THE ANALYSIS AND DEVELOPMENT OF A CASCODE SWEEP GENERATOR CIRCUIT

by

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A Thesis Submitted to the Faculty of the

DEPARTMENT OF ELECTRICAL ENGINEERING

In Partial Fulfillment of the Requirements
For the Degree of

MASTER OF SCIENCE

In the Graduate College

UNIVERSITY OF ARIZONA

1958

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Chapter 1.

INTRODUCTION

A linear time base circuit provides a waveform which exhibits a linear voltage variation with time. This type of circuit has extensive application in laboratory oscilloscopes, radar and television indicators, in precision time measurements, and in time modulation. This type of waveform is customarily referred to as a sweep waveform even though it is used in many applications other than the horizontal deflection of an electron beam.

In order to classify the cascode sweep generator, the different types of sweep generators may be divided into four general types.

The first type is the gas discharge or the thyratron sweep generator. It is particularly noted for its simplicity of operation. This circuit uses the special characteristic of a thyratron to automatically charge and discharge a sweep capacitor. The method of obtaining linearity is to use only a small linear portion of the total capacitor charging curve by discharging the capacitor before it can be fully charged. The linearity of this sweep depends upon the RC charging time constant of the sweep capacitor and the duration of the sweep voltage. Perfect linearity cannot be obtained by this method because some exponential curvature in the charging curve of the capacitor always exists.

A multivibrator type of sweep generator is the second type of sweep circuit. This type of generator uses a monostable multivibrator to gate a switch tube which controls the charging and discharging of the sweep capacitor. The method of obtaining linearity in this type of circuit is identical with that described for the thyratron sweep generator. Here again, the linearity is dependent upon the RC charging time constant of the sweep capacitor and the duration of the sweep voltage.

The third type of sweep generator is similar to the second type except that the constant current characteristics of a pentode or transistor are used in conjunction with the switch tube in order to obtain linear charging of the sweep capacitor. Here the linearity is dependent upon the constant current characteristics of the pentode or transistor. This circuit is capable of generating a much larger sweep voltage than the first and second type and still have very good linearity.

The fourth type of sweep generator is classed as those circuits employing a method of feedback to linearize the sweep output. The Miller integrator and the bootstrap circuit are of this type. These circuits are both more complicated than those of the first three types. The Miller integrator, used in conjunction with a switch tube, uses a feedback integrator in a way that maintains the voltage across the charging resistor constant. Usually a cathode follower is used in conjunction with the feedback integrator in order to reduce the flyback time.

The bootstrap cathode follower type of circuit is used in conjunction with a switch tube. The basic principle involved is using positive feedback in a way that maintains the voltage across the charging resistor constant. Both the Miller integrator and the bootstrap cathode follower are basically very similar. The linearities obtainable in these two circuits are by far the best of the four different types, but the circuits are also more complex.

The cascode sweep generator circuit of this thesis is classed as a multivibrator type of sweep generator. The linearity of the sweep is also dependent upon the RC charging time constant of the sweep capacitor and the time duration of the sweep. This circuit uses a new and different type of multivibrator circuit in conjunction with a sweep capacitor, thereby eliminating the need for a separate switch tube. The circuit can be made either free running or monostable by a small resistor adjustment. The simplicity of the circuit results in low design costs and still retains many advantages of more complicated circuits.

A practical transistor circuit which operates on the same principle as the tube circuit has been designed by the author. A complete analysis of both circuits is presented in this thesis.

Chapter 2.

ANALYSIS OF THE TUBE CIRCUIT

2.1 INTRODUCTION

The purpose of this chapter is to describe the circuit operation and to perform a detailed analysis of the circuit. The analysis will include equivalent circuits and design equations.

2.2 PRINCIPLE OF OPERATION

The circuit shown in figure 2.1 employs regenerative switching similar to a basic multivibrator circuit. The sweep generator can be made to operate either driven (monostable) or recurrent (free running), by making a small variation in R_4. To explain the operation of the circuit, it is assumed that the circuit is adjusted for monostable operation. The circuit has three states of operation, the normal state, unstable state, and the quasi-stable or timing state.

In its normal state, V_1 is conducting and V_2 is cutoff by the voltage drop across R_4. See figure 2.3 for the equivalent circuit. With V_2 cutoff, there is no current flow in R_2, which means that V_1 is essentially biased for operation at zero volts. When a negative trigger of large enough magnitude is applied to the cathode of V_2, V_2 starts to conduct. The circuit now operates in an unstable state, which causes regenerative switching to the timing state. The equivalent circuit for the unstable state is shown in figure 2.2. The loop gain must be greater than one to assure that the circuit switches from the normal state to the timing state. The loop gain, or switching time, can be determined by any standard method of analysis.[1]

After the regenerative switching is complete, the circuit is in its timing state, V_1 off and V_2 on. The equivalent circuit for the timing state is shown in figure 2.4. As seen from figure 2.4, the only source of B+ for V_2 is the charge on capacitor C_2. This capacitor discharges through the circuit and biases V_1 into cutoff. V_1 remains cutoff until such time as discharge current through R_2 is small enough that e_{g1} will no longer hold V_1 cutoff. During this time, capacitor C_1 will charge exponentially toward Ebb and provide a sweep output voltage. The duration of the sweep (T_s), is controlled by the discharge time constant of C_2. This time is such that only a small portion of the charging curve for C_1 is used and thereby a linear sweep is attained.

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[1] J. Millman and H. Taub, Pulse and Digital Circuits, McGraw-Hill Book Co., Inc., 1956, pp. 147-152

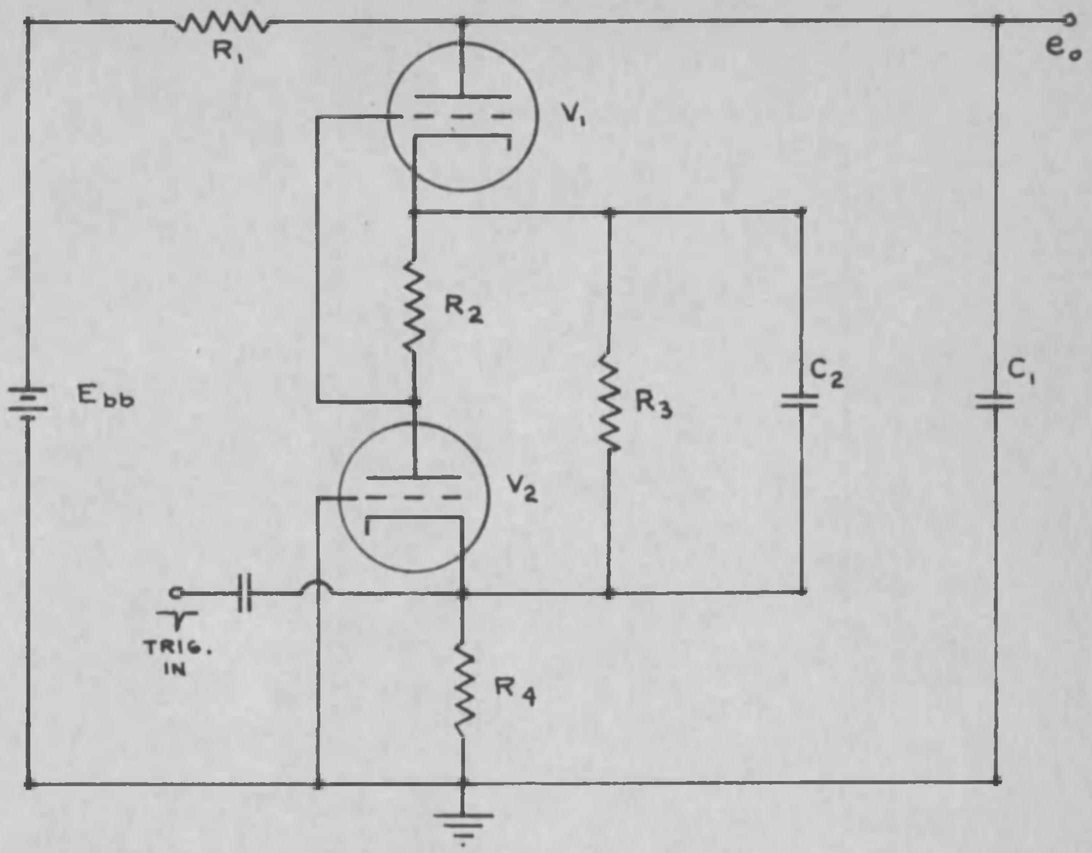


FIG. 2.1 CASCODE SWEEP GENERATOR CIRCUIT

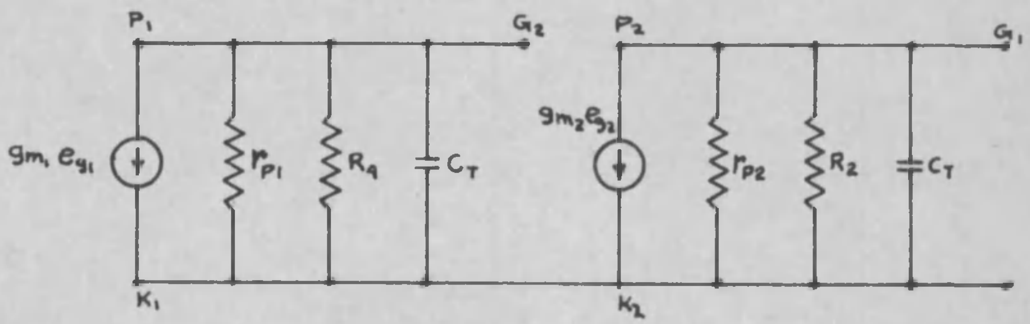


FIG. 2.2 UNSTABLE STATE A.C. EQUIVALENT CIRCUIT

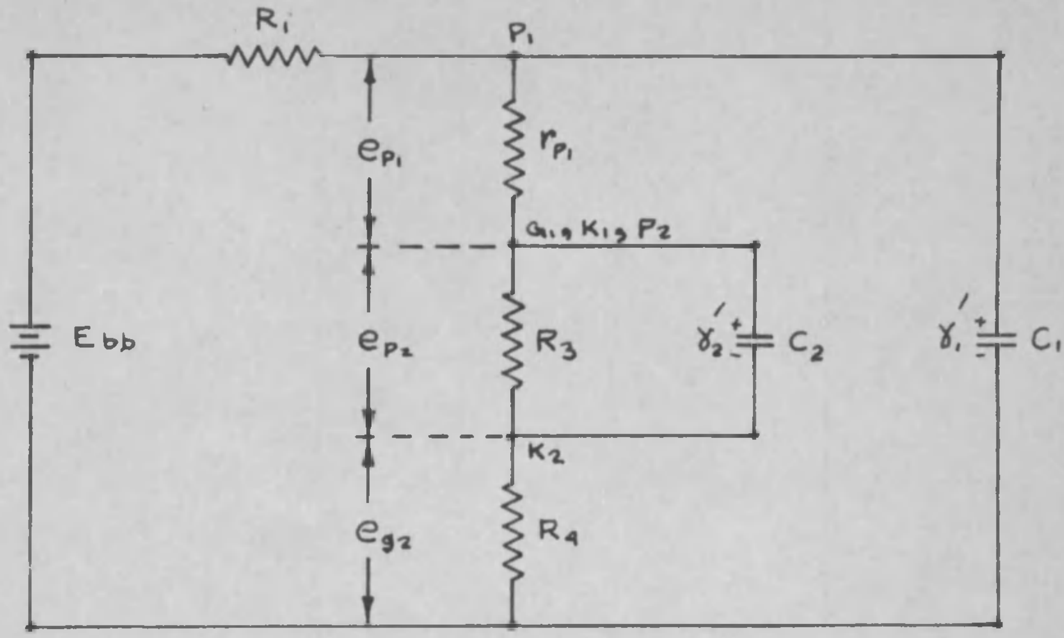


FIG. 2:3 NORMAL STATE EQUIVALENT CIRCUIT
V_1 ON, V_2 OFF

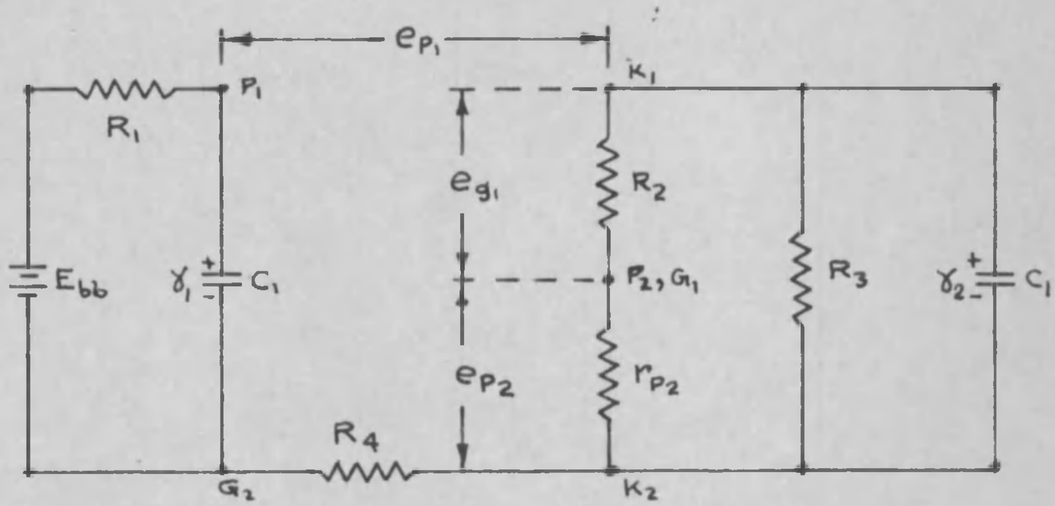


FIG. 2.4 TIMING STATE EQUIVALENT CIRCUIT
V_1 OFF, V_2 ON

At the end of the timing state, the circuit will self trigger into the unstable state. Regenerative switching action again takes place and the circuit switches to the normal state. The instant the circuit reaches the normal state, capacitors C_1 and C_2 discharge and charge respectively to their steady state condition. This interval of time is known as the flyback time and is discussed in more detail in section 2.4.

2.3 DERIVATION OF THE DESIGN EQUATIONS

Assume that the switching time (from state to state) is fast enough so that the charges on C_1 and C_2 do not change during this interval. Then, the initial charges on C_1 and on C_2 in the timing state equal the final charges on C_1 and C_2 in the normal state. Correspondingly, the initial charges in the normal state equal the final charges in the timing state. From the normal state,

$$\gamma_1 = \frac{E_{bb}(r_{p1} + R_3 + R_4)}{R_1 + r_{p1} + R_3 + R_4} \qquad (2.1)$$

$$\gamma_2 = \frac{E_{bb} R_3}{R_1 + r_{p1} + R_3 + R_4} \qquad (2.2)$$

From the timing state,

$$\gamma_1' = E_{bb} - (E_{bb} - \gamma_1)\, \varepsilon^{-T_s/R_1C_1} \qquad (2.3)$$

$$\gamma_2' = \gamma_2\, \varepsilon^{-T_s/R_2'C_2} \qquad (2.4)$$

Where T_s is the time duration of the sweep and

$$R_2' = \frac{R_2\, r_{p2}}{R_2 + r_{p2}} \qquad (2.5)$$

The amplitude of the sweep output voltage, e_o, will be the total change in voltage on C_1 during the timing state. Therefore, from the timing state equivalent circuit,

$$e_o = e_{c1} - \gamma_1 \qquad (2.6)$$

$$e_{c1} = E_{bb} - (E_{bb} - \gamma_1)\, \varepsilon^{-t/R_1C_1} \qquad (2.7)$$

$$e_o = (E_{bb} - \gamma_1)(1 - \varepsilon^{-t/R_1C_1}) \qquad (2.8)$$

Substituting equations (2.1) into (2.8) and reducing we have,

$$e_o = \frac{E_{bb}\, R_1\, (1 - \varepsilon^{-t/R_1C_1})}{R_1 + r_{p1} + R_3 + R_4} \qquad (2.9)$$

From the figure 2.4 it can be seen that the plate voltage of the cutoff tube V_1 is not constant, but varies with time. This also means that the voltage required to cut off V_1 also varies. To determine this cutoff voltage, the following approximation is made:

$$E_{co_1} = \frac{E_{p_1}}{\mu_{co_1}} \qquad (2.10)$$

where E_{co} is the cutoff voltage of V_1, E_{p1} equals the plate voltage of V_1 and μ_{co_1} is the cutoff Mu of V_1. It should be noted that this approximation is applicable only to triodes. The amount of error resulting from use of the approximation depends upon the specific tube type. Figure 2.5 shows how μ_{co_1} is determined from the plate characteristics, and also indicates the amount of error involved in the approximation. From figure 2.4, the time required for e_{g1} to equal the cutoff voltage of E_{co1} is the duration of the timing state. Therefore,

$$e_{g_1} = E_{co_1} = \frac{E_{p_1}}{\mu_{co_1}} \qquad (2.11)$$

From figure 2.4,

$$e_{g_1} = \frac{R_2 \gamma_2}{R_2 + r_{p_2}} \qquad (2.12)$$

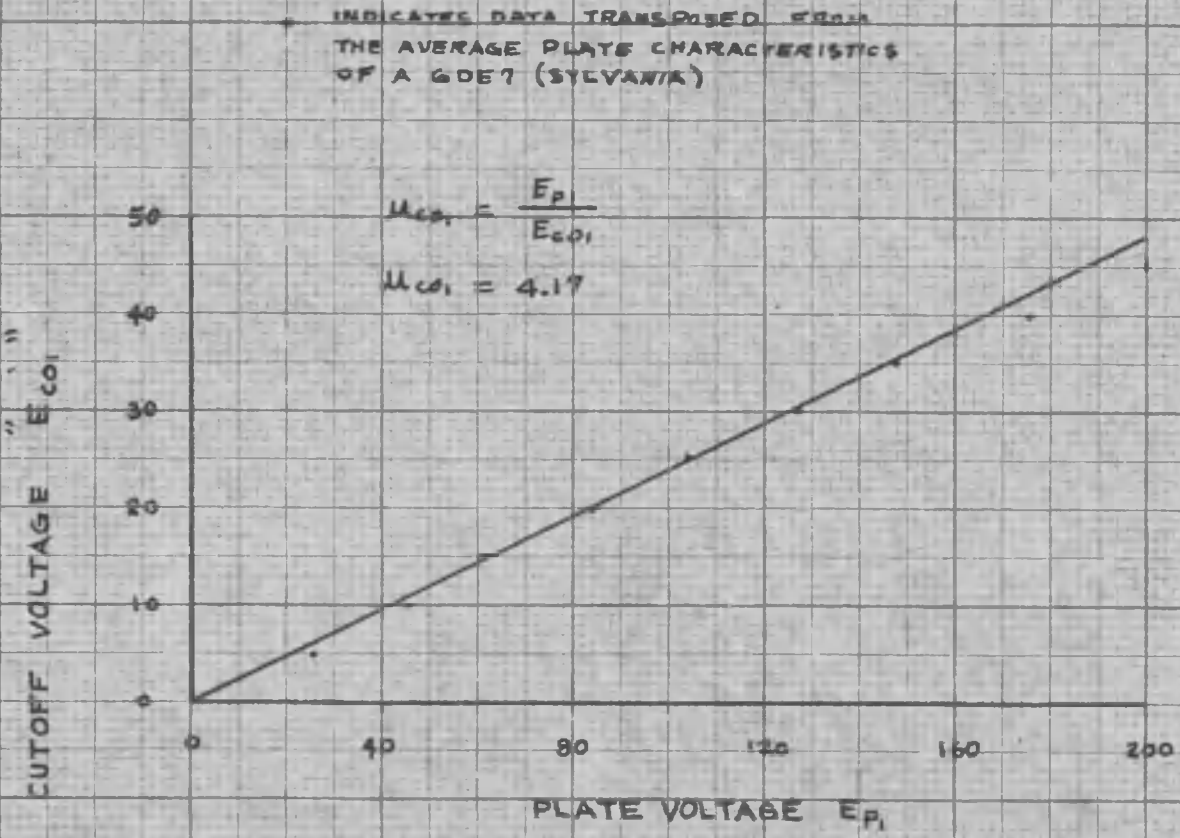


FIG. 2.5 CUTOFF Mu "μ_{co_1}" FOR 6DE7

Substitute equation (2.2) in (2.12).

$$e_{g_1} = \frac{R_2 R_3 E_{bb}}{(R_2 + r_{p_2})(R_1 + r_{p_1} + R_3 + R_4)} \varepsilon^{-t/R_2'C_2} \tag{2.13}$$

From figure 2.4,

$$e_{p_1} = E_{bb} - (E_{bb} - \gamma_1)\varepsilon^{-t/R_1C_1} - \gamma_2 \varepsilon^{-t/R_2'C_2} \tag{2.14}$$

Substitute equations (2.1) and (2.2) in (2.14) and reduce.

$$e_{p_1} = E_{bb} - \frac{E_{bb} R_1 \varepsilon^{-t/R_1C_1}}{R_1 + r_{p_1} + R_3 + R_4} - \frac{E_{bb} R_3 \varepsilon^{-t/R_2'C_2}}{R_1 + r_{p_1} + R_3 + R_4} \tag{2.15}$$

Substitute equations (2.13) and (2.15) in (2.11) and reduce.

$$\varepsilon^{t/R_2'C_2} = \frac{R_3\left[R_2(\mu_{co_1} + 1) + r_{p_2}\right]}{(R_2 + r_{p_2})\left[R_1\left(1 - \varepsilon^{-t/R_1C_1}\right) + r_{p_1} + R_3 + R_4\right]} \tag{2.16}$$

An equation for the sweep voltage linearity will now be set up. From equation (2.9)

$$e_o = \frac{E_{bb} R_1}{R_1 + r_{p_1} + R_3 + R_4}\left(1 - \varepsilon^{-t/R_1C_1}\right) \tag{2.9}$$

it is seen that the linearity is wholly dependent upon the term

$$\left(1 - \varepsilon^{-t/R_1C_1} \right)$$

Using a standard method analysis[2] to analyse the linearity of this sweep output voltage, the following equation results:

$$\varepsilon_d = \frac{T_s}{8 R_1 C_1} \tag{2.17}$$

Here e_d is the displacement error, and T_s is the duration of the sweep. The displacement error is defined as the maximum difference between the actual sweep voltage and a linear sweep which passes through the beginning and end points of the actual sweep. (see figure 2.6). In equation form,

$$\varepsilon_d \equiv \frac{(e_s - e_s')\,max}{E_s}$$

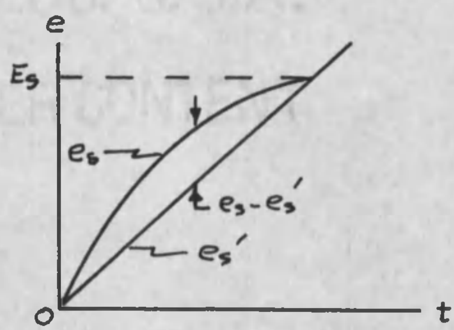


FIG. 2.6 DISPLACEMENT ERROR DEFINITION

[2] J. Millman and H. Taub, Pulse and Digital Circuits, McGraw-Hill Book Co., Inc., 1956, pp. 147-152

To change e_d into percent multiply by 100.

Further reduction of the equation for e_o can be accomplished by approximation using the following series:

$$\left(1-\varepsilon^{-t/R_1C_1}\right) = \frac{t}{R_1C_1} - \frac{1}{2}\left(\frac{t}{R_1C_1}\right)^2 + \frac{1}{6}\left(\frac{t}{R_1C_1}\right)^3 - - - - -$$

When the deviation from linearity is very small, the above equation can be approximated by the first term of the series with little error. Therefore,

$$\left(1-\varepsilon^{-t/R_1C_1}\right) \cong \frac{t}{R_1C_1} \tag{2.18}$$

Substitute equation (2.18) into (2.9) and (2.16).

$$e_o = \frac{E_{bb}\ R_1}{R_1 + r_{p1} + R_3 + R_4}\left(\frac{t}{R_1C_1}\right) \tag{2.19}$$

$$\varepsilon^{t/R_2'C_2} = \frac{R_3\left[R_2(\mu_{co1}+1) + r_{p2}\right]}{(R_2 + r_{p2})\left[R_1\left(\frac{t}{R_1C_1}\right) + r_{p1} + R_3 + R_4\right]} \tag{2.20}$$

Equations (2.19) and (2.20) are applicable to any and all sweep durations. Therefore, let

$$t = T_s \tag{2.21}$$

Substitute equation (2.21) into (2.19) and (2.20)

$$E_o = \frac{8\, E_{bb}\, R_1\, \mathcal{E}_d}{R_1 + r_{p1} + R_3 + R_4} \tag{2.22}$$

$$\varepsilon^{T_s/R_2'C_2} = \frac{R_3\left[R_2(\mu_{co1}+1) + r_{p2}\right]}{(R_2 + r_{p2})(8R_1\mathcal{E}_d + r_{p1} + R_3 + R_4)} \tag{2.23}$$

From equation (2.17)

$$T_s = 8\,\mathcal{E}_d\, R_1 C_1 \tag{2.24}$$

Therefore,

$$\varepsilon^{\frac{8R_1C_1\mathcal{E}_d}{R_2'C_2}} = \frac{R_3\left[R_2(\mu_{co1}+1) + r_{p2}\right]}{(R_2 + r_{p2})(8R_1\mathcal{E}_d + r_{p1} + R_3 + R_4)} \tag{2.25}$$

or,

$$T_s = 8R_1C_1\mathcal{E}_d = R_2'C_2 \ln\left[\frac{R_3\left[R_2(\mu_{co1}+1) + r_{p2}\right]}{(R_2 + r_{p2})(8R_1\mathcal{E}_d + r_{p1} + R_3 + R_4)}\right] \tag{2.26}$$

By a process of trial and error substitution, equation (2.26) can be satisfied. Thus, when all values for equation (2.26) have been determined, the sweep duration is varied by changing C_1 and C_2. The linearity and output voltage amplitude should remain constant throughout the entire sweep range, as long as the ratio between C_1 and C_2 remains constant. In order for the circuit to be monostable and not free running, it must be assured that V_2 is cutoff during the entire normal state. For this condition,

$$e_{g_2} > E_{co_2} \tag{2.27}$$

From figure 2.3,

$$e_{g_2} = \frac{E_{bb} R_4}{R_1 + r_{p_1} + R_3 + R_4} \tag{2.28}$$

In the case of a triode,

$$E_{co_2} = \frac{E_{p_2}}{\mu_{co_2}} \tag{2.29}$$

Again from figure 2.3,

$$E_{p_2} = \frac{E_{bb} R_3}{R_1 + r_{p_1} + R_3 + R_4} \tag{2.30}$$

Substitute equations (2.28), (2.29), (2.30) into (2.27) and reduce.

$$\mu_{co_2} R_4 > R_3 \tag{2.31}$$

2.4 DISCUSSION OF THE FLYBACK TIME

As stated previously, the time required for C_1 and C_2 to recover to their steady state condition represents the flyback time. It is desirable to reduce the flyback time as much as possible. Because of the complexity involved, an equation for the flyback time has not been derived. Instead, a simplified discussion should be adequate for this analysis. Examination of the normal state equivalent circuit shows a second order system.

Based on the particular circuit parameters used in this circuit, the flyback voltage across C_1 can be represented by the following general equation:

$$e_{F.B.} = A_0 + A_1 \varepsilon^{-t/T_1} + A_2 \varepsilon^{-t/T_2}$$

From the equation, it is seen that the system has two different time constants, T_1 and T_2. In the present case, T_2 is large compared to T_1, and A_1 is large compared to A_2.

Then for all practical purposes, the third term of the equation can be neglected. It then follows that the flyback time constant is essentially T_1. T_1 is primarily depemdent upon $(r_{p1} + R_4)\, C_1$. This quantity should be as small as possible if a reduction in flyback time is important.

2.5 DISCUSSION OF CIRCUIT PARAMETERS

The value of R_1 is determined primarily from the equation,

$$E_o = \frac{8\, Ebb\, R_1\, e_d}{R_1 + r_{p1} + R_3 + R_4} \qquad (2.22)$$

From this equation, R_1 must be much larger than $(r_{p1} + R_3 + R_4)$ in order to maximize the sweep output voltage. If this is true, the output is practically dependent upon Ebb and the displacement error e_d. There are definite practical limitations in how large R_1 should be made. R_1 must remain small in comparison with the plate resistance of the tube in cutoff, in order for the equivalent circuits to be correct.

The input impedance of the next stage must be large in comparison with R_1 in order to prevent loading. R_1 indirectly affects the flyback time, because it has a direct effect on C_1. As previously discussed, the smaller the value of C_1 the smaller the flyback time. From equation (2.17), if the displacement error is held constant, an increase in R_1 causes a corresponding decrease in C_1.

The value of R_2 is determined primarily from the equation (2.26). As can be seen, increasing R_2 increases the sweep duration T_s. It should also be noted that R_2 does not affect the output voltage E_o. Therefore, this makes a convenient parameter to vary in order to change the sweep duration. There are practical limitations in how large R_2 can be made without other complications. Because R_2 is the load resistor of V_2, it has a definite effect on the speed with which the circuit switches from state to state. The recovery time of the circuit is also affected.[3] A general rule is to limit R_2 to a range near the value of the plate resistance of the tube.

The value of R_3 has an effect on the flyback voltage, the sweep duration, and the amplitude of the sweep output voltage. R_3 affects the flyback voltage because it directly affects R_4. In order for the circuit to be monostable, the conditions of equation (2.31) must be satisfied.

$$\mu_{co2} R_4 > R_3$$

If μ_{co2} is constant, then an increase in R_3 has to be balanced by an increase in R_4. By examination of equation (2.22) it is readily seen that an increase

[3] J. Millman and H. Taub, Pulse and Digital Circuits, McGraw-Hill Book Co., Inc., 1956, pp. 150-156

in R_3 causes a decrease in E_o.

In summarizing, an increase in R_3 causes E_o to decrease, flyback time to increase, and T_s to increase. Correspondingly, a decrease in R_3 causes E_o to increase, flyback time to decrease, and T_s to decrease.

Certain considerations are necessary in choosing the proper tube types for V_1 and V_2. For V_1, the plate resistance is of prime importance and must be as small as possible. This keeps the flyback time at a minimum. μ_{co1}, as described earlier in the chapter, should be as high as possible. The interelectrode capacitance should be as small as possible. A high perveance triode is the most practical type of tube. For V_2, a high Mu triode with low interelectrode capacitance is preferable. With high gain, the switching time and recovery time decreases. This in turn increases the high frequency range of the generator.

The value of Ebb depends primarily upon the amplitude of the sweep output voltage necessary.

Based on the preceding discussion, and upon experimentation, the following component values have been chosen. The tube used is a dual triode with dissimilar sections. One section is a high perveance triode, while the other is only a medium Mu triode. A high Mu triode would be more desirable, but for space limitations a 6DE7 might be preferable.

$R1 = 50K$ $R_4 = 760$ ohms

$R_2 = 14.5K$ V_1 = Section 2 of 6DE7

$R_3 = 10K$ V_2 = Section 1 of 6DE7

$e_d = .0125$

From plate characteristic curves and tube data, the tube parameters are:

$r_{p1} = 985$ ohms $\mu_{co1} = 4.17$

$r_{p2} = 10.2K$ $\mu_{co2} = 13.8$

2.6 SUMMARY OF THE DESIGN PROCEDURE OF THE TUBE CIRCUIT

1. Specify the design requirements of the circuit.
   a. Sweep duration
   b. Sweep linearity (displacement error)
   c. Sweep output voltage amplitude
2. Choose the types of tubes to be used.
3. From the plate characteristic curves of the tubes determine the cutoff Mu and the plate resistance at zero grid bias.
4. Keeping section 2.4 thoroughly in mind, adjust the values of the circuit until equation (2.26) has been satisfied.
5. Substitute in equation (2.31) to see that this is also satisfied. If not, adjust R_4 and go back to step 4.
6. Solve for the sweep output amplitude from equation

(2.26). If $C_1 = C_2$, then the sweep duration is easily adjusted by varying C_1 and C_2.

Chapter 3

ANALYSIS AND DEVELOPMENT OF THE TRANSISTORIZED CIRCUIT

3.1 INTRODUCTION

The purpose of this chapter is to discuss some of the development problems of the transistor circuit, as well as to analyse the circuit. This circuit is designed to operate on the same principle as the tube circuit. After extended study and experimentation, the circuit shown in figure 3.1 was developed.

3.2 PRINCIPLE OF OPERATION

Operating as a monostable generator, the circuit has 3 stable states, the normal, unstable, and timing state. In the normal state, T_1 conducts and T_2 is cutoff. The voltage drop across R_4 must be larger than that across R_7. The base to emitter junction is then reversed biased, keeping T_2 in cutoff. When a trigger is applied causing T_1 to conduct, this causes the base to ground voltage of T_1 to drop low enough so that the base to emitter junction of T_1 is reversed biased. This cuts off T_1 and the circuit is in its timing state. See figure 3.3 for the timing state equivalent circuit.

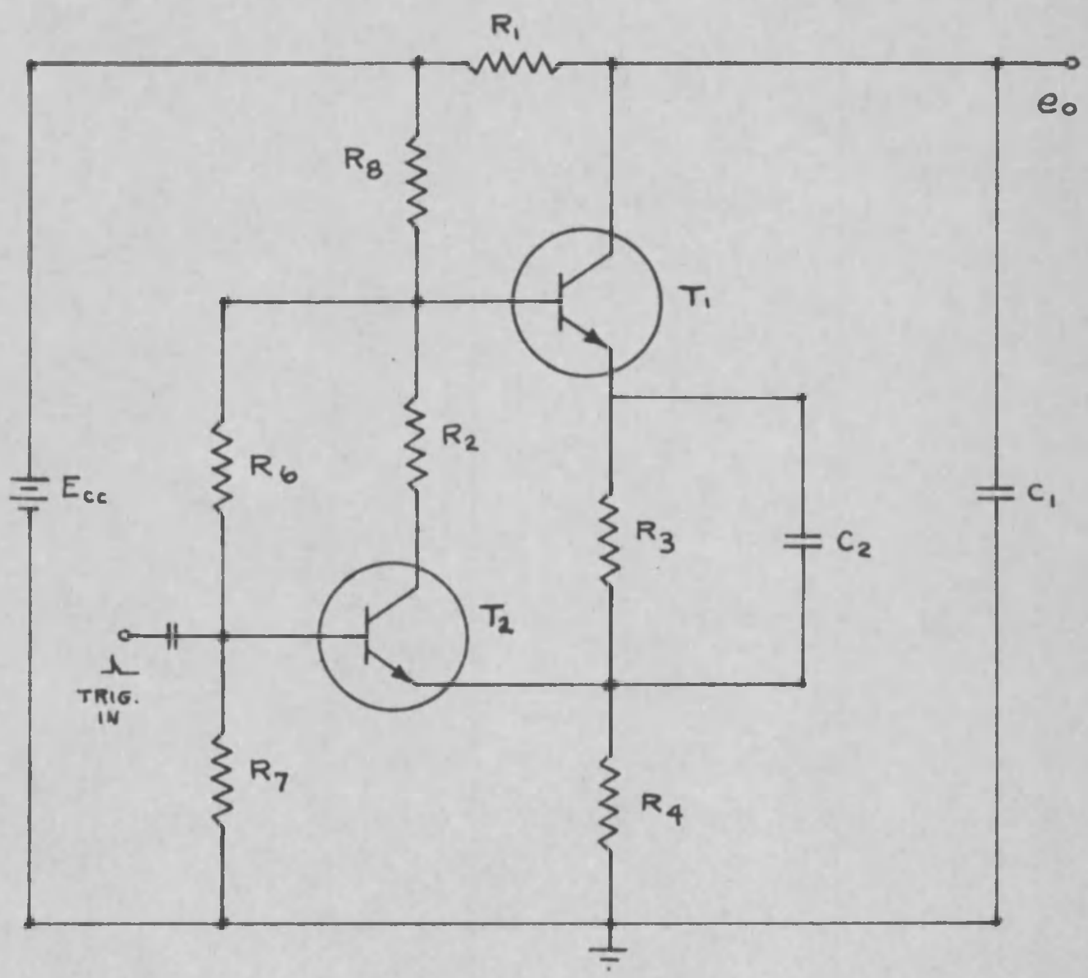


FIG. 3.1 CASCODE SWEEP GENERATOR CIRCUIT

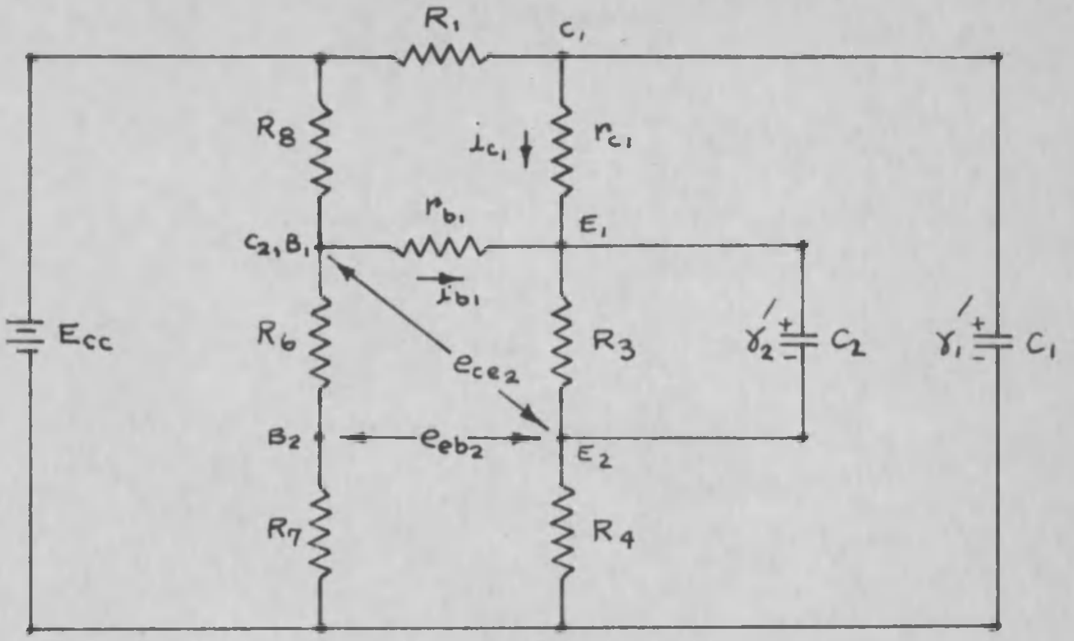


FIG. 3.2 NORMAL STATE EQUIVALENT CIRCUIT
T_1 ON, T_2 OFF

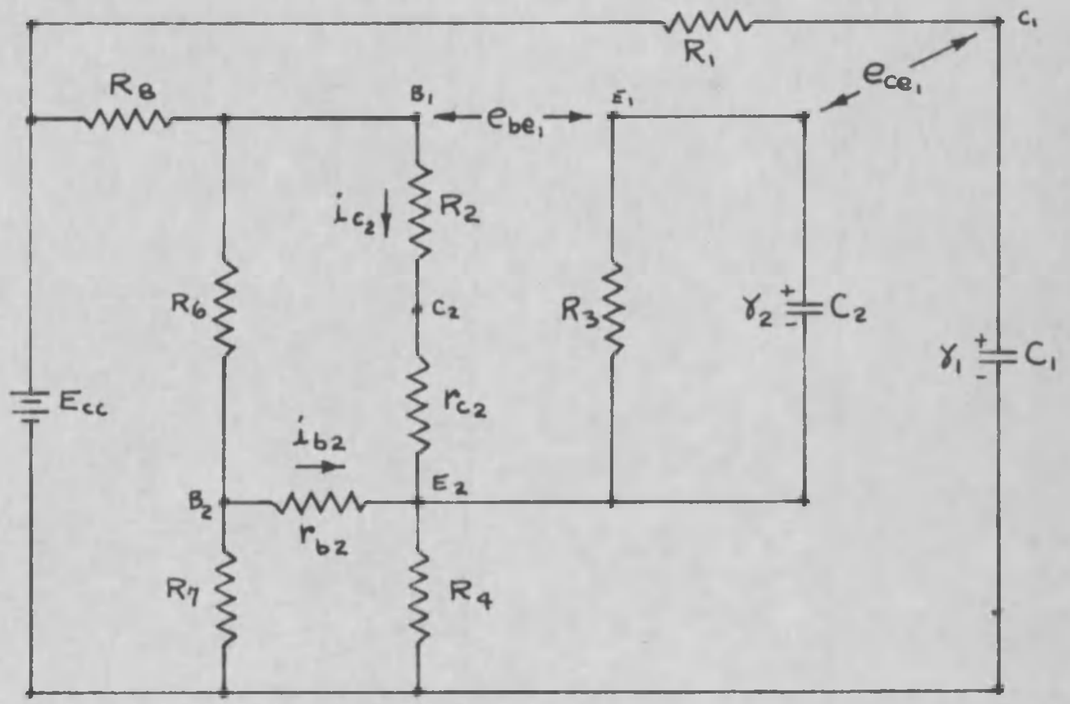


FIG. 3.3 TIMING STATE EQUIVALENT CIRCUIT
T_1 OFF, T_2 ON

The circuit remains in this state until C_2 discharges enough so that the base to ground voltage of T_1 equals the emitter to ground voltage of T_1, or in other words, until the base to emitter voltage reaches zero. When this condition is met, T_1 starts to conduct and the circuit returns to its normal state. The a-c equivalent circuit for the unstable state is shown in figure 3.4.

The basic operational difference between the tube and transistor circuit is that the collector of T_2 is biased from a fixed voltage developed by a voltage dividing network. This voltage remains constant during the timing state. In the tube circuit, V_2 receives its plate voltage primarily from the charge on C_2. This voltage varies during the sweep duration. All in all, the two circuits are very similar.

3.3 DERIVATION OF THE DESIGN EQUATIONS

By examination of the equivalent circuits for the normal and timing state, equations (2.8), (2.17), (2.18), and (2.24) derived for the tube circuit also apply to the transistor circuit. An equation for γ_1 is all that is necessary to solve for the equation of the sweep output voltage. The sweep duration is the interval of time it takes e_{be_1} to reach zero volts. Therefore, an equation for e_{be_1}, set equal to zero yields the sweep duration T_s. This requires an equation for γ_2 which can

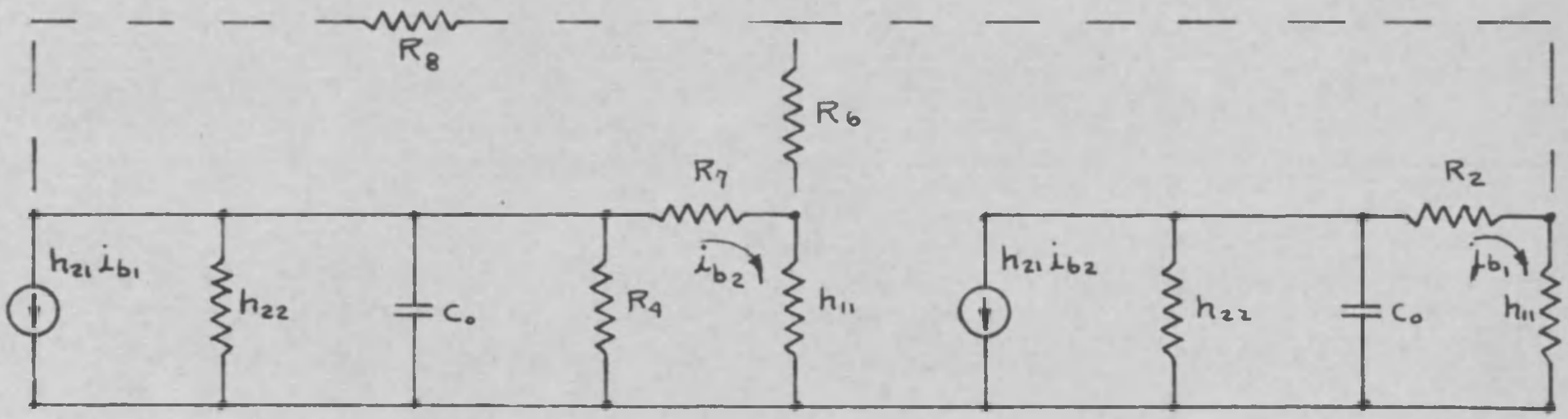


FIG. 3.4 UNSTABLE STATE A.C. EQUIVALENT CIRCUIT

be solved from the normal state equivalent circuit.

In addition to these two equations, an equation for e_{eb_2} in the normal state must be known in order to assure monostable operation.

When using transistors, it is necessary to know what the base current and collector currents are in order to determine the proper bias condition. Transistor biasing is covered in detail in section 3.4.

Also, in this case, equations for the collector to emitter voltage must be known in order to assure that the maximum ratings are not exceeded.

Making the approximations,

$$r_{b1}(R_8+R_6+R_7) \ll (R_6+R_7)(R_1+R_8),$$

$$r_{b1} \ll R_1, R_8, (R_6+R_7), (R_3+R_4) \quad \text{and that}$$

e_{ce1} is negligible, the following equations can be derived from the normal state equivalent circuit using determinants or any other method.

$$\gamma_1 = \frac{(R_6+R_7)(R_1+R_8)(R_3+R_4)\,E_{cc}}{D_1} \tag{3.1}$$

where,

$$D_1 = R_1 R_8 (R_3+R_4+R_6+R_7) + (R_1+R_8)(R_3+R_4)(R_6+R_7) \tag{3.2}$$

$$\gamma_2 = \frac{(R_1+R_8)(R_6+R_7)\,R_3\,E_{cc}}{D_1} \tag{3.3}$$

If, $e_{be_1} << e_{R_3}$

$$e_{ce_2}(max) = \gamma_2 \qquad (3.4)$$

also if, $r_{b_1}(R_1+R_3+R_4) << (R_3+R_4)(R_1+R_8)$

$$e_{eb_2} = \frac{E_{cc}(R_1+R_8)\left[R_4(R_6+R_7) - R_7(R_3+R_4)\right]}{D_1} \qquad (3.5)$$

e_{eb2} must be positive for monostable operation.

$$i_{b_1} = \frac{E_{cc}\left[R_1(R_6+R_7) - R_8(R_3+R_4)\right]}{D_1} \qquad (3.6)$$

$$i_{c_1} = \frac{E_{cc}\left[R_8(R_3+R_4+R_6+R_7) + r_{b_1}(R_6+R_7)\right]}{D_1} \qquad (3.7)$$

From the equivalent circuit of the timing state the following equations may be derived:

$$i_{b2} = \frac{E_{cc}(R_2R_7 - R_4R_6)}{D_2} \qquad (3.8)$$

where,

$$D_2 = (R_4+R_7+r_{b2})\left[R_2(R_6+R_8) + R_6R_8\right]$$

$$+ R_4R_7(R_2+R_6+r_{b2}) + r_{b2}(R_4R_6+R_2R_7) \qquad (3.9)$$

$$i_{c_2} = \frac{E_{cc}\left[R_6(R_4+R_7+r_{b2}) + R_7r_{b2}\right]}{D_2} \qquad (3.10)$$

$$e_{R2} = \frac{E_{cc}\, R_2 \left[R_2 (R_4 + R_7 + r_{b2}) + R_7\, r_{b2} \right]}{D_2} \tag{3.11}$$

$$e_{c2} = \gamma_2\, \varepsilon^{-t/R_3 C_2} \tag{3.12}$$

When T_2 is operating in the saturation region, r_{c2} is very small. Therefore, it has been assumed that r_{c2} is a short circuit.

The circuit switches from the timing to the normal state when,

$$e_{c2} = e_{R2} \tag{3.13}$$

Therefore by substituting equations (3.12) and (3.11) into (3.13) and solving for T_s,

$$T_s = R_3 C_2 \ln \gamma_2 / e_{R2} \tag{3.14}$$

But,

$$T_s = 8\, \varepsilon_d\, R_1 C_1 \tag{2.24}$$

is also applicable to the transistor circuit. Therefore,

$$T_s = 8\, \varepsilon_d\, R_1 C_1 = R_3 C_2 \ln \gamma_2 / e_{R2} \tag{3.15}$$

If $C_1 = C_2$ then,

$$8\, \varepsilon_d\, R_1 = R_3 \ln \gamma_2 / e_{R2} \tag{3.16}$$

Substitute equations (2.18), (2.21), (2.17) into equation (2.8).

$$E_o = \left(E_{cc} - \gamma_1\right) 8 \varepsilon d \qquad (3.17)$$

Substitute equation (3.1) into (3.15) and reduce.

$$E_o = \frac{8 \varepsilon d \, E_{cc} \, R_1 R_8 \left(R_3 + R_4 + R_6 + R_8\right)}{D_1} \qquad (3.18)$$

Assuming ER_4 is negligible compared to γ_1',

$$e_{ce_1}(max) = \gamma_1' - e_{R_2} \qquad (3.19)$$

where;

$$\gamma_1' = E_o + \gamma_1 \qquad (3.20)$$

3.4 TRANSISTOR BIASING

T_2 uses a straight-forward biasing system with current feedback stabilization. The bias on this transistor must be accurately adjusted for operation barely into saturation during the timing state. This is extremely important if fast switching and sharp transition between the timing and normal state is desired[4]. From the bias design equations in section 3.3 and the curves

[4]J. Millman and H. Taub, Pulse and Digital Circuits, McGraw-Hill Book Co., Inc., 1956, p. 597

of the input and output characteristics (figures 3.5 through 3.10) proper biasing can be determined.

The biasing of transistor T_2 is a little more complicated. From the equivalent circuit of the normal state, it is seen that the collector supply voltage varies during the flyback portion of the normal state. This voltage is large at the beginning of the flyback and relatively small at the end. Because of this varying voltage, it is difficult to bias the transistor so that it is operating barely into saturation throughout the entire normal state.

At first, it might seem that a self bias dependent upon the collector voltage is the most desirable type. This type of bias is not practical when a transistor is operating in saturation where the collector voltage is small. This also does not make a good switch with a bias resistor between the collector and base. Therefore, a bias with current feedback stabilization has been used. With this form of bias, i_{b1} remains relatively constant throughout the normal state. Now the question is, how can this bias be used efficiently? There are two methods. The first is to adjust the bias so that saturation is reached at the end of the flyback time. When this is done, T_1 operates as a constant current generator causing C_1 to discharge linearily

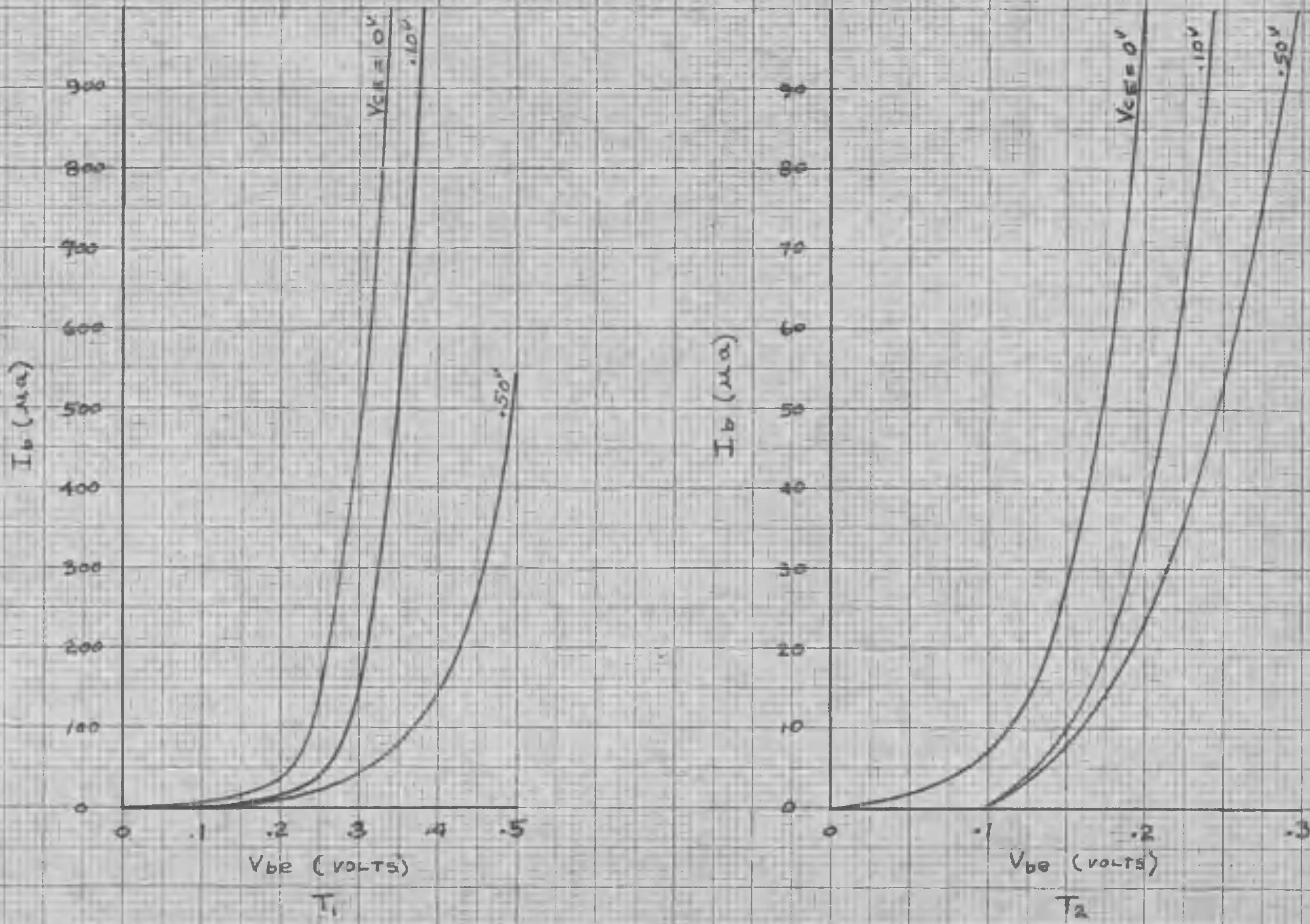


FIG. 3.5 INPUT CHARACTERISTICS FOR 2N167 TRANSISTOR

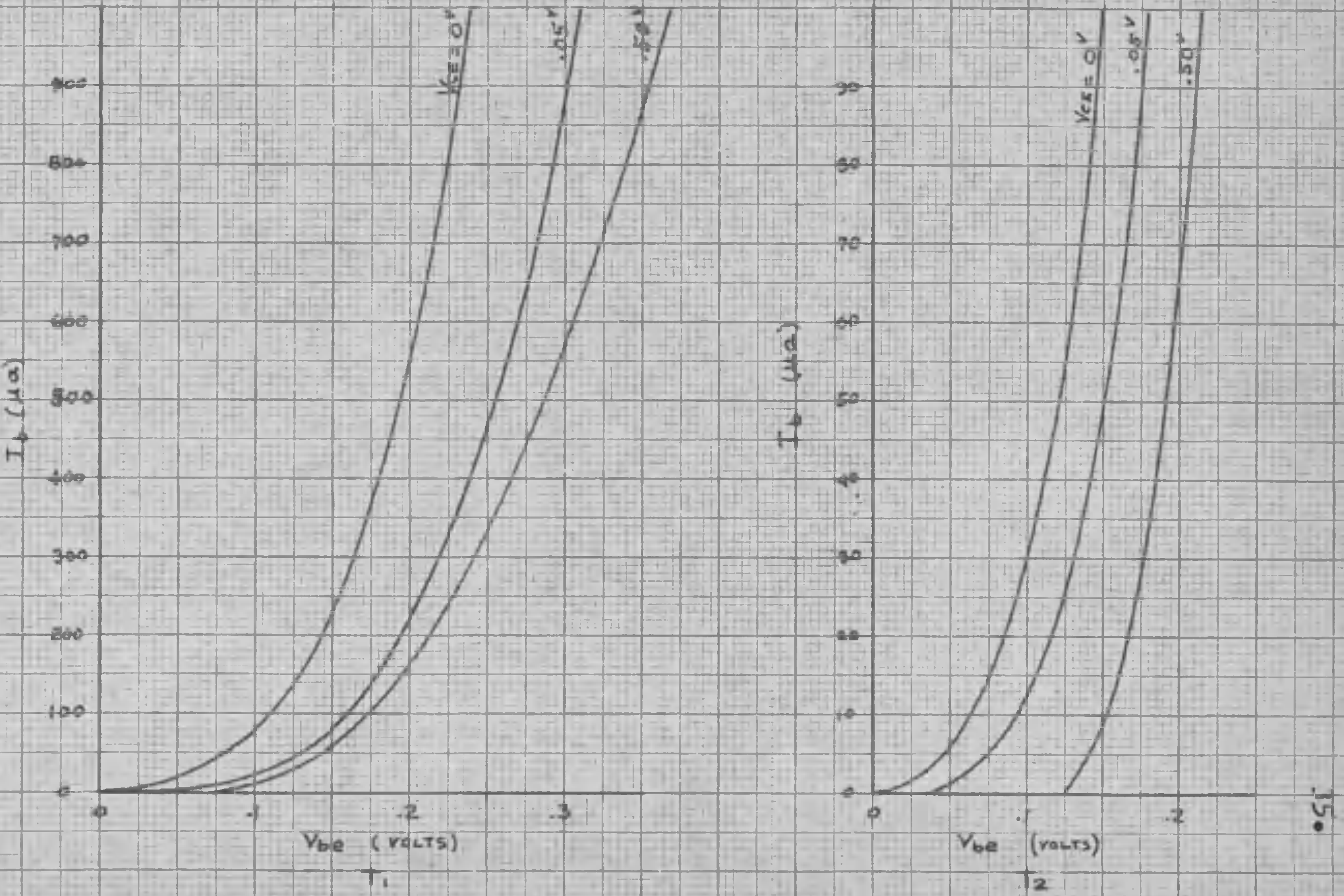


FIG 3.6 INPUT CHARACTERISTICS FOR 2N35 TRANSISTOR

35

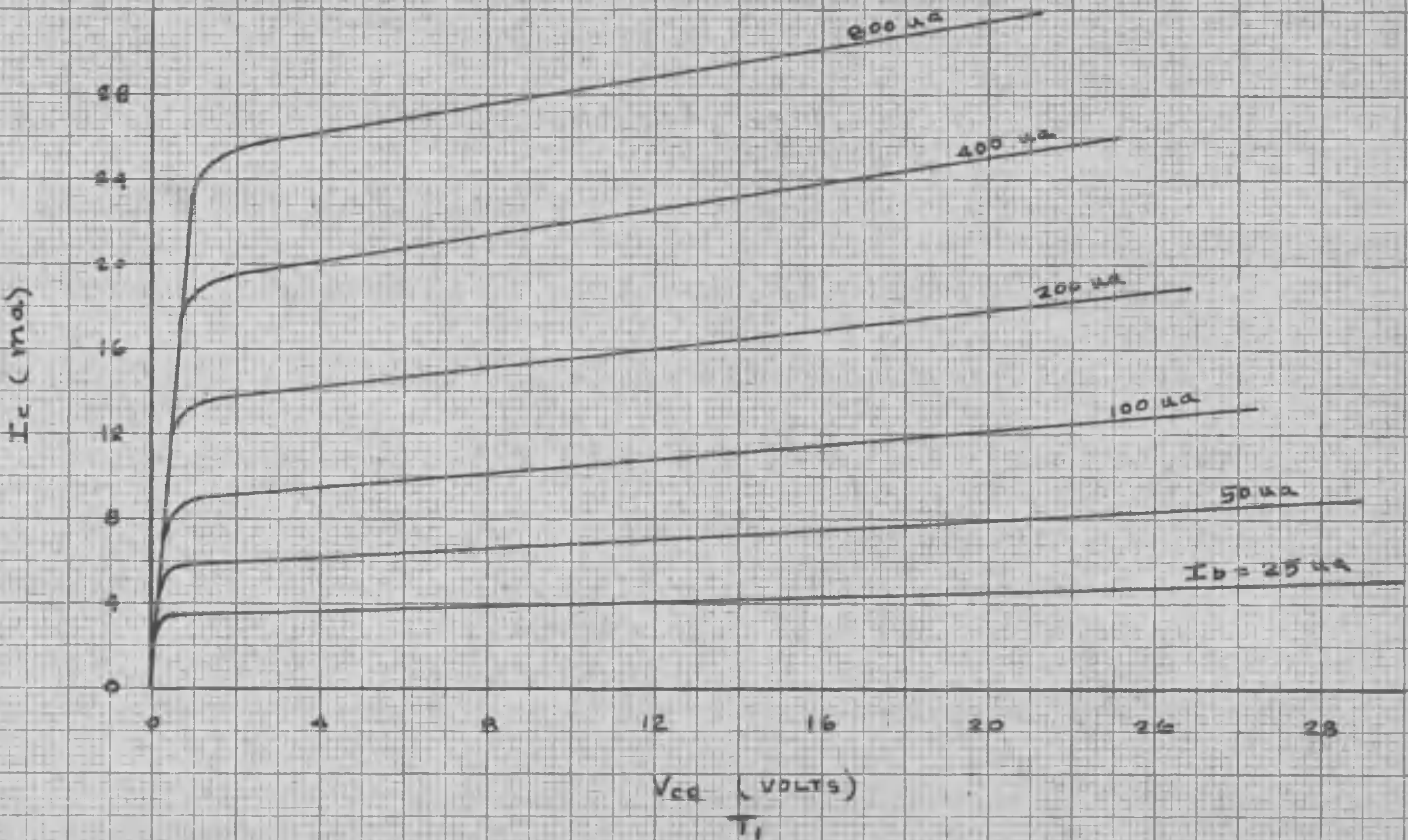
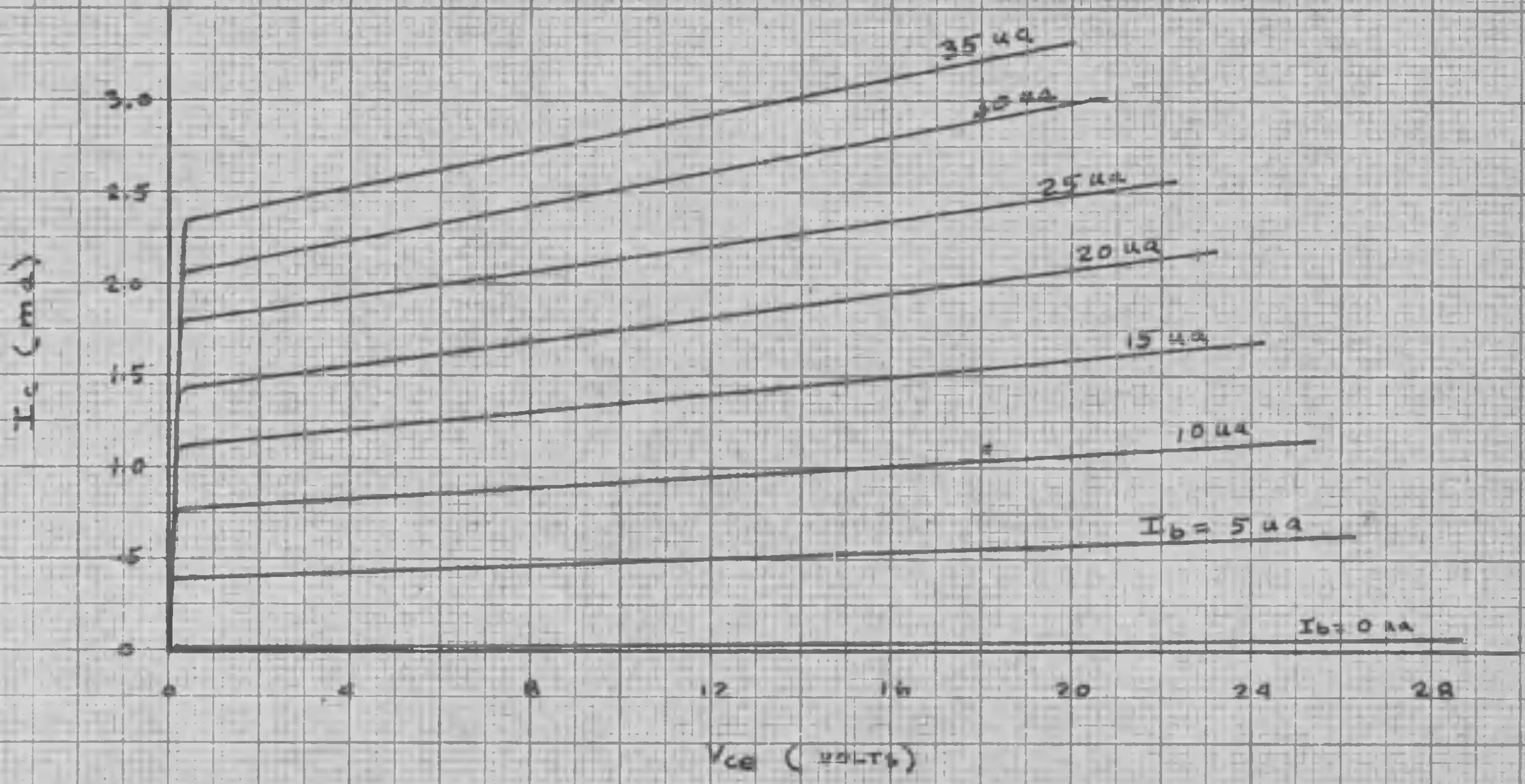


FIG. 3.7 OUTPUT CHARACTERISTICS FOR 2N167 TRANSISTOR



T2

FIG 3.8 OUTPUT CHARACTERISTICS FOR 2N167 TRANSISTOR

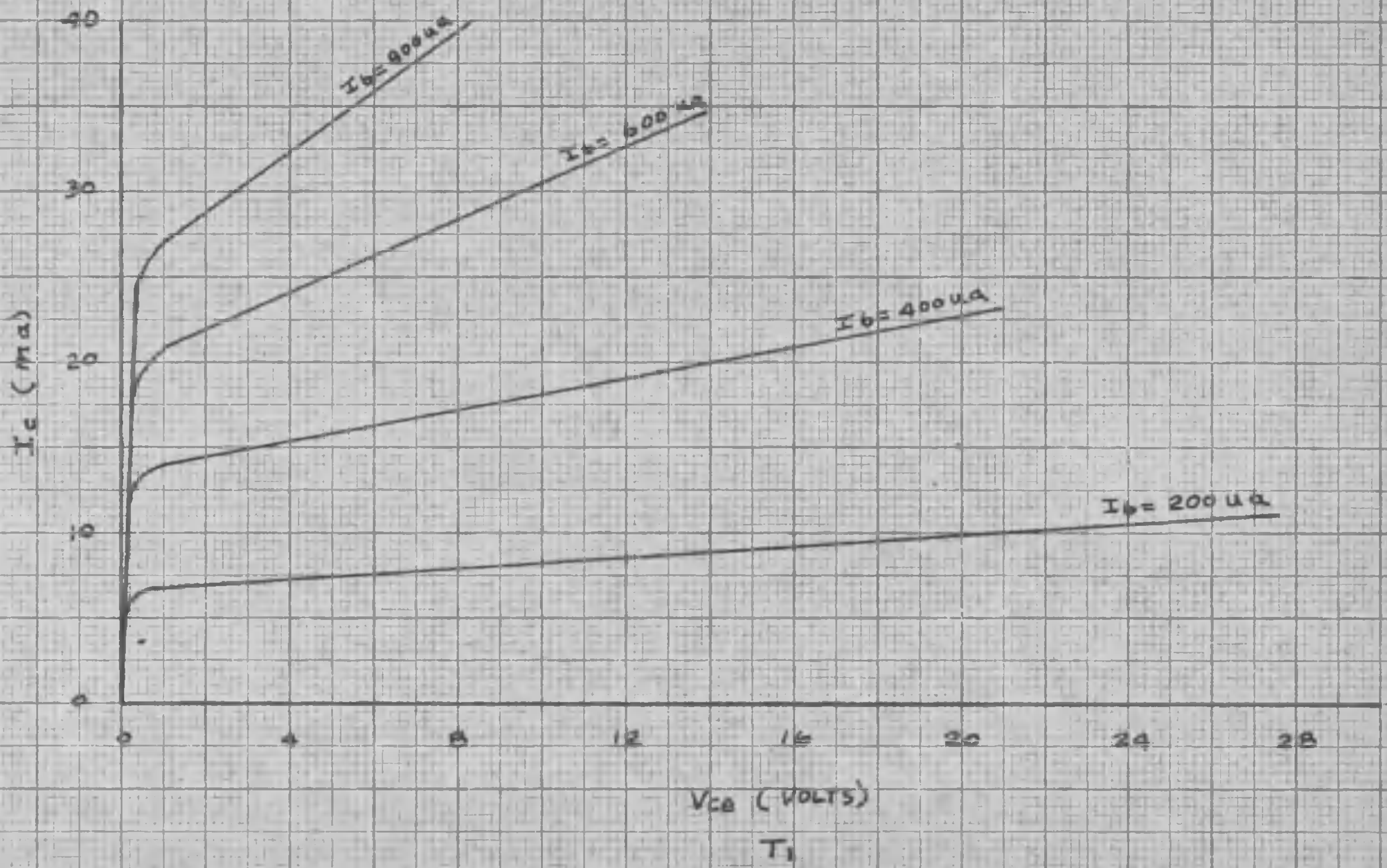


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FIG 3.9 OUTPUT CHARACTERISTICS FOR 2N35 TRANSISTOR

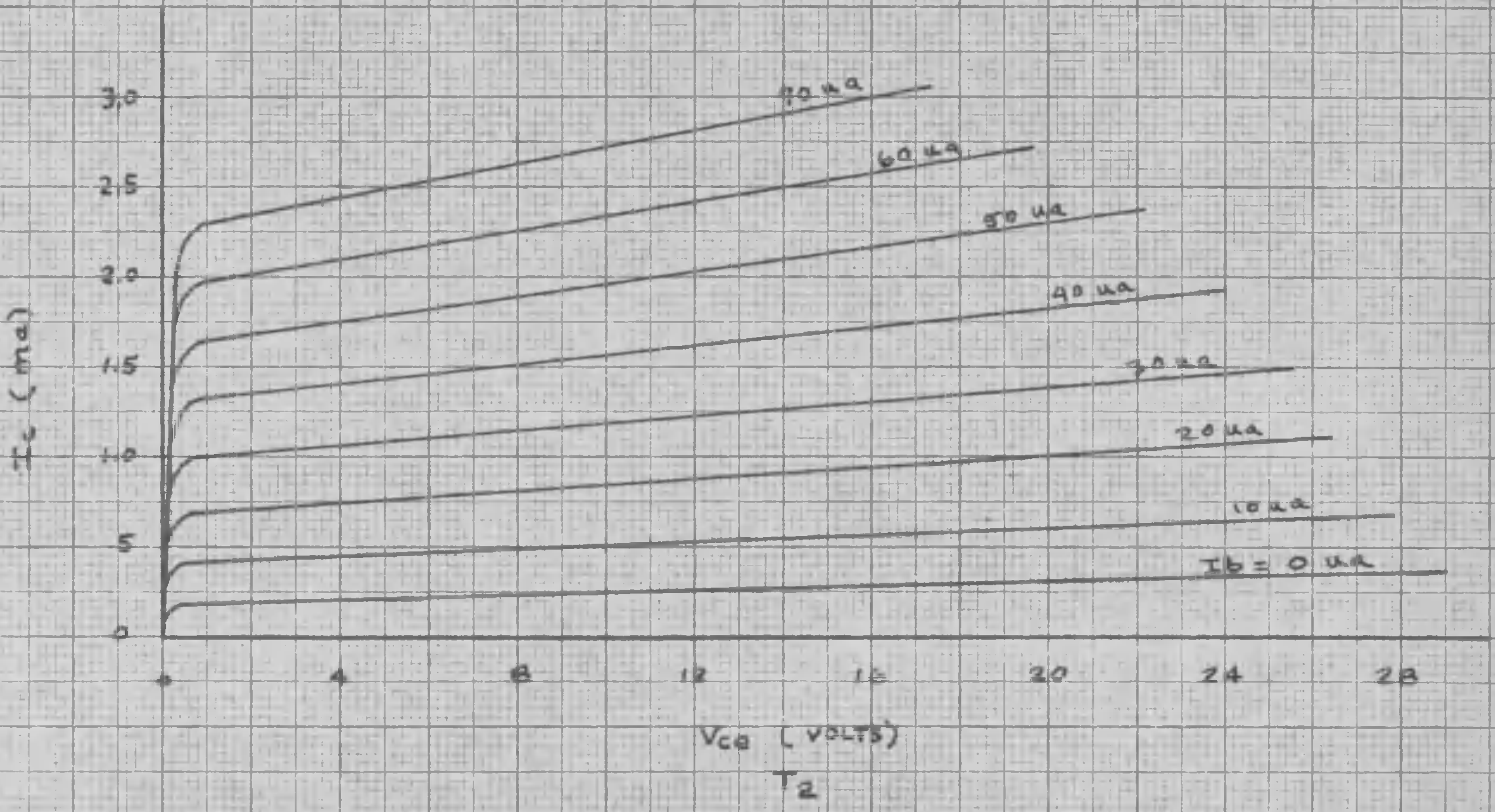

FIG 3.10 OUTPUT CHARACTERISTICS FOR 2N35 TRANSISTOR

and C_2 to charge linearily. This method provides fast triggering from normal to timing state. However, in a case where good holdoff characteristics are important, this method might not be the most preferable. By holdoff characteristics is meant the ability of the circuit to accept high repetition rate triggers without affecting the duration of the sweep output voltage.

The second method is to adjust the bias so that saturation is reached a short time before the end of the flyback time. Because the circuit is oversaturated during a small portion of the flyback time, there is a certain amount of delay in coming out of saturation when a trigger is applied. This seems to improve the holdoff characteristics of the circuit. This method also decreases the flyback time to a small degree.

In biasing T_2, care must be taken in order not to exceed the maximum collector dissipation for any appreciable time. The amount of time that the collector dissipation can be exceeded should be small in comparison to the thermal time constant of the transistor.[5] This thermal time constant is the time it takes a transistor to recover to normal operating temperatures when the maximum dissipation has temporarily been exceeded.

Using the bias equations in section 3.3 and the

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[5] Richard B. Hurley, "Transistor Data for Logical Circuit Design", Electronic Industries and Tele Tech., Vol. 16, October, 1957, pp. 60-61 and pp. 165-166

input and output characteristics shown in figures 3.5 through 3.10, the proper biasing can now be determined.

3.5 DISCUSSION OF CIRCUIT PARAMETERS

The value for R_1 should be as large as possible, with certain limitations. The same limitations that applied to the tube circuit also apply to the transistor circuit. Briefly, these are input impedance of the following stage and the resistance of transistor T_1 in cutoff.

The value of Ecc is largely determined by the maximum collector to emitter voltage of the transistors being used. See equations (3.4) and (3.19).

Resistors R_6, R_7, and R_8 are primarily biasing resistors. Their values depend on the bias operating points of the transistors. However, the value of R_7 has an important effect on the circuit switching speed. The effect can be determined by analysing the equivalent circuit for the unstable state. See figure 3.4. Neglecting the effect of R_6 and R_8 for reasons of simplicity, and assuming the R_4 is much smaller than $1/h22$, the following equation is derived.

$$h_{21}\left(\frac{1/h_{22}}{1/h_{22} + R_2 + h_{11}}\right) h_{21}\left(\frac{R_4}{R_a + R_7 + h_{11}}\right) = 1$$

If the loop gain is larger than one, then regenerative switching takes place. The larger the loop gain the faster the switching speed. Therefore, if R_2 is made relatively small in comparison with 1/h22 and R_7 small in comparison with R_4, then the loop gain is increased.

The size of R_7 largely determines the values for R_6 and R_8. R_2 and R_3 have an effect on the sweep duration as seen from equation (3.15). A decrease in R_2 or R_3 decreases T_s. R_3 also has a large effect on the bias of T_1.

The values for C_1 and C_2 are determined by the sweep duration, the displacement error, and the size of R_1.

The following factors are important in choosing the proper transistors for the circuit. The maximum collector to emitter voltage should be as high as possible. The reason for this is that this voltage largely determines the maximum obtainable sweep output voltage from the circuit. The transistor should have good high frequency characteristics (high α_{co} and low capacitance). The forward short circuit current amplification factor (h_{21}) should also be as large as possible.

B ased upon the preceeding discussion and experimentation, the following component values have been chosen.

$Ecc = 168.5$	$R_6 = 27$ K
$R_1 = 50$ K	$R_7 = 680$
$R_2 = 19$ K	$R_8 = 88$ K
$R_3 = 10.01$ K	$e_d = .0125$
$R_4 = 240$	$T_1 = T_2 = $ 2N167

From the characteristic curves (figures 3.5 through 3.10) and manufacturers' information, the following transistor parameters were used for 2N167.

$r_{b1} = 800$	V_{ce} (Maximum) 30 volts
$r_{b2} = 3.75$ K	$h_{11} = 5$ K
$1/h_{22} = 37$ K	$h_{21} = 80$

3.6 SUMMARY OF THE DESIGN PROCEDURE OF THE TRANSISTOR CIRCUIT

1. Specify the design requirement of the circuit.
2. Select the type of transistors to be used.
3. Choose the bias voltage Ecc.
4. Check for proper bias conditions using the input and output characteristics together with the following bias equations: Equations (3.6), (3.7) and (3.8), (3.10). See section 3.3.
5. Keeping section 3.4 thoroughly in mind, make the proper circuit adjustment to satisfy equations (3.5), (3.15) and (3.16)

6. Substitute into equations (3.4 and (3.17) to assure that maximum transistor voltages are not exceeded.
7. Solve for the output sweep amplitude using equation (3.18).
8. Solve for the sweep duration using equation (2.24).

Chapter 4.

EXPERIMENTAL AND THEORETICAL DATA

4.1 INTRODUCTION

This chapter is sub-divided into two parts. The first is on the tube circuit and the second on the transistor circuit. The purpose of this chapter is to verify the major design equations derived in chapters 2 and 3, and to describe the characteristics of each circuit. Also, through various experiments, the accuracy and limitations of these equations will be shown. Curves are used to show experimental and theoretical data, and photographs are used to show the various different wave forms. A Tektronix type 514 AD oscilloscope was used to measure and display the various waveforms.

4.2 EXPERIMENTAL RESULTS OF THE TUBE CIRCUIT

In the tube circuit, the circuit was first adjusted for a sweep duration of 100 μsec, thus determining the resistance and capacitance values shown below. (See figure 2.1)

E_{bb} = 250 volts $\qquad$ R_4 = 760

R_1 = 50 K $\qquad$ $C_1 = C_2$ = 0.02 μfd

$R_2 = 14.5K$ $e_d = 0.0125$

$R_3 = 10K$ V_1 = Section 2 of 6DE7

V_2 = Section 1 of 6DE7

These circuit values, unless otherwise stated on the data, were used throughout the experiments on the tube circuit.

Figure 4.1 indicates a very good stability of the sweep duration over a wide variation in Ebb. This more or less verifies equation (2.26), which indicates sweep duration is independent of Ebb. The error shown is probably due to the change in tube parameters over the wide range of Ebb.

Figure 4.2 indicates the accuracy of equation (2.22) for the sweep output voltage. The error here is again probably due to the change in tube parameters over the wide range in Ebb.

The waveforms shown in figure 4.3 indicate amplitude and shapes of the various circuit voltages. This figure is used as a standard to compare the effects of various circuit changes.

In figure 4.4 the only change in the circuit was to change the trigger repetition rate from 5.8 kc to 100 kc. The purpose of this was to show the holdoff capabilities of the circuit. As can be seen, the effect on the sweep duration and the sweep amplitude was negligible.

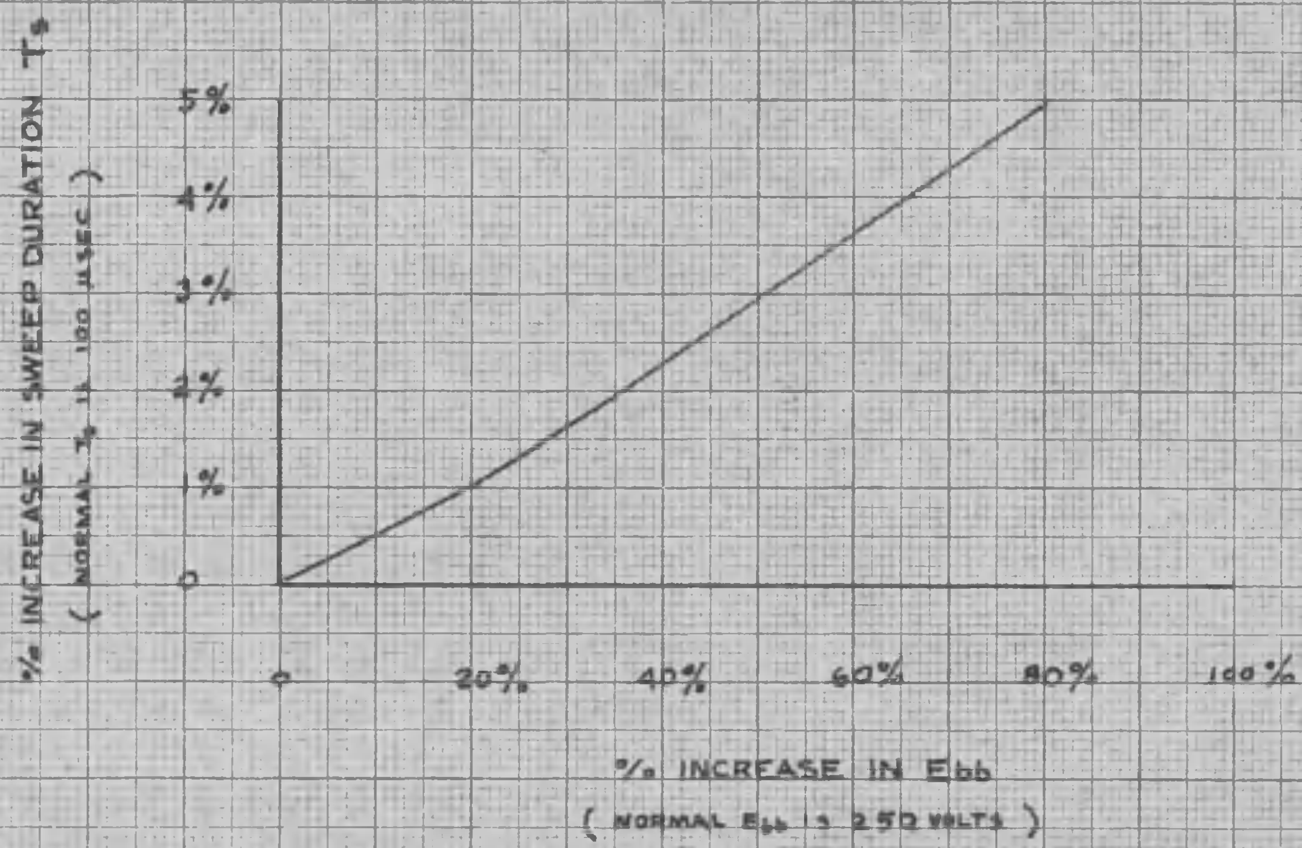


FIG. 4.1
SWEEP DURATION STABILITY

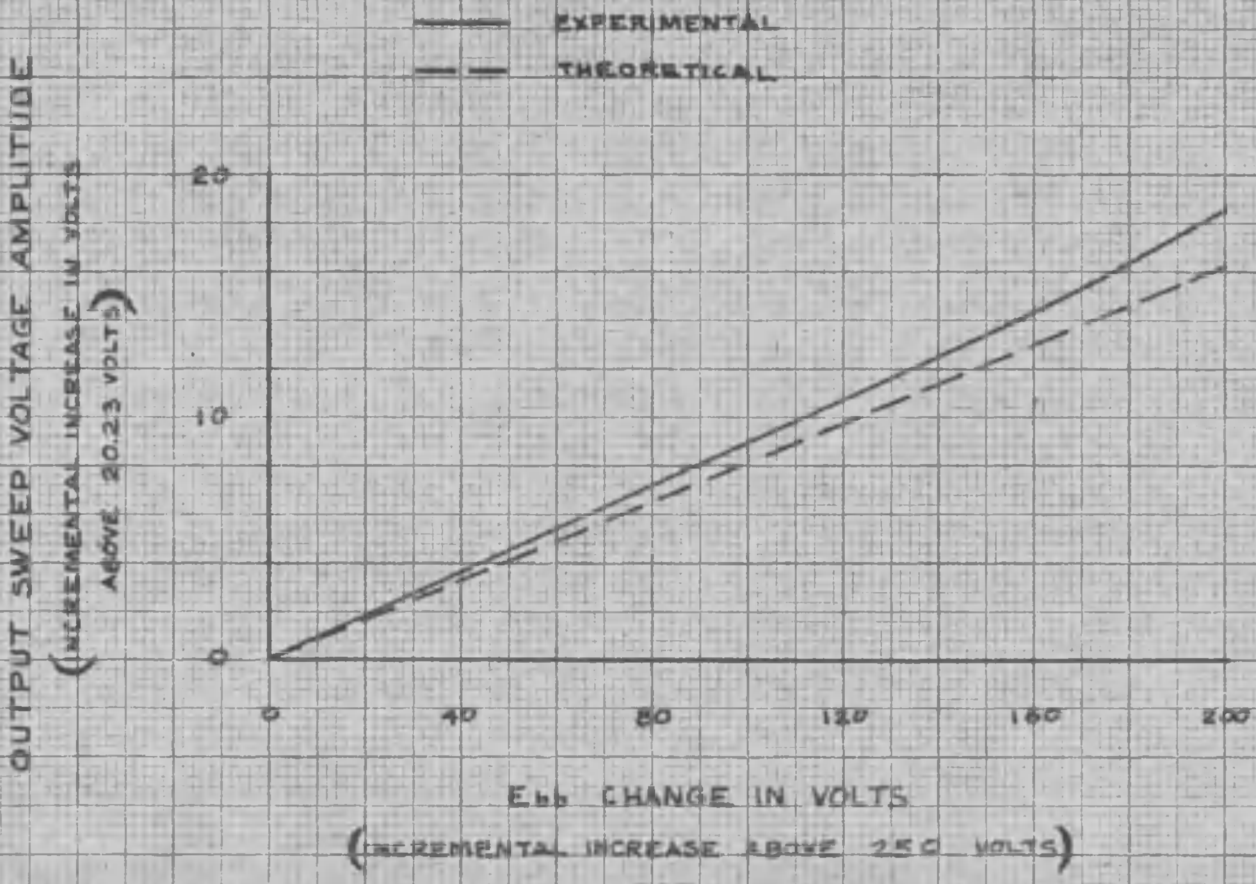


FIG. 4.2
CHANGE IN OUTPUT SWEEP VOLTAGE WITH
VARIATION IN Ebb

Note the 100 kc triggers feeding in on all the waveforms except the plate to ground voltage of V_1, which represents the output voltage of the circuit. This indicates good isolation from the trigger circuit.

In figure 4.5, the only change in the circuit was to eliminate the trigger voltage and adjust R_4 for free running operation. The purpose of this was to show the difference between a monostable and free running operation. As can be seen, the change in the sweep duration and sweep amplitude is negligible. Note the slight difference in switching between normal and timing state. The flyback time, in free running operation depends largely upon the value of R_4.

In figure 4.6, C_1 and C_2 were changed to 0.20 μfd and the trigger repetition rate was adjusted to 580 cps. The purpose of this was to note any deviation from equation (2.26) and the change in wave shapes in a low frequency range. Equation (2.26) shows that the change in sweep duration is linear with respect to change in C_1 and C_2. Therefore, with these particular values, the sweep duration should theoretically be 1000 μsec. As can be seen from the photographs, the sweep duration is very close to 1000 μsec. This corresponds to the slight decrease in sweep amplitude due to the fact that the duration is a little less than the theoretical value.

Note the sharp transitions between states of operation. The small error involved is probably due to the fast switching time between states as compared to the sweep duration.

In figure 4.7, C_1 and C_2 were changed to 0.005 μfd and the trigger repetition rate changed to 20 kc. The purpose of this experiment is to show the effects of the circuit in a high frequency range. The sweep duration is a little more than 25 μsec, which accounts for the slight sweep amplitude increase. Notice the rounding effect on the peak of the output. This is due to slow switching time with respect to the sweep duration. The output, however, retains its linearity up to the rounded peak of the waveform.

4.3 CHARACTERISTICS OF THE TUBE CIRCUIT

The input impedance of the circuit is approximately 400 ohms. The circuit can be made monostable or free running by adjusting R_4. This affords excellent synchronization qualities and the ability of the circuit to trigger from very small trigger voltages.

The input impedance of the following stage must be high in order to retain the calculated linearity of the sweep voltage. The sweep output amplitude is directly proportional to Ebb and the displacement error. With a displacement error of 1.25% and Ebb equal to 450

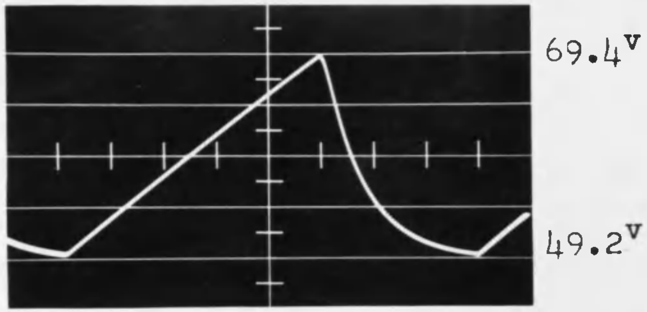
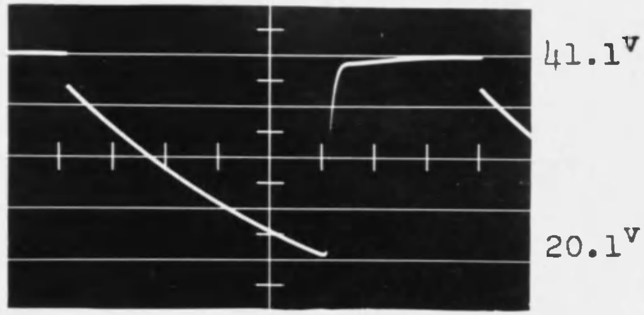


Plate to Gnd. Voltage of V_1



Cathode to Gnd. Voltage of V_1

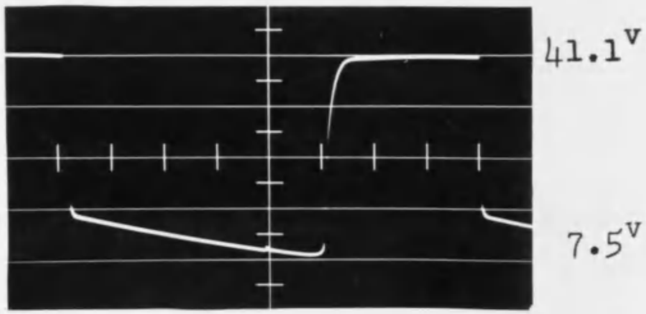
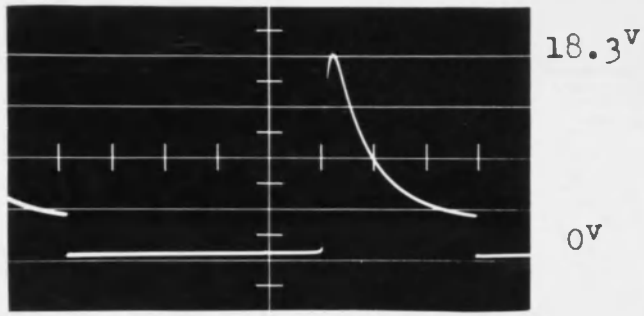


Plate to Gnd. Voltage of V_2



Cathode to Gnd. Voltage of V_2

FIG. 4.3 WAVEFORMS FOR THE TUBE CIRCUIT

Time calibration	20 usec/div
Trigger Rep. Rate	5.8 kc
C_1 equals C_2	.02 ufd

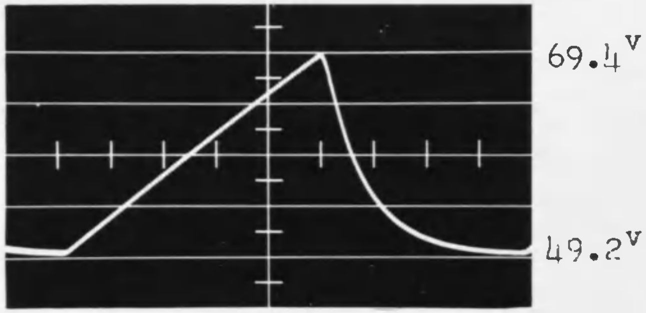
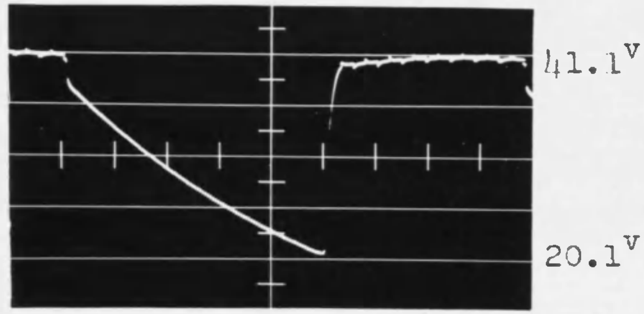


Plate to Gnd. Voltage of V_1



Cathode to Gnd. Voltage of V_1

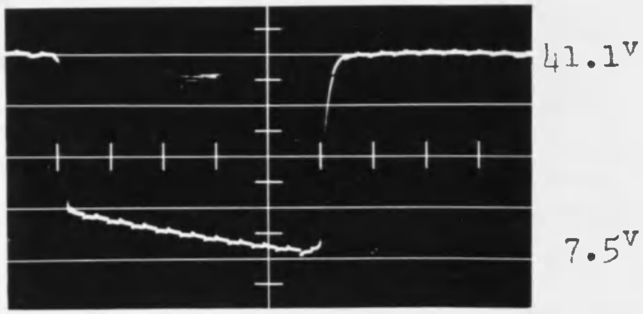
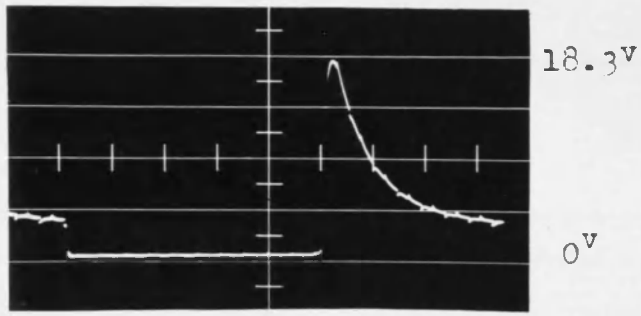


Plate to Gnd. Voltage of V_2



Cathode to Gnd. Voltage of V_2

FIG. 4.4 WAVEFORMS FOR THE TUBE CIRCUIT

Time calibration	20 usec/div
Trigger Rep. Rate	100 kc
C_1 equals C_2	.02 ufd

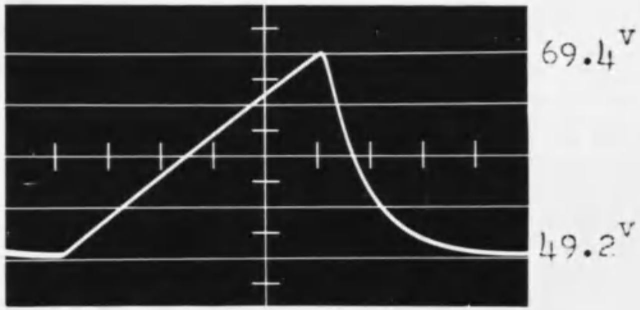
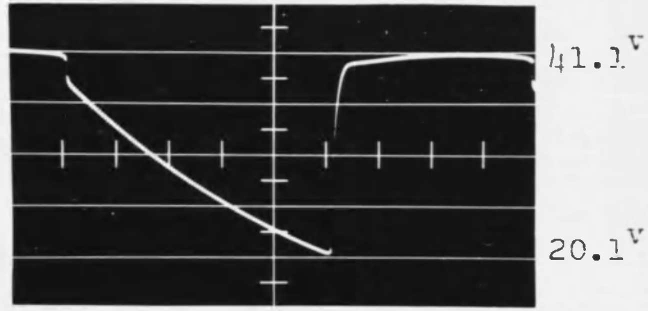


Plate to Gnd. Voltage of V_1



Cathode to Gnd. Voltage of V_1

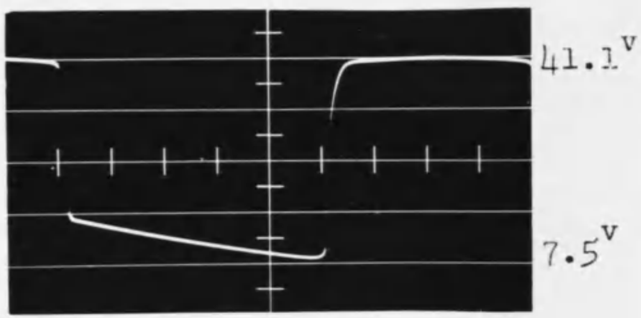
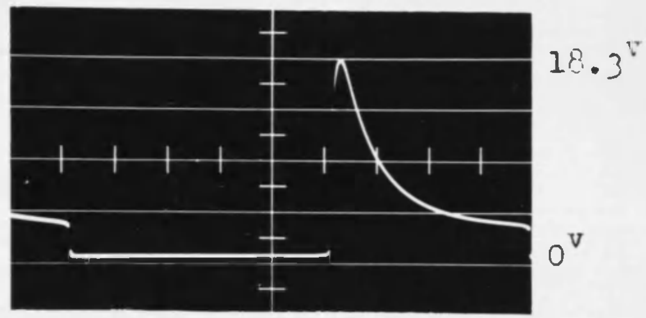


Plate to Gnd. Voltage of V_2



Cathode to Gnd. Voltage of V_2

FIG. 4.5 WAVEFORMS FOR THE TUBE CIRCUIT

Time calibration 20 usec/div
Free Running R_4 = 745 ohms
C_1 equals C_2 .02 ufd

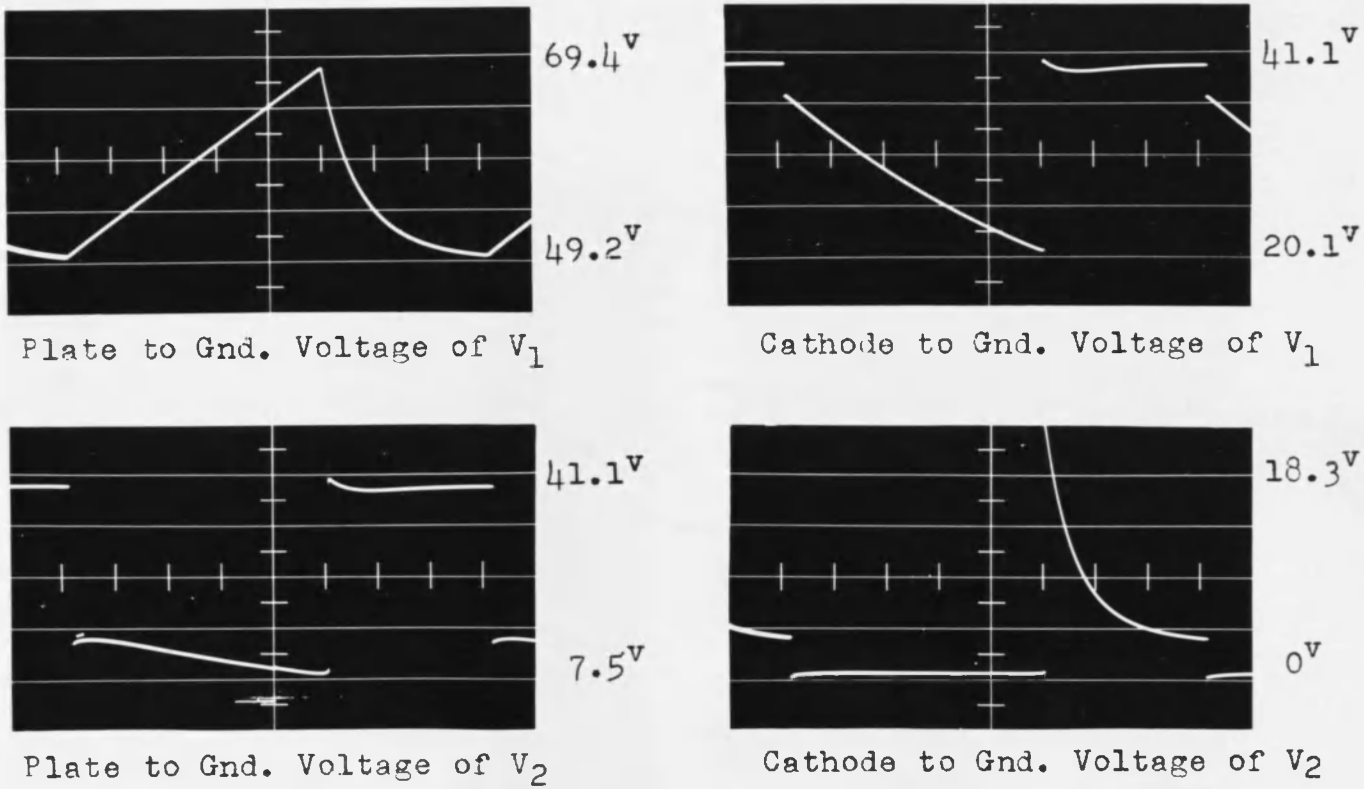


FIG. 4.6 WAVEFORMS FOR THE TUBE CIRCUIT

Time calibration	200 usec/div
Trigger Rep. Rate	.58 kc
C_1 equals C_2	.20 ufd

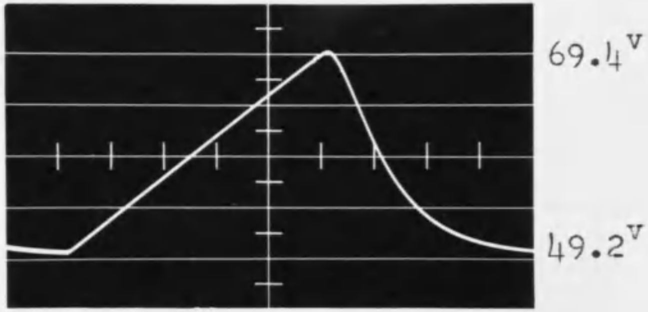
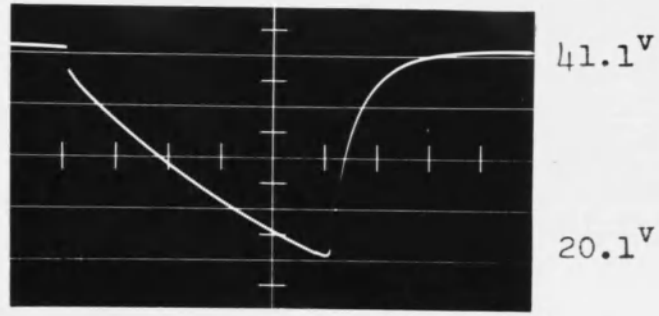


Plate to Gnd. Voltage of V_1



Cathode to Gnd. Voltage of V_1

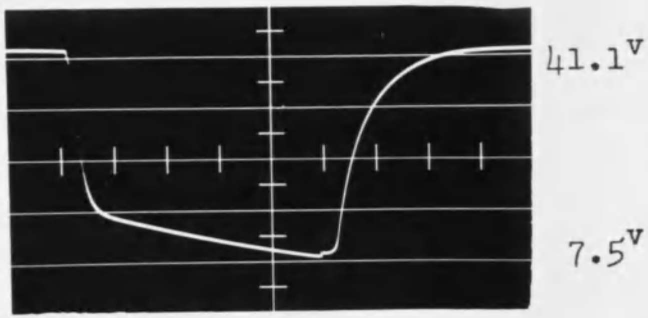
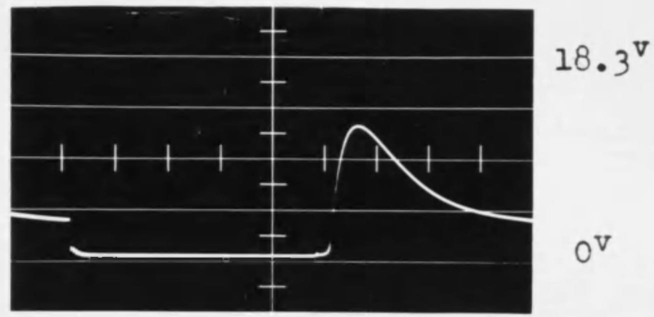


Plate to Gnd. Voltage of V_2



Cathode to Gnd. Voltage of V_2

FIG. 4.7 WAVEFORMS FOR THE TUBE CIRCUIT

Time calibration	5 usec/div
Trigger Rep. Rate	20 kc
C_1 equals C_2	.005 ufd

volts, the sweep output amplitude for this circuit was 38.75 volts. The flyback time for this circuit was approximately 60% of the sweep duration.

The sweep duration stability of the circuit is excellent as seen from figure 4.1. The range of sweep durations of the circuit is quite wide. The low frequency limitation of the circuit is not actually known, but a sweep duration of 10 msec was obtained with good stability. It is felt that much longer sweep durations can be obtained with good stability. A sweep duration of 20 usec was attained with good holdoff characteristics. If the holdoff characteristics are neglected, sweep durations in the range of 5 usec are possible.

4.4 EXPERIMENTAL RESULTS OF THE TRANSISTOR CIRCUIT

In the transistor circuit, the circuit was first adjusted for a sweep duration of 100 usec, thus determining the resistance and capacitance values shown below. (See figure 3.1)

$E_{cc} = 168.5$ volts	$R_6 = 27K$
$R_1 = 50$ K	$R_7 = 680$
$R_2 = 19K$	$R_8 = 88$ K
$R_3 = 10.01$ K	$e_d = 0.0125$
$R_4 = 240$	$T_1 = T_2 = 2N167$
	$C_1 = C_2 = 0.02$ ufd

These circuit values, unless otherwise stated on the data, were used throughout the experiments on the transistor circuit.

Figure 4.8 indicates the sweep duration stability variation in Ecc. This is to verify equation (3.14), which indicates T_s is independent of Ecc. Note that with a 6% change in Ecc there was no noticable change in sweep duration. At 6% change, error started to appear. This error was probably due to enough change in transistor bias to change the transistor circuit parameters. This is a good indication of the range of stability of the biasing circuit.

Figure 4.9 indicates the accuracy of equation (3.16) for the sweep output voltage. The error here is probably due to the change in transistor circuit parameters as the result of a change in bias.

Figure 4.11 shows the effects of changing the transistors from a 2N167 to a 2N35. The reason for this was to show the difference between a relatively high frequency transistor (2N167) and a low frequency transistor (2N35). This also shows the ability of the circuit to tolerate different transistors disregarding biasing. From the photographs it is seen that the switching and transition between states is slower using the 2N35 transistor. This difference is most evident from the

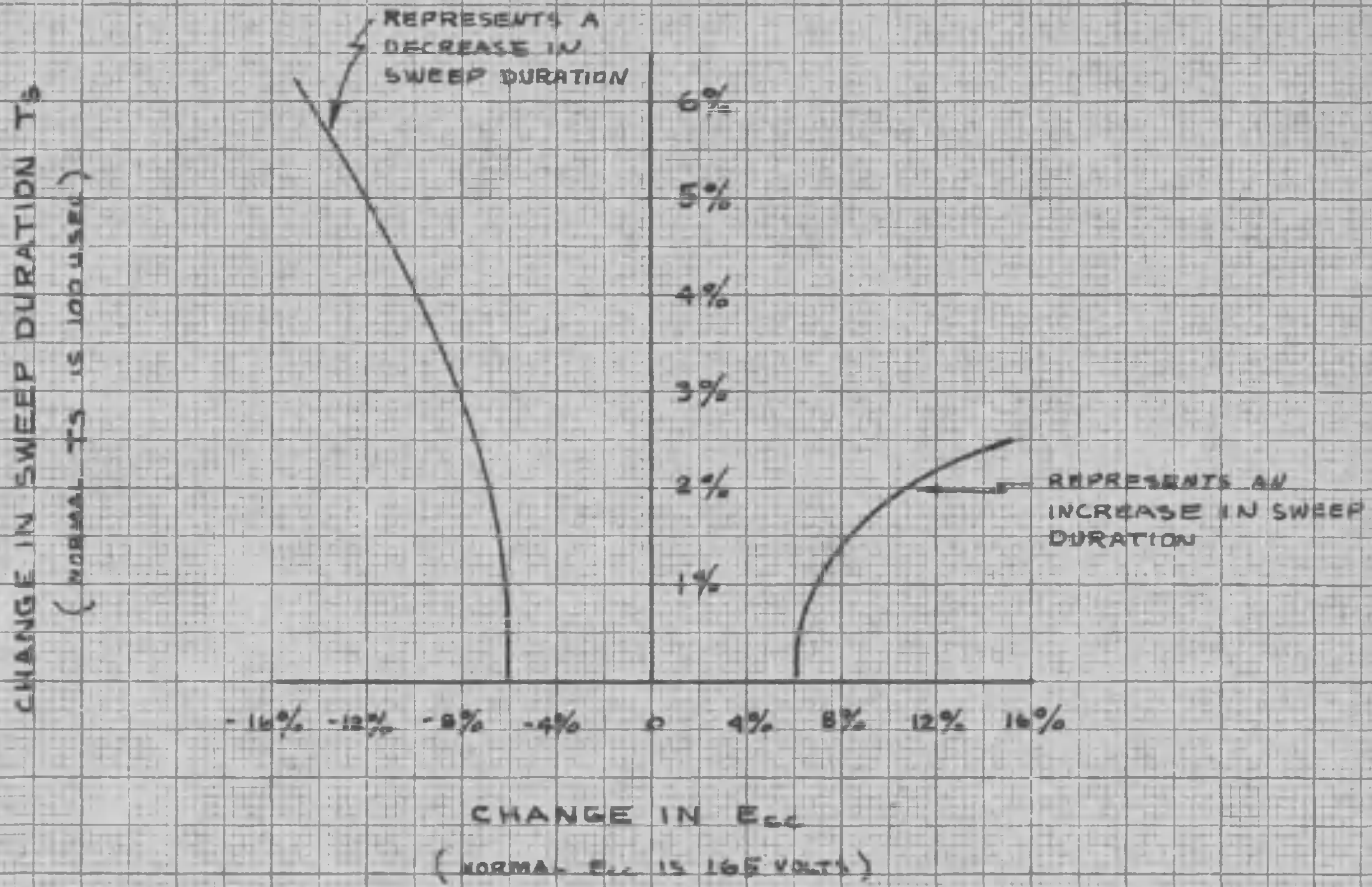


FIG. 4.8

SWEEP DURATION STABILITY

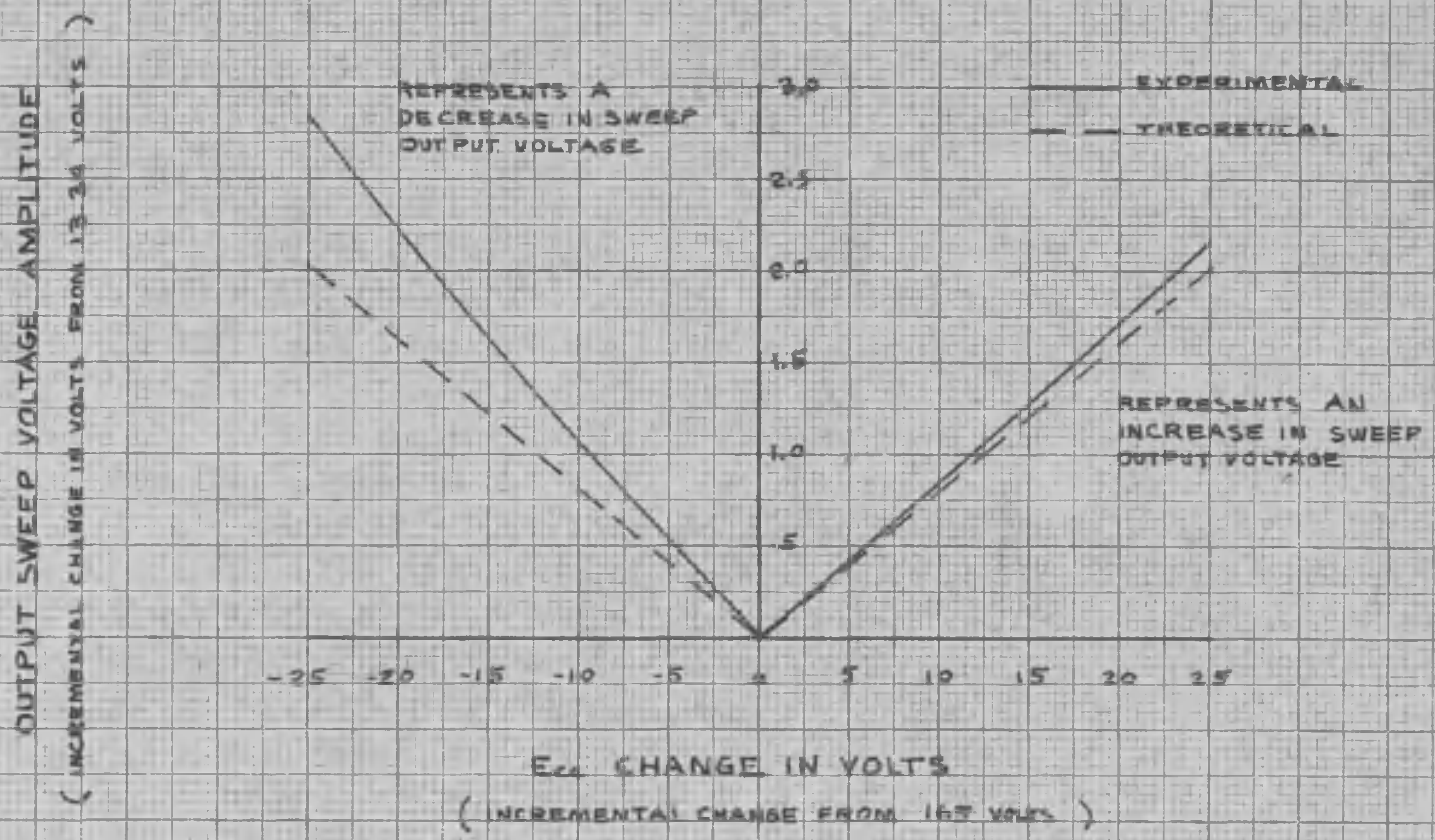
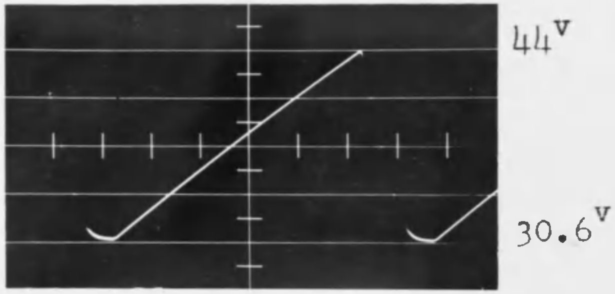
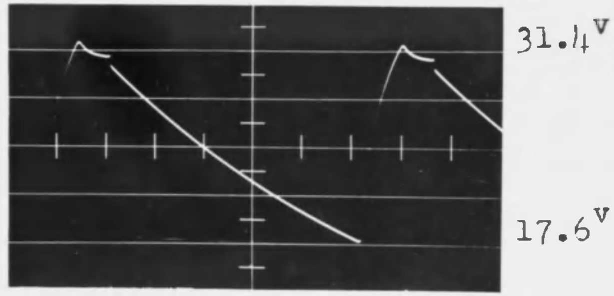


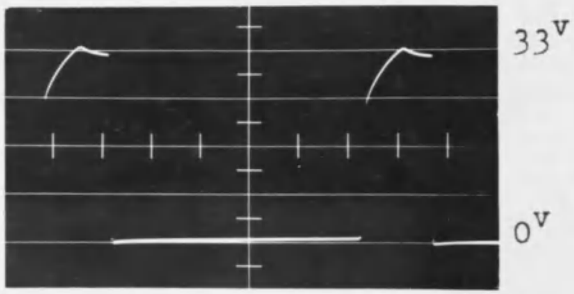
FIG. 4.9
CHANGE IN OUTPUT SWEEP VOLTAGE WITH
VARIATION IN E_{cc}



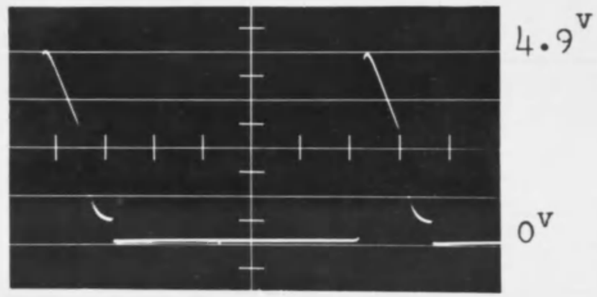
Collector to Gnd. Voltage of T_1



Emitter to Gnd. Voltage of T_1



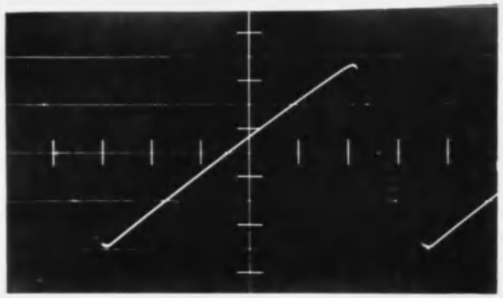
Collector to Gnd. Voltage of T_2



Emitter to Gnd. Voltage of T_2

FIG. 4.10 WAVEFORMS FOR TRANSISTOR CIRCUIT

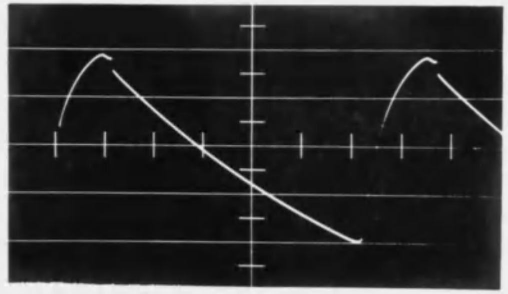
Time calibration	20 usec/div
Trigger Rep. Rate	6.8 kc
Transistor type	2N167
C_1 equals C_2	.02 ufd



44^V

30.6^V

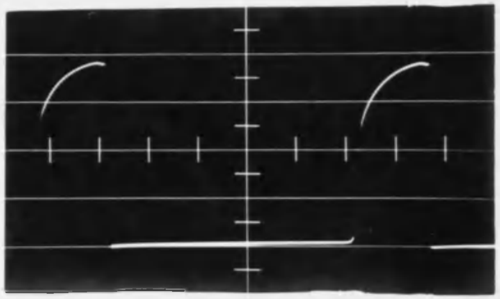
Collector to Gnd. Voltage of T_1



31.4^V

17.6^V

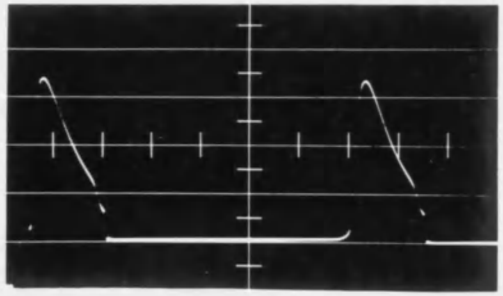
Emitter to Gnd. Voltage of T_1



33^V

0^V

Collector to Gnd. Voltage of T_2



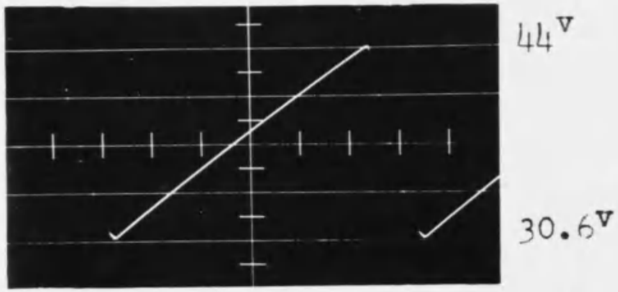
4.9^V

0^V

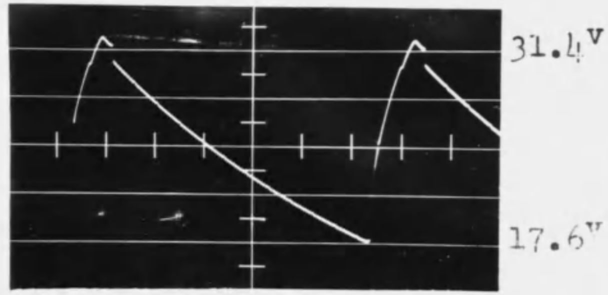
Emitter to Gnd. Voltage of T_2

FIG. 4.11 WAVEFORMS FOR TRANSISTOR CIRCUIT

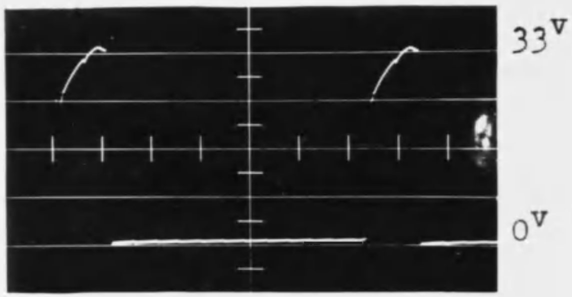
Time calibration	20 usec/div
Trigger Rep. Rate	6.8 kc
Transistor type	2N35
C_1 equals C_2	.02 ufd



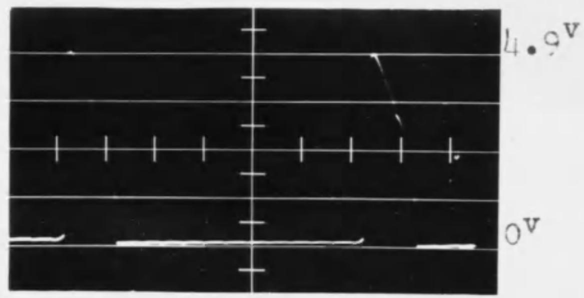
Collector to Gnd. Voltage of T_1



Emitter to Gnd. Voltage of T_1



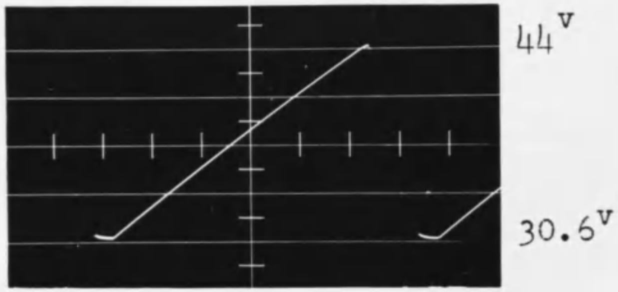
Collector to Gnd. Voltage of T_2



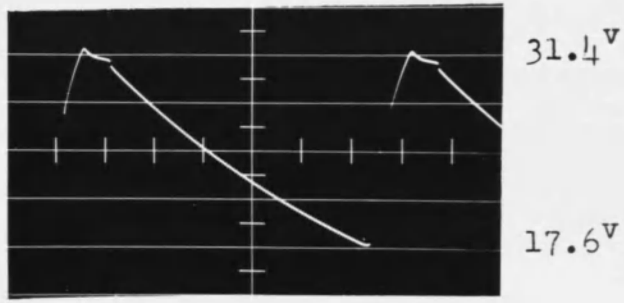
Emitter to Gnd. Voltage of T_2

FIG. 4.12 WAVEFORMS FOR TRANSISTOR CIRCUIT

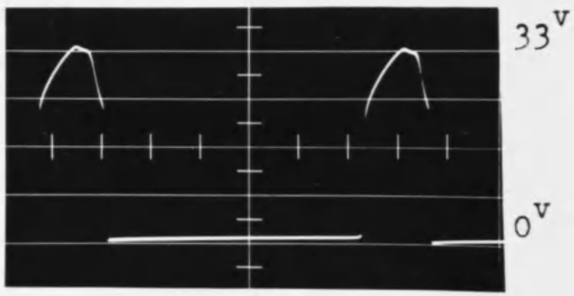
Time calibration	20 usec/div
Trigger Rep. Rate	100 kc
Transistor Type	2N167
C_1 equals C_2	.02 ufd



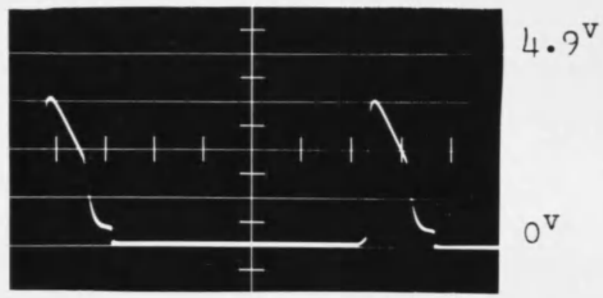
Collector to Gnd. Voltage of T_1



Emitter to Gnd. Voltage of T_1



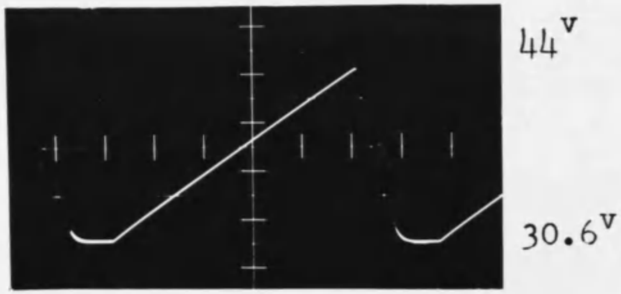
Collector to Gnd. Voltage of T_2



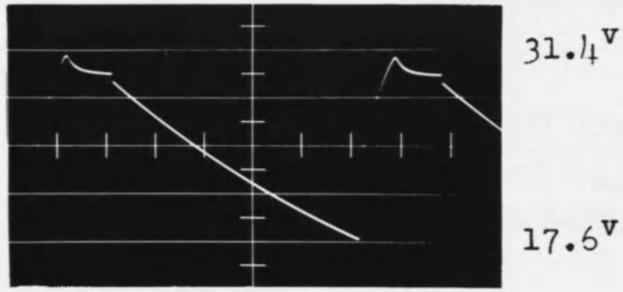
Emitter to Gnd. Voltage of T_2

FIG. 4.13 WAVEFORMS FOR TRANSISTOR CIRCUIT

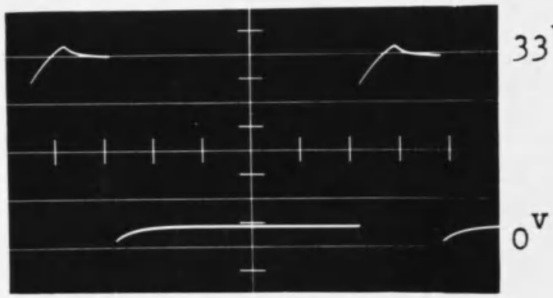
Time calibration	20 usec/div
Free Running	R_4 = 180 ohms
Transistor type	2N167
C_1 equals C_2	.02 ufd



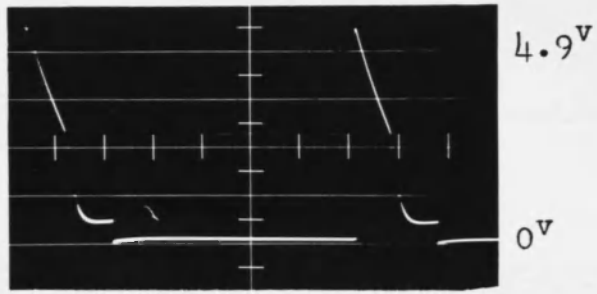
Collector to Gnd. Voltage of T_1



Emitter to Gnd. Voltage of T_1



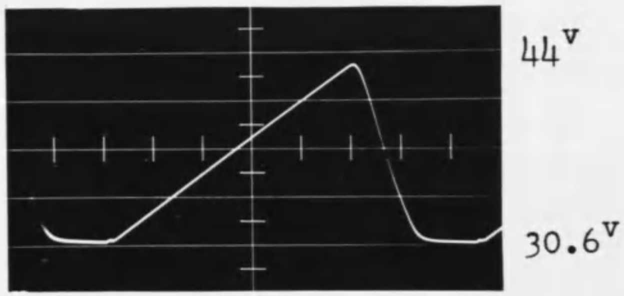
Collector to Gnd. Voltage of T_2



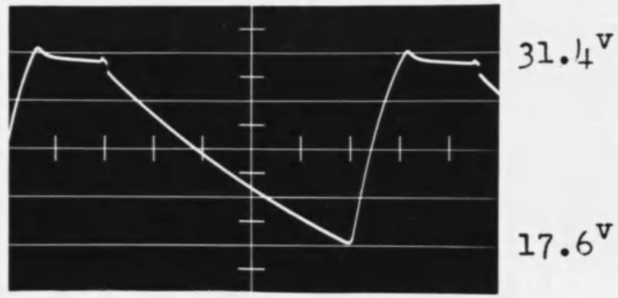
Emitter to Gnd. Voltage of T_2

FIG. 4.14 WAVEFORMS FOR TRANSISTOR CIRCUIT

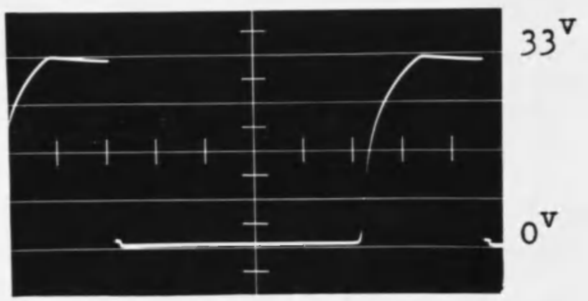
Time calibration	200 usec/div
Trigger Rep. Rate	.68 kc
Transistor type	2N167
C_1 equals C_2	.20 ufd



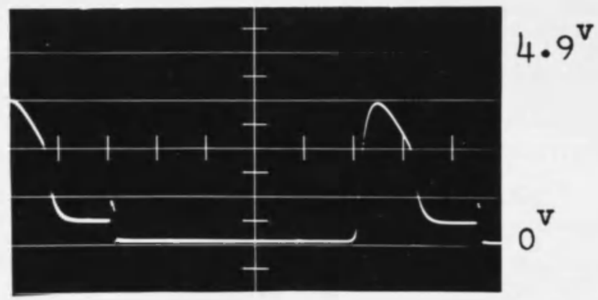
Collector to Gnd. Voltage of T_1



Emitter to Gnd. Voltage of T_1



Collector to Gnd. Voltage of T_2



Emitter to Gnd. Voltage of T_2

FIG. 4.15 WAVEFORMS FOR TRANSISTOR CIRCUIT

Time calibration	5 usec/div
Trigger Rep. Rate	25 kc
Transistor type	2N167
C_1 equals C_2	.005 ufd

waveforms of the collector to ground voltage of T_2. Also note that the sweep amplitude and duration were not greatly affected.

Figure 4.12 displays the holdoff capabilities of the circuit by changing the repetition rate from 6.9 kc to 100 kc. From the photos, it is seen that the change in sweep duration and amplitude is negligible.

In figure 4.13, the trigger voltage was omitted and the circuit adjusted for free running operation. From the photographs, it is seen that the sweep amplitude and duration increased a small amount. This is probably due to change in bias of transistor T_2. The change in R_4 from 240 ohms to 180 ohms causes T_2 to operate further into saturation. Note the rounding corner on the emitter to ground voltage of T_2 as it switches from the normal to the timing state. This effect becomes more and more pronounced the further the transistor is operated into saturation.

Figure 4.14 displays the circuit effects in the low frequency range. C_1 and C_2 were changed to 0.20 μfd and the repetition rate was adjusted to 680 cps. From the photographs it is seen that the sweep duration is 1000 μsec, but the sweep amplitude has decreased a small amount. The switching time between states is very fast as compared to the sweep duration.

Figure 4.15 displays the circuit effects in the high frequency range. C_1 and C_2 were adjusted to 0.005 μfd and the trigger repetition rate was adjusted to 25 kc. The sweep duration was decreased a small amount from the theoretical 25 μsec. value. The sweep amplitude likewise decreased. Note the rounding of the peaks and the switching time in comparison with the sweep duration. Note the delay involved in triggering the circuit from the normal state to the timing state. This is most evident in the waveform of emitter to ground voltage of T_2. This delay is probably due to T_1 operating into saturation. This causes a certain delay in coming out of saturation.

4.5 CHARACTERISTICS OF THE TRANSISTOR CIRCUIT

The input impedance of the circuit is approximately 500 ohms. The circuit can be made monostable or free running by adjusting R_4. This affords excellent synchronization qualities and the ability of the circuit to trigger from small trigger voltages.

The input impedance of the following stage must be high in order to retain the calculated linearity of the sweep voltage. The sweep amplitude is directly proportional to Ecc and the displacement error, however the maximum amplitude is limited to the maximum collector to emitter voltages of the transistor being used. With a maximum collector to emitter voltage of 30 volts, a displacement error of 1.25%, Ecc equal to 167.5 volts; the sweep output voltage of the circuit was 13.4 volts. The flyback time for the circuit is approximately 25% of the sweep duration.

The sweep duration stability of the circuit is excellent as seen from figure 4.8. The range of sweep duration of the circuit is quite wide. The low frequency of the circuit is not actually known, but a duration of 10 msec was obtained with good stability. A factor which might affect the low frequency limitation of this circuit is the thermal time constant of the transistor. This is true only if the transistor is biased in a region exceeding the maximum collector dissipation during the flyback state. Refer to section 3.3.

A sweep duration of 20 μsec was attained with good holdoff characteristics. If holdoff characteristics are neglected, sweep durations in the range of 5 μsec are possible.

Chapter 5.

SUMMARY

A step by step design procedure has been outlined in section 2.5 and section 3.6. If great care is taken in determining the respective tube and transistor parameters used in the circuits, the sweep output duration for the tube and transistor circuit can be calculated to within 3% and 1% respectively. The sweep output voltage amplitudes of the tube and transistor circuit can be calculated to within 2% and 10% respectively.

In comparing the tube circuit with the transistor circuit, each circuit has certain advantages over the other. The advantages of the tube circuit over the transistor circuit as as follows: (1) the design is less complex in the tube circuit, (2) a much larger sweep voltage amplitude is attainable in the tube circuit, and (3) the sweep duration of the tube circuit remains constant over a much broader change in B+ voltage.

The advantages of the transistor circuit over the tube circuit are as follows: (1) no filament voltage is necessary, (2) the power comsumption is much less in the transistor circuit, (3) the physical space requirements of the transistor circuit are much less, (4) the flyback time is approximately 1/3 that of the tube circuit. For comparison purposes, the transistor circuit may be compared to a transistor sweep circuit described in the July, 1957, issue of Electronic Equipment.[6] Both circuits use two transistors, and are similar in that they both are of the multivibrator sweep type. The amplitude of the sweep voltage generated in both circuits is limited by the maximum collector to emitter voltage of the transistors used. Using the same maximum collector to emitter voltage and the same linearity in each circuit, the sweep circuit developed for this thesis is capable of generating approximately seven times the sweep output amplitude of the other circuit. The flyback time of the two circuits is approximately the same. Sweep duration range cannot be compared because the range of the comparison circuit is not known. Together with the excellent sweep duration stability and the relatively large sweep voltage amplitude, it is felt that the circuit developed for this thesis has a great deal more to offer.

In conclusion, it is felt that the most important feature of the tube circuit is the stability of the sweep duration with large changes in B+ voltage. The most important feature of the transistor circuit is the relatively large sweep amplitude compared to other transistor circuits and its excellent sweep duration stability.

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[6] Joseph Chernof, "Design Features of a Transistor Sweep Circuit", Electronic Equipment, Vol. 5, July, 1957

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