, the impulse generator is tripped first and the oscillograph time-base is initiated by means of a pick-up aerial from the electrical disturbance so produced. In order to have the time-base start before the arrival of the impulse, a length of co-axial cable is introduced between the oscillograph and the test object. This cable acts to delay the impulse to the deflection plates.

The simplest method of tripping the impulse generator consists of setting the lowest spark-gap to flash over at a predetermined charging voltage. The stages are then charged until this flash-over occurs. The disadvantages of this scheme are that the accuracy of the charging voltage is subject to the vagaries of the first spark gap, and that the instant of the discharge is not accurately predictable[1].

An electrical tripping system overcomes the disadvantages of the preceding tripping method. This method employs a three-electrode gap as the first spark gap. The center electrode is grounded through a large resistance and the charging voltage is applied between the two outer electrodes which are adjusted to withstand this voltage by a small ~~minimum~~ margin. The discharge is effected by a pulse applied to the central electrode.

2.3 PROPOSED TRIGGERING SYSTEM

The block diagram of the suggested system is shown in figure 2.1. It is suggested that two variable delay circuits be used to provide the required delay rather than either of the two preceding methods. In this proposed system, the impulse generator is manually tripped by

[1] F. S. Edwards, A. S. Husbands, and F. R. Perry. "The Development and Design of High-Voltage Impulse Generators", The Proceedings of the Institution of Electrical Engineers, vol. 98, part 1, May 1951, p. 155.

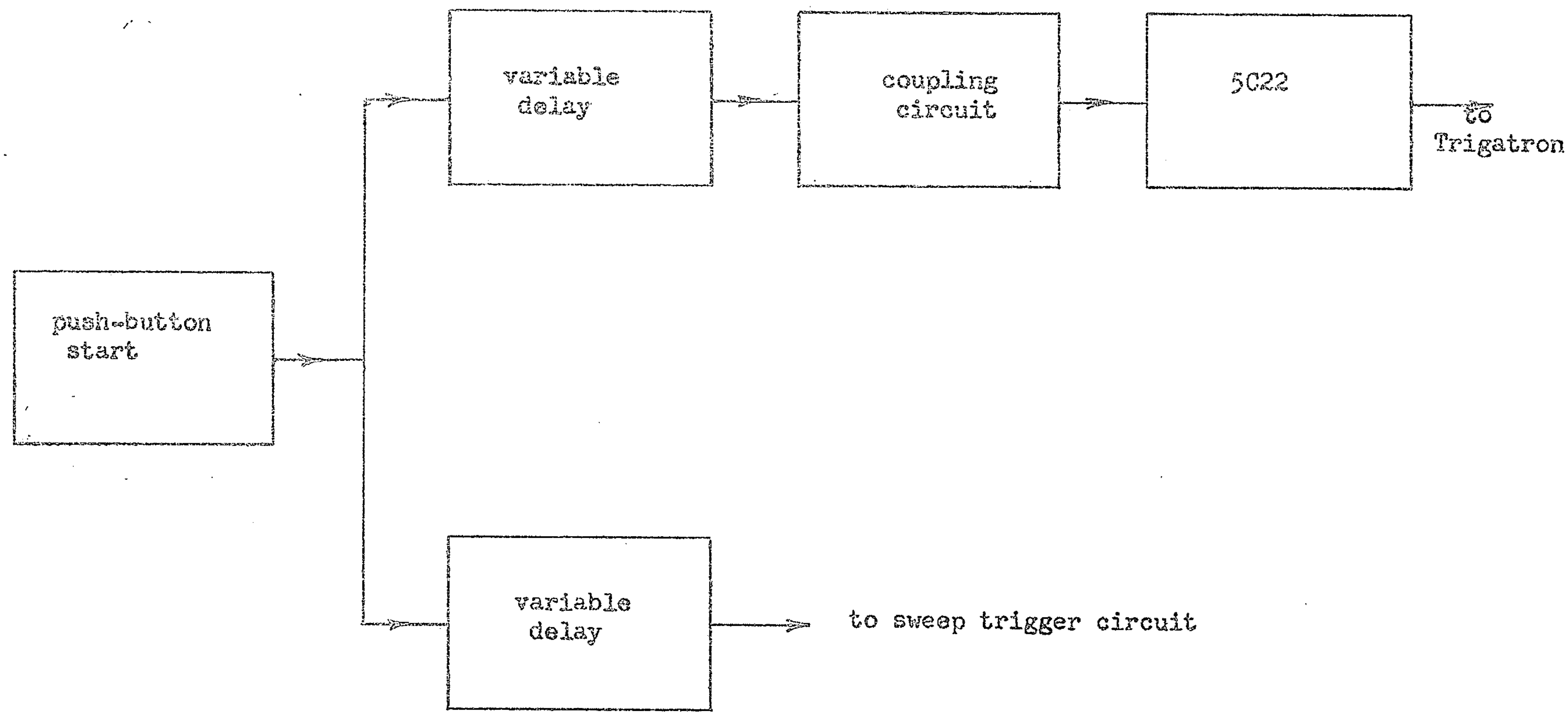


FIGURE 2.1

PROPOSED TRIGGERING SYSTEM

a push button. This push button circuitry generates a positive pulse which is fed into two variable delay circuits. The output of one delay triggers the time-base directly while the output from the second delay is used as a basis for generating a high voltage pulse to trigger the impulse generator. With this method it would be possible to produce any desired delay between the start of the time-base and the impulse waveform. This is particularly advantageous when dealing with transients of various durations. A standard 1.5x40 usec full wave would require a delay of about 1 usec between the start of the sweep and the waveform whereas a "chopped" waveform would require a delay of approximately 0.2 usec. In this particular application a fixed delay would prove a definite disadvantage. One method of obtaining a variable time delay is with the use of a tapped delay line. With the tapped delay line it is possible to obtain a number of different fixed delays. A continuously variable delay may be obtained with either a phantastron type of circuit or a monostable multivibrator. By using two variable delays it is possible to generate a short as well as a long delay thus increasing the flexibility of this type of system over the system with a fixed delay.

A much improved version of the three-electrode gap method of tripping the impulse generator is called the "Trigatron"[2]. Figure 2.2 illustrates a "Trigatron" gap. The high voltage electrode is a sphere and the grounded hemisphere has a hole cut in it into which projects a

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[2] J. D. Craggs and J. M. Meek. <u>High Voltage Laboratory Technique</u> (London: Butterworths Scientific Publications, 1954) pp. 147 and 180.

metal rod. A glass tube is fitted over the rod electrode and is surrounded by a metal foil which is connected to the grounded hemisphere. A trigger pulse is applied to the rod electrode, causing the main gap to break down at a voltage appreciably lower than that required to cause its break-down in the absence of the tripping pulse. The function of the glass tube is to promote corona discharge round the rod as this causes photoionization in the annular gap and the main gap and consequently facilitates their rapid break-down. The "Trigatron" gap is suggested as the best means of tripping the impulse generator.

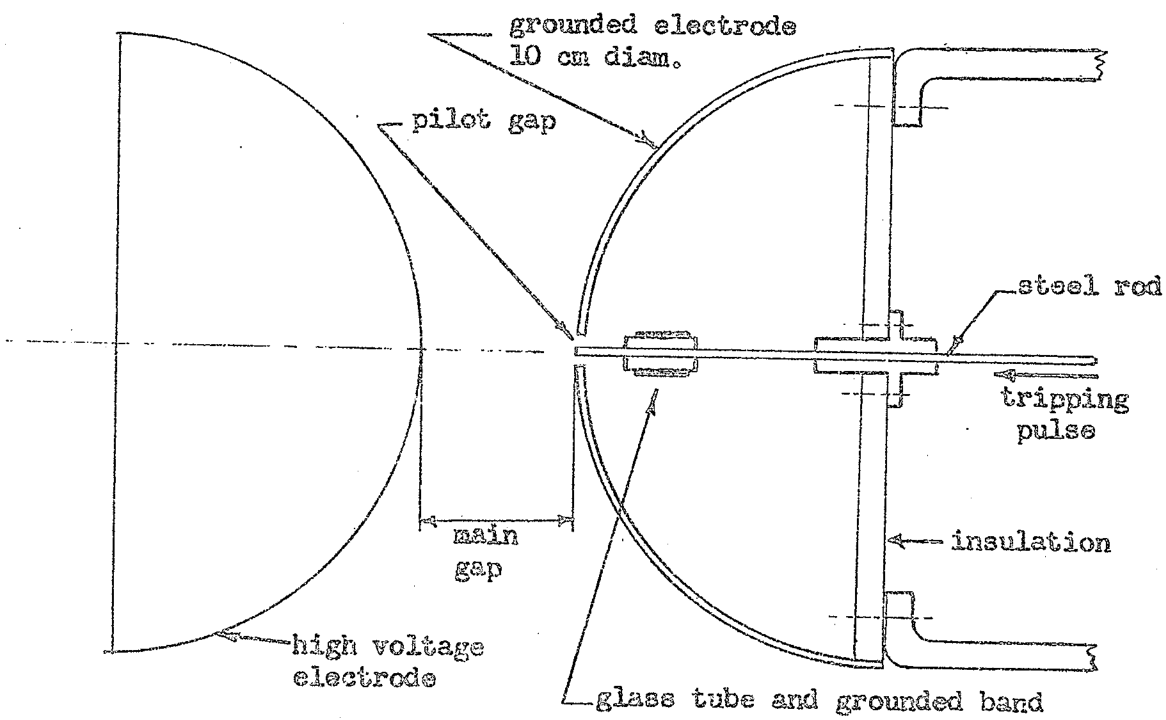


FIGURE 2.2

TRIGGERED GAP ASSEMBLY

A minimum trigger pulse of 5 kv is required by the "Trigatron" to successfully trip the impulse generator. A suggested method of generating this pulse from the low voltage output pulse of the delay circuit is shown in figure 2.3. The low voltage pulse from the delay circuit is fed into a pretriggering or coupling stage which generates a positive 300 volt pulse. This stage is identical in operation to the sweep triggering circuit shown in figure 3.5. The 300 volt pulse is used to trigger the 5C22 hydrogen thyratron. This thyratron generates a positive 8 kv pulse across a 50 ohm resistor with a duration of 5 usec and is fed to the trigger electrode of the "Trigatron". The diode prevents the 8 kv pulse at the grid of the 5C22 from being fed back into the control circuits.

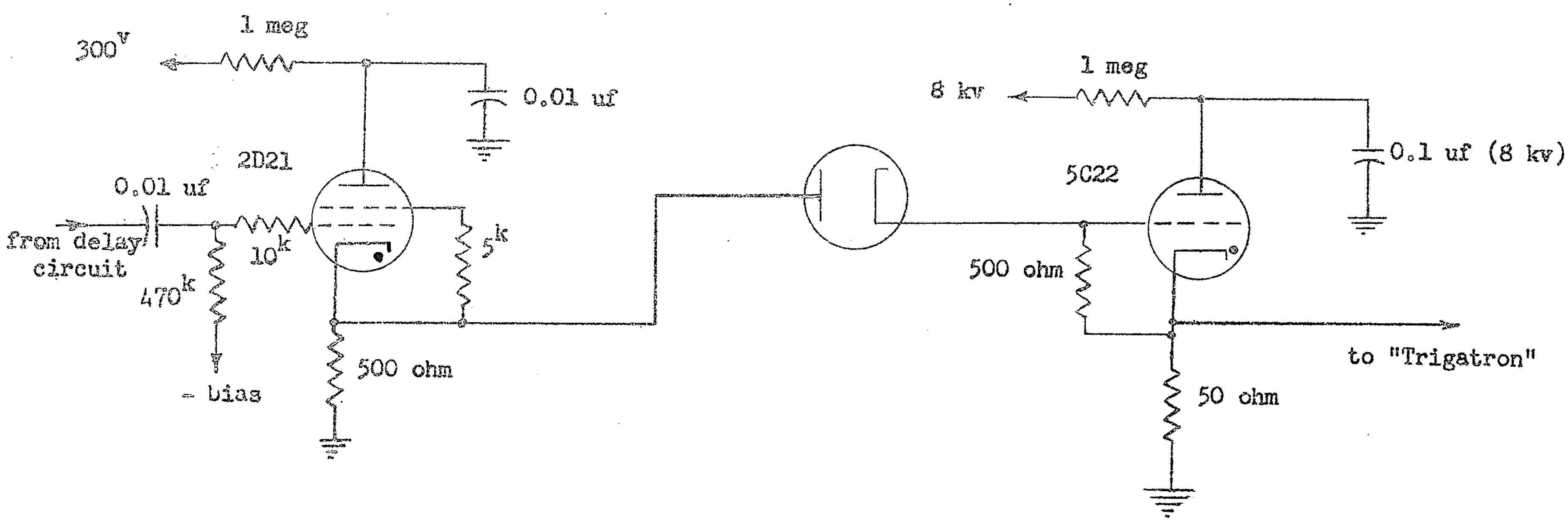


FIGURE 2.3

"TRIGATRON" TRIGGER GENERATOR

CHAPTER III

GENERAL OSCILLOGRAPH CIRCUITRY

3.1 INTRODUCTION

This chapter describes the circuits which are directly associated with the oscillograph such as; the time-base, unblanking pulse, positioning, and cathode-ray tube potential divider circuits. The general operation of the oscillograph is shown with the aid of the block diagram of figure 3.1.

3.2 THE CATHODE-RAY TUBE

A DuMont type K-1068 cathode-ray tube, which was designed for high writing speeds, was chosen as the basis for this oscillograph. The manufacturer's tube data is given in Appendix A and the post-deflection acceleration principle is described in detail in Appendix B.

3.3 CATHODE-RAY TUBE POTENTIAL DIVIDER

This potential divider supplies the accelerating potential to the second anode as well as potentials for the focus and bias electrodes. The circuit was designed to satisfy the tube manufacturer's requirements (see Appendix A). The 3,000 volt accelerating potential between the second anode and the cathode is larger than the recommended design value (2,000 volts) however, it is still below the maximum allowable value (3,500 volts). The schematic circuit diagram of this divider is shown in figure 3.2. The tube V_{29} (6AL5) forms a clamping

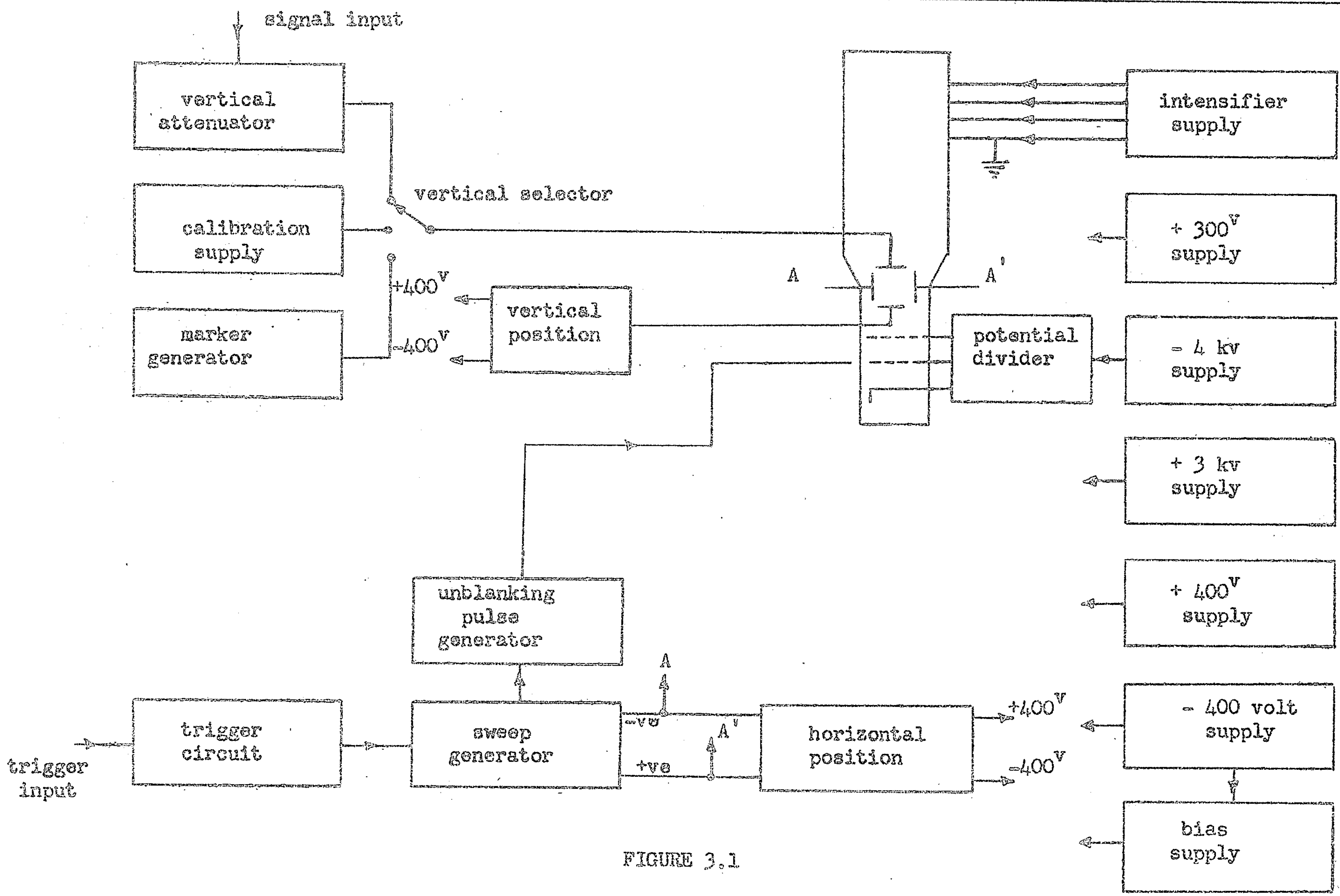


FIGURE 3.1

BLOCK DIAGRAM OF THE OSCILLOGRAPH

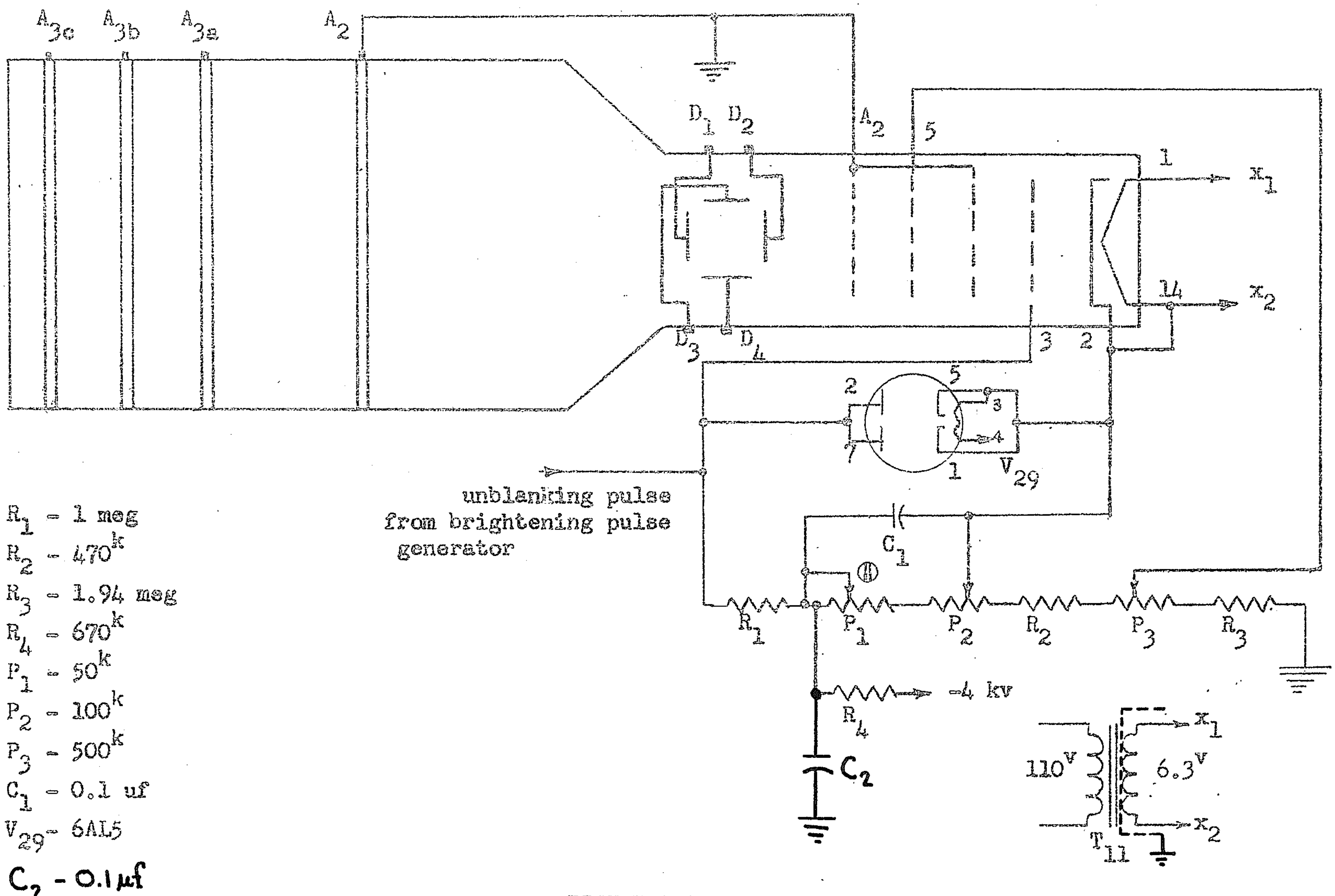


FIGURE 3.2

CATHODE-RAY TUBE POTENTIAL DIVIDER

circuit which limits the positive potential of the control grid with respect to the cathode to zero volts dc. Without this protection, an unblanking pulse, whose magnitude is larger than the bias level set with the intensity adjustment P_2, would drive the grid positive. Severe burning of the cathode-ray tube screen results when the tube is operated with the control grid at a potential above that of the cathode. The resistor R_4 acts as a dropping resistor which reduces the -4,000 volt potential to -3,150 volts. It should be noted that the transformer T_{11} must be insulated to operate at 4,000 volts dc.

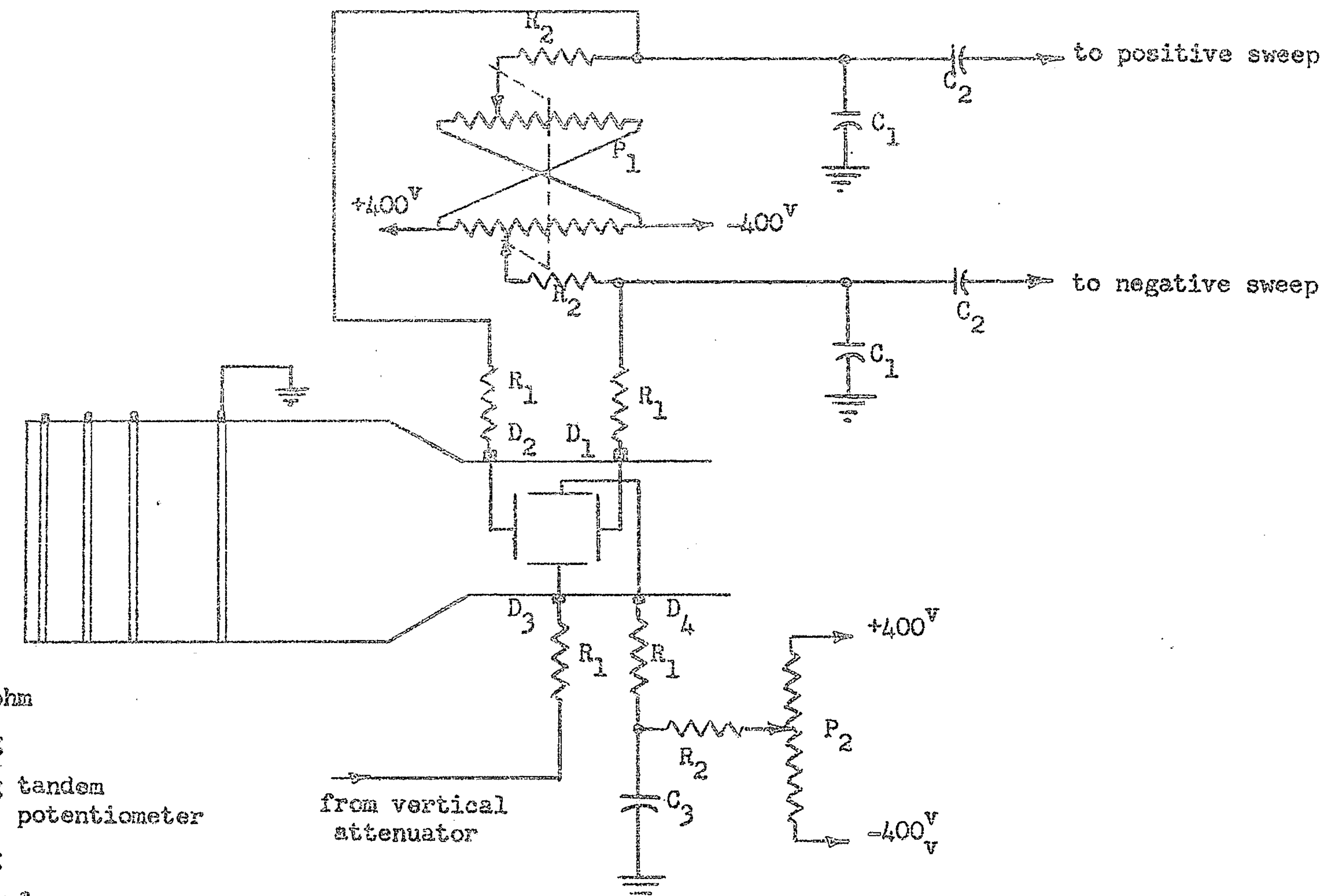
3.4 DEFLECTION PLATE CIRCUITRY

The general circuitry associated with the deflection plates is shown in figures 3.3 and 3.4. This includes the vertical and horizontal positioning potentials, a vertical signal attenuator, and switching arrangements.

Balanced horizontal positioning potentials are required in order to keep the average potential between the deflection plates close to zero. These equal but opposite horizontal positioning potentials are applied to the horizontal deflection plates from the wiper arms of the tandem potentiometer, P_1. For vertical deflection, one plate is of necessity grounded because the transient to be displayed is generated with respect to ground potential. The vertical positioning potential is obtained from the potentiometer P_2. The resistors R_1 in figure 3.3 are placed as close to the deflection plates as possible in order to damp out any high frequency oscillations which may occur due to the tuned circuit formed by shunt capacitance and lead inductance.

A beam position indicator is a valuable component in any oscillograph. One possible method of obtaining such an indicator is with the use of a simple high impedance vacuum tube voltmeter circuit. The potential at the wiper arm of the positioning potentiometer is measured with this meter and the scale calibrated to read the beam position rather than the potential of the wiper arm. The one disadvantage of this scheme is that the meter scale calibration is true for only a fixed value of intensifier voltage since the deflection sensitivity is dependent on the post-deflection acceleration potential to a certain degree. This indicator is left as a valuable suggestion which should be included if, in practice, the tube is operated with one fixed value of intensifier voltage.

The vertical attenuator was designed to match a 75 ohm co-axial cable which transmits the signal from the voltage divider to the oscillograph. In figure 3.4, 10 - 100% of the input signal, in ten percent steps, is made available by the attenuator which provides the oscillograph with a 75 ohm input impedance thus matching the cable to the oscillograph. This attenuator should be placed as close as possible to the deflection plates so as to minimize the shunt capacitance between the wiring and the ground point.



R_1 - 100 ohm

R_2 - 1 meg

P_1 - 3 meg tandem potentiometer

P_2 - 3 meg

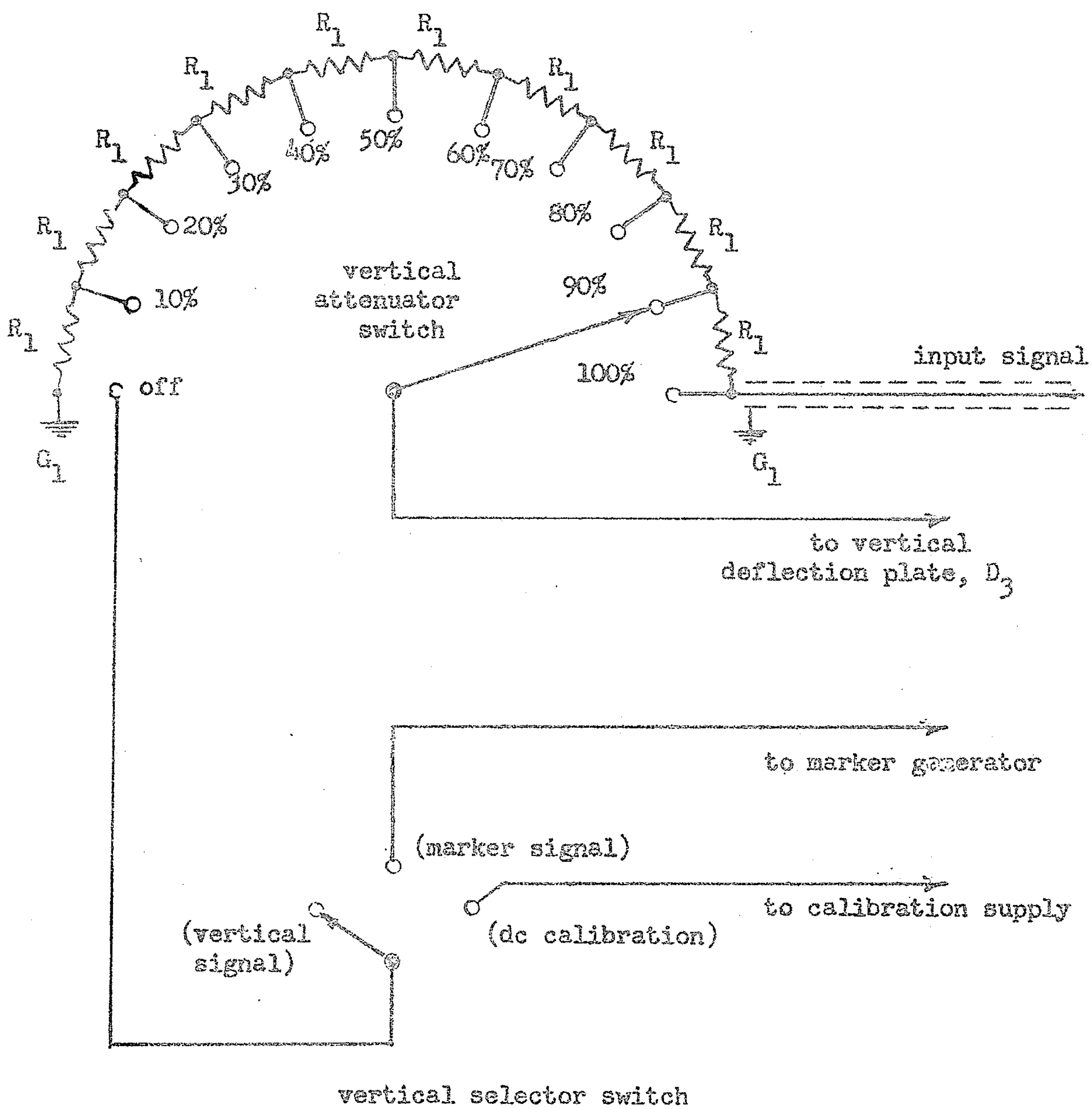
C_1 - 0.25 uf

C_2 - 0.03 uf (1600^V dc)

C_3 - 0.1 uf

FIGURE 3.3

BEAM POSITIONING CIRCUIT



R_1 - 7.5 ohms, 5% (noninductive)

Note: signal grounds return to one point

FIGURE 3.4

VERTICAL DEFLECTION SYSTEM SWITCHING

3.5 SWEEP CIRCUIT

(A) REQUIREMENTS

When examining or recording single transients, it is desirable to have the time-base deflection occur only when a transient signal appears at the vertical deflection system. This single sweep reduces the tendency of a continuous base line to obscure the desired display.

A linear time-base is not essential provided that a calibration signal can be recorded on a subsequent trace. It must be assumed that the sweep speed remains constant from trace to trace for this method of calibration to have any meaning. In practice, a departure from linearity over the trace of up to thirty per cent does not produce an undue amount of distortion and can be tolerated in many applications[1].

Due to the type of cathode-ray tube used, the required magnitude of the time-base waveform is considerably larger than those of the conventional tubes. A low deflection sensitivity is always associated with a large accelerating voltage even with the use of the post-deflection acceleration principle. In most cases this low deflection sensitivity necessitates the use of a wide-band video amplifier for vertical signal amplification; however, in this application, the magnitude of the transient is sufficiently large that an attenuator is required rather than an amplifier. If the waveform of the sweep voltage

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[1] H. A. Prime and P. Ravenhill. "The Design of a High-Speed Oscillograph", Journal of Scientific Instruments, vol. 27, no. 7 July 1950. Page 192.

is a rising exponential, the magnitude should be about 3,000 volts in order to satisfy the low deflection sensitivity requirements and to keep the sweep as linear as possible.

A push-pull deflection system is preferable in high quality recording in order to keep the average potential between the horizontal deflection plates as close to zero as possible. If this potential is not zero, severe astigmatism will result.

The sweep speeds required, range from approximately 1 usec to 250 usec for a 5 inch screen.

In the photography of transient phenomena, some form of beam brightening or beam control is required so that the electron beam is on only for the duration of the transient or some other time which can be predetermined. This beam control prevents the confusion produced by a visible return trace of the beam and is essential to avoid film fogging. To implement this beam control, the sweep circuit should provide a pulse that can be used to generate an unblanking pulse.

(B) <u>SWEEP CIRCUIT</u>

The schematic diagram of the sweep circuit[2] is shown in figure 3.5. When using a push-pull type of deflection, each sweep waveform must be in the order of 1,500 volts in order to keep the time-base as linear as possible. This requirement, and the fact that at the highest possible

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[2]H. A. Prime and P. Ravenhill, *op. cit.*

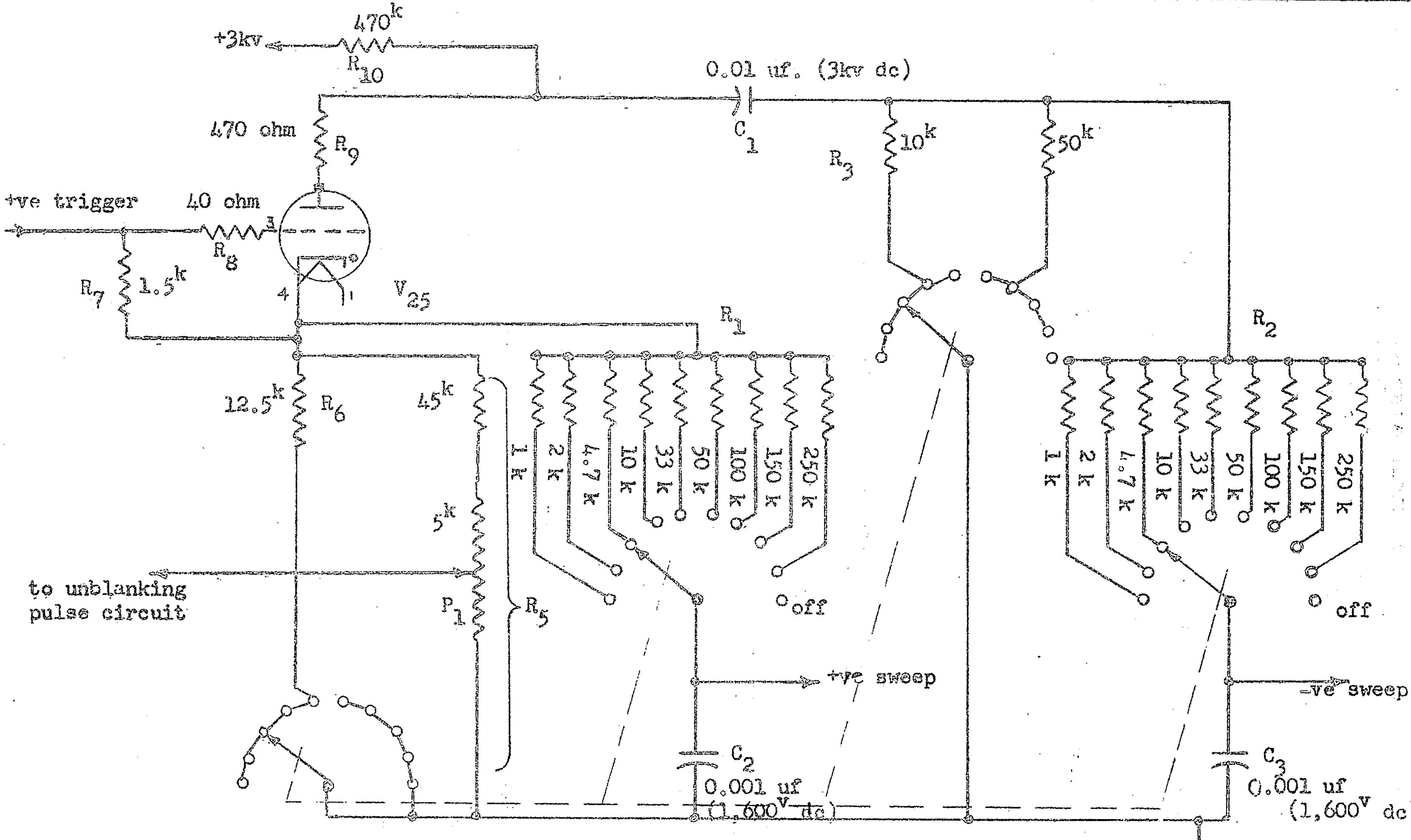

FIGURE 3.5

SWEEP CIRCUIT

sweep speeds, peak currents of more than one ampere are required to charge the stray capacitances associated with the deflecting plates to 1,500 volts in the requisite time, can be met more readily by a thyratron circuit than by a vacuum-tube circuit, because of the higher current rating of the gas-filled tube, other ratings being comparable. A hydrogen thyratron (type 3C45) was used as the time-base switching tube because of its inherent advantages of low ionization time (0.6 usec), the small effect of temperature variations, positive grid control characteristic, and small firing delay. This tube has a deionization time of 25 usec, but this characteristic is of no consequence considering that the sweep circuit is not operated repetitively. The peak current rating of the 3C45 is 35 amps and the maximum plate voltage is 3,000 volts. A minimum trigger pulse of 150 volts across an impedance of 1,500 ohms is required to fire the thyratron. The duration of this pulse should not be less than 4 usec.

In the quiescent state, the capacitor C_1 is charged through R_{10} from the +3 kv supply. When a positive trigger pulse is applied to the grid of the thyratron, it is fired and capacitor C_1 discharged through the hydrogen thyratron V_{25} into the two integrating circuits, C_2R_1 and C_3R_2 connected across R_5 and R_3 respectively. The push-pull time-base voltage waveform is taken across the two capacitors C_2 and C_3 and is exponential. A partial analysis of this sweep generator is given in appendix D. Because of the high charging voltage, that is about +3 kv, only a part of the complete exponential wave is required to sweep the tube. This keeps the departure from linearity below

30 per cent. The resistors R_5 and R_3 discharge C_2 and C_3 during the quiescent period. The ganged switching of the resistors as shown in figure 3.5 enables the time-base speed to be varied, with the values as indicated, over the range of 1 to 250 usec. A fraction of the positive-going pulse generated across R_5 is used to drive a cathode follower which provides an unblanking pulse to the control grid of the cathode-ray tube. The magnitude of this pulse is adjusted by the potentiometer P_1.

The heater of the 3C45 is connected internally to the cathode; thus the transformer must have an insulation rating of at least 1,500 volts. In order that the positive and negative sweep waveform may be well balanced, this transformer must have a low shunt capacitance to ground, that is, in the order of 20 pf. If a low shunt capacitance transformer is not available it would be possible to compensate for this stray capacitance by adding a small capacitor across R_3.

The resistors R_8 and R_9 are grid and plate stoppers and are placed as close as possible to the grid and plate respectively.

3.6 SWEEP TRIGGERING CIRCUIT

The hydrogen thyratron used in the sweep circuit requires a minimum positive trigger voltage of 150 volts which should have a fast rate of rise (in the order of 3 kv/usec) in order to keep the firing delay minimum and to reduce jitter. The sweep triggering circuit shown in figure 3.6 generates the required positive trigger from a low voltage triggering source and also acts to isolate the sweep circuit from any pre-triggering circuitry. It provides a low-impedance trigger pulse which has a rate of rise of approximately 6 kv/usec, an amplitude of 300 volts and a duration of about 5 usec. A low voltage positive trigger of about 10 volts is applied to the circuit through C_1, triggering the shield grid thyratron, V_{27} (2D21). The negative bias controls the amplitude of this positive trigger voltage. In the quiescent state, the capacitor C_3 is charged to +300 volts through the resistor R_4 from the 300 volt supply. When the tube is fired, C_3 is discharged through the tube and the resistor R_3. A positive pulse with a magnitude of 300 volts and a time constant of approximately R_3C_3 is taken across R_3 and applied to the plate of V_{28}. This positive pulse is passed through V_{28} (6AU4) to the grid of the hydrogen thyratron V_{25} (3C45) in the sweep circuit. When V_{25} fires, the grid attains a potential of nearly 1,500 volts; thus the diode, V_{28}, should have a peak inverse voltage rating of approximately this figure. As the cathode of V_{28} is connected to one side of the heater, the filament supply requires to be insulated adequately for the peak-cathode voltage (1,500 volts). A certain portion of the positive pulse at the grid of V_{25} will be fed back through V_{28} due to the cathode to plate capacitance (8.5 pf). The capacitor C_2 connected across the plate of V_{28} shunts this positive pulse to ground. The tube V_{28} (6AU4) was chosen for its peak-inverse voltage rating as well as for its ability to handle large peak currents. The presence of R_5 is a practice recommended by the tube manufacturer which limits any transient screen current to a safe value. The resistor R_2 acts to limit the grid current which is required to charge the grid to cathode capacitance.

3.7 UNBLANKING PULSE CIRCUIT

As stated previously in section 3.5A, it is desirable that the current in the electron beam of the cathode-ray tube have a maximum value only during the actual time of the sweep. This practice prevents damage to the cathode-ray tube screen as well as facilitating the recording of high-speed single transients. This control of the electron beam is achieved in practice by the application of a negative pulse to the cathode or a positive pulse to the control grid, either of which should have a duration equal to that of the time sweep[3]. It is sometimes convenient to obtain such a pulse by differentiating the time sweep voltage but such a procedure may result in the early part of the signal being lost at high sweep speeds (that is less than 1 usec) because of the inherent time delay in the brightening pulse amplifier[4]. Alternately, the time sweep may be triggered from the brightening pulse amplifier[5]. This latter approach may have a serious disadvantage in that an undue delay in the operation of the time-base may result in background illumination with a corresponding reduction in overall contrast.

In the design of this sweep circuit, the former method of obtaining a brightening or unblanking pulse was used primarily because

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[3] D. R. Hardy, B. Jackson, and R. Feinberg. "The Recording of High Speed Single Stroke Electrical Transients", Electronic Engineering, vol. 28, 1956, Jan. p.8 and Feb. p.75.

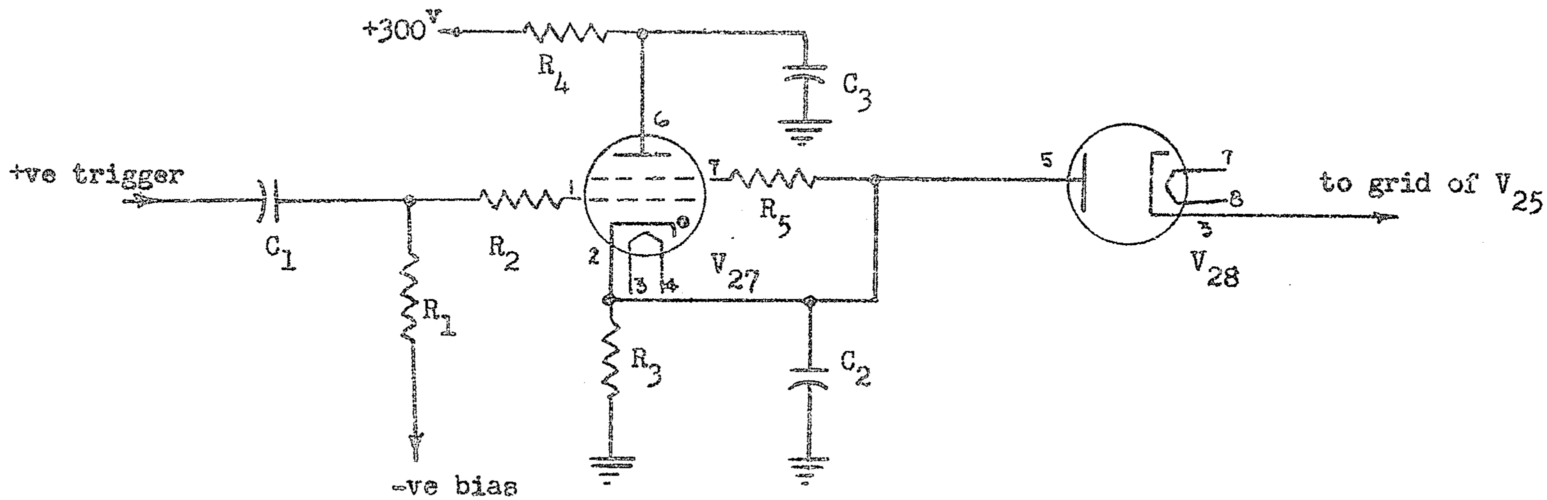
[4] D. R. Hardy. "A High Speed Transient Recorder", Journal of Scientific Instruments, vol. 29, no. 8, Aug. 1952, p. 241.

[5] H. A. Prime and P. Ravenhill, loc. cit.

the fastest sweep speed of interest was 1 usec. A cathode follower, rather than an amplifier, was used as the brightening pulse generator. The use of the cathode follower in place of the more conventional pulse amplifier has the merit of saving approximately 60 ma since it takes virtually no current in the quiescent condition whereas an amplifier would be biased "on". The high input impedance (low input capacitance) of the cathode follower has a negligible loading effect on the sweep generator. The schematic circuit diagram of the unblanking pulse generator is shown in figure 3.7.

The tube V_{26} (6AG5) is a triode-connected pentode which features low input and output capacitances and a relatively high g_m. A fraction of the differentiated positive-going sweep waveform that is generated across R_5 in figure 3.5 is fed into V_{26} via C_1 in figure 3.7. The grid leak resistor R_1 is connected to a tap on the cathode resistor, R_2-R_3; the bias voltage appearing across R_2. The output voltage is taken from the cathode of V_{26} and applied to the control grid of the cathode-ray tube via the blocking capacitor C_2.

In the quiescent state, the cathode-ray tube electron beam is cut off due to the negative bias applied to the control grid by the intensity adjustment. At the start of the sweep, the brightening pulse drives the control grid to zero bias thus unblanking the electron beam. The duration of the brightening pulse is equal to the duration of the sweep waveform. One oscillogram is presented in appendix D which illustrates the relative shape of the brightening pulse to the sweep waveform over the duration of interest and another which shows the overall shape of both waveforms.



V_{27} = 2D21

V_{28} = 6AU4

C_1 = 200 pf

C_2 = 50 pf (3 kv dc)

C_3 = 0.01 uf

R_1, R_4 = 470^k

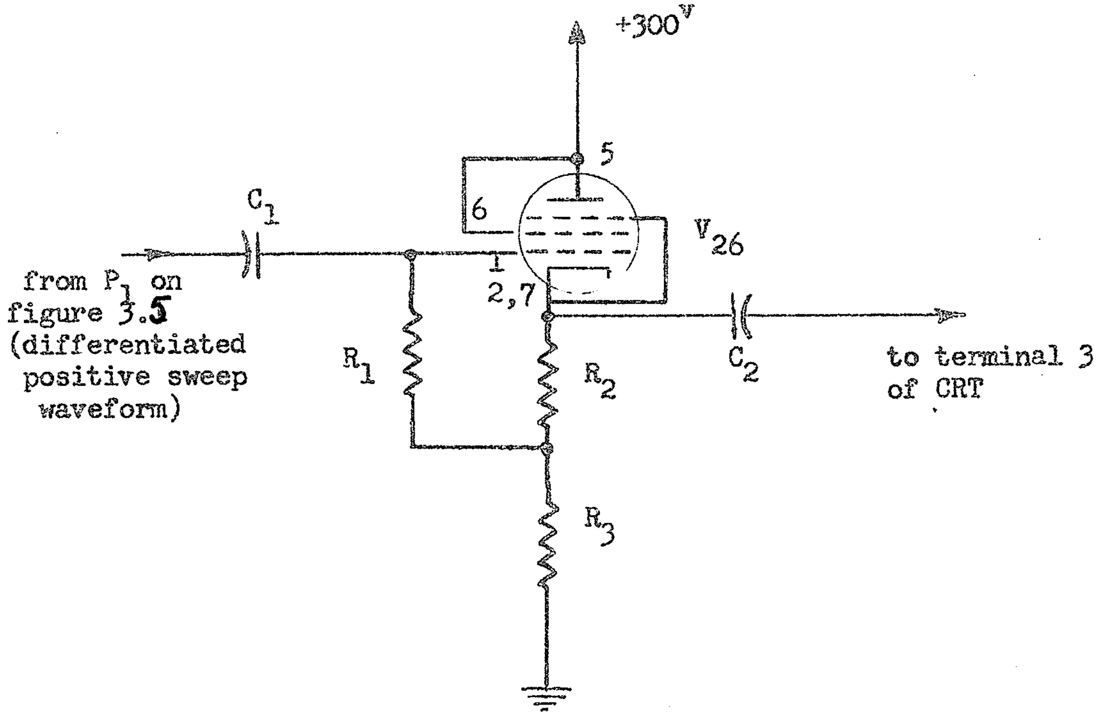
R_2 = 10^k

R_3 = 500 ohm

R_5 = 5^k

FIGURE 3.6

SWEEP TRIGGERING CIRCUIT



V_{26} - 6AG5
R_1 - 470^k
R_2 - 3^k
R_3 - 50^k
C_1 - 0.02 uf
C_2 - 0.01 uf (4^{kv} dc)

FIGURE 3.7

BRIGHTENING PULSE GENERATOR

CHAPTER IV

TIME-BASE CALIBRATION

4.1 INTRODUCTION

This chapter describes the various methods of calibrating an oscillograph time-base and presents the design of the chosen method of calibration.

4.2 METHODS OF CALIBRATION

(A) INTENSITY MODULATION

Intensity modulation or bright-dark modulation is one method of calibrating a time-base and was considered in the design of this oscillograph. Intensity modulation consists of applying a signal of known frequency, usually a series of narrow negative pulses, to the control grid of the cathode-ray tube. The electron beam is cut off for the duration of the negative pulse thus producing a dark spot on the screen of the tube. For sweep speeds in the order of a microsecond, intensity modulation requires a short negative pulse in the order of a millimicrosecond: thus intensity modulation was considered not practical. Further, a pulse repetition frequency, in the order of a megacycle, would be required. A certain loss of detail in the information displayed is inherent in intensity modulation methods. Intensity modulation, as a means of time-base calibration for this oscillograph, was rejected because of foregoing disadvantages. It is most satisfactory when used on a time-base which is nearly linear and is of relatively slow speed.

(B) TIMING SIGNAL

The second means of calibration considered consists of a time-base marker voltage applied directly to the vertical deflection plates. When this method is used in the calibration of a time-base the waveform of interest is recorded on one trace and the timing waveform on a subsequent trace. This method was chosen because of its simplicity and also because there were really no serious disadvantages associated with its use, apart from the fact that the timing waveform does occupy a certain portion of the available screen area. Theoretically, any sweep speed may be calibrated by this method while the intensity modulation method is only applicable for slower sweep speeds.

The waveshape of the calibrating signal is immaterial provided that its frequency is known. In high-voltage impulse testing it is common practice to use a sinusoid as a timing waveform for the oscillograph. It was decided to use a sine wave because the extra circuitry involved to obtain short pulses or "pips" to reduce the screen area used by the marker voltage was considered not to be justified.

4.3 MARKER GENERATOR

(A) REQUIREMENTS

The requirements of the marker generator or time-base calibration voltage generator are as follows: the deflection sensitivity of the cathode-ray tube is about 40 volts dc per inch per kilovolt of predeflection accelerating potential; thus the output from any marker generator must have a peak-to-peak amplitude of at least 100 volts. The frequency range necessary is from 25 kc to 10 mc and the frequencies necessary to cover this range satisfactorily are: 25 kc, 50 kc, 100 kc, 200 kc, 500 kc, 1 mc, 2 mc, 5 mc, and 10 mc. It is important that the time-base calibration be as accurate as possible; thus the frequency stability of the generator must be high.

(B) METHODS

There are many possible means of obtaining a marker voltage. However, only two methods were seriously considered in this design. One method applies the simple ringing circuit which is commonly used as a range measuring circuit in radar sets. The basic circuit is shown in figure 4.1. The triode tube is normally conducting with a steady current I_o flowing through the tube and inductance. At time t_1, a negative gate is applied to the circuit which drives the tube into cut off and causes the tank circuit to oscillate at a frequency of $1/2\pi\sqrt{LC}$ cycles per second. Theoretically, the initial amplitude of oscillation is $2I_o\sqrt{L/C}$ volts peak-to-peak. If a high Q coil is used, an output whose amplitude is fairly constant is obtained, provided that the gate duration is not more than ten times the period of oscillation. At time t_2, the negative gate ceases, the tube conducts, and a new oscillation is started but since the equivalent resistance of the conducting tube shunts the tank circuit, this oscillation is quickly damped.

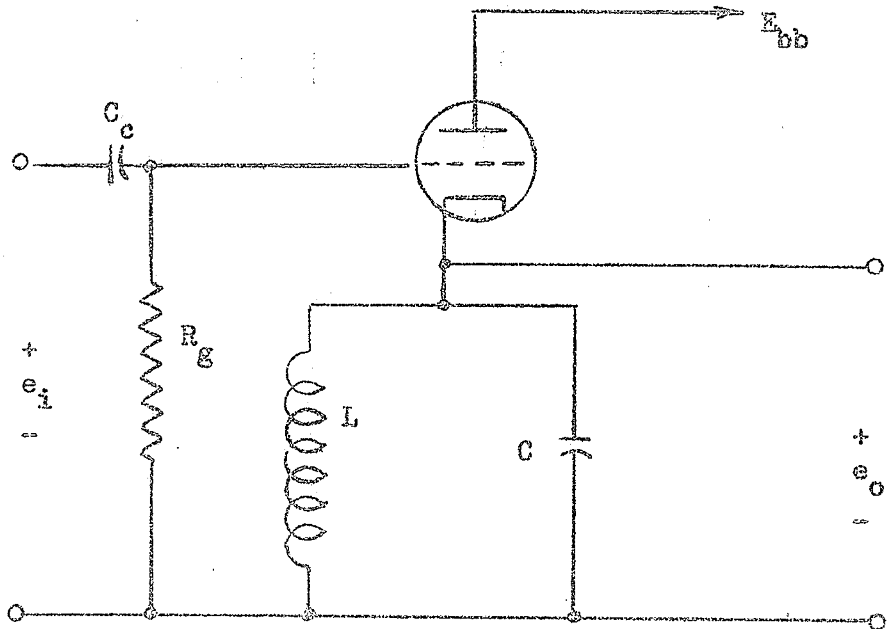


FIGURE 4.1

RINGING CIRCUIT

The one advantage of this circuit is its sinusoidal output whose amplitude satisfies the amplitude specification. However, there are some disadvantages. The frequency of oscillation is only as stable as the elements in the tuned circuit. Another disadvantage is the requirement of a gating generator which is triggered at the beginning of the sweep. The amplitude of this gate waveform must be larger than the peak value of the output voltage of the ringing circuit in order to keep the tube in the cut off region of operation. The decrease in amplitude of the output is another undesirable quality of the circuit. However, a negative resistance oscillator could be used along with the ringing tank circuit to produce oscillations of constant amplitude. The power transferred to the tank circuit from the oscillator offsets the losses in the tank circuit due to the associated resistances and this makes the oscillations continue with undiminished amplitude. Although the basic circuit is fairly simple, this method of marker voltage generation was rejected because of the foregoing disadvantages.

The second marker generator considered uses a standard oscillator as the basic circuit. The standard oscillator has a constant output, requires no gating generator, and, if it is crystal controlled, then the frequency stability is greatly improved (see appendix C). This circuit possesses none of the disadvantages of the ringing circuit. Various harmonics of this oscillator frequency were obtained with frequency multiplying circuits and subharmonics with sinusoidal frequency dividing circuits. Theoretically, it is possible to generate any harmonic or subharmonic of the standard oscillator frequency. The standard oscillator was designed to operate at a frequency within the desired frequency range.

4.4 SINUSOIDAL FREQUENCY MULTIPLIERS AND DIVIDERS

(A) SINUSOIDAL FREQUENCY MULTIPLIER

Frequency multiplying circuits were used in the marker generator in order to obtain frequencies higher than the standard oscillator frequency.

The two essentials of frequency multiplying are; the generation of a waveform which is synchronized with the input and contains the desired harmonic of the input frequency and, the selection of the desired harmonic with a filter. Harmonic generation consists of applying a sine wave to a device with a nonlinear transfer function so as to generate a nonsinusoidal output containing harmonics of the input. Three common harmonic generating devices are diodes, iron-core inductances and nonlinear amplifiers. The class C amplifier is the most flexible and practical of the three for the present purpose.

If a pentode amplifier tube is used with a large bias and a large grid drive, the pulse of plate current is nearly rectangular. If I is the peak value of the assumed square pulse of current starting at $-\alpha$ and ending at $+\alpha$ and if f_o is the input frequency, then the plate current i(t) can be shown by a Fourier Analysis to be

$$i(t) = 2I\left[\frac{\alpha}{2\pi} + \frac{\sin\alpha}{\pi}\cos 2\pi f_o t + \ldots + \frac{\sin n\alpha}{n\pi}\cos 2\pi n f_o t\right]$$

From the previous formula, the current of the n^{th} harmonic is

$$i_n(t) = \frac{2I\sin n\alpha}{n\pi}\cos 2\pi f_o t$$

The value of $i_n(t)$ is maximum when $\sin n\alpha = 1$, thus if α is controlled then the maximum value of any harmonic can be generated. However, the amplitude of the n^{th} harmonic is $\frac{2I}{n\pi}$. This indicates the difficulty involved in generating large outputs of a high harmonic.

The output voltage of the filter of the n^{th} harmonic is

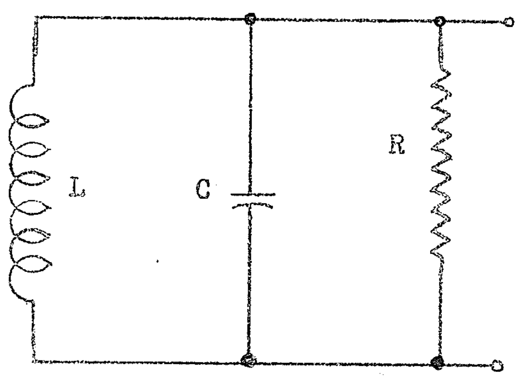
$$v_n(t) = \frac{2I\sin n\alpha}{n\pi}\cos(2\pi f_o t + \text{Arg } Z(nf_o)) \cdot \left| Z(nf_o)\right|$$

Where $Z(nf_o)$ is the transfer function of the filter. The most desirable filter is one having a large impedance at the n^{th} harmonic and zero impedance at all other frequencies.

This requirement can be met by any type of bandpass filter however, the simple LC tuned circuit is the most commonly used filter; in frequency multipliers. This LC filter is used as a load impedance for the class C amplifier as shown in figure 4.3, the impedance being maximum at the n^{th} harmonic. Figure 4.2 illustrates an idealized parallel resonant circuit. It is assumed that the shunt resistance R represents all the losses in the circuit. It can be shown that the impedance of the circuit in figure 4.2 has a maximum value at the resonant frequency, that is when

$$nf_o = f_r = 1/2\pi\sqrt{LC}$$

At this frequency the parallel LC combination offers an infinite impedance to the input and the resulting impedance is just equal to R.



R represents all losses in the circuit

L is the inductance of a high Q coil

C represents the capacitance of the capacitor forming the tank circuit

FIGURE 4.2

IDEALIZED PARALLEL RESONANT CIRCUIT

The amplitude of the impedance at a frequency f in the neighborhood of resonance is given by

$$Z = \frac{R}{\sqrt{1 + Q^2\left(\frac{f}{f_r} - \frac{f_r}{f}\right)^2}}$$

If the only requirement of the output waveform is that it be sinusoidal, then the resonant circuit should have a high value of Q as shown in the above equation.

The basic multiplier circuit is shown in figure 4.3. A high g_m pentode is the most suitable tube for a frequency multiplier circuit. Triodes could be used but their higher cutoff voltage and lower plate resistance make them less efficient as a multiplier tube. Since the tube plate resistance shunts the tank circuit when the tube is conducting, a low plate resistance, such as in a triode, has an adverse effect on the effective Q of the circuit. The triode tube, because of its larger cut off voltage, requires a much larger grid drive than the pentode tube for successful operation. The circuit in figure 4.3 requires an input which is much larger than the cut off voltage of the tube. The bypassed cathode resistor is included as a means of protecting the tube in case the grid drive fails. The screen potential may be used as a means of controlling the output amplitude especially if attenuation is desired. The grid-leak bias is self regulating and consequently, the bias automatically tends to adjust itself to the peak value of the grid drive, thereby eliminating the need for a critical adjustment of the excitation voltage. This method of obtaining grid bias is often used in oscillators and r-f amplifiers. The positive portion of the incoming signal initially raises the grid to a potential which is positive with respect to the cathode, causing grid current to flow which charges the coupling capacitor C_c. The discharge of C_c through R_g is made very nearly constant by making the time constant C_cR_g long compared to the period between the positive peaks of the input waveform. When R_g is made large, sufficient bias will be developed so that grid current will flow only during the most positive peak of the incoming signal.

The LC tank circuit is tuned to the desired harmonic of the input signal and the output is taken from the plate of the multiplier tube.

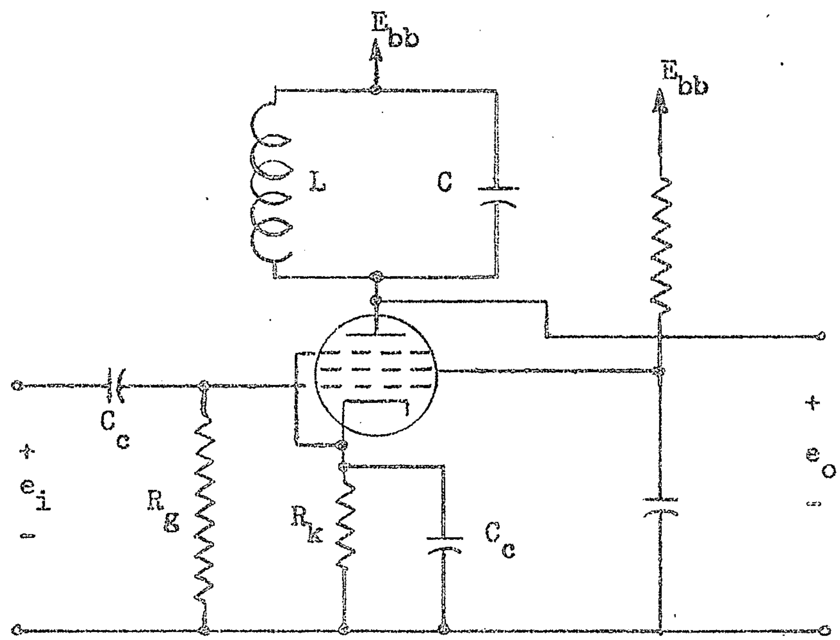


FIGURE 4.3

BASIC MULTIPLIER CIRCUIT

(B) SINUSOIDAL FREQUENCY DIVIDERS

Frequency dividing circuits were used in the marker generator to generate frequencies below the standard oscillator frequency. The frequency divider is analogous to a multiplier in that a synchronized waveform must be generated which is rich in the desired subharmonic. The desired frequency must be selected from the waveform with a filter.

There are many methods of generating subharmonics of a control frequency. However, the more practical methods consist of synchronization of oscillators; the oscillator frequency being the desired frequency and the synchronizing frequency being the control frequency. There are two types of oscillators used—regenerative and relaxation oscillators. The regenerative oscillator has good frequency stability but synchronization sometimes presents problems. The relaxation oscillator is relatively unstable in frequency but is much easier to synchronize. A third type of divider operates regeneratively and also includes modulation and frequency multiplying circuits.

Two types of frequency dividers were considered in the design of the marker generator— a synchronized relaxation oscillator and a divider using regeneration and modulation. The block diagram of the latter method is shown in figure 4.4 and a possible circuit is presented in figure 4.5.

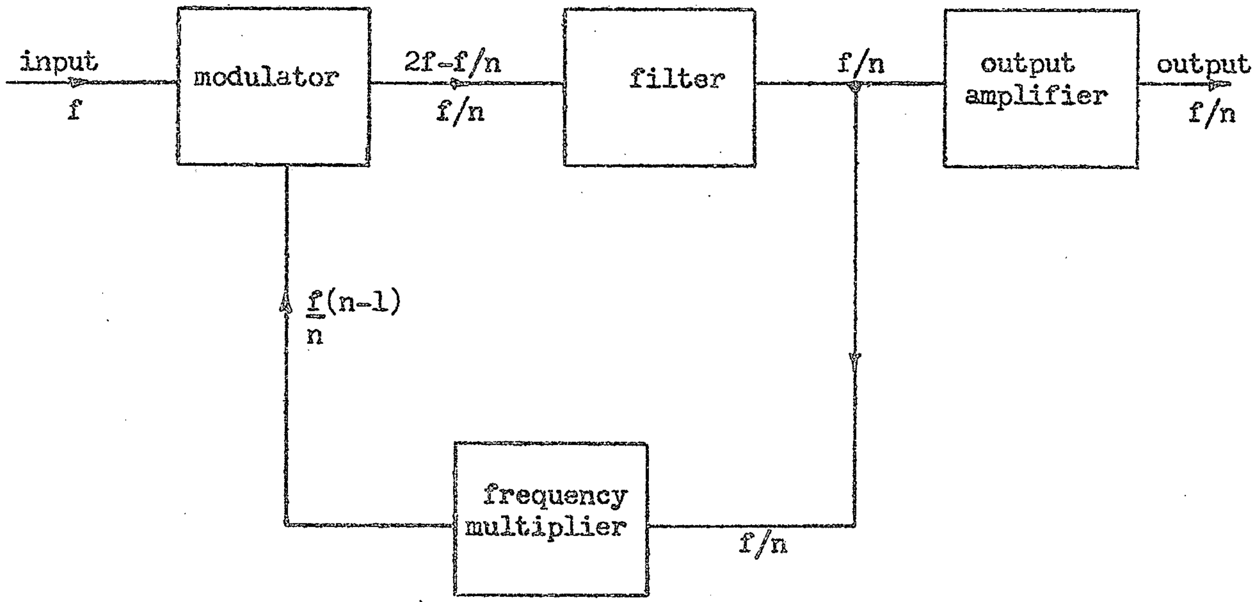


FIGURE 4.4

BLOCK DIAGRAM OF A FREQUENCY DIVIDER

In figure 4.5, the control frequency f is fed into the first grid of the modulator tube whose plate is connected to the L_1C_1 tuned circuit which is tuned to f/n. It is assumed that a frequency f/n exists at the output of L_1C_1. This signal is applied to the output amplifier and also to the grid of the frequency multiplying generator. This harmonic generator multiplies the input frequency f/n by n-1 thus producing an output f-f/n. This frequency, selected by the L_2C_2 tuned circuit, is impressed on the third grid of the modulator tube through C_3. In the modulator tube, intermodulation between this component and the original input frequency produces the sum and difference frequencies, that is 2f-f/n and f/n. The filter L_1C_1 selects the latter frequency which was assumed to exist. This circuit, once started, is self-sustaining.

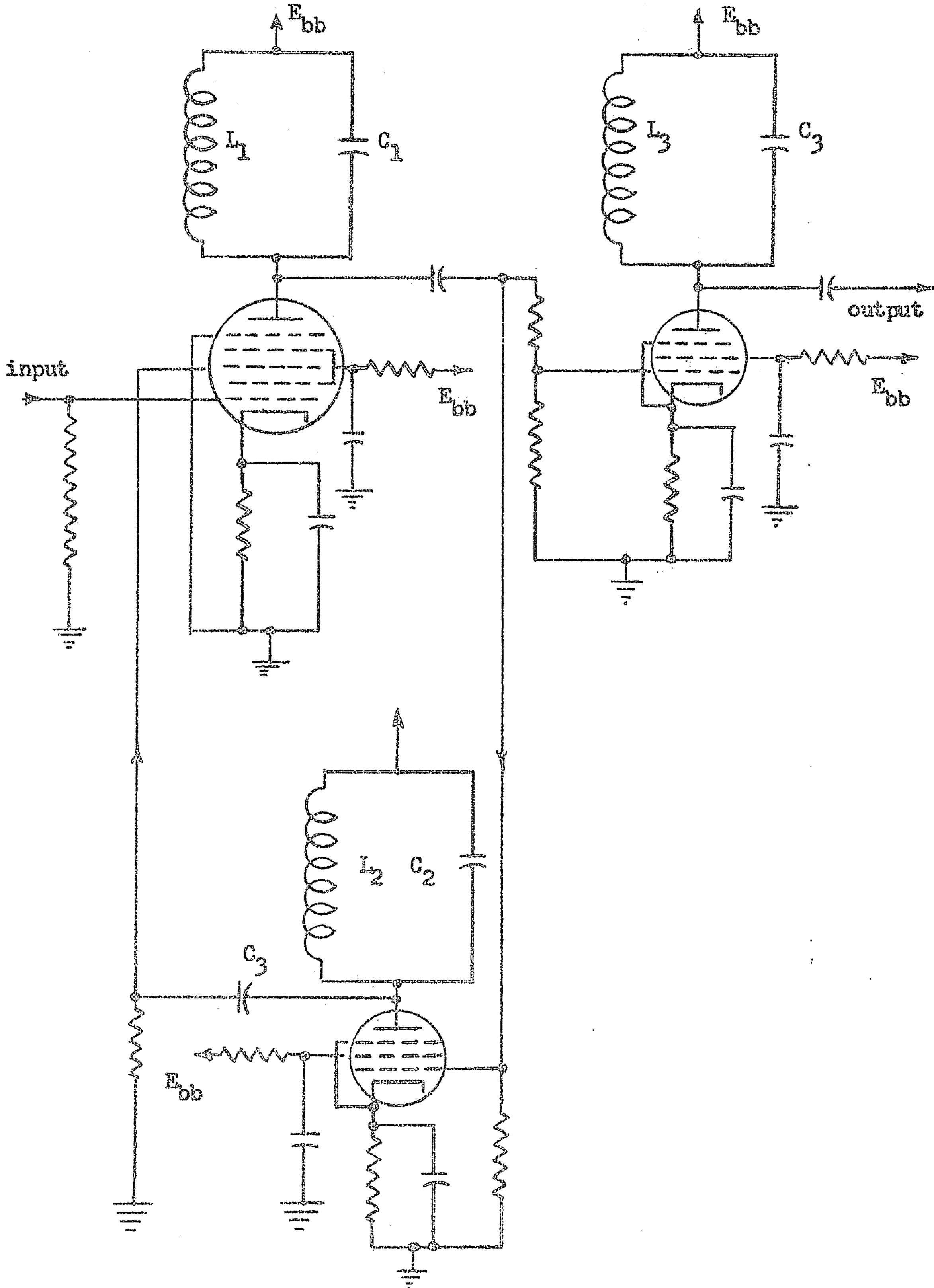


FIGURE 4.5

FREQUENCY DIVIDING CIRCUIT

The second type of frequency divider considered consists of a synchronized relaxation oscillator. The type of relaxation oscillator considered was an astable multivibrator as shown in figure 4.6. The synchronizing signal is applied to the grid of the first tube and the output taken from the plate of the second tube. The output from an astable multivibrator is basically a square wave which contains very small amounts of even harmonics. It is quite a simple matter to select the fundamental frequency of the multivibrator output with a filter. The filter considered was a class C amplifier. The multivibrator output is used as the grid drive and the tank circuit tuned to select the fundamental of the grid excitation.

There are certain disadvantages to the use of relaxation oscillators as a method of frequency division. Relaxation oscillators are limited to about 1 mc at which frequency they are difficult to synchronize. They continue to give an output in the absence of all synchronization. If too large a division is attempted in one step, then the oscillator may get out of synchronism due to tube ageing and changes in supply voltages and parameter values. The only disadvantage of the first type of divider described is its relative complexity. One very good characteristic of the first type is, that in the absence of a synchronizing signal, the output ceases.

The relaxation oscillator type of divider was chosen chiefly because of its simplicity. The division was kept to a factor of two; thus there was no danger of the oscillator getting out of synchronism. The frequencies required from the dividers were below 100 kc. Thus no difficulty was encountered in the multivibrator design. The only remaining disadvantage was the fact that there might be an output even in the absence of the synchronizing signal. However, in this application, this fact is really not a disadvantage because if the synchronizing signal is not available to the divider then the oscillator frequency is also unavailable as a marker waveform.

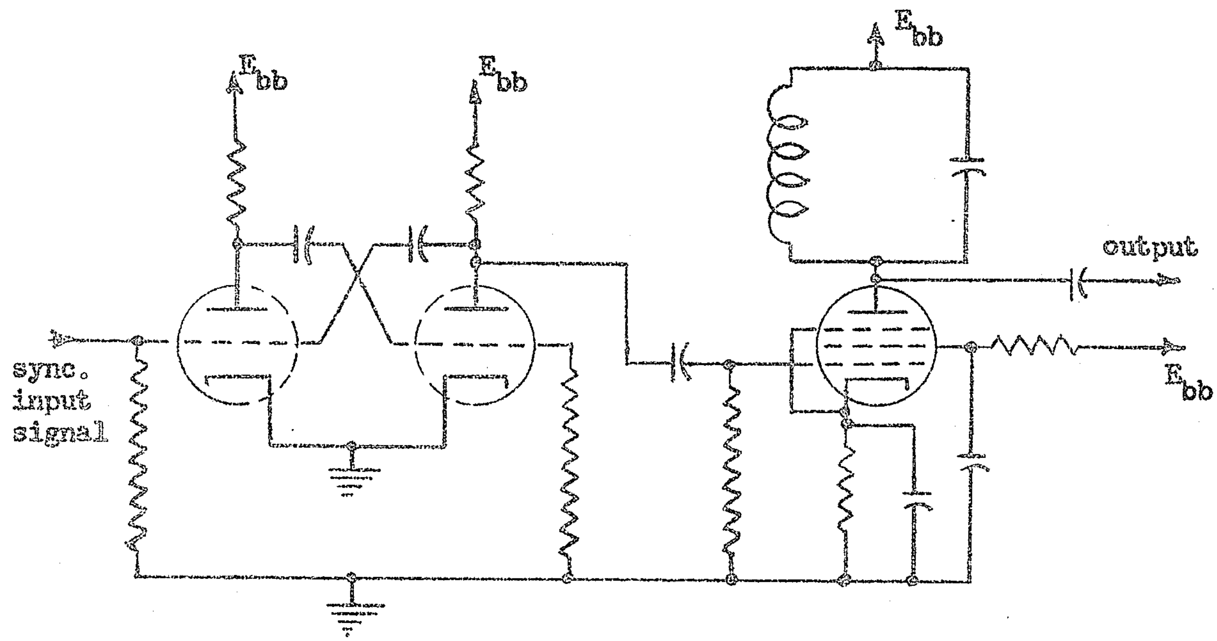


FIGURE 4.6

RELAXATION OSCILLATOR FREQUENCY DIVIDER

4.5 DESCRIPTION OF THE DESIGNED MARKER GENERATOR

The block diagram of the marker generator is shown in figure 4.7. The 100 kc crystal controlled oscillator drives a frequency doubler and a frequency quintupler, thus generating the required frequencies ~~as follow~~; 100 kc, 200 kc, and 500 kc. The 100 kc oscillator also produces the synchronizing signal for the 50 kc astable multivibrator. This multivibrator drives an amplifier which is tuned to 50 kc in order

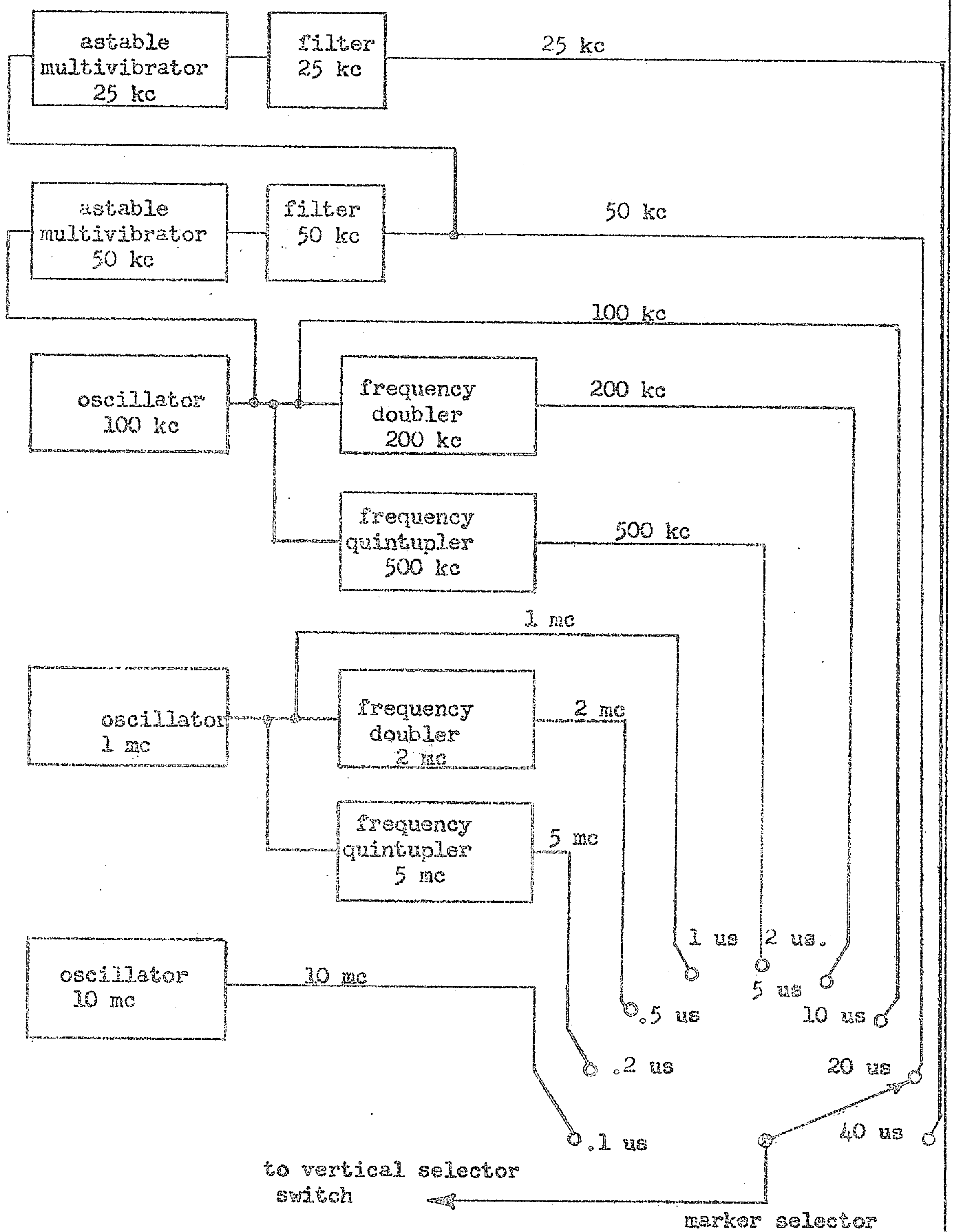


FIGURE 4.7

BLOCK DIAGRAM OF MARKER GENERATOR

to obtain the sinusoidal 50 kc output. This output is used as a synchronizing signal for the 25 kc generator which is similar in operation to the 50 kc generator. The 1 mc, 2 mc, and 5 mc generators are similar in form to the 100 kc, 200 kc, and 500 kc generators respectively. The 10 mc generator is another crystal controlled oscillator.

The circuit diagram of the 100 kc oscillator is shown in figure 4.8. Basically, the circuit is an electron-coupled oscillator which combines an oscillator and power amplifier in one tube envelope. This allows power to be coupled from the power amplifier tank circuit without affecting the frequency stability of the oscillator. This circuit was recommended by the crystal manufacturer for frequencies from 70 to 200 kc[1]. The type of crystal used was an HC6/U meaning hermetically sealed and plated. The crystal is either gold or silver plated, the plating acting as the electrodes. The crystal is spring mounted and sealed under vacuum or the container is filled with nitrogen or helium. The manufacturer states that these crystals should be only used in circuits designed for them as the maximum current rating limits the crystal to 10 milliwatts power dissipation. The capacitance of a crystal controlled oscillator circuit is important since the frequency of a crystal will be lowered when capacitance is added to the circuit. This crystal has been calibrated at 100 kc when operated in a circuit with a capacitance of 32 pf. A recommended circuit was used to preserve the 0.005 per cent tolerance stated by the manufacturer. The only minor change made in this circuit was the addition of a tuned circuit

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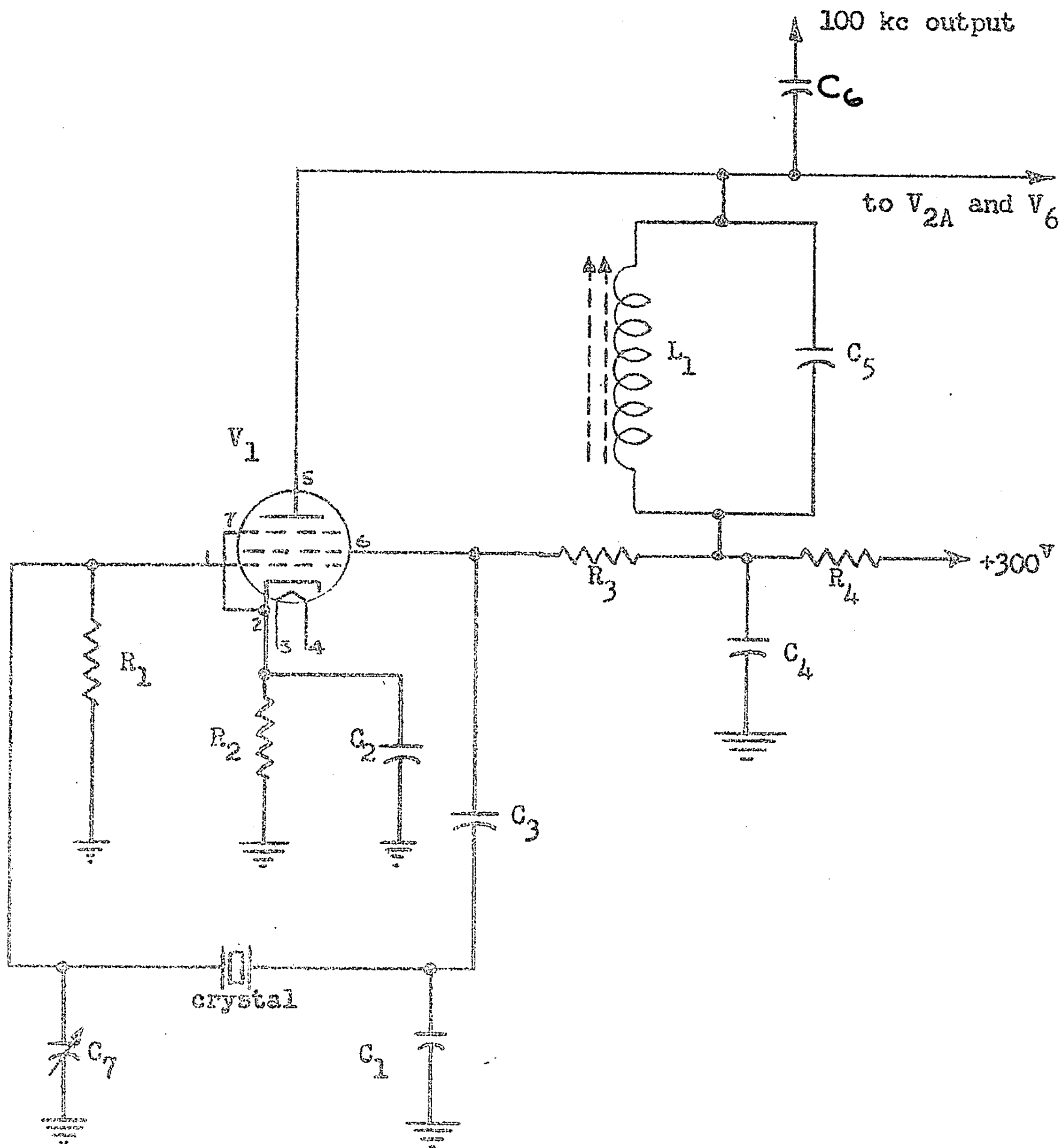
[1]"Texas Crystals" catalog no. 860.



crystal - 100 kc
R_1 - 1 meg
R_2 - 470 ohm
R_3 - 47^k
R_4 - 25^k
C_1 - 100 pf
C_2 - 0.01 uf

C_3 - 0.002 uf
C_4 - 0.01 uf
C_5 - 200 pf
C_6 - 100 pf
C_7 - 50 pf
L_1 - 7-13.5 mh
V_1 - 6BH6

FIGURE 4.8

100 kc CRYSTAL CONTROLLED OSCILLATOR

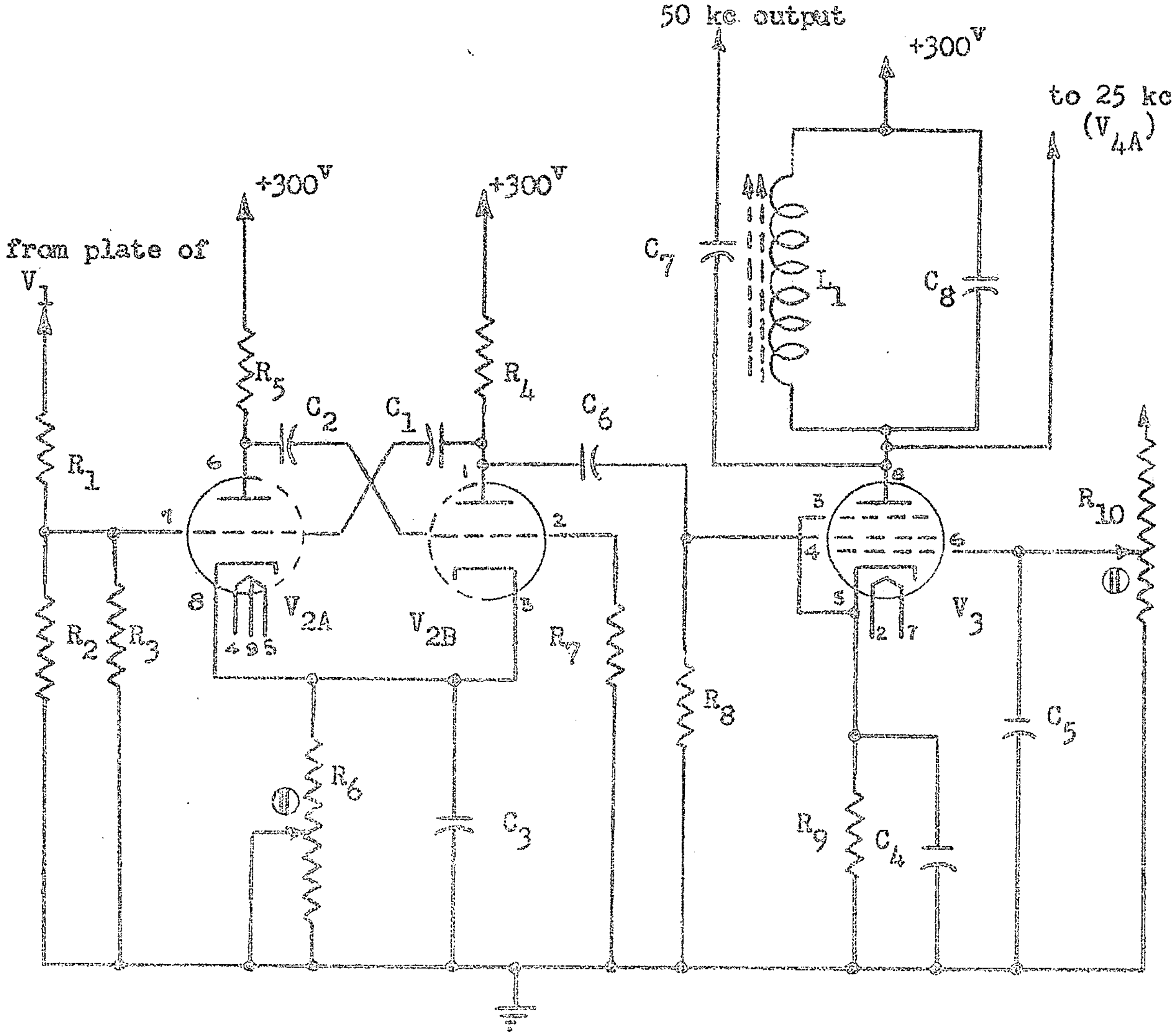
L_1C_5 in place of a 5.6^k resistor in the plate circuit. This change was made to increase the amplitude of the output and to generate a better sine wave. The tank circuit is tuned to 100 kc and the output is taken from the plate of V_1 through C_6.

The circuit diagram of the 50 kc generator is shown in figure 4.9. The waveform generator consists of an astable multivibrator, designed to operate at approximately 50 kc. The resistor R_6 provides for a fine frequency adjustment. A fraction of the output from the plate of V_1 of figure 4.8 is fed to the grid of V_{2A} via R_1 and R_3. This 100 kc sinusoid acts as a synchronizing signal which locks the multivibrator to 50 kc. The square wave output of the multivibrator is taken from the plate of V_{2B} and is fed to the grid of the class C amplifier via C_6. This class C amplifier acts as a filter which selects the fundamental component from the 50 kc square-wave multivibrator output. The output of the 50 kc sinusoidal generator is obtained from the plate of V_3 via C_7. The tank circuit of the class C amplifier is tuned to the fundamental of the square wave.

The circuit diagram of the 25 kc generator is shown in figure 4.10. The operation is identical to that of the 50 kc generator. The synchronizing signal is obtained from the plate of V_3 and is applied to V_{4A} through R_1 and R_3. The multivibrator is designed to operate at approximately 25 kc with R_6 acting as a fine frequency adjustment. The amplifier tuned circuit L_1C_8 is tuned to 25 kc and the sinusoidal 25 kc output is taken from V_5 via C_7.

The 200 kc generator is shown in figure 4.11. This generator consists of a frequency doubling circuit and a cathode follower which separates the 100 kc oscillator from the frequency doubler. The cathode follower isolates the crystal oscillator from load changes in the output circuit thereby insuring optimum operating conditions for the oscillator. The potentiometer P_2 controls the level of the input signal into the cathode follower and P_3 provides a variable biasing arrangement to maintain the grid of V_{6A} above ground. The output from the cathode follower is fed into the multiplier tube via C_4. The multiplier circuit consisting of V_7 is similar to the circuit described previously in section 4.3. The tube is a 6AC7 which has a high value of g_m. The bypassed cathode resistor R_5 acts to limit the tube current in case of failure of the grid drive. The C_2P_1 combination provides the screen potential and P_1 is adjusted to give an output of approximately 100 volts peak-to-peak. The tank circuit L_1C_1 is tuned to the required output frequency, that is 200 kc.

The 500 kc generator is shown in figure 4.12 and is identical in operation to the 200 kc generator. The 1mc crystal controlled oscillator is shown in figure 4.13 and is basically the same as the 100 kc oscillator. This circuit was recommended by the crystal manufacturer for frequencies from 0.8 to 20 mc. The 10 mc oscillator is shown in figure 4.15 and is the same as the 1 mc oscillator with the exception of the plate potential. In order to get an output in the vicinity of 100 volts peak-to-peak the plate supply voltage was increased to 300 volts. The tank circuit of each oscillator circuit is tuned to the desired frequency. The 2 mc and 5 mc generators are shown in figure 4.14. These are identical in operation to the 200 and 500 kc generators. It was found that isolating amplifiers were not required between the 1 mc oscillator and the 2 mc and 5 mc generators.

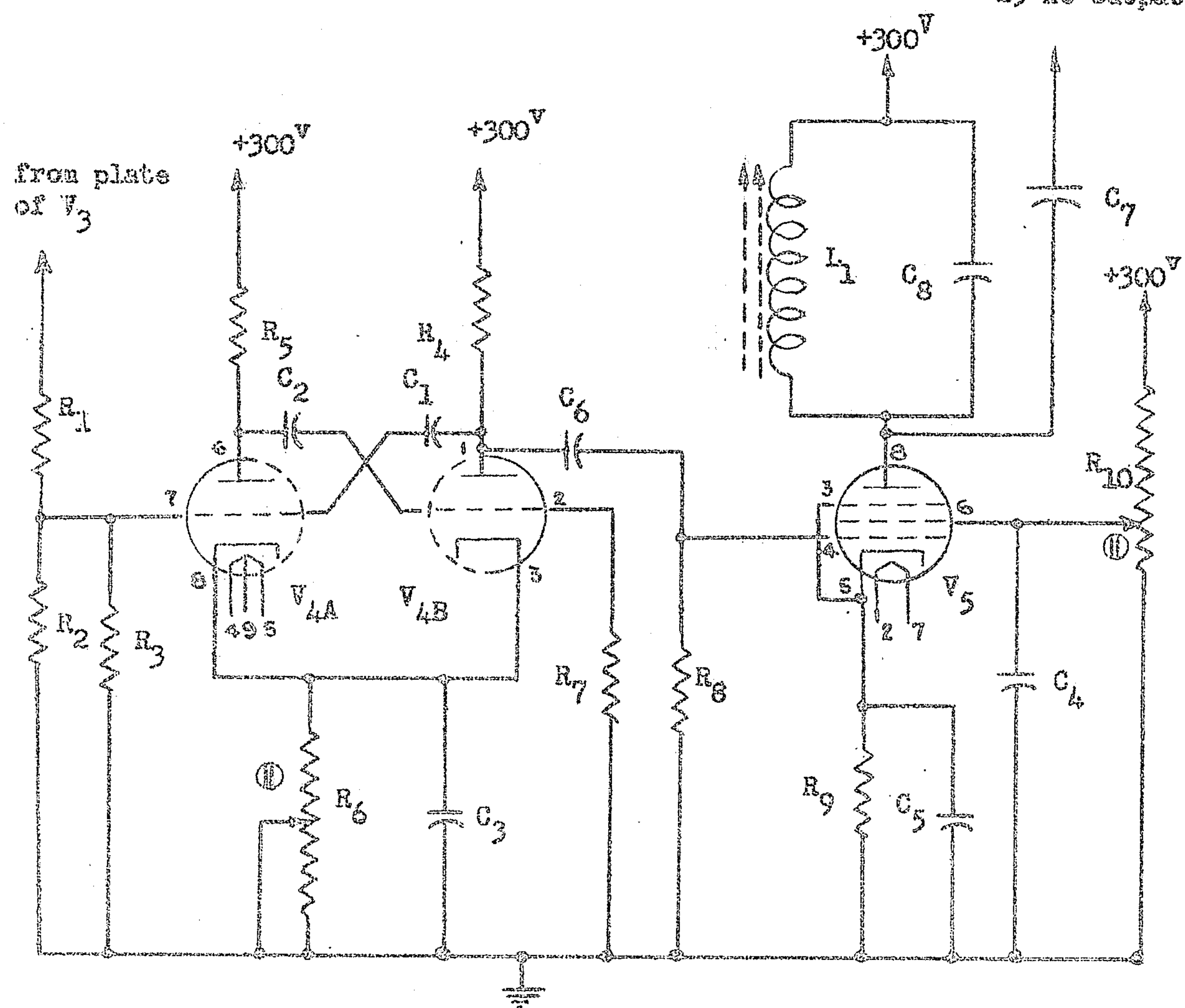


R_1, R_2 – 330^k
R_3, R_7 – 100^k
R_4, R_5 – 50^k
R_6 – 10^k
R_8 – 1 meg
R_9 – 100 ohm
R_{10} – 150^k
C_1, C_2 – 68 pf

C_3 – 0.5 uf
C_4, C_5, C_6 – 0.05 uf
C_7 – 0.01 uf
C_8 – 660 pf
L_1 – 13–25 mh
V_2 – 12AU7
V_3 – 6AC7

FIGURE 4.9

50 kc GENERATOR

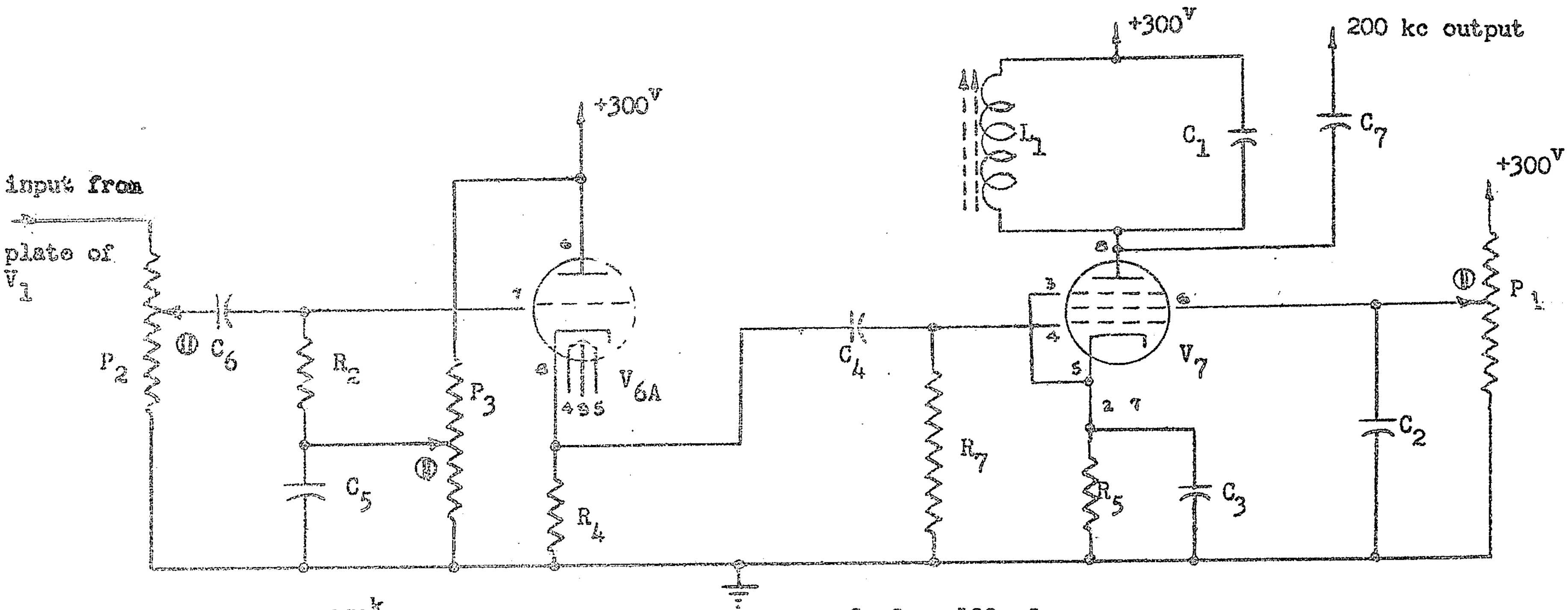


V_4 – 12AU7
V_5 – 6AC7
R_1, R_2 – 330^k
R_3, R_7 – 100^k
R_4, R_5 – 47^k
R_8 – 1 meg
R_9 – 150 ohm
R_{10} – 150^k

C_1, C_2 – 100 pf
C_3, C_5 – 1 uf
C_4, C_6 – 0.05 uf
C_7 – 1000 pf
C_8 – 3300 pf
L_1 – 13-25 mh

FIGURE 4.10

25 kc GENERATOR

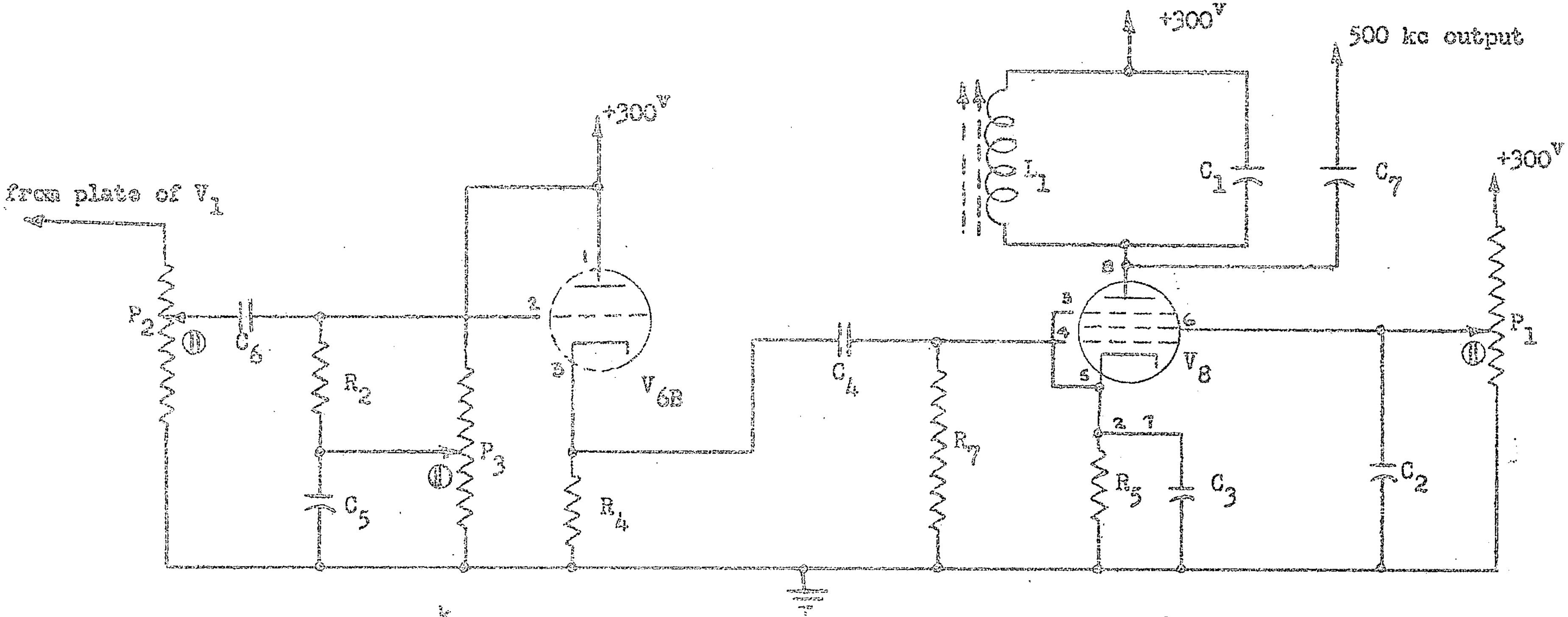


P_1 - 150^k
P_2, P_3 - 500^k
R_2 - 470^k
R_4 - 47^k
R_5 - 150 ohm
R_7 - 1 meg
C_1 - 330 pf.
C_2, C_5 - 0.01 uf

C_4, C_6 - 100 pf
C_3 - 0.1 uf
C_7 - 0.01 uf
L_1 - 1.5-3 mh
V_{6A} - ½ 12AU7
V_7 - 6AC7

FIGURE 4.11

200 kc GENERATOR



P_1 – 150^k

PP_2, P_3 – 500^k

R_2 – 470^k

R_4 – 47^k

R_5 – 150 ohm

R_7 – 1 meg

C_1 – 330 pf

C_2, C_5 – 0.01 uf

C_4, C_6 – 100 pf

C_3 – 0.1uf

C_7 – 0.01 uf

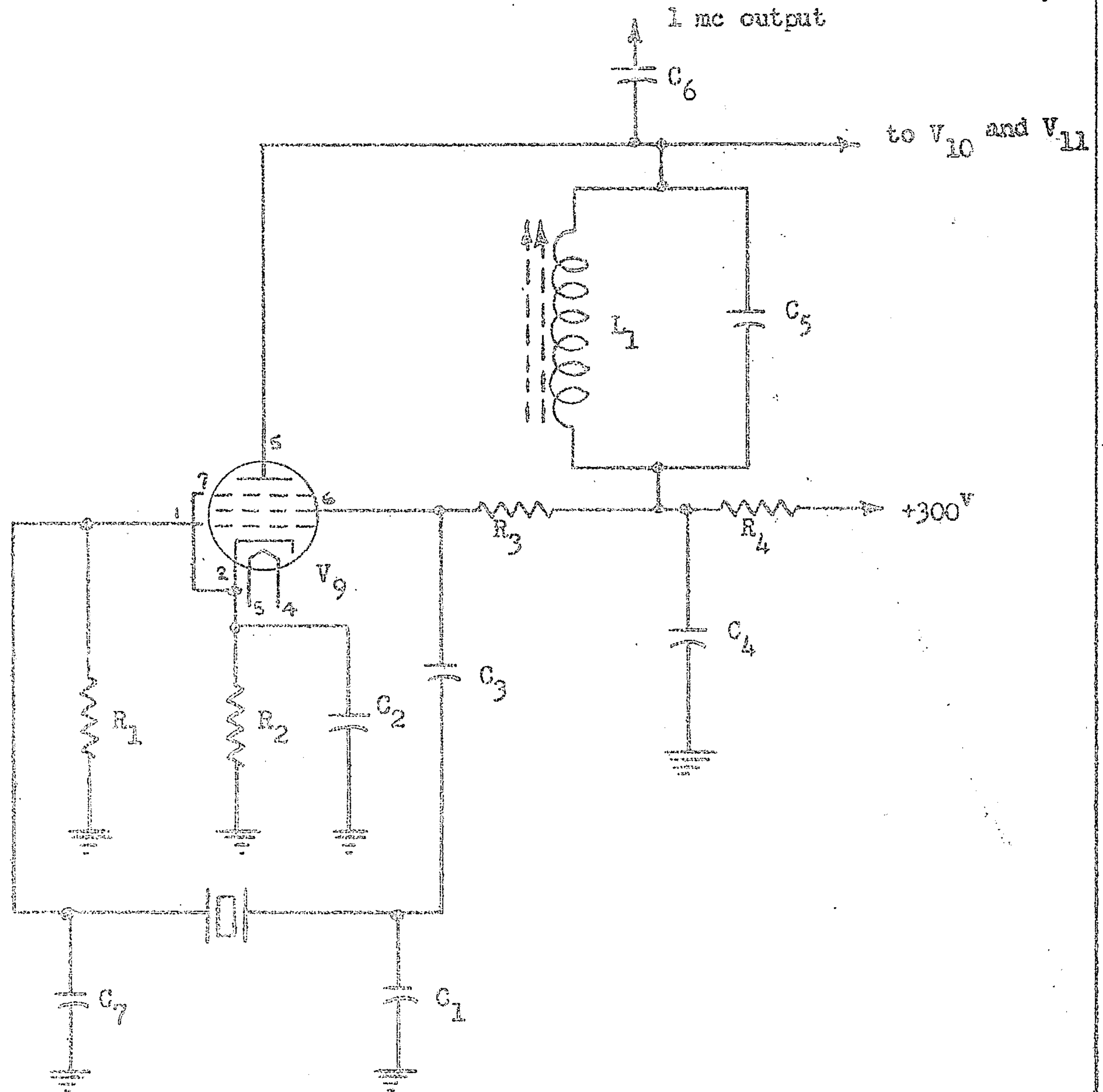
L_1 – 200–500 uh

V_{6B} – $\frac{1}{2}$ 12AU7

V_8 – 6AC7

FIGURE 4.12

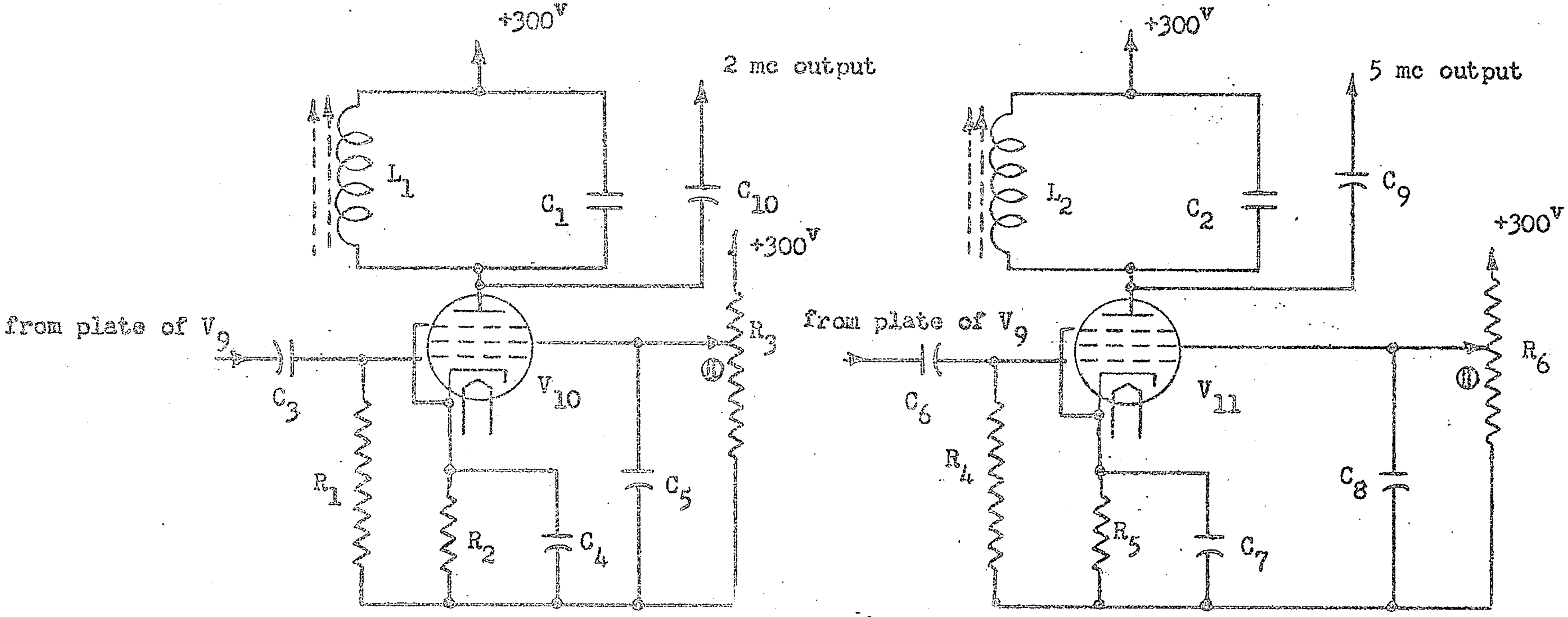
500 kc GENERATOR



crystal - 1 mc
R_1 - 18^k
R_2 - 470 ohm
R_3 - 47^k
R_4 - 25^k
C_1, C_7 - 68 pf
C_2 - 0.01 uf
C_3 - 0.002 uf
C_4 - 0.01 uf
C_5 - 200 pf
C_6 - 100 pf
L_1 - 105-200 uh
V_9 - 6BH6

FIGURE 4.13

1 mc CRYSTAL CONTROLLED OSCILLATOR

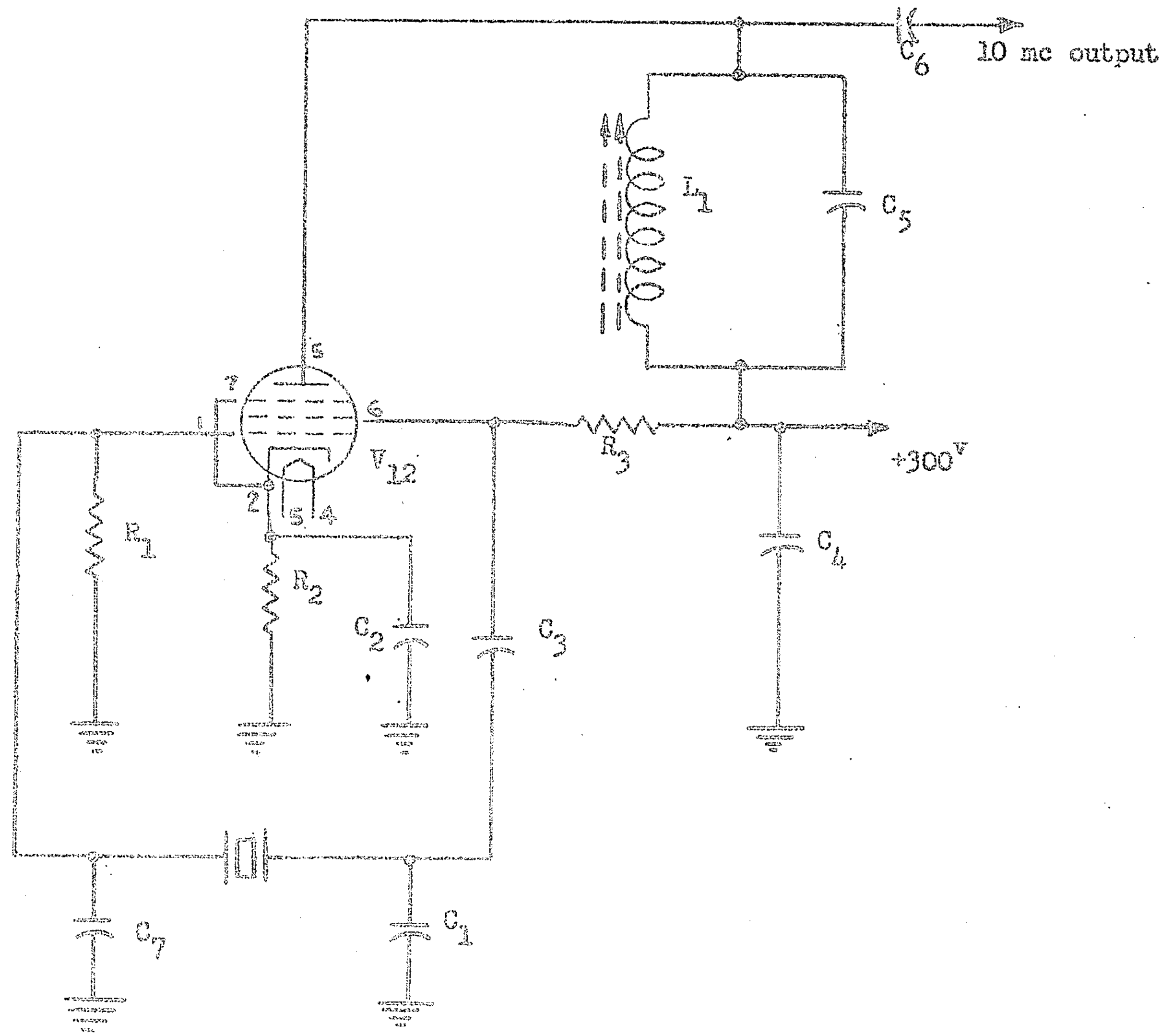


R_1, R_4 = 1 meg
R_2, R_5 = 150 ohm
R_3, R_6 = 150^k
L_1 = 45-75 uh
L_2 = 6.8-10.8 uh
V_{11}, V_{10} = 6AC7

$C_1, C_3, C_6, C_9, C_{10}$ = 100 pf
C_2 = 120 pf
C_4 = 0.01 uf
C_5, C_8 = 0.002 uf
C_7 = 0.005 uf

FIGURE 4.14

2 mc and 5 mc GENERATORS



crystal – 10 mc
R_1 – 1 meg
R_2 – 470 ohm
R_3 – 47^k
C_1, C_7 – 68 pf
C_4, C_2, C_3 – 0.002 uf

C_6 – 100 pf
L_1 – 0.7–1.1 uh
V_{12} – 6BH6

FIGURE 4.15

10 mc CRYSTAL CONTROLLED OSCILLATOR

4.6 OSCILLOGRAMS OF THE MARKER GENERATOR VOLTAGES

The output voltage from each generator was adjusted by varying the screen potential of the class C amplifier tubes or by detuning the LC plate circuit slightly. Each waveform is approximately 100 volts peak-to-peak. The vertical scale on all the oscillograms is 100 volts per cm and the time-base is as indicated on the oscillograms.

25 kc 50 usec/cm

50 kc 20 usec/cm

100 kc 10 usec/cm

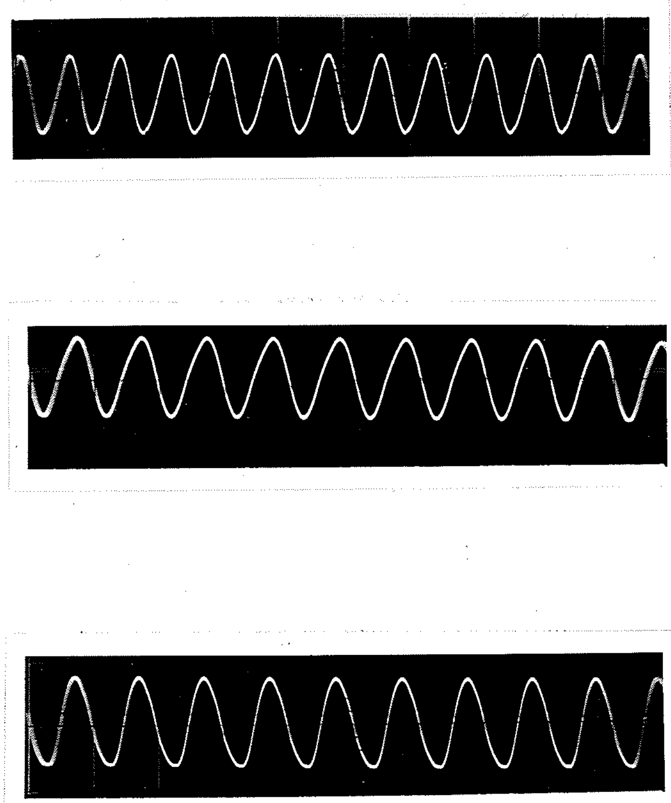


FIGURE 4.16

MARKER GENERATOR OSCILLOGRAMS

VERTICAL CALIBRATION: 100 VOLTS/CM

200 kc — 5 usec/cm

500 kc — 2 usec/cm

1 mc — 1 usec/cm

2 mc — 0.5 usec/cm

5 mc — 0.2 usec/cm

10 mc — 0.1 usec/cm

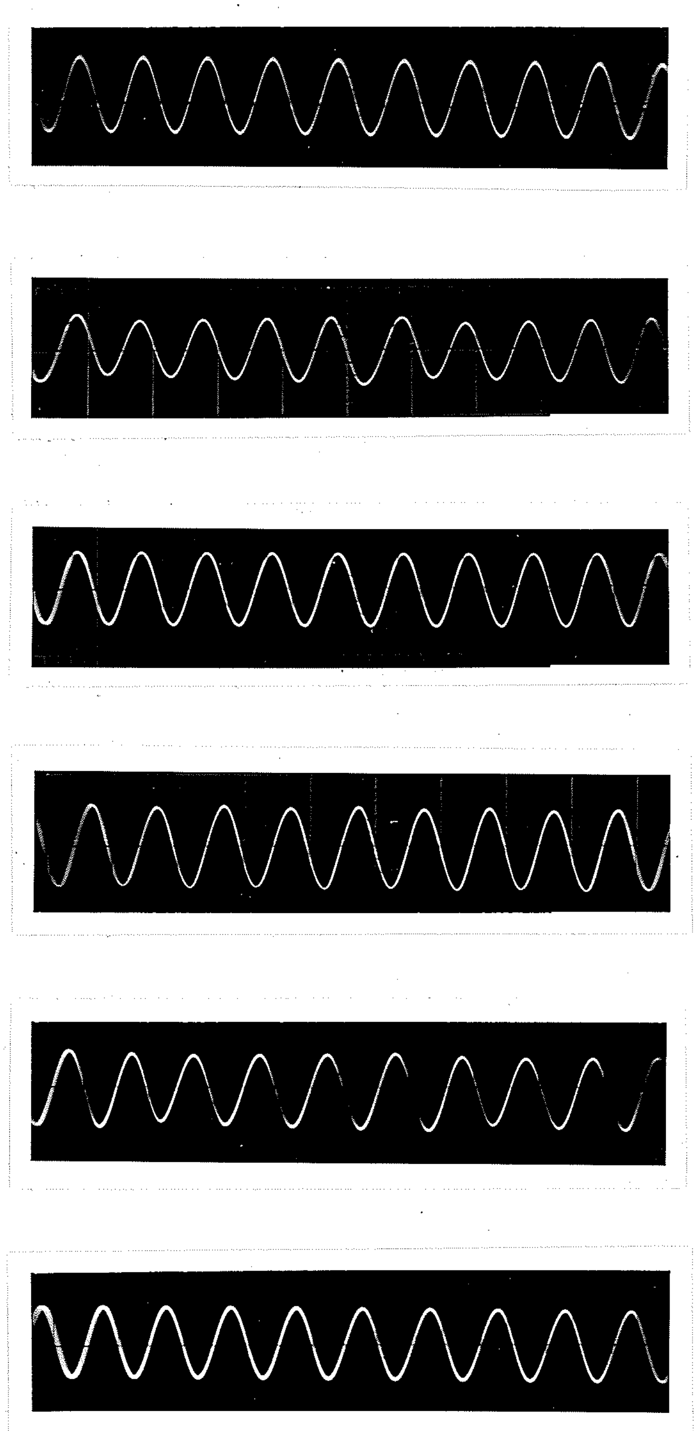


FIGURE 4.17

MARKER GENERATOR OSCILLOGRAMS

VERTICAL CALIBRATION: 100 VOLTS/CM

CHAPTER V

POWER SUPPLIES

5.1 INTENSIFIER POWER SUPPLY

The cathode-ray tube used in the oscillograph requires an intensifier or post-deflection acceleration potential in the order of +20,000 volts dc. The current drain from this supply is low and the load is constant; thus there is no need of a regulated supply. A separate constant voltage supply will eliminate line variations. A peak-to-peak ripple of 0.5 per cent is a reasonable design figure for power supplies used in electrostatic-tube applications[1].

A dc potential of 20 kv may be obtained through the use of a standard power-line frequency or a r-f power supply. When space and weight are at a premium, the r-f supply has a distinct advantage over the 60 cycle supply due to the smaller components used in the input-voltage filtering. However, the r-f supply requires special filtering and shielding of the input and output leads. Because of its relative simplicity, the availability of parts, and because space and weight were not at a premium, the standard 60 cycle supply was chosen.

The type of supply used was a full wave voltage doubler, the schematic circuit diagram of which is shown in figure 5.1. The transformer T_3 allows the output to be varied from 0 to 23,000 volts dc.

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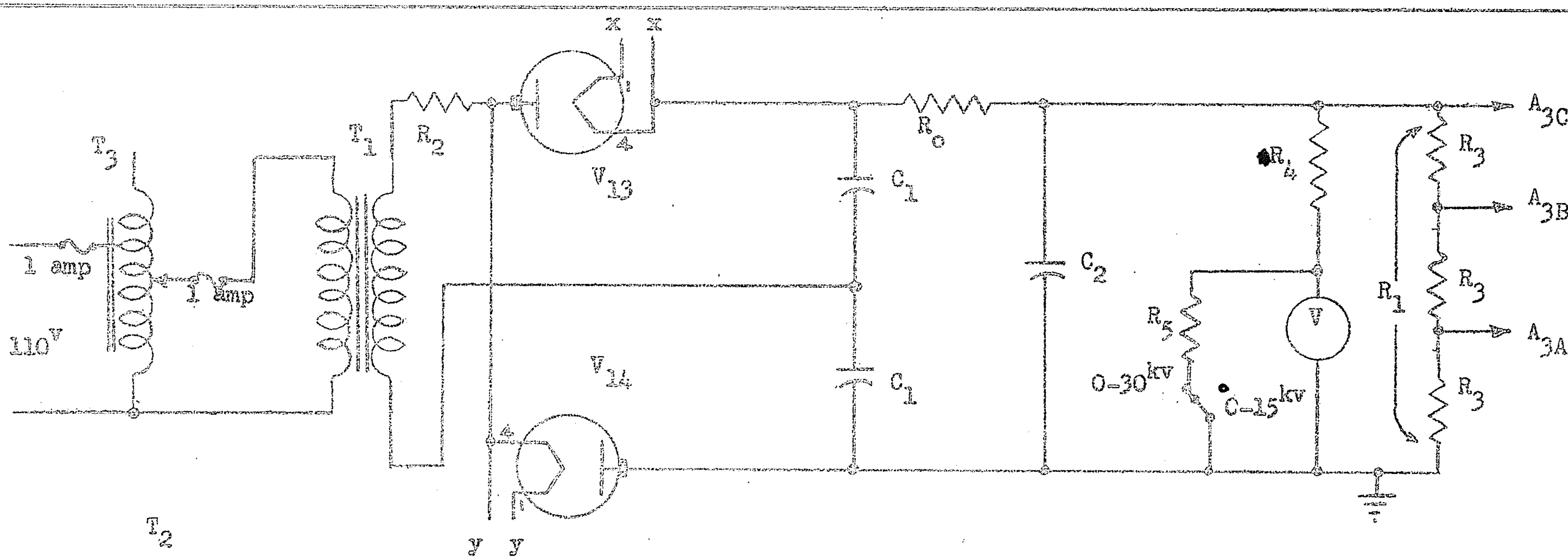
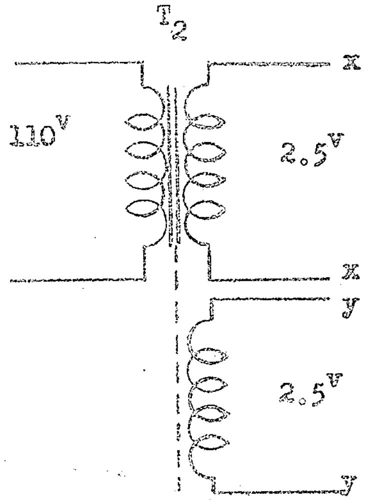
[1] T. Soller, M. A. Starr, and G. E. Valley, Jr., (editors). Cathode Ray Tube Displays (Vol. 22 of the Radiation Laboratory Series, New York: McGraw Hill Book Co. 1949) p. 166.





V_{13}, V_{14} = 878-A
R_o = 10 meg
R_1 120 meg
R_2 = 5^k
R_3 = 40 meg
R_4 = 600 meg 3%
R_5 = 20^k shunt, 1%

C_1 = 0.01 uf 30 kv dc
C_2 = 0.02 uf 15 kc dc
T_1 = $110^V/7000^V$
T_2 = $110^V/2.5^V$
T_3 = $110^V/0\text{-}130^V$
V - meter, 25 ua F.S., 20^k

FIGURE 5.1

FULL WAVE VOLTAGE DOUBLER

The filaments of the two tubes must be supplied by two windings of a well insulated transformer since each filament is at a different voltage with respect to ground potential. With the use of the voltage doubler circuit only half the transformer-secondary voltage and half the filter capacity of the standard half-wave rectifier is required.

The 878-A rectifier tube is a common high voltage rectifier tube which has a peak inverse voltage rating of 20,000 volts. The tube is meant to be used in a low current drain supply since the rated continuous dc output current is given as 5 ma. The resistor R_2 acts as a protective device which limits the peak current that can flow in the tubes to restore the charge on the capacitors C_1. The resistor R_1 acts as a potential divider as well as a bleeder resistor.

For the two section filter as shown in figure 5.1, the theoretical percentage ripple voltage is given by the following equation[2];

$$\mathcal{E}_r = \frac{T_1}{T_2}\left[\frac{1}{T_1 f(1 - e^{-1/ft})} + \ln f T_1 (1 - e^{-1/fT}) - 1\right]$$

Where $T_1 = R_o C_1$

$T_2 = R_1 C_2$

p = ratio of discharge time per cycle to period of ripple

$\doteq 1$

t = discharge time, (p/f)

---

[2]Soller, *ibid.*, p. 167

Using the component values of figure 5.1 and assuming $p \doteq 1$, $\xi_r = 0.3$ per cent.

The post deflection accelerating potential is applied to the cathode-ray tube terminals A_{3A}, A_{3B}, and A_{3C} from the potential divider R_1. These terminals are shown in figure 3.2. This total potential at terminal A_{3C} is measured with the meter to about 3 per cent accuracy as shown in figure 5.1. The meter has two ranges; 0-15 kv and 0-30 kv.

5.2 +3,000 AND -4,000 VOLT POWER SUPPLIES

The -4,000 volt dc power supply is used to supply the voltage divider which provides [redacted] accelerating, bias, and focus electrode potentials as shown in figure 5.2.

The +3,000 volt dc power supply is used as the plate supply for the sweep circuit. Both supplies use the same transformer as shown in figure 5.2. The simple half-wave rectifier is adequate for both these supplies since the current drain is very low. The filter-circuit requirements are met by the two-section RC-filter. The resistor R_1 acts as a current limiting resistor during the initial charging of the condenser. The filament transformer used for V_{16} must be well insulated since the cathode, hence the heater, is operating at about 4,000 volts with respect to the ground potential. The potential divider R_3 enables +3,000 volts to be obtained from the +4,000 volts available from V_{16}.

5.3 D.C. CALIBRATION SUPPLY

The dc calibration supply provides a continuously variable positive or negative dc potential to the vertical deflection plates of the cathode-ray tube so that the vertical axis of the screen may be calibrated. The schematic circuit diagram of this half-wave rectifier is shown in figure 5.3. By means of the autotransformer T_7 the voltage available from this power supply may be varied from ± 0-1,000 volts dc. The dpdt switch is used to change from a positive potential to a negative potential. The meter measures this calibration voltage to within 1 per cent accuracy. There are four voltage ranges; 0-30, 0-100, 0-300, and 0-1000 volts dc.

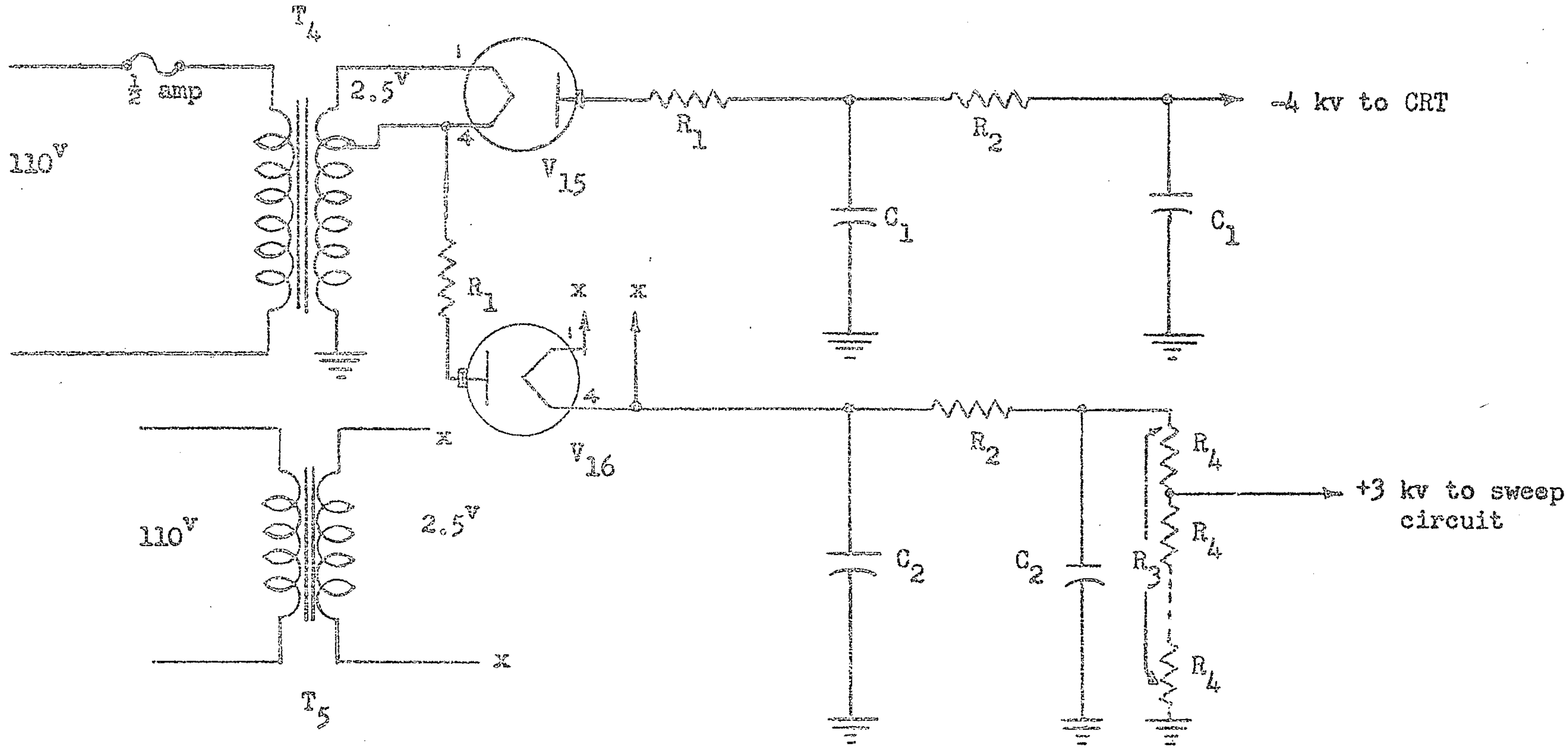
5.4 H AND V POSITIONING POWER SUPPLIES

These two power supplies, +400 and -400 volts dc, are used to supply the horizontal and vertical positioning potentials. As in most of the power supplies used in this oscillograph, the simple half-wave rectifier satisfies the low current drain requirements from these supplies. The schematic circuit diagram is shown in figure 5.4. The negative power supply is also used to supply a small negative bias to the sweep triggering circuit. The potentiometer P_4 provides a variable bias supply from 0 volts to approximately -50 volts dc.

5.5 +300 VOLT REGULATED POWER SUPPLY

This supply is a voltage regulated power supply which supplies the anode voltage for the marker generator. A regulated supply was used to provide a low output impedance, stability against input voltage variations and output load changes, and ripple reduction. The schematic circuit diagram is shown in figure 5.5[3]. This regulated supply provides a dc voltage of +300 volts, with 1 per cent regulation and a maximum load current of about 200 mamp. The unregulated power source consists of a conventional full-wave rectifier circuit filtered by a single section choke and capacitor filter.

[3] See "Handbook of Preferred Circuits of Navy Aeronautical Electronic Equipment", prepared by the National Bureau of Standards. The regulator circuit used is the "NBS preferred circuit no. 3A".



V_{15}, V_{16} - 2X2

R_1 - 1^k

R_2 - 250^k

R_3 - 3.76 meg

R_4 - 470^k

C_1 - 0.5 uf (5 kv dc)

C_2 - 0.25 uf (5 kv dc)

T_4 - $110^V/1100^V$

T_5 - $110^V/2.5^V$

FIGURE 5.2

+3 kv AND -4 kv SUPPLIES

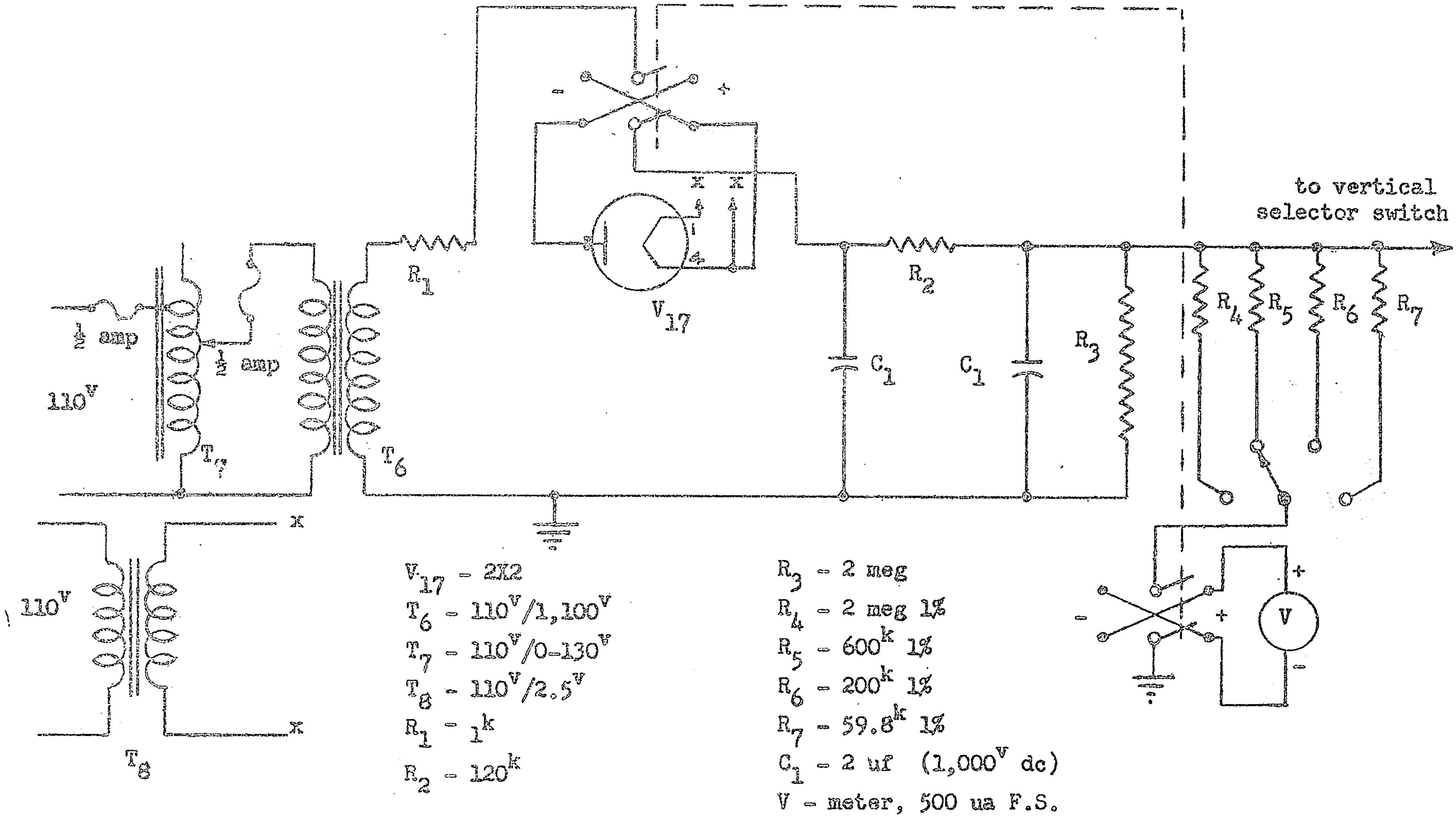
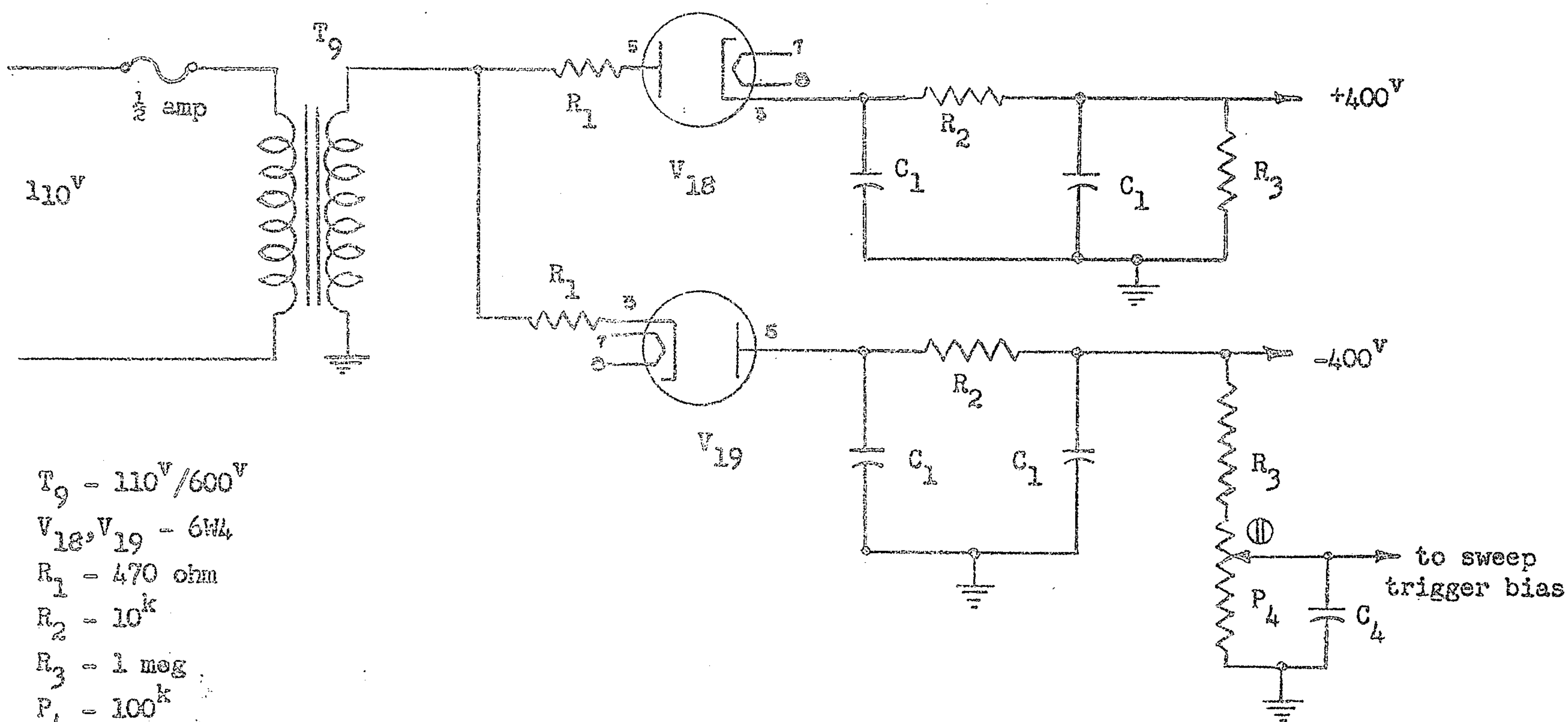


FIGURE 5.3

CALIBRATION SUPPLY



T_9 - $110^V/600^V$

V_{18}, V_{19} - 6W4

R_1 - 470 ohm

R_2 - 10^k

R_3 - 1 meg

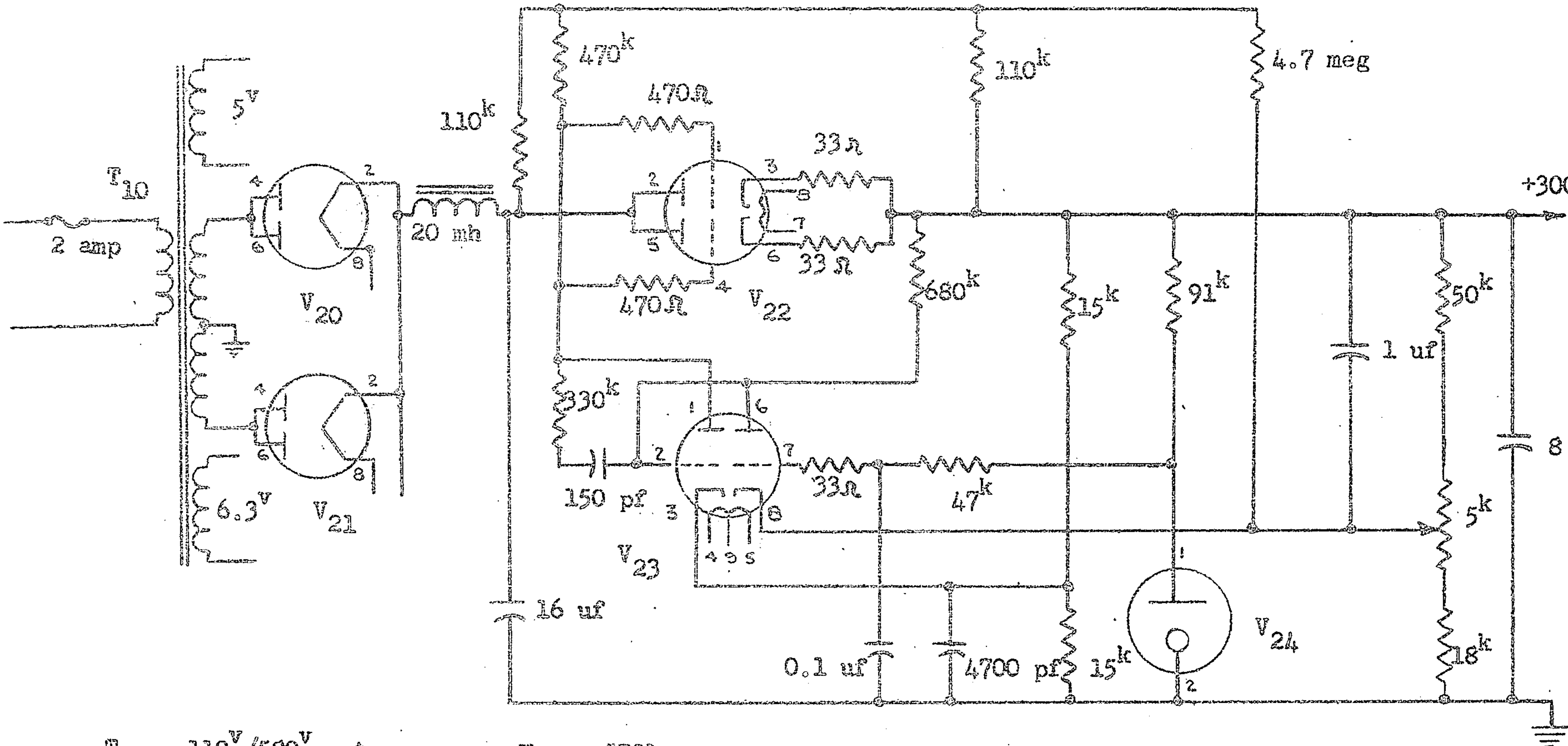
P_4 - 100^k

C_1 - 2 uf (600^V dc)

C_4 - 1 uf

FIGURE 5.4

SHIFT POTENTIAL SUPPLIES



T_{10} - $110^V/580^V$ c.t.
V_{20}, V_{21} - 5R4
V_{22} - 6080
V_{23} - 5751
V_{24} - 85A2

FIGURE 5.5

+300 VOLT REGULATED POWER SUPPLY

CHAPTER VI

RESULTS AND CONCLUSIONS

6.1 RESULTS

The oscillograph was tested with the use of an impulse generator simulator. This generator produces simulated impulse generator waveforms. In normal applications, it is operated recurrently but for this case it was operated so as to produce a single waveform. The block diagram of this simulator is shown in figure 6.1. The positive trigger pulse from the simulator was used to trigger the time-base of the oscillograph.

The standard 1.5x40 full wave is presented in figure 6.2. The vertical axis was calibrated by means of the calibration power supply. Three base lines were used to calibrate this axis; 0, 250, and 500 volts. The horizontal axis was calibrated with a 100 kc signal from the marker generator.

A "chopped" full wave is shown in figure 6.3. The vertical calibration is identical to that of figure 6.2; however, a 1 mc sinusoid was used to calibrate the time-base.

6.2 CONCLUSIONS

From the results obtained in section 6.1 (figures 6.2 and 6.3), it is expected that the circuitry developed will perform satisfactorily when the high voltage generator becomes available for use.

The final oscillograph will contain three cathode-ray tubes so that three oscillograms of one impulse test may be made simultaneously. One cathode-ray tube may record a voltage oscillogram and the other two tubes may be used to obtain two current oscillograms: One current oscillogram would be taken at a fast sweep speed and the second at a slower sweep, comparable to the sweep speed used for the voltage oscillogram.

A large vertical steel cabinet will be constructed to house all power supplies in the bottom racks, with the marker generator above these supplies. Three sweep generators on separately removable plug-in chassis are on the next rack and the three corresponding cathode-ray tubes are mounted above these generators. This flexibility will permit easy interchange of sweep generators to provide faster, slower, or possibly linear sweeps which may beome necessary at some future date. A possible front panel layout is shown in figure 6.4.

The above assembly work, is, of course not part of this thesis, and will be done by the Technical Staff of the Electrical Engineering Department, probably during the summer of 1961.

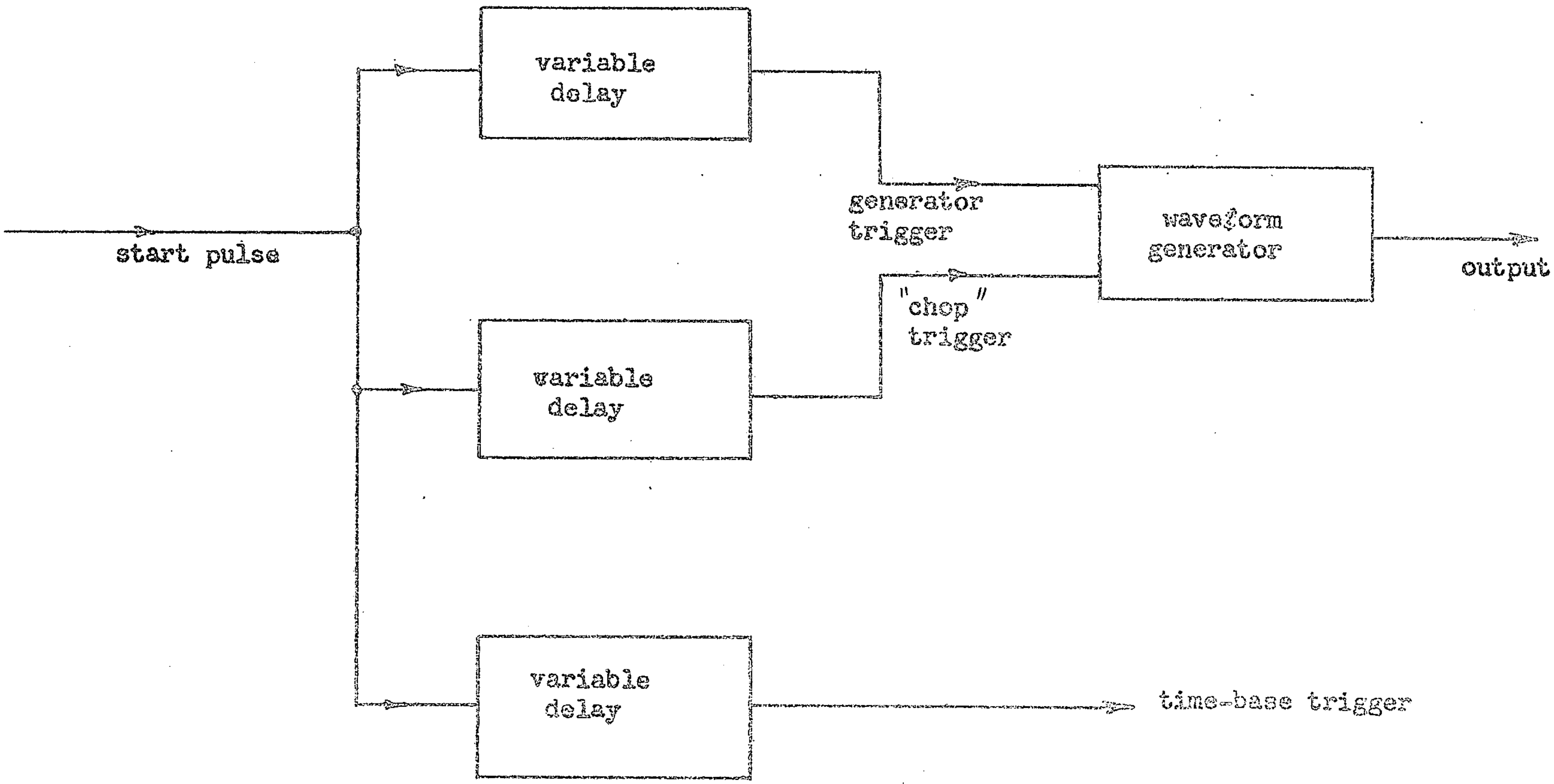


FIGURE 6.1

BLOCK DIAGRAM OF IMPULSE GENERATOR SIMULATOR

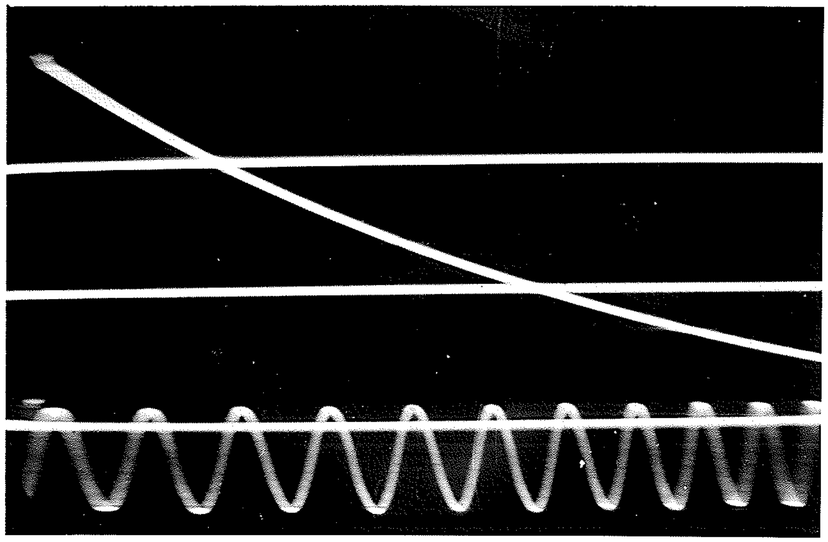


FIGURE 6.2

1.5x40 FULL WAVE

VERTICAL CALIBRATION: 0, 250, 500 VOLTS
HORIZONTAL CALIBRATION: 100 kc

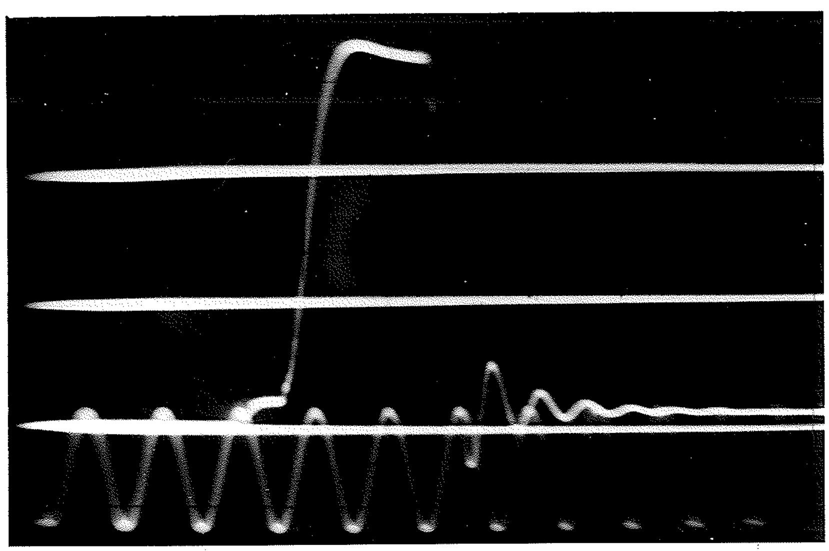
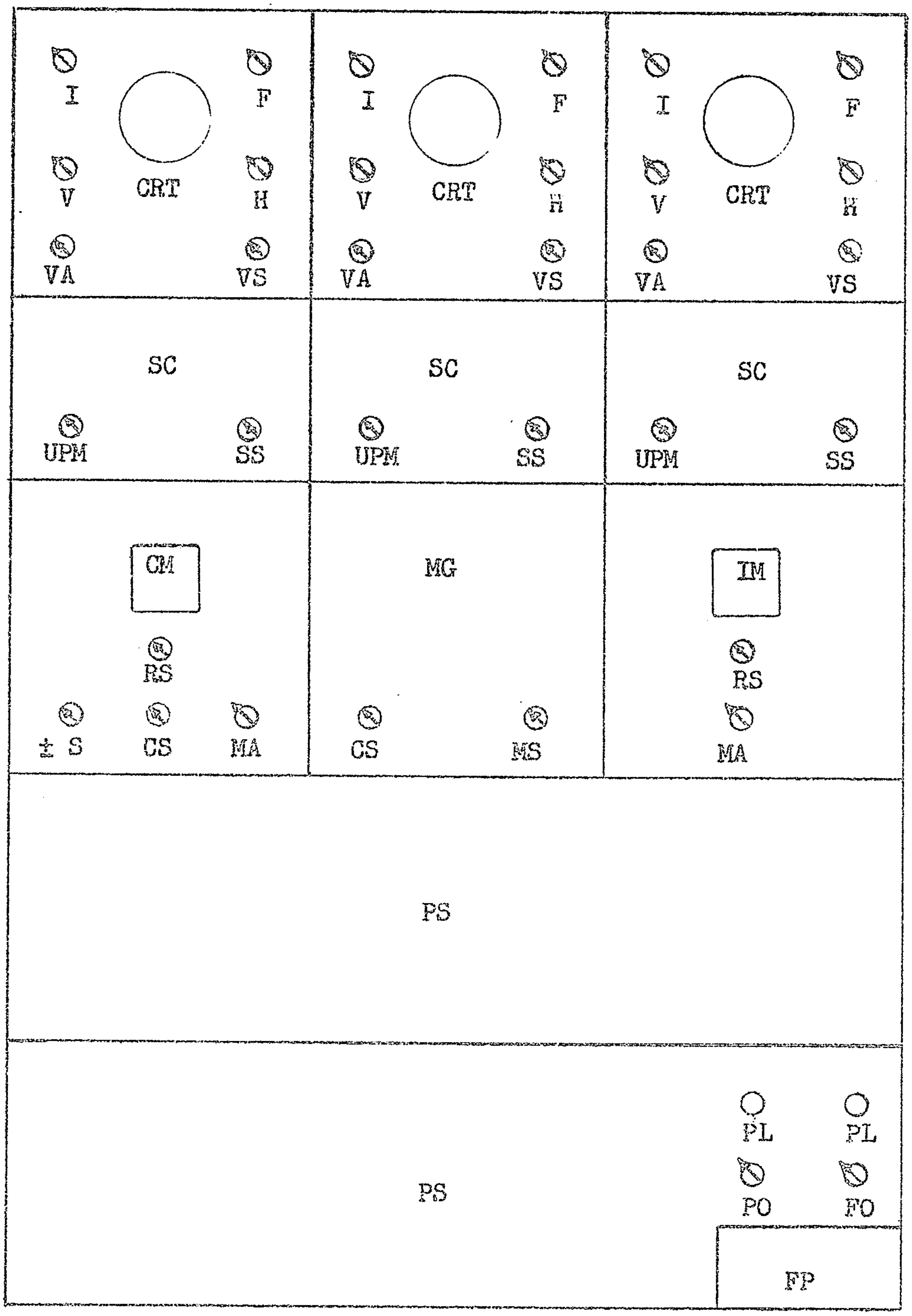


FIGURE 6.3

"CHOPPED" 1.5x40 FULL WAVE

VERTICAL CALIBRATION: 0, 250, 500 VOLTS
HORIZONTAL CALIBRATION: 1 mc



See next page for key of symbols.

FIGURE 6.4

PROPOSED FRONT PANEL LAYOUT

KEY OF SYMBOLS USED IN FIGURE 6.4

CRT - cathode-ray tube

F - focus adjust

I - intensity adjust

H - horizontal position

V - vertical position

VA - vertical attenuator switch

VS - vertical selector switch

SC - sweep circuit

UPM - unblanking pulse magnitude adjust

SS - sweep speed selector

CM - calibration voltage meter

MG - marker generator

IM - intensifier voltage meter

± S - plus-minus switch

RS - range selector

CS - channel selector

MA - magnitude adjust

MS - marker voltage selector

PS - power supplies

PO - power on

FO - filaments on

PL - pilot lamp

FP - fuse panel

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APPENDIX

APPENDIX A

DUMONT K1068 CATHODE-RAY TUBE

The type K1068 cathode-ray tubes are high voltage tubes which incorporate an intensifier subdivided into several steps. This feature permits the use of much higher overall accelerating voltages with deflection sensitivities only slightly less than heretofore obtainable in low voltage cathode-ray tubes. Operation with intensifier to second anode voltage ratios as high as ten to one are made possible by the multiband feature. This tube has a flat face, cylindrical body, and the deflection plate and anode connections are made through the neck instead of through the base. Short deflection plate leads and special shielding between the deflection plates facilitates high frequency operation. The gun is designed to draw negligible focusing electrode current

GENERAL CHARACTERISTICS

ELECTRICAL DATA

heater voltage	6.3 volts
heater current	0.6 ± 10% amps
focusing method	electrostatic
deflecting method	electrostatic
phosphor	no. 2
fluorescence	green
phosphorescence	green

persistence	long

MECHANICAL DATA

overall length	$16\frac{3}{4} \pm 3/8$ ins
greatest diameter of bulb	$5\frac{1}{4} \pm 3/32$ ins
minimum useful screen diameter	$4\frac{1}{4}$ ins
bulb contacts (recessed small ball caps)	J1-22
neck contacts (small ball caps)	J1-25
base (medium shell diheptal 12-pin)	B12-37
basing	14F

base alignment:

1D2 trace aligns with pin no. 5 and the tube axis	± 10°
positive voltage on D1 deflects the beam approximately towards pin number 5	
angle between 3D4 and 1D2 traces	90 ± 3°

bulb contact alignment:

J1-22 contacts align with 1D2 trace	± 10°
J1-22 contacts on the same side as pin number 5	

MAXIMUM RATINGS (design center values)

anode no. 3 voltage (accelerator high voltage electrode)	25,500 volts dc max.
anode no. 2 voltage[1]	3,500 volts dc max.
ratio anode no. 3 to anode no. 2 voltage	10 max.

[1] Anode no. 2 and grid no. 2, which are connected together within the tube, are referred to herein as anode no. 2.

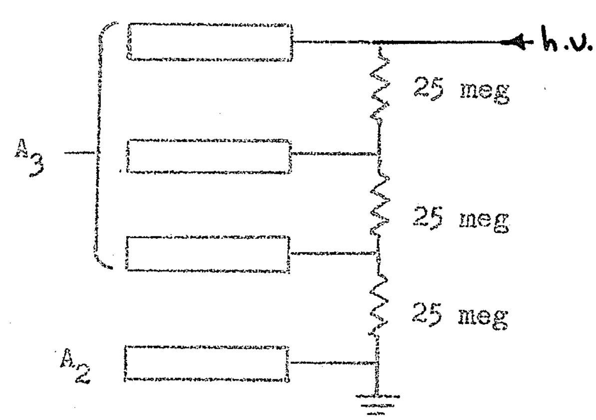
anode no. 1 voltage	1,550 volts dc max
grid no. 1 voltage	
negative bias value	125 volts dc max
positive bias value	0 volts dc max
positive peak value	2 volts max
peak heater cathode voltage[2]	
heater negative with respect to cathode	125 volts dc max
heater positive with respect to cathode	125 voltsdc max
peak voltage between anode no. 2 and any deflection electrode	1,200 volts max

TYPICAL OPERATING CONDITIONS

for anode no. 3 voltage of[3]	10,000	20,000 volts
for anode no. 2 voltage of	2,000	2,000 volts

[2]Cathode should be returned to one side or to the mid-tap of the heater transformer winding.

[3]This voltage should be equally divided over the three intensifier electrodes. The following is a suggested method of connection. The two A_2 terminals must be connected together.



anode no. 1 voltage for focus	362 to 695	362 to 695 volts
grid no. 1 voltage[4]	-30 to -90	-30 to -90 volts
deflection factors:		
D1 and D2	102 to 154	140 to 210 volt dc/in
D3 and D4	97 to 145	131 to 197 volts dc/in
anode no. 1 voltage for focus		18.1% to 34.8% of E_{b2}
grid no. 1 voltage[4]		1.5% to 4.5% of E_{b2}
anode no. 1 current for any operating condition		-50 to +10 uamps
deflection factors:		
no third anode or $E_{b2} = E_{b3}$		
D1 and D2		30 to 45 volts dc/in/kv of E_{b2}
D3 and D4		30 to 45 volts dc/in/kv of E_{b2}
for E_{b3} = twice E_{b2}		
D1 and D2		36 to 54 volts dc/in/kv of E_{b2}
D3 and D4		36 to 54 volts dc/in/kv of E_{b2}
spot position (undeflected)[5]		within 20 mm square

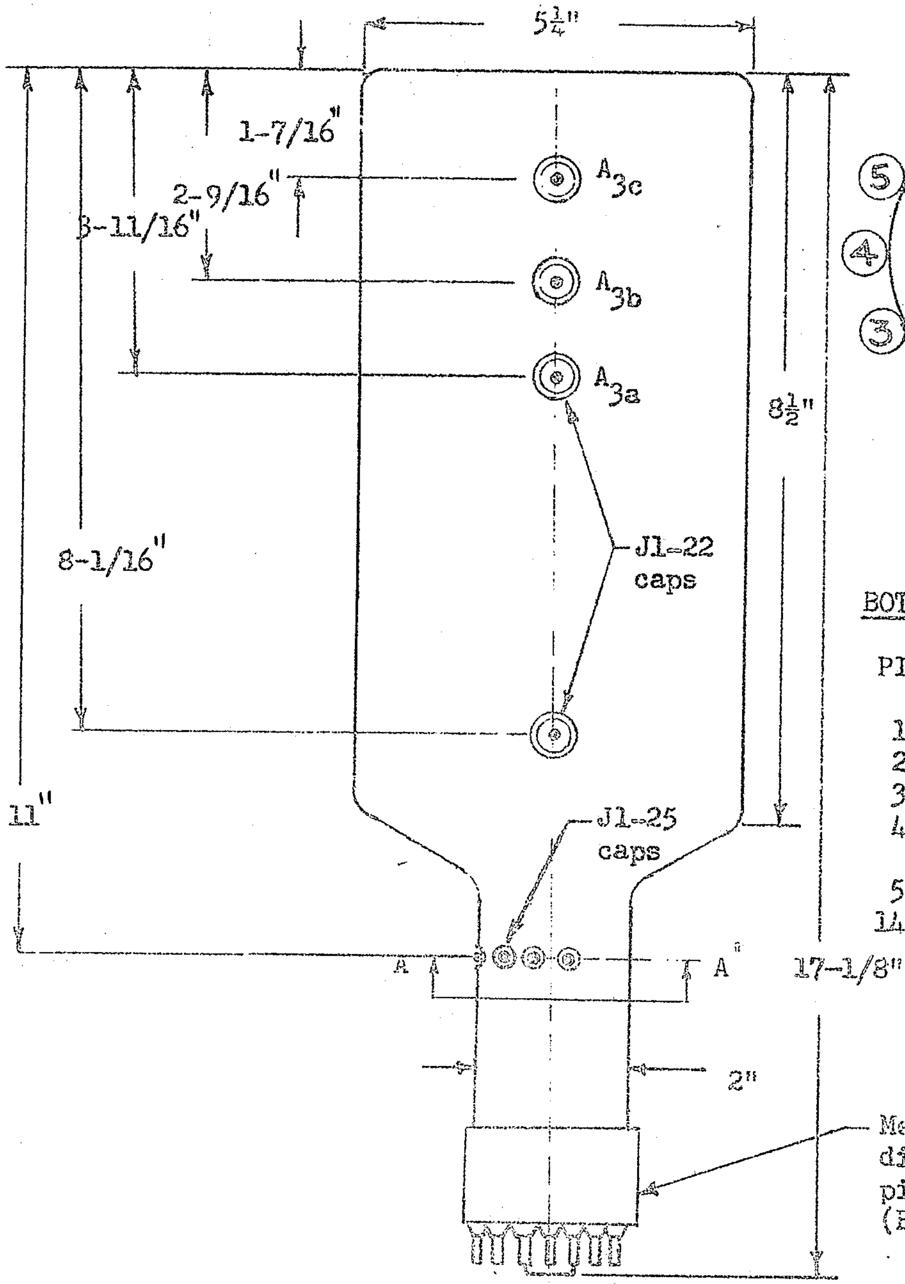
MAXIMUM CIRCUIT VALUES

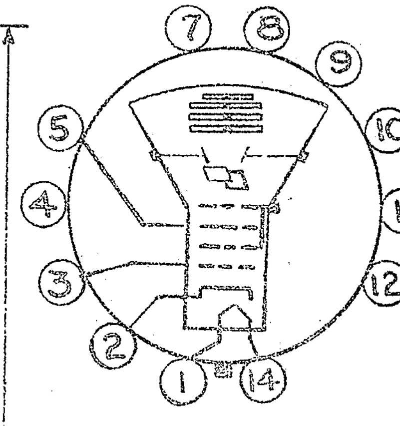
grid no. 1 circuit resistance	1.5 megs max
resistance in any deflecting electrode circuit[6]	5 megs max

[4]Visual extinction of undeflected focused spot.

[5]With $E_{b3} = 10{,}000^v$, $E_{b2} = 2{,}000^v$ and E_{b1} adjusted for focus.

[6]It is recommended that the deflecting electrode circuit resistances be approximately equal.





TUBE K-1068

BOTTOM VIEW OF BASE

PIN NO.	ELEMENT
1	heater
2	cathode
3	grid no. 1
4	internal connection
5	anode no. 1
14	heater

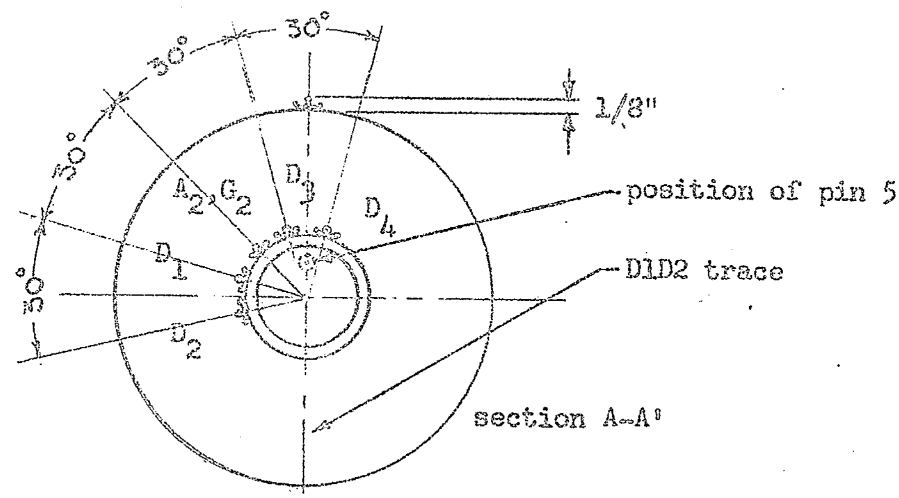


FIGURE A.1

APPENDIX B

POSTDEFLECTION ACCELERATION CATHODE-RAY TUBES

The deflection sensitivity of the cathode-ray beam for a given deflecting potential varies inversely with the accelerating potential. Consequently, a small accelerating voltage produces better sensitivity than a large accelerating voltage. This would seem to indicate that low values of accelerating potential would be more advantageous than high values because of the increase in deflection sensitivity. However, the light output of a phosphor or the luminosity of the spot on the screen increases roughly as the square of the voltage through which the electrons have been accelerated. The energy of the electron beam is directly dependent on the accelerating potential. As a result, a compromise must be reached between luminosity and deflection sensitivity.

In order to obtain a useful intensity of trace when very fast single sweeps are to be observed or fast writing speeds are to be obtained, the energy of the cathode-ray beam must be considerably higher than in the customary tube types. This increase in beam energy could be obtained by increasing the beam current, however, this usually results in making the spot size too large. A better method of increasing the beam energy is by raising the accelerating voltage. This procedure would result in a decrease in deflection sensitivity in customary tube types. However, by the use of the postdeflection-acceleration principle, a high luminosity and a reasonable deflection sensitivity can be obtained. This principle consists of including an additional electrode near the screen of the cathode-ray tube. This additional electrode is called the intensifier, postaccelerating electrode, or third anode. The accelerating potential is applied to the first anode and the second anode is kept at the same potential so that the deflection occuring between the deflection plates is controlled solely by the deflecting potential.

In postdeflection-acceleration cathode-ray tubes, the accelerating potential is chosen to give adequate sensitivity and the postaccelerating potential gives the increased intensity of the spot on the screen. Each particular application of the cathode-ray tube has its own sensitivity and intensity requirements. In postdeflection-acceleration cathode-ray tubes, it may be necessary to change the shape of the conventional cathode-ray tube and also, divide the postaccelerating potential among several bands located near the screen. This is particularly common in tubes that have a ratio of intensifier to accelerating potential greater than two. The change in tube shape reduces the distortions that are introduced by the nonaxial components of electric field caused by the intensifier. The insulation problems incurred in postaccelerating cathode-ray tubes due to the high voltages are sometimes minimized by applying part of the total accelerating voltage below ground potential and the rest above ground potential.

APPENDIX C

QUARTZ CRYSTALS

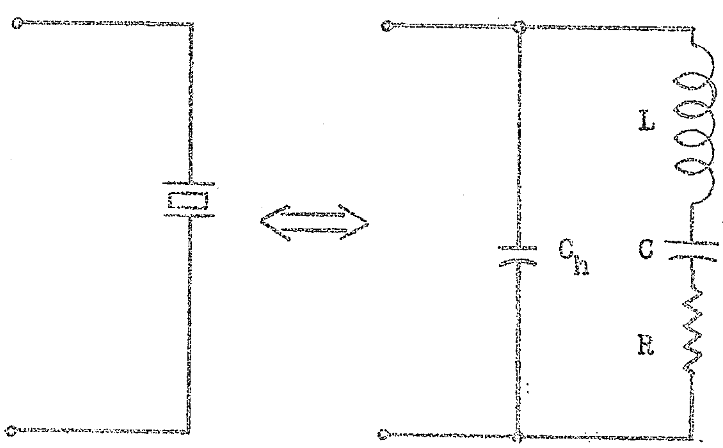
When a high degree of frequency stability over long periods of time is desired, the usual resonant circuit of an oscillator is replaced by a mechanically vibrating piezoelectric quartz crystal. An alternating voltage applied across the quartz crystal causes the crystal to vibrate and, if this frequency approximates the mechanical resonant frequency of the crystal, the vibrations will be intense.

The electrical equivalent circuit of a vibrating quartz crystal is shown in figure C.1. The series LRC combination represents the crystal while the capacitance C_h represents the electrostatic capacitance between the crystal electrodes when the crystal is not vibrating. Typical values of these parameters are shown in figure C.1. The Q of such a resonant circuit is very high. Values in excess of 20,000 are very common and in special cases values over 1,000,000 have been observed. The small value of C/C_h causes the coupling between the crystal resonator and the external circuit to be inherently small. The quartz crystal resonator also has a remarkably high ratio of inductance to capacitance.

The impedance characteristic of the equivalent circuit is shown in figure C.2. At the frequency f_1 which makes $2\pi f_1 L = 1/2\pi f_1 C$, the branch LCR is in series resonance, and so offers a low resistive impedance at the external terminals. At a slightly higher frequency f_2 the inductive reactance of the series branch is equal to the capacitive reactance of C_h; thus the crystal is in parallel resonance. At this parallel resonant frequency, the crystal offers a high impedance at the external terminals. The difference between these two frequencies is usually less than one percent.

Production techniques have decreased the crystal temperature coefficient to a negligible value over a given temperature range. The effects of crystal ag●ing are decreased by mounting the crystal in a vacuum or in an inert gas. Resonant frequencies ranging from a few hundred cycles to about 30 mc can be obtained from crystals directly and frequencies up to 100 mc may be obtained by overtone operation of the crystal.

A crystal controlled oscillator uses a crystal to control the frequency by so locating the crystal in the circuit that the crystal equivalent electrical network becomes part of the resonant circuit which would normally control the frequency.



$L = 3.3$ h
$C = 0.042$ pf
$C_h = 5.8$ pf
$Q = 23,000$

FIGURE C.1

EQUIVALENT CIRCUIT OF A QUARTZ CRYSTAL

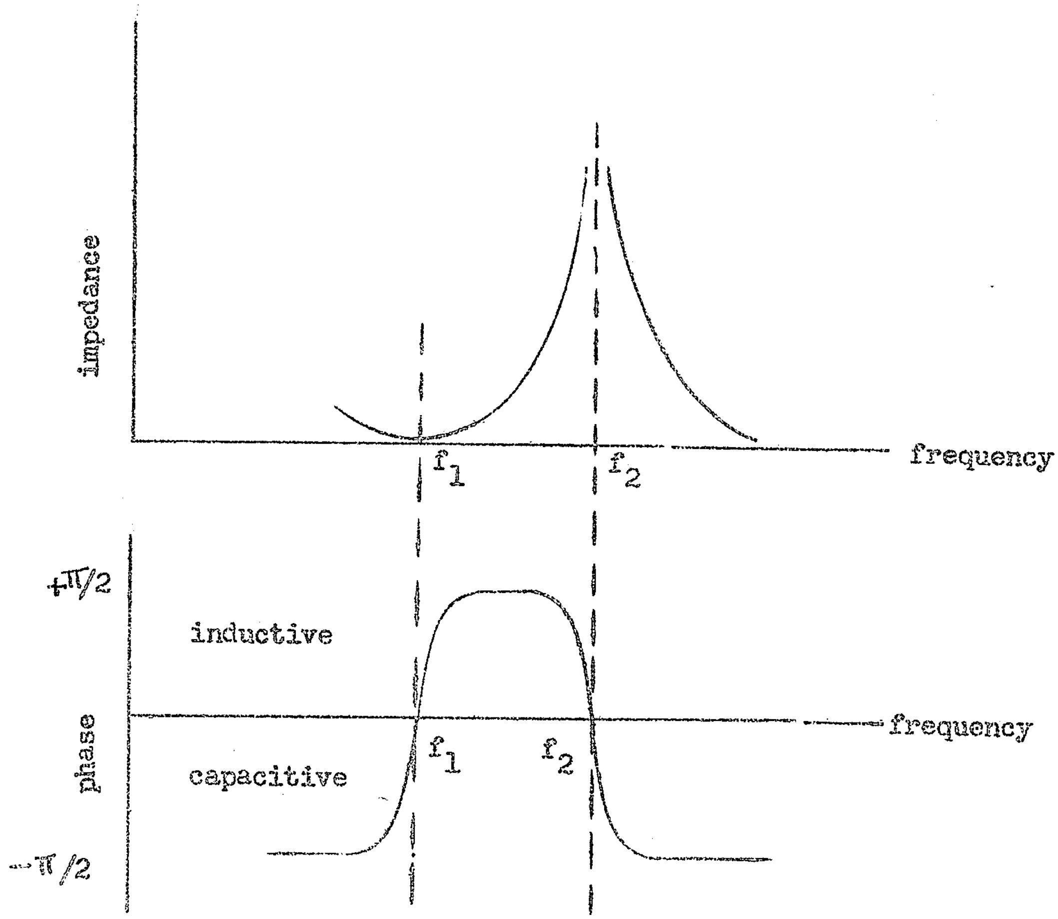


FIGURE C.2

IMPEDANCE CHARACTERISTICS

APPENDIX D

THE SWEEP CIRCUIT

The basic circuit diagram of the sweep circuit is shown in figure D.1. A waveform, positive with respect to ground potential, is obtained from point "A" and its negative is obtained from point "B". The equivalent circuit of figure D.1 is shown in figure D.2. This circuit neglects the tube drop and any tendency for the capacitor C to recharge through R_2 from E during the discharge of C through the $R_1R_3C_1$ circuit.

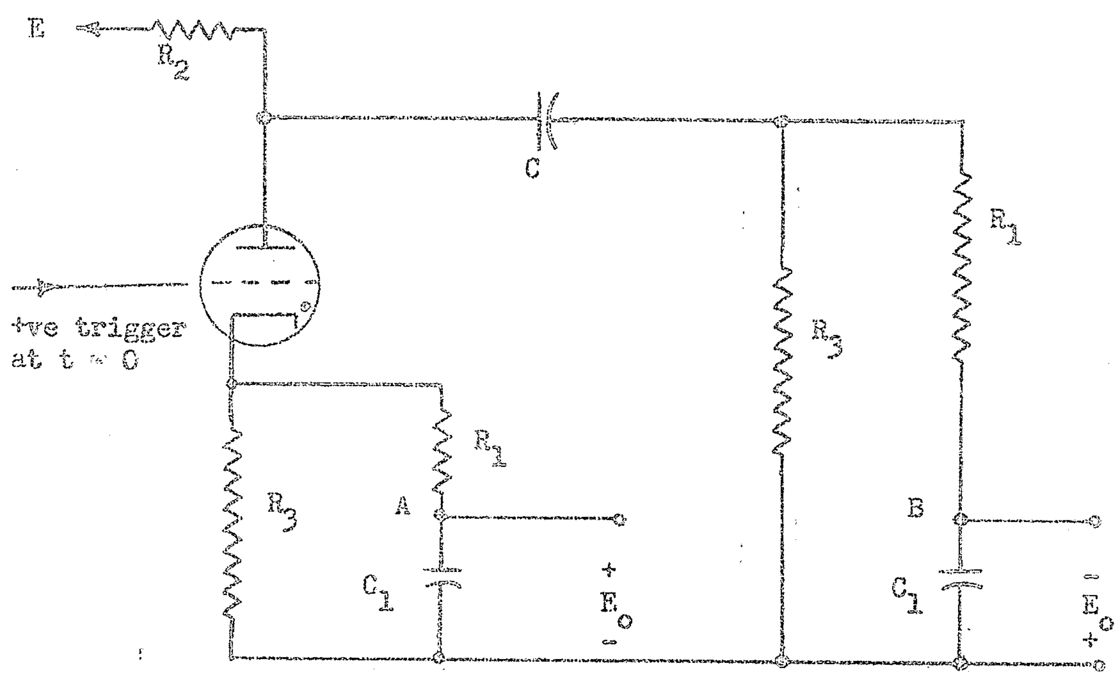


FIGURE D.1

BASIC SWEEP CIRCUIT

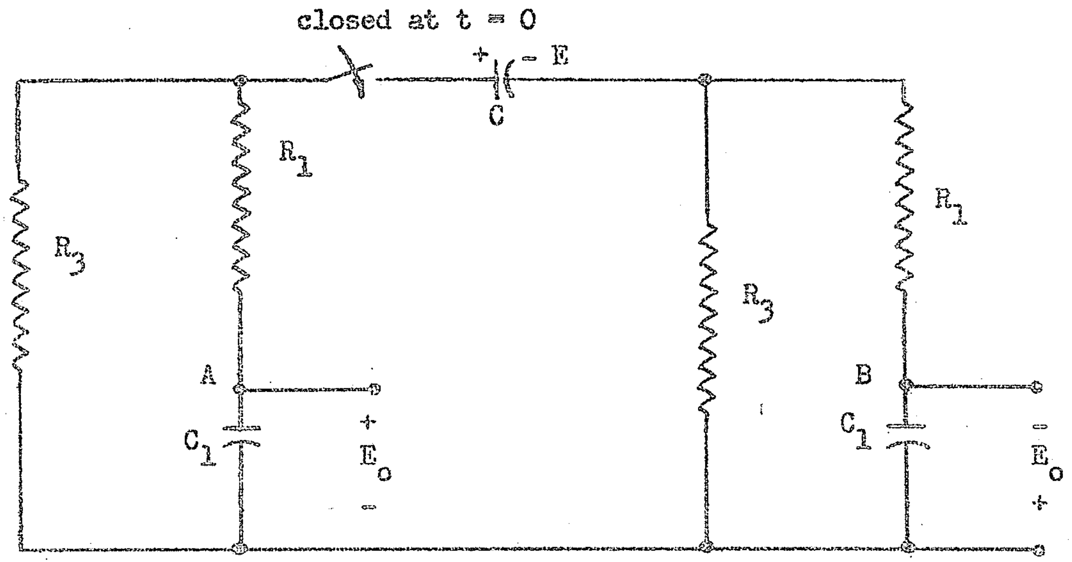


FIGURE D.2

EQUIVALENT CIRCUIT OF THE SWEEP CIRCUIT

From a mesh current analysis of figure D.2, it can be shown that the Laplace Transform of the output voltage E_o at point "A" is

$$E_o(s) = \frac{E}{C_1 R_1} \cdot \left[\frac{1}{s^2 + s\left(\frac{1}{2CR_3} + \frac{1}{2CR_1} + \frac{1}{C_1 R_1}\right) + \frac{1}{2CC_1R_1R_3}} \right] \quad \text{.....(D.1)}$$

and taking the inverse Laplace Transform

$$E_o(t) = \frac{K}{2\omega_n\sqrt{\zeta^2-1}} \left[e^{-(\zeta\omega_n - \omega_n\sqrt{\zeta^2-1})t} - e^{-(\zeta\omega_n + \omega_n\sqrt{\zeta^2-1})t} \right] \quad \text{......(D.2)}$$

where $K = \dfrac{E}{2C_1R_1}$

$$\omega_n = \sqrt{\frac{1}{2CC_1R_1R_3}}$$

$$\zeta = \frac{1}{2}\left[\sqrt{\frac{C_1R_1}{2CR_3}} + \sqrt{\frac{C_1R_3}{2CR_1}} + \sqrt{\frac{2CR_3}{C_1R_1}}\right]$$

If $\zeta > 1$ the first term in equation D.2 has a negligible effect on the rise time of $E_o(t)$. Also, if $R_3 \gg R_1$ and $C \gg C_1$ then the maximum value of $E_o(t)$ is E/2 as shown by

$$\frac{K}{2\omega_n\sqrt{\zeta^2-1}} \doteq \frac{E}{\frac{C_1R_1}{CR_3} - \frac{C_1}{C} - 2}$$

$$\doteq E/2$$

The validity of equation D.2 and its underlying assumptions were checked by comparing an oscillogram of $E_o(t)$ from point "A" in figure D.1 and a plot of equation D.2. The oscillogram was taken with the sweep circuit operating on a "single shot" basis and is presented in figure D.4. When the sweep was operated on a recurrent basis, the magnitude of the sweep waveform decreased drastically. As an example, when the sweep circuit was operated continuously at 60 cycles per second, the maximum magnitude of the sweep waveform was 740 volts using the same values of E, R_1, R_3, C, and C_1 as used in the "one-shot" oscillogram of figure D.3. An enlarged tracing of the "one-shot" oscillogram was obtained with the use of a photographic enlarger. Equation D.2 was evaluated with the same values of E, R_1, R_3, C, and C_3 and was plotted to the same scale as the enlarged oscillogram. The results are illustrated in figure D.3. From this diagram, the overall error is seen to be approximately 5 per cent.

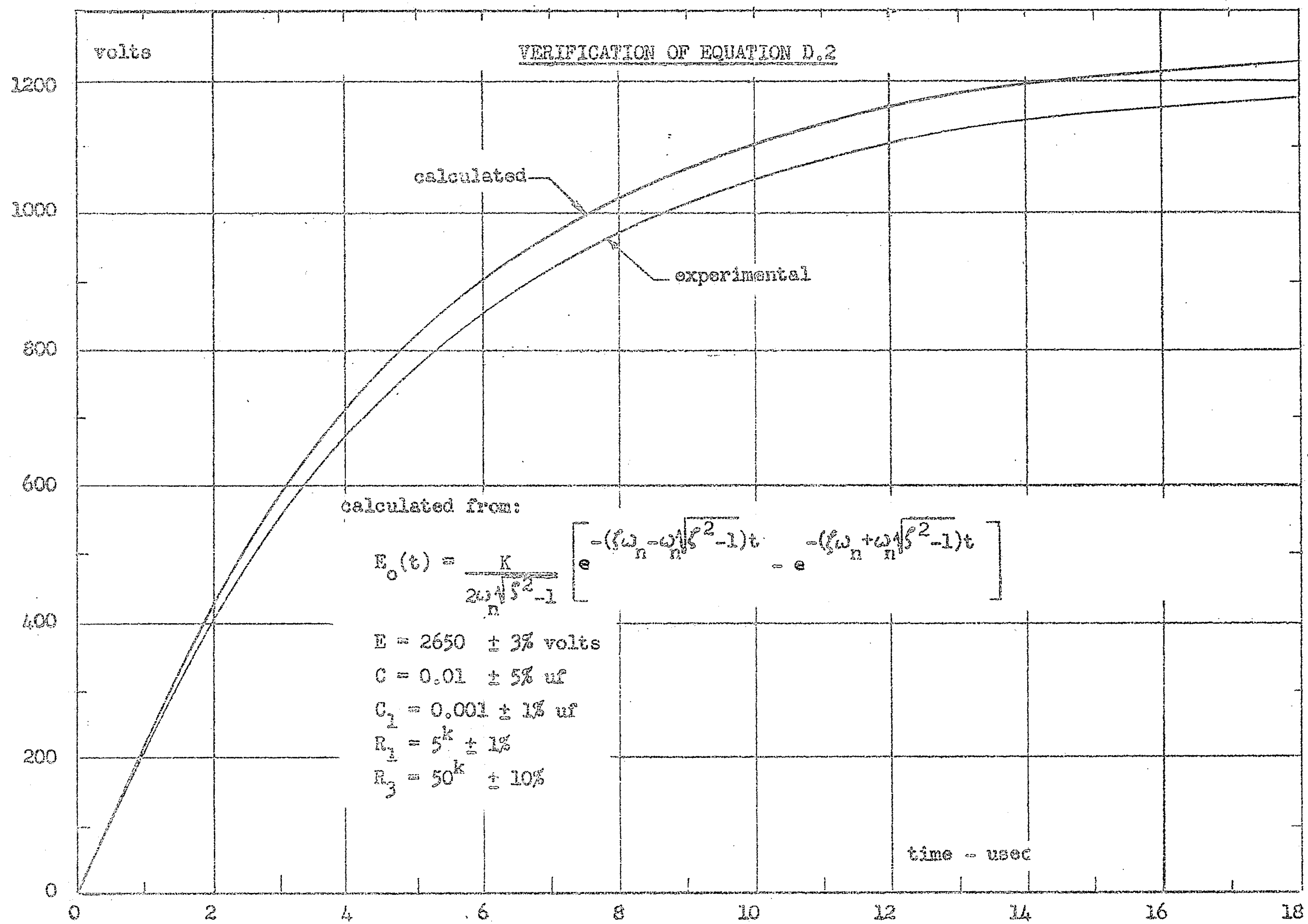

volts
VERIFICATION OF EQUATION D.2
1200
1000
800
600
400
200
0
calculated
experimental
calculated from:
$$E_o(t) = \frac{K}{2\omega_n\sqrt{\zeta^2-1}}\left[e^{-(\zeta\omega_n-\omega_n\sqrt{\zeta^2-1})t} - e^{-(\zeta\omega_n+\omega_n\sqrt{\zeta^2-1})t}\right]$$
$E = 2650 \pm 3\%$ volts
$C = 0.01 \pm 5\%$ uf
$C_1 = 0.001 \pm 1\%$ uf
$R_1 = 5^k \pm 1\%$
$R_3 = 50^k \pm 10\%$
time - usec
0
2
4
6
8
10
12
14
16
18

FIGURE D.3

A "Tektronix Type 533" oscilloscope was used to obtain the oscillogram and the accuracy of both the vertical calibrator and the horizontal deflection system was listed by the manufacturer as ± 3 per cent. It would appear that stray capacitance and inductance have a minor effect on the rise time of the sweep waveform.

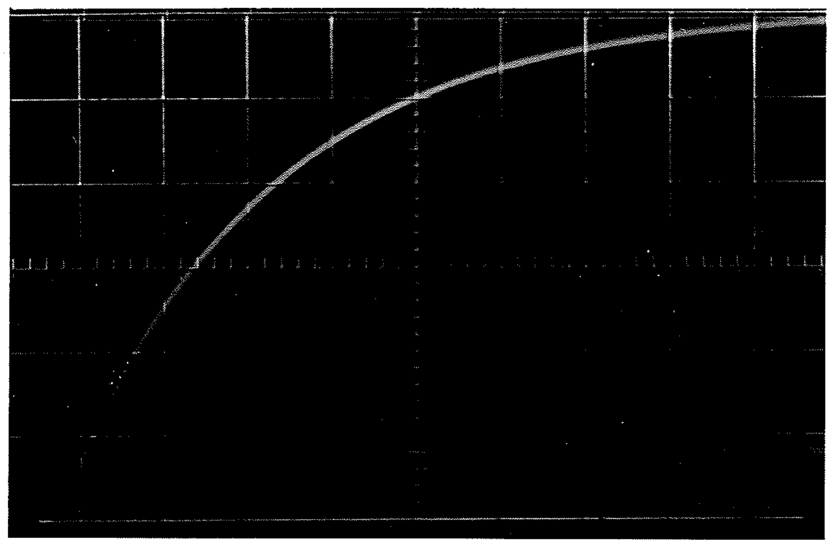


FIGURE D.4

POSITIVE SWEEP WAVEFORM

VERTICAL SCALE: 200 V/cm

HORIZONTAL SCALE: 2 usec/cm

The following oscillograms were taken when the sweep circuit was operated recurrently at 60 cycles per second. The "Tektronix Type 533" oscilloscope was used to obtain these oscillograms. Figure D.5

presents an oscillogram which shows the relative shape of the brightening pulse and the positive sweep waveform. The horizontal scale is 2 usec/cm and the vertical scale is 20 volts/cm and 200 volts/cm for the brightening pulse and the sweep waveform respectively. The values of the sweep circuit components were set as follows; $R_1 = 5^k$, $R_3 = 50^k$, $C_1 = 0.001$ uf, and $C = 0.01$ uf (see figure D.1). From figure D.5, it is seen that the amplitude of the brightening pulse is relatively constant over the duration of the sweep waveform. Figure D.6 shows the overall shape of the brightening pulse and the sweep voltage waveform. This oscillogram illustrates the history of the discharge of capacitor C in the sweep circuit. The horizontal scale is 0.5 msec/cm and the vertical scale is 200 volts/cm and 20 volts/cm for the sweep waveform and brightening pulse respectively. It is seen that both waveforms have the same overall shape. In figure D.7, the two sweep waveforms are illustrated. From an examination of these oscillograms it can be seen that the sweep waveforms are well balanced. The sweep circuit component values were set as follows; $R_1 = 5^k$, $R_3 = 10^k$, $C_1 = 0.001$ uf, and $C = 0.01$ uf (see figure D.1)

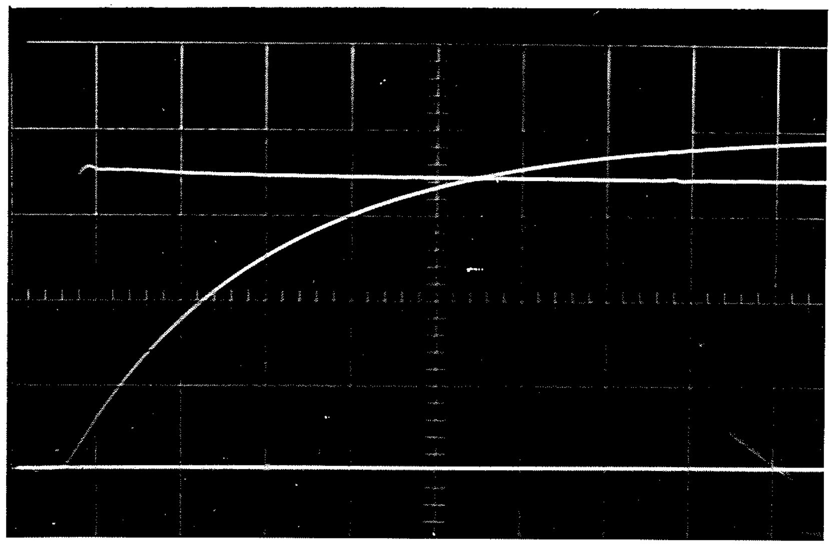


FIGURE D.5

BRIGHTENING PULSE AND SWEEP WAVEFORM

HORIZONTAL SCALE: 2 usec/cm
VERTICAL SCALE: 20 volts/cm - BRIGHTENING PULSE
200 volts/cm - SWEEP WAVEFORM

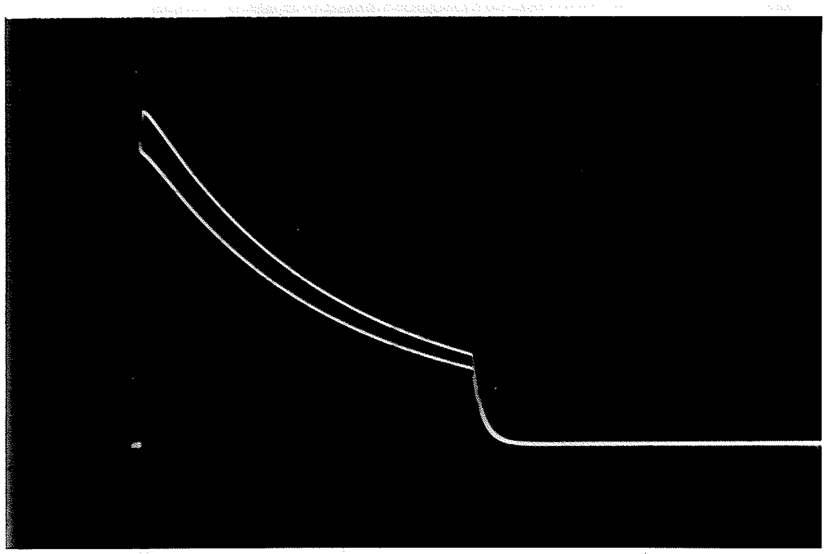
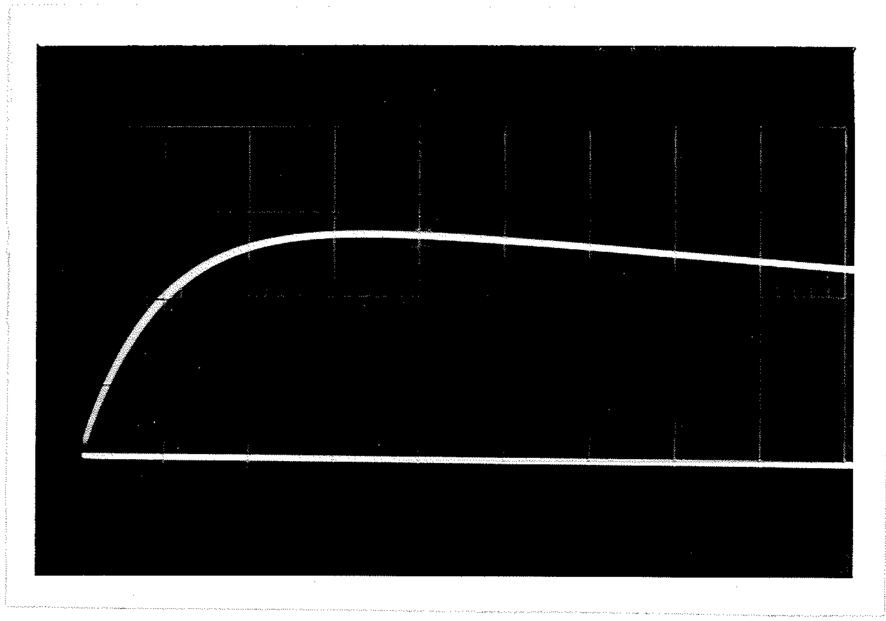


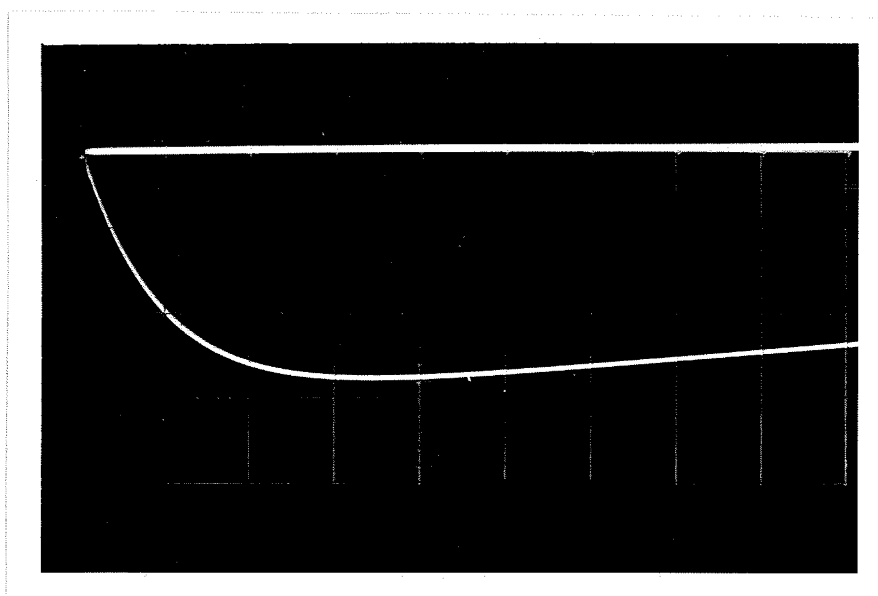
FIGURE D.6

BRIGHTENING PULSE AND SWEEP WAVEFORM

HORIZONTAL SCALE: 0.5 msec/cm
VERTICAL SCALE: 20 volts/cm - BRIGHTENING PULSE
200 volts/cm - SWEEP WAVEFORM



POSITIVE SWEEP VOLTAGE WAVEFORM



NEGATIVE SWEEP VOLTAGE WAVEFORM

VERTICAL SCALE: 375 volts/cm
HORIZONTAL SCALE: 5 usec/cm

FIGURE D.7