Circuit Concepts

SWEEP GENERATOR CIRCUITS

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TEKTRONIX

CIRCUIT CONCEPTS

CONTENTS

INTRODUCTION

The cathode-ray oscilloscope (CRO) is a measuring device that displays electrical phenomena. Ordinarily, the phenomenon is displayed as a visual plot of the instantaneous voltage of a wave as a function of time. In most cases, the time-reference (base) is supplied by the instrument itself.

The sweep generator develops a sawtooth voltage, which after being processed in the horizontal amplifier, is applied to the horizontal deflection plates of the CRT, producing the horizontal sweep. When the deflection factor (voltage required to produce a linear unit of beam displacement) and the sawtooth voltage rate-of-rise are known quantities, the horizontal dimensions of any waveform traced on the CRT face plate can be converted to units of time.

If precise time measurements are to be made, it is essential that equal increments of horizontal distance represent equal increments of time. If the displacement of the electron beam is at all times proportional to the voltage applied to the horizontal deflection plates, it follows that the sweep voltage must change at a linear rate.

In this text we shall discuss various techniques of generating this linear sweep function voltage and various control devices used in Tektronix oscilloscopes.
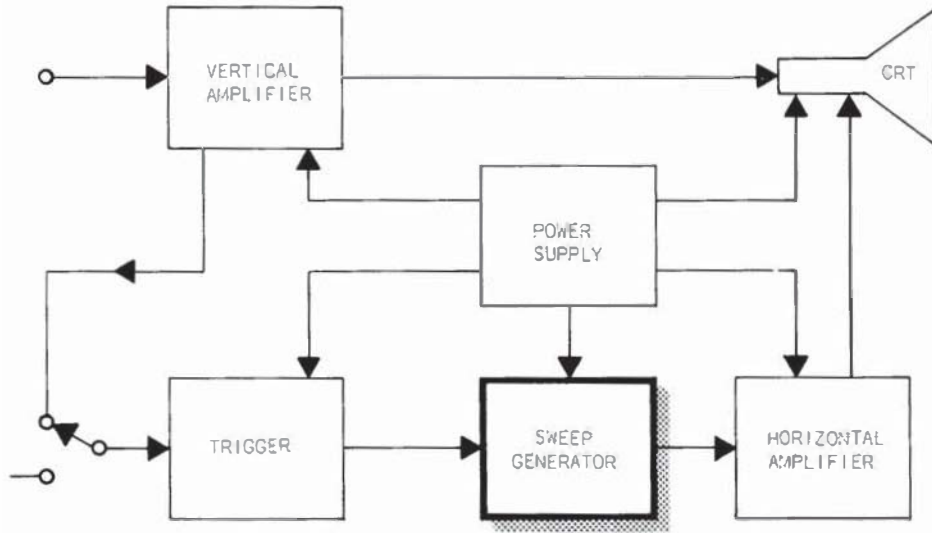
1

SAWTOOTH WAVEFORM CHARACTERISTICS

the
sawtooth

A sawtooth waveform is described by a voltage or current which starts at a given level, rises at a linear rate to some maximum value (either positive or negative), and then falls abruptly to the original level (Fig. 1-1). The rising portion of the sawtooth is called the *ramp* (A). In the development of a sweep function, the *linearity* of the ramp (or sweep) is of prime concern. The falling portion of the sawtooth is known as the *retrace* or "flyback" portion (B). The period between retrace and ramp-start is called the *holdoff* period (C). The voltage or current level from which the ramp makes its excursions is usually referred to as the *start level*. It will be advantageous to discuss these characteristics before we examine the circuits which generate the sawtooth.

LINEARITY

When we consider that elapsed time measurements are to be made from the time base of the CRT display, we must give careful consideration to the sawtooth voltage (current) that will be used to establish this time base.

sawtooth
voltage
versus spot
position

Fig. 1-1 illustrates the relationship of the sawtooth voltage to the position of the CRT spot. If we assume that the leading edge (or ramp portion) of the sawtooth voltage waveform is a straight line that produces a truly linear sweep, the spot position would conform to the equation: $X = KT$, where X is the amplitude of the voltage (current) which produces horizontal deflection -- K is a constant -- and T is the time from start of sweep.

If we apply very accurate time marks and adjust the
sweep speed of the oscilloscope properly, the display
can be made to resemble that shown in Fig. 1-1. Here
the relationship between the horizontal position of
the CRT spot and the ramp voltage is clearly revealed.
If the ramp is absolutely linear, each pulse will
exactly coincide with one of the graticule markings.
A perfectly linear ramp is difficult, if not
impossible, to generate. For all practical purposes
however, the sweep ramps generated by the state-of-
the-art sweep generators can be regarded as linear.

Although the concept of sweep linearity is a simple
one, quantitative descriptions of this characteristic
differ widely. Before comparing some of these
descriptions, it is essential that the difference
between sweep linearity and *sweep accuracy* be
established.

sweep
accuracy

In this text, the term "sweep accuracy" will mean
"the deviation of an observed sweep rate from a
standard timing reference rate," and will be
expressed in terms of *rate error* as a percentage of
full scale deflection. (*Full scale* in this context
is usually defined as the center 8 divisions of a
10-division graticule.) In Fig. 1-2, line (a)
represents the rate-of-rise (V_R/T) of the reference
voltage. Line (b) represents the rate-of-rise (V_0/T)
of the observed voltage. Rate error (e_r) is equal to
$V_R - V_0$. The expression for sweep accuracy thus
becomes $\dfrac{V_R - V_0}{V_R}$ (100). Fig. 1-2 illustrates the
relationship of the quantities just described to the
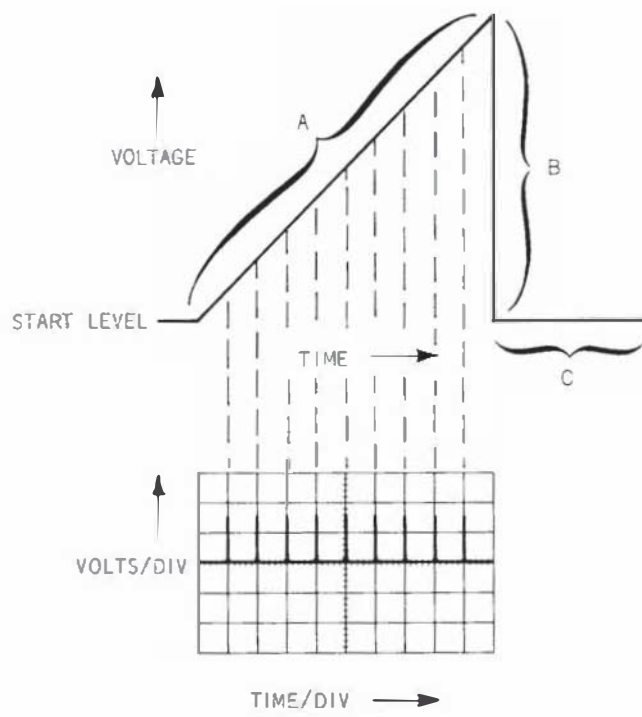appearance of an oscilloscope display.

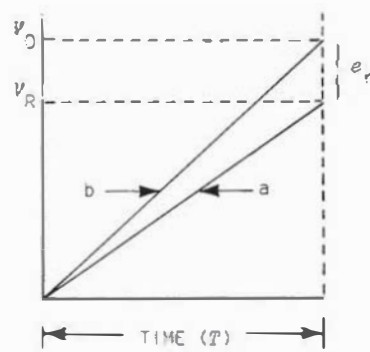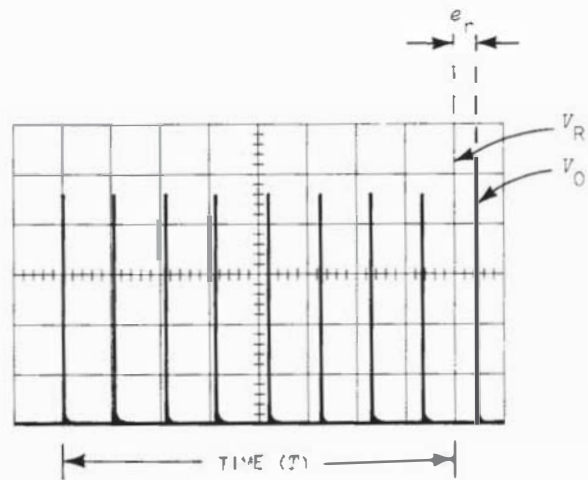Fig. 1-1.   Sawtooth characteristics.

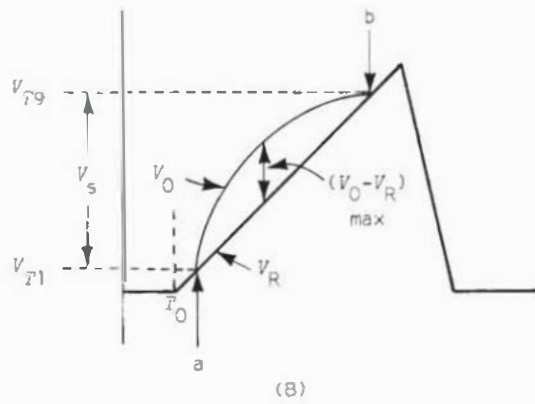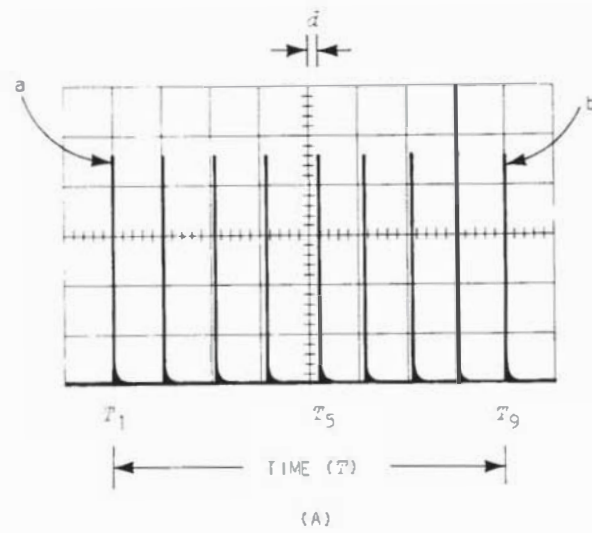Fig. 1-2. Sweep accuracy $= \dfrac{V_R - V_O}{V_R}$.

6

(A)

(B)

Fig. 1-3. Displacement error $\frac{d}{T}$ (100).

Let us now compare the concept of sweep accuracy with that of sweep linearity. According to the definition presented in the previous paragraph, the sweep shown in Fig. 1-3B exhibits 0% error, since its beginning and termination lie on the reference sweep. There is obviously something wrong with the sweep however, since the time markers in the oscilloscope display (Fig. 1-3A) do not coincide with every graticule mark. Any measurement of signal duration made at the center of the graticule would obviously include a positive error. The error, of course, is due to the sweep's nonlinearity (shown highly exaggerated).

displacement error

One definition of nonlinearity makes use of the quantities labeled in Fig. 1-3B. Linearity is described in terms of *displacement error* and is expressed:

$$e_d = \frac{(V_0 - V_R) \text{ max}}{T}$$

Where: e_d = displacement error

V_0 = Observed ramp voltage

V_T = "Reference" ramp voltage

$T = V_{T9} - V_{T1}$

In actual practice, the same results are obtained by measuring the maximum departure from linearity directly from the oscilloscope display, and dividing again by the quantity T. (Note that neither d nor T need be converted to units of time or voltage, but may be expressed in terms of graticule divisions.) It is customary to convert the fraction obtained by this equation to a percentage. Thus displacement error equals $\frac{d}{T}$ (100).

Nonlinearity may also be expressed in terms of *deviation from average* (Fig. 1-4). In this case, it is necessary to measure the total distance (in graticule divisions) between two specified time markers and divide this quantity by the number of intervening marker *intervals*. This calculation yields the average distance between markers (t_{av}). Next, measure the distance between the two most widely separated adjacent markers (t_d). The expression for nonlinearity, again a percentage, then becomes

$$\frac{t_{av} - t_d}{t_{av}} (100).$$

A third method of measuring sweep nonlinearity is illustrated in Fig. 1-5. Sweep time is again adjusted as shown. To obtain the necessary data, first measure the distance between the two most widely separated adjacent markers (t_f). Subtract the distance between the two adjacent markers closest together (t_c), and divide the difference $(t_f - t_c)$ by the length of the trace segment under consideration (T). Thus the difference measurement becomes

$$\frac{t_f - t_c}{T} (100).$$

The technique used to determine the accuracy of a displayed sweep should provide the person making incremental time measurements with information that is most meaningful to him.
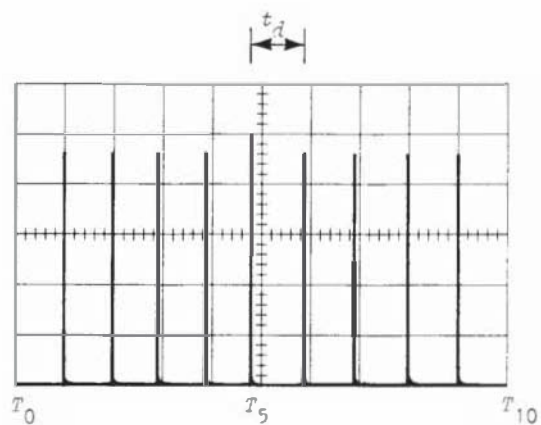
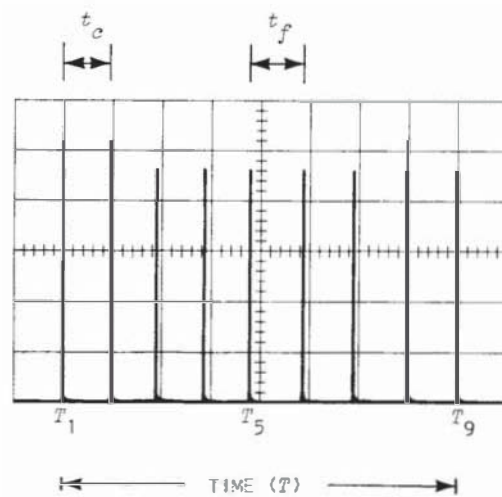Fig. 1-4. Deviation from average $= \dfrac{t_{av} - t_d}{t_{av}} (100)$.

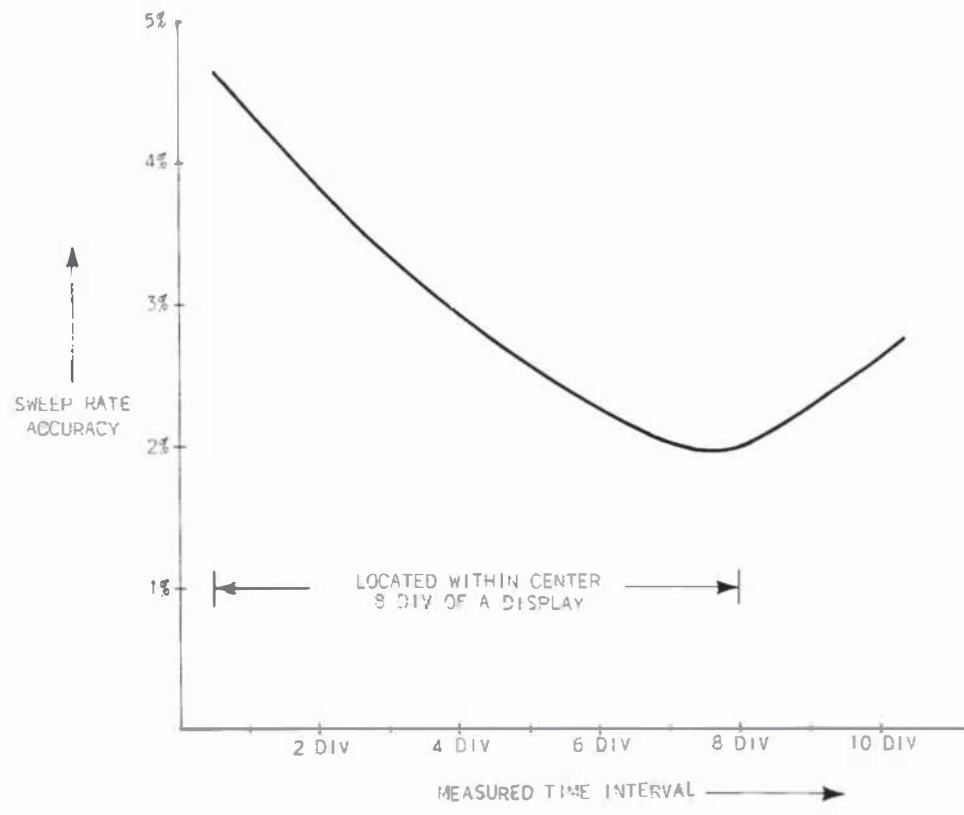Fig. 1-5. Difference Measurement, $\dfrac{t_f - t_c}{T} (100)$.

Fig. 1-6. Sweep rate accuracy versus
measured time intervals.

A recent trend is to state sweep accuracy in terms
of defined increments of a display, thus combining
sweep rate accuracy and linearity in one statement.

combining
rate
accuracy and
linearity
statements

A graph showing how sweep accuracy may change as a
function of a specific measured time interval is
illustrated in Fig. 1-6. For example, a time base
may be accurate to within 3% over a 4-division
interval of a centered 8-division display. If a 2-
division interval is used when making a measurement,
the accuracy may deteriorate to 4%.

Resolution errors are not considered in Fig. 1-6, but
must be considered as the interval becomes shorter.
It is estimated that statements of sweep accuracy
become unreliable when attempting to measure distances
of .5 mm or less.

RETRACE (RAMP RECOVERY)

Upon termination of the ramp portion of the sawtooth
voltage, the sawtooth waveform enters the *retrace* or
ramp-recovery period (part B in Fig. 1-1). Ideally,
the ramp should recover (and restore the electron
beam to its original starting point) in zero time,
since no display occurs during that period. (In
most oscilloscopes, arrangements are made to cut off
the CRT electron beam during retrace time.) A short
retrace time also assures maximum sweep duty cycle
and therefore a brighter trace. For these and other
reasons, retrace time is an important consideration
in the generation of a practical sweep function voltage.

HOLDOFF

Portion "C" of the sawtooth waveform, Fig. 1-2, is a
predetermined period that is commonly referred to as
holdoff. Holdoff time is purposely designed into the
more sophisticated sweep generators to insure the
stabilization of all circuitry that is associated
with the sweep generator. This circuitry
stabilization is necessary for many reasons, some of
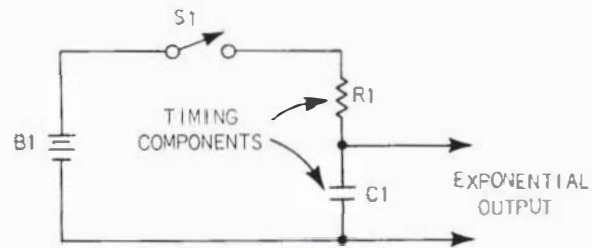which will be discussed in the forthcoming chapters.

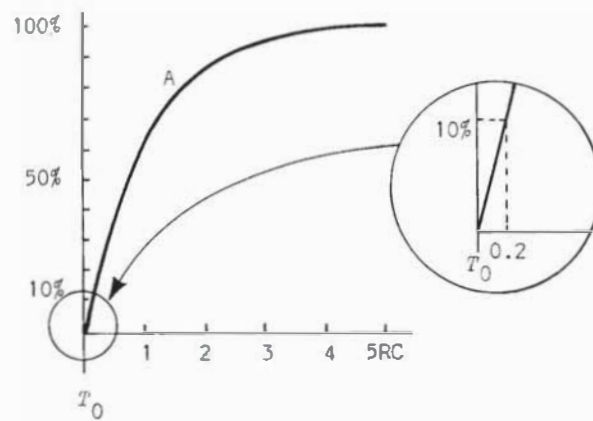Fig. 2-1.   Exponential RC circuit.

Fig. 2-2.   RC exponential curve.

2

THE GATED-CLAMP-TUBE GENERATOR

time
constant

Basically, all sawtooth voltages are generated by charging a capacitor through a resistance. The *rate-of-charge* is determined by the *time constant* of the circuit, which is defined as the time required to charge a capacitor to 63.2% of the total applied voltage. The time constant is calculated from the equation, T = RC. For all practical purposes, after 5 RC periods, the capacitor will become fully charged to the total applied voltage.

exponential
sweep

A basic exponential sweep circuit is illustrated in Fig. 2-1. If switch S1 is closed for a sufficient period, the capacitor will charge to the applied voltage of B1. The rate-of-charge is determined by the product of R and C. Fig. 2-2 shows a typical charge curve, as a plot of percentage-of-charge versus elapsed time.

Due to the exponential characteristic of the capacitor voltage, the total waveform would not provide a linear time base for the CRO. A close examination of curve "A" in Fig. 2-2 will reveal, however, that the first portion of the curve, (corresponding to 10% or less of $E_{applied}$), is reasonably linear. If switch S1 were opened at approximately .2 RC, the resulting waveform could be useful as a sweep function voltage. This method of ramp derivation will be described in detail in the following discussion of the Gated-Clamp-Tube Generator.
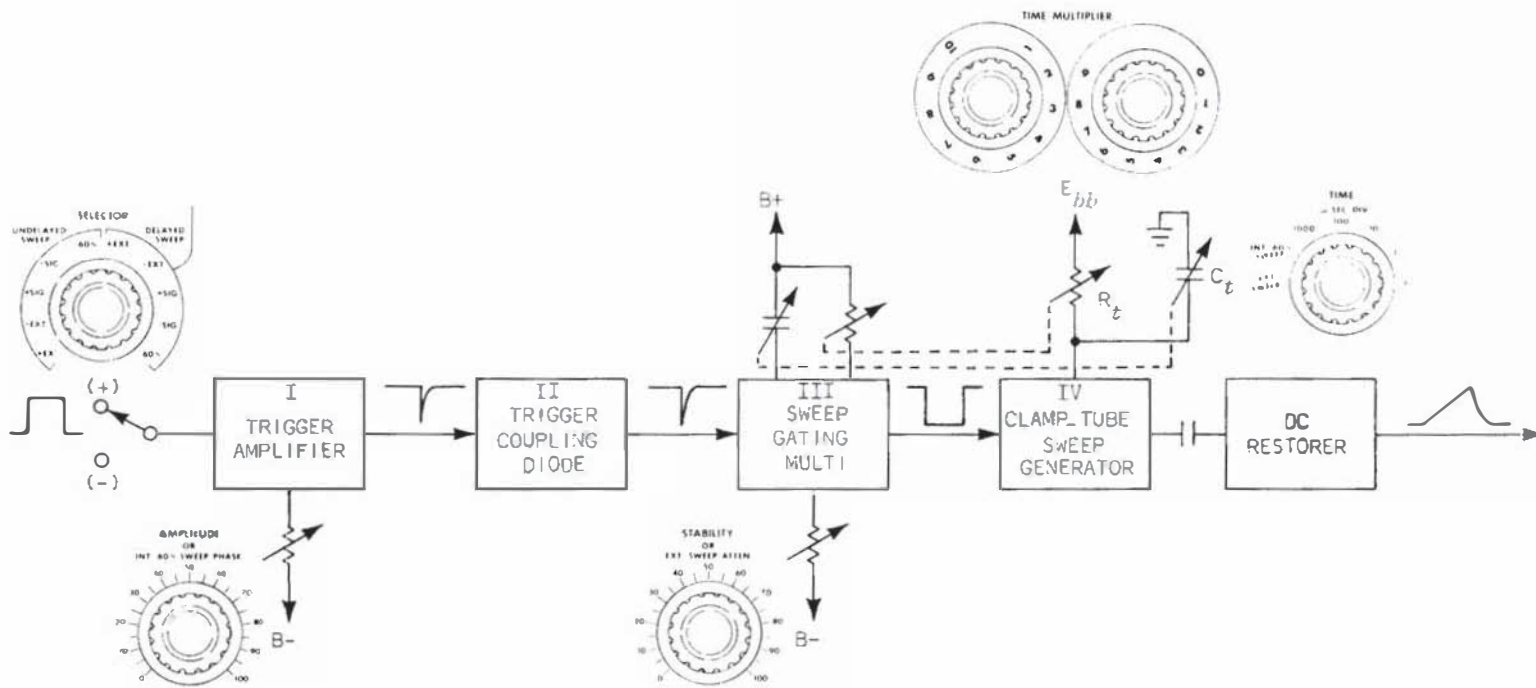
Fig. 2-3.

The block diagram in Fig. 2-3 includes a basic gated-clamp-tube generator and its control circuits. The sweep generator block consists of an active device, in this case a high-g_m vacuum tube, and associated circuitry. The vacuum tube will function as the switch (S1), discussed in Fig. 2-1. Referring to Fig. 2-4A, control-grid divider R1 and R2 establish a fixed bias very close to zero volts. Quiescently, the tube conducts heavily. The plate voltage (E_2) is very close to the cathode potential. Upon receiving the negative gate voltage (e_{in}) at time T_0, the plate voltage starts to rise toward the supply voltage E_{bb}. Due to the circuit configuration, the plate voltage rises at an *exponential rate*, determined by the product of R_t and C_t.

charge
amplitude

The *amplitude* of charge across C_t is determined by the *duration* of the gating pulse and the *rate-of-charge* of C_t (See Fig. 2-4B). If the gating-pulse duration is very short with respect to the $R_t \cdot C_t$ time constant, the charge across C_t will be small compared to the supply voltage. This ratio of sawtooth voltage to power supply voltage constitutes one of the disadvantages of this technique. To get a practical sawtooth voltage amplitude, the power supply voltage must be very high. The sawtooth voltage e_{Ct} can be expressed as:

$$e_{Ct} = \frac{E \cdot t}{R_t C_t} \quad \text{where } E = E_{bb} - E_2$$

The percentage of charge can be expressed as:

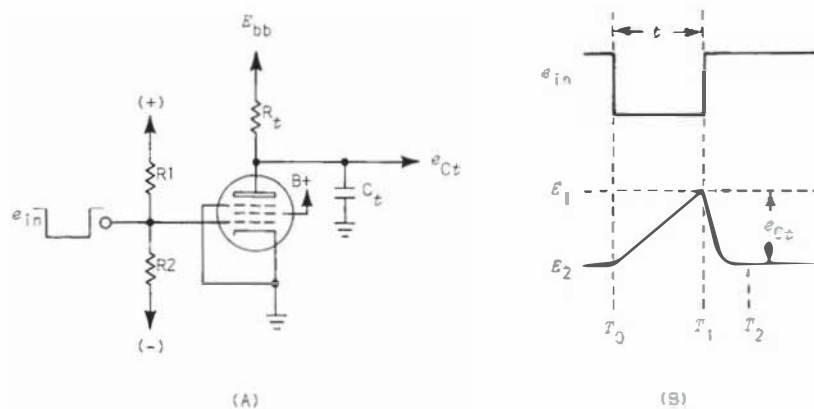$$\% \text{ Charge} = \frac{e_{Ct}}{E_{bb}} (100)$$

(A)   (B)

Fig. 2-4.

As was stated previously, if the percentage of charge
on C_t is maintained at 10% of the applied voltage,
the ramp will be reasonably linear.

Upon termination of the negative gate pulse
(Fig. 2-4A), the tube rapidly returns to its
original zero-bias state. The effective r_p of the
tube, under zero-bias conditions, is very small,
and provides a relatively low impedance path for the
timing capacitor discharge current during retrace
time. The time constant of the circuit (Fig. 2-4B)
during retrace $(T_2 - T_1)$ is the product of the r_p of
the zero biased-vacuum tube and the capacitance
(C_t) of the timing capacitor. If the control grid
is held in its quiescent state, the plate voltage
will also return to its quiescent level (E_2) and be
ready to receive another gate pulse.

It is important to note at this point, that no steps
have been taken to insure that the sweep will return
to its quiescent condition before another gate pulse
arrives. A technique for assuring complete ramp
recovery is clearly desirable. In the chapters to
follow we shall examine some of the methods by
which this "holdoff" period is established.

gate pulse
width versus
ramp
amplitude

It should be apparent that there is an interdependent
relationship between the timing components of the
gating-multi and the RC timing components of the
sweep generator. If the SWEEP TIME/DIV is made
variable by permitting changes in the values of R_t
and C_t, it follows that in order to maintain the
required *ratio* of gate-pulse-width to ramp rate-of-
rise, the values of the timing components in the
gating-multi must be changed proportionately.

Component values vary within a given tolerance and as a result, the ratio of gate-pulse duration and ramp rate-of-rise varies. The effect of this variation is a noticeable change in *sweep length* on the CRT display, as SWEEP TIME/DIV is changed. To compensate for sweep-length variations, the gate-pulse duration is extended so that the sweep terminates off the face of the CRT. This technique insures adequate sweep length to allow measurements within the entire graticule area at all sweep speeds.

sweep
length

Having completed discussion of the sweep generator block in the block diagram of Fig. 2-3, we will now discuss the relationship of the remaining circuits to the sweep generator.

Previous discussion of the sweep generator indicated that the "clamp-tube" generator is a *driven device*. That is, the sweep generator remains in a given quiescent state indefinitely unless a "gate voltage" is applied. The gate voltage contains all of the information required by the sweep generator to produce sawtooth voltage.
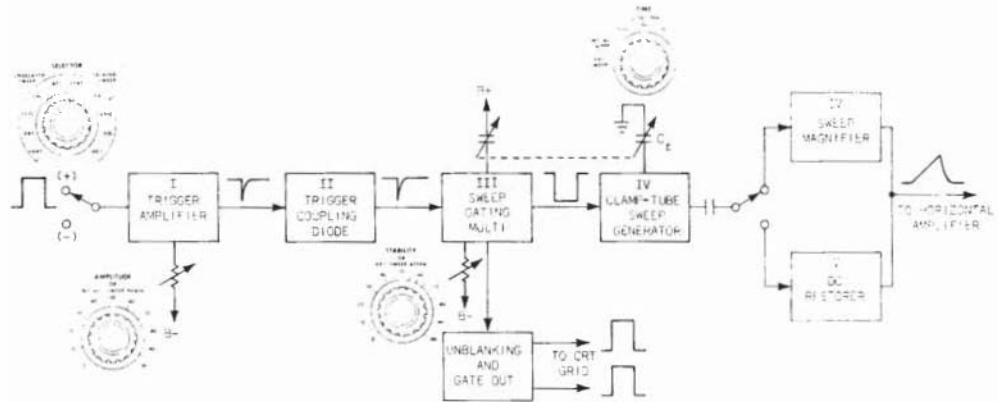
Fig. 2-5.

gated-clamp-
tube sweep
generator

Block III in Fig. 2-5 represents a Sweep-Gating
Multivibrator. Its primary function is to convert
trigger signals of various shapes into gate pulses
of controlled duration. These gate pulses are then
used to: (1) gate the sweep generator, and (2)
unblank the CRT during sweep time.

Fig. 2-6 is a schematic diagram of a simplified
gated-clamp-tube sweep generator and its associated
circuitry. We shall refer to this illustration while
discussing the operation of the sweep-gating
multivibrator.

Fig. 2-6. Gated-clamp-tube sweep generator.

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The gating-multi operates as a *monostable* device. The mode of operation is dictated by the requirement of the triggered sweep. The multi may also be operated in an *astable* or free-running configuration. The astable mode is useful if a synchronized sweep becomes necessary. Adjustment of the STABILITY control determines the mode of operation of the gating multivibrator.
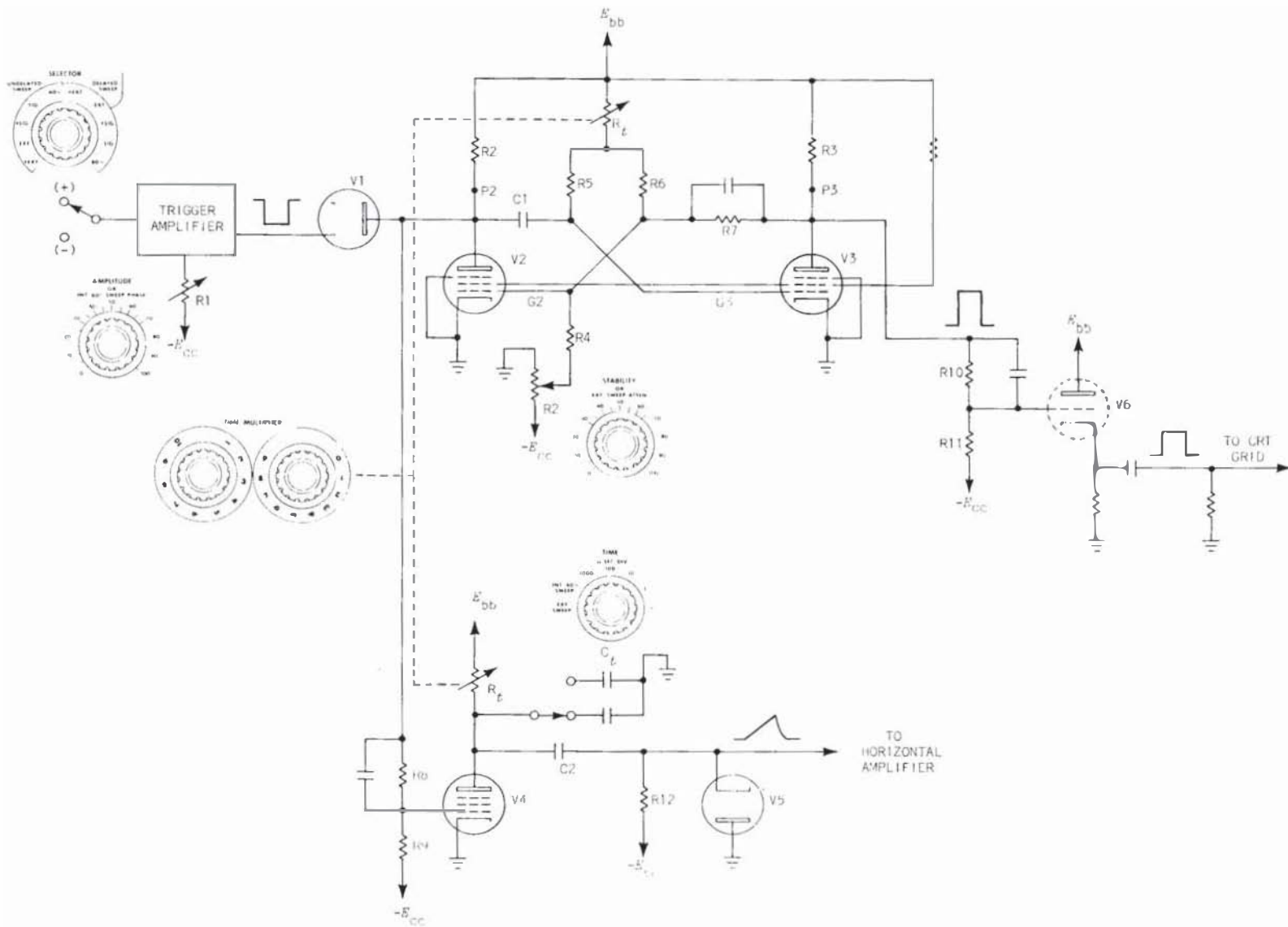
Prior to the application of a trigger signal, the gating-multi must be in a "ready" state. This state requires that V2 be *off* and V3 *on* (i.e., V3 is in clamp) due to its zero bias. While referring to Fig. 2-6 and the ladder diagram of Fig. 2-7, we will describe the operation of the gating-multi and the effects of its operation upon the associated circuitry.

After a trigger signal is processed by the trigger circuits, it attains the proper polarity and magnitude required by the sweep-gating multi. The trigger is applied to the coupling diode (V1) at time T_0 (Fig. 2-7), forward biasing V1. The impulse is coupled through C1 to the control grid of V3. V3 begins a rapid transition toward cutoff, which results in a positive step voltage at P3. The positive step is impressed upon the control grid (G2), through the divider R7, R4 and R2, driving G2 into clamp or to a point within its grid base.* The voltage at P2 falls, carrying G3 with it, insuring cut-off of V3. The multi is now in its quasi-stable state and totally nonreceptive to incoming triggers that may be present.

gating
operation

The duration of this quasi-stable state is determined by the RC time constant of C1, R5, R_t and the r_p of V2 during its conducting period. The period of the quasi-stable state is also the period of sweep, since the negative excursion at P2 is impressed upon the control grid of V4. This negative "gate" brings V4 out of clamp and allows C_t to begin charging toward the supply, starting ramp generation.

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*The region of grid bias between that which causes plate current cut-off and that which causes grid current flow.

In Fig. 2-7, during the period $T_1 - T_0$, as C1 charges, the voltage at G3 will rise exponentially toward E_{bb} through R5, R_t and the dynamic plate resistance (r_p) of V2. At time T_1, the control grid (G3) rises to a voltage that allows V3 to come into conduction. The increased plate current in V3 results in a negative-going voltage at P3. The negative-voltage step is DC coupled to G2 and begins to cut off V2. As a result of the decreasing plate current of V2, the voltage at P2 starts toward E_{bb}, causing G3 to rise to a point that places V3 in a state of clamp. The tube V3, in clamp, establishes V2 in cutoff and the multi is once again in the "ready" state.

AC
unblanking

Prior to sweep generation, the INTENSITY control is normally set to establish a near cut-off condition to the CRT electron beam. At time T_0, when the plate of V3 was rising toward E_{bb}, the control grid of V6 (the unblanking cathode follower) was carried with it. Through cathode-follower action, V6 couples a positive voltage step, or *unblanking pulse* to the CRT grid via a DC blocking capacitor, and is of the proper amplitude and duration to allow sufficient beam intensity during sweep time. Due to AC coupling the unblanking pulse, it is necessary to readjust the INTENSITY control each time the SWEEP TIME/DIV is changed. The intensity change is caused by changing
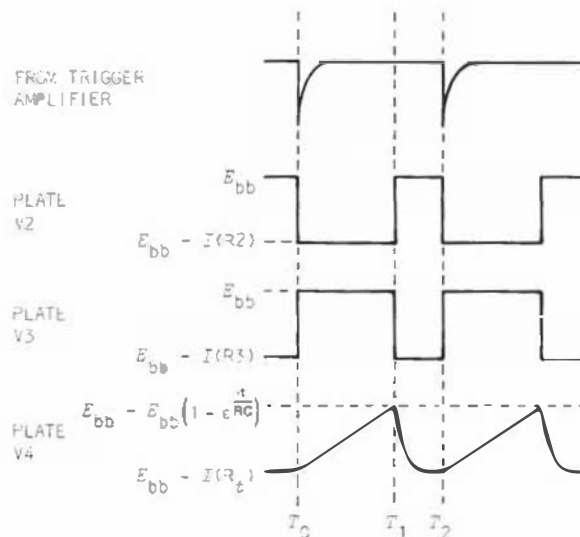

Fig. 2-7.

the duration of the pulse that is driving a circuit
having a fixed RC time constant. This results in an
average DC level change at the control grid of the
CRT. This method of unblanking caused some degree
of inconvenience to the operator, and as a result,
more sophisticated unblanking methods were to be
developed in later instruments.

Having discussed one complete cycle of operation of
the gating multi, and its relationship to the
associated circuitry in the sweep generator block
diagram, several conclusions can be drawn. The
gating multi:

(1)   Provides control of the sweep period
      (sweep length).

(2)   Generates the unblanking signal.

(3)   Converts various-shaped trigger signals
      to defined pulses of controllable
      duration, suitable for controlled
      sweep generation.

DC
restoration
(ramp-start
reference)

The circuit in Fig. 2-6, consisting of V5, C2, and
R12, is a DC restorer, or clamping circuit. The
circuit provides a fixed starting reference for the
ramp voltage.

Under static conditions, diode V5 is conducting and
the voltage at its cathode will be about -.5 volts
(the forward voltage drop of V5 when conducting).
The time constant C2 and R12 is long enough to insure
undistorted signal transfer at the slowest sweep speed.

When a positive-going ramp voltage is applied to the
coupling capacitor C2, the current through R12
develops a positive voltage which cuts off V5. The
ramp continues to run up for the period of the gate
pulse. At the end of sweep time, the applied voltage
falls rapidly toward the original starting point.
The negative-going voltage appears across the
nonconducting clamp diode until the turn-on voltage
is reached (-.5V). When V5 conducts, no appreciable
voltage change occurs at its cathode (due to the
diode's low forward resistance), even though the
applied voltage may continue to fall. The junction
of V5 and R12 is now *clamped* to that voltage level
established by the forward-biased clamp diode.

The method of DC restoration just described establishes the fixed ramp-start level necessary to assure a relatively stable beam position, independent of sweep speed variations.

The gated-clamp-tube generator has some apparent disadvantages:

(1) The lack of complete ramp recovery assurance, i.e., holdoff;

(2) Less than desirable sweep linearity;

(3) Large power-supply voltage requirements, necessary to provide practical sawtooth voltage amplitudes;

(4) The DC restorer, having a definite low-frequency cut-off point, limits the minimum sweep speed obtainable;

(5) Inability to control and adjust sweep length;

(6) The inconvenience of having to maintain a fixed ratio of sweep-generator timing components to gate-multi timing components.

In the chapter to follow, we will discuss some considerations that brought some modifications and refinements to the gated-clamp-tube generator.
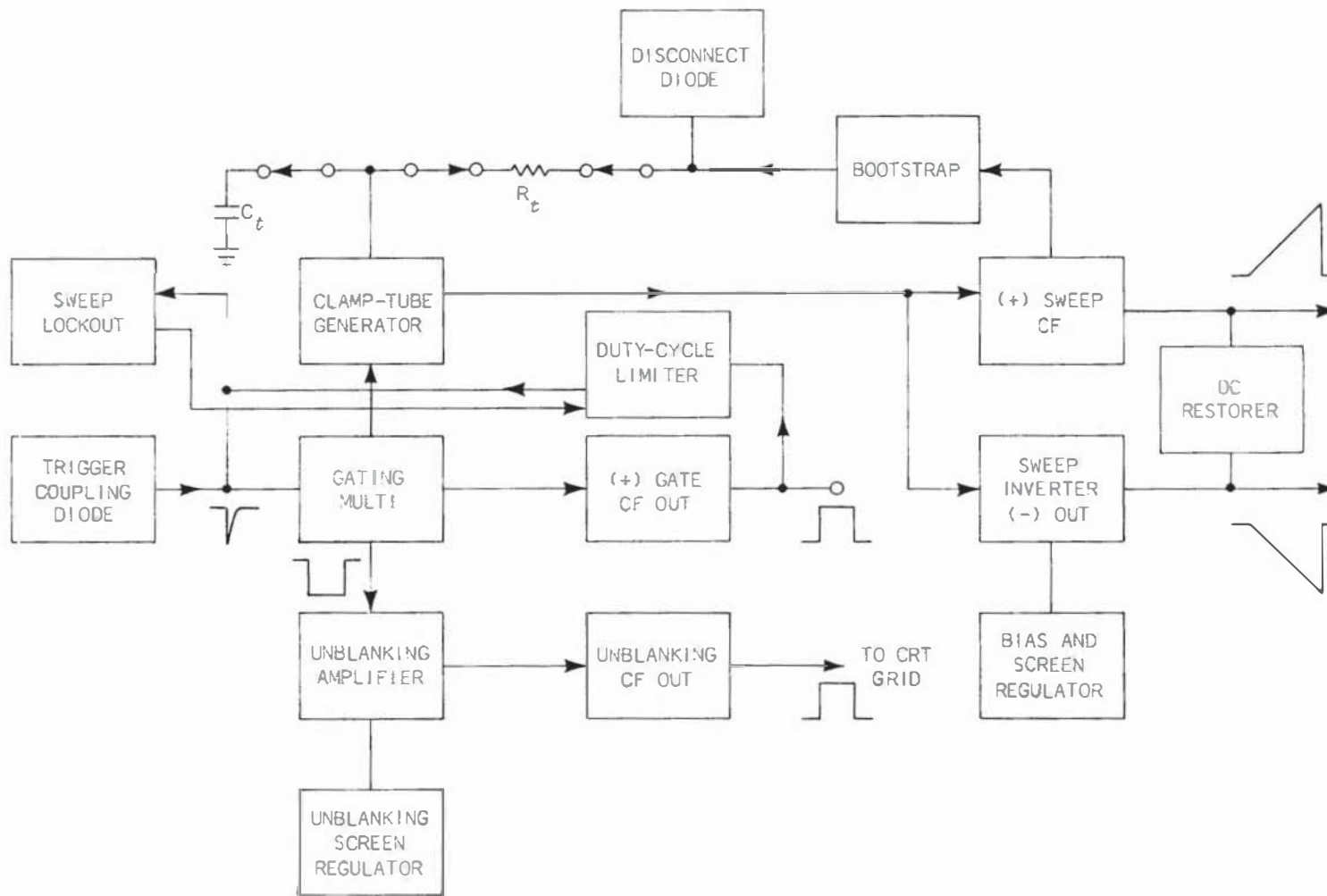
Fig. 3-1. Bootstrap sawtooth generator and associated circuits.

3

THE BOOTSTRAP SWEEP GENERATOR

feedback

One method of improving the quality of sawtooth
generation is to incorporate feedback. Feedback
will provide a means of supplying a relatively
constant current to the charging circuits, in
addition to supplying a larger sawtooth-output
voltage for a given power supply voltage. In the
following paragraphs we will discuss the provision
of *positive feedback* as an attempt to overcome some
of the disadvantages found in the gated-clamp-tube
generator.

We are aware that when attempting to charge a
capacitor to some voltage, through a resistance, the
voltage across the capacitor becomes *series opposing*
to the applied voltage. Current in the external
circuit is reduced as the capacitor becomes fully
charged. This results in a nonlinear voltage rate-
of-change. One method of overcoming this exponential

constant
I_{Ct}

characteristic would be to cause the supply voltage
to change *exactly* as the capacitor voltage changes,
thus maintaining a constant current in the capacitor.
One circuit that operates under this principle of
positive feedback is commonly referred to as the
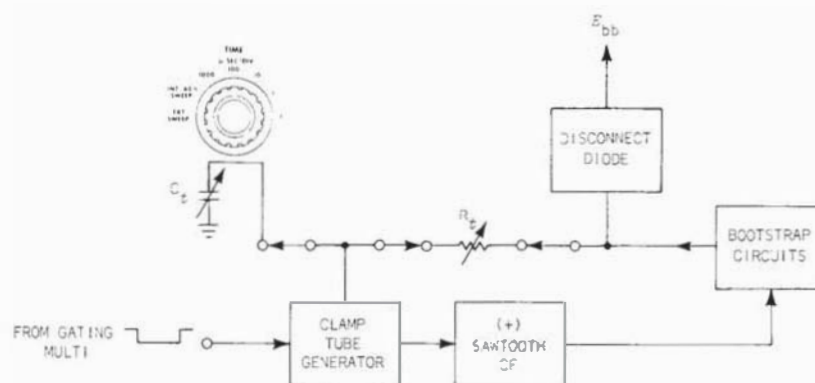Bootstrap Sweep Generator. (See block diagram in
Fig. 3-2).

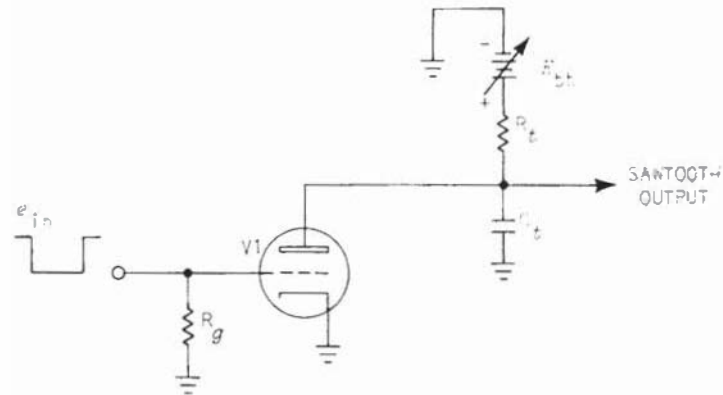Fig. 3-2. Basic bootstrap generator.

Fig. 3-3. Basic bootstrap generator.

Fig. 3-3 is a schematic diagram of the familiar clamp-tube generator, with the addition of a fictitious variable power supply. If a negative gate signal were applied to V1, interrupting plate current flow to R_t, the timing capacitor (C_t) would begin to charge toward E_{bb}. Referring to the ladder diagram in Fig. 3-5, at time T_1 (+), the capacitor would charge to some voltage ($\bar{X}$). If we could increase the supply voltage by "X" amount at the same time, we would realize a constant-voltage drop across R_t, resulting in a constant charging current to capacitor C_t. As was stated previously; supplying a constant charging current to a capacitor results in a linear voltage rate-of-change.

Fig. 3-4 represents a simple bootstrap sweep generator. Triode V1, R_t and C_t make up the clamp-tube generator. Triode V2, R_k and C_b make up the feedback circuit that provides the bootstrap effect. The circuit operation is as follows: Prior to gating V1, the sweep generator is established in its quiescent state. V1 is biased at zero volts, supplying a heavy plate current to R_t and V3. The voltage levels are as indicated in Fig. 3-5. At time T_1, the grid of V1 is driven sufficiently negative to cut off V1. The timing capacitor (C_t) starts to charge through R_t toward the supply voltage. As the voltage E_{Ct} rises, the grid of V2 is carried with it. Assuming for the present that V2 has unity gain, its cathode will rise a like amount. At time T_2, the sum of E_{Rk} and E_{Cb} exceeds the power-supply voltage, cutting off V3. The bootstrap capacitor

(C_b), charged to $E_{bb} - E_{V3}$, now becomes the power
supply for the charging circuit. The lower plate of
C_b rises with the cathode of V2, and if we assume
that the voltage across C_b remains constant, the
upper plate of C_b must increase by the same amount.
The charging current path can be traced from the
upper plate of C_t, through R_t to the positive plate
of C_b, from the lower plate of C_b, through V2, to the
V2 plate supply. To maintain this current during
ramp run-up time, the plate supply of V2 must be
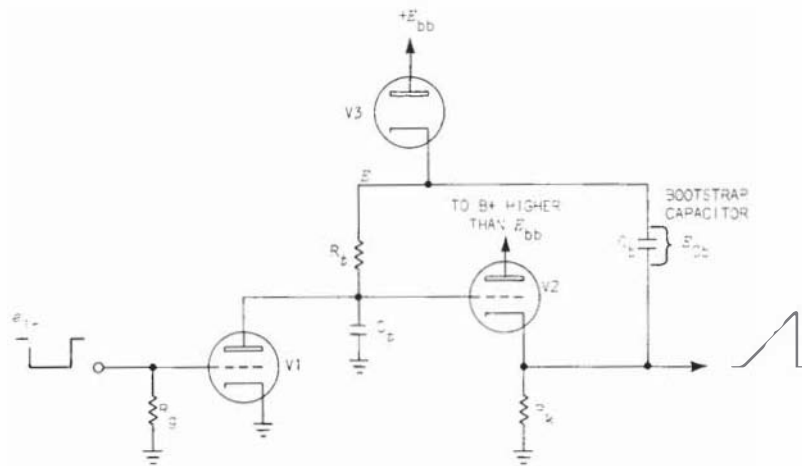considerably higher than E_{bb}.
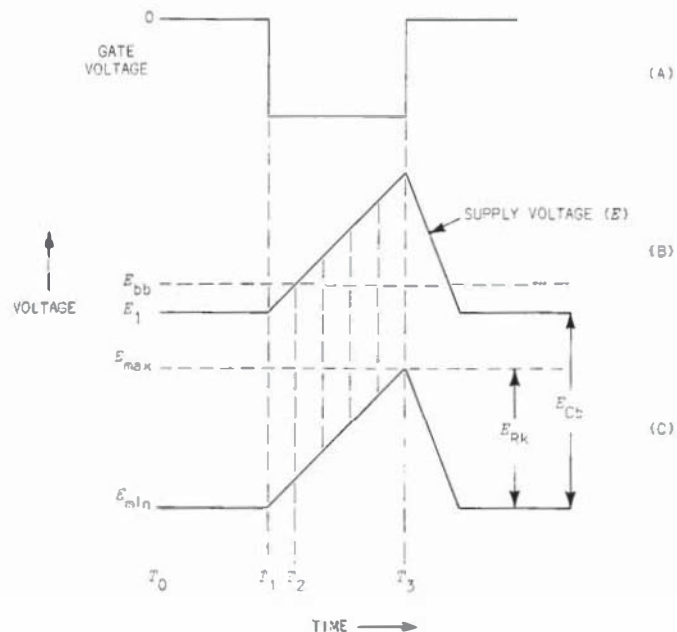
Fig. 3-4. Bootstrap sawtooth generator.

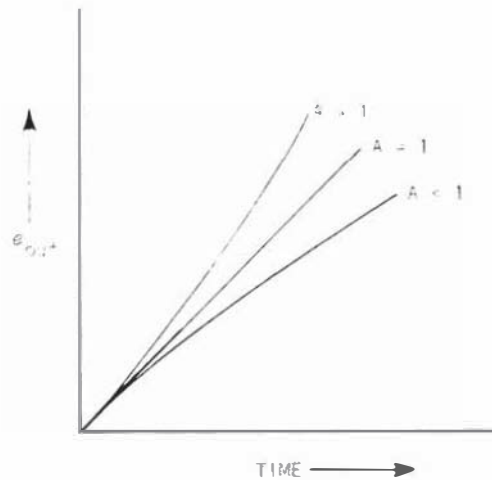Fig. 3-5. Bootstrap generator voltage waveforms.

Fig. 3-6. Slope deviation related to
amplifier loop gain.

Upon termination of the negative gate pulse, V1 comes
back into saturation, allowing C_t to discharge
rapidly through the low r_p of V1. The abrupt voltage
change across C_t is coupled through V2 and C_b causing
V3 to come into conduction, restoring C_b to its
initial condition. The discharge path for C_b is
through V3, R_k, back to the lower plate of C_b.

Due to the positive feedback action just described,
the voltage at the cathode of V3 increased by the
same amount as E_{Ct}. We can conclude then, that we
have maintained a constant voltage drop across R_t.
By Ohm's Law, a constant voltage across R_t implies
a constant charging current, satisfying the basic
requirement for linear ramp generation.

loop gain
errors

The previous discussion was based on the assumption
that several conditions were ideal. For example,
the loop gain, (total gain around the feedback loop),
was assumed to be unity. Any deviation from unity
gain in the bootstrap generator will contribute to
ramp nonlinearity. See Fig. 3-6.

C_b to
C_t ratio

Notice also that in actual operation, the charging
current tends to discharge C_b. Recalling that during
ramp run-up, the voltage across C_b must remain
constant to assure absolute linearity, it is clear
that the tendency of the charging current to
discharge C_b would contribute to ramp nonlinearity.

To offset this tendency, C_b should be much larger than C_t. As a rule-of-thumb, the ratio of C_t to C_b should be about 20. The bootstrap capacitor should not be too large however, lest circuit-restoration time be increased, causing the generator to become sensitive to changes in repetition rate. For a mathematical treatment of linearity, refer to Millman and Taub, *Pulse and Digital Circuits*, pp 230-235.

Fig. 3-3 shows V1 as a triode. This configuration was used simply for ease of illustration. In actual practice, the switching tube, in both the clamp-tube and bootstrap generator, is a high G_m pentode. The pentode is used mainly because of its constant-current characteristics. For example, for given screen grid and control grid-to-cathode potentials, the plate current of a pentode is relatively independent of plate voltage variations, except when operating at relatively low voltages. A typical example is type 6AG7 power pentode tube. With a grid bias of -2 volts and a screen potential of 150 volts, a change in plate voltage from 300 volts to 200 volts causes a plate current change from about 40 mA to 38 mA. If this current change were not relatively constant, the linearity of the ramp would suffer considerably.

pentode
switching
tube

Serious considerations should be given to the bootstrap generator in terms of reliability. Tube aging and line voltage variations (affecting heater temperature) contribute largely to linearity errors in the bootstrap generator. If the loop gain of the bootstrap generator were to change by only 1%, the ramp linearity would change by a factor of 2.

Many of the disadvantages of the clamp-tube-generator have been overcome in bootstrap generator. Through incorporation of positive feedback, it is possible to design a circuit that generates a relatively large-amplitude sawtooth waveform with good linearity, using practical power-supply voltages. The bootstrap generator can be made to generate sawtooth voltages of sufficient amplitude to drive the CRT deflection plates directly, thereby eliminating the need of the horizontal amplifier. Elimination of the horizontal amplifier permits much faster sweep speeds, limited only by the ability of the generator to drive CRT deflection plate

impedance. The bootstrap generator has been utilized in Tektronix instruments to provide sweep speeds (unmagnified) ranging from 20 μs to 2 ns/div.

duty cycle
limiter

The block diagram in Fig. 3-1 represents an entire sweep generator and its control and output circuits. The block labeled "Duty Cycle Limiter" will be discussed in the following paragraphs.

The duty-cycle limiter is included in some bootstrap sweep generator systems to automatically limit the duty cycle of the sweep, thereby compensating for possible over-dissipation of circuit components, should the sweep repetition rate increase significantly.

In Fig. 3-7, the plate of V2 is connected to the control grid of that portion of the multi (A) that drives the clamp tube in the sweep generator. A complimentary output of the gating multi (B), is applied to the control grid of V2, through resistive-voltage divider R6, R7, and R8. The divider R1 and R2 sets the control grid of V1 to some fixed DC voltage. Potentiometer (R7) is adjusted to a voltage slightly more negative than that voltage at the grid of V1, allowing V1 to conduct the current I_{R3}. Since V2 is cut off, the sensitivity of the multi can be set by adjusting R10. (See circuit description of sweep-gating multi in Chapter 2.) An incoming trigger pulse switches the multi from its stable state to its quasi-stable state. A negative pulse is generated at point (A), cutting off the clamp tube in the sweep generator, allowing the ramp to begin its rise. At the time point (A) goes negative, point (B) steps positive.

For repetitive sweeps, these positive voltage steps are applied to the control grid of V2 through the divider R6, R7, and R8. C1 will charge to the *average* voltage at the junction of R5 and R7. For example, assume that the positive pulses present at the top end of the grid divider have peak amplitudes of 20 V. As C1 charges toward 20 volts, the voltage at the grid of V2 will reach a level more positive than that voltage at the control grid of V1. Current through R3 will then switch from V1 to V2. The voltage at the junction of R4 and R9 will become more negative, reverting the gating multi to its stable state. The multi, in its stable state, clamps the sweep generator, terminating the sweep. The period of the sweep then, is determined by the RC time constant of C1 · R5. The duration of the sweep generator's quiescent period is determined by the *same* RC time constant, but the voltage present for discharge is considerably less than that present at charge time, so the period required for C1's discharge is much greater than that required for its charge. This ratio of charge-to-discharge time of C1 determines the duty cycle of the sweep.
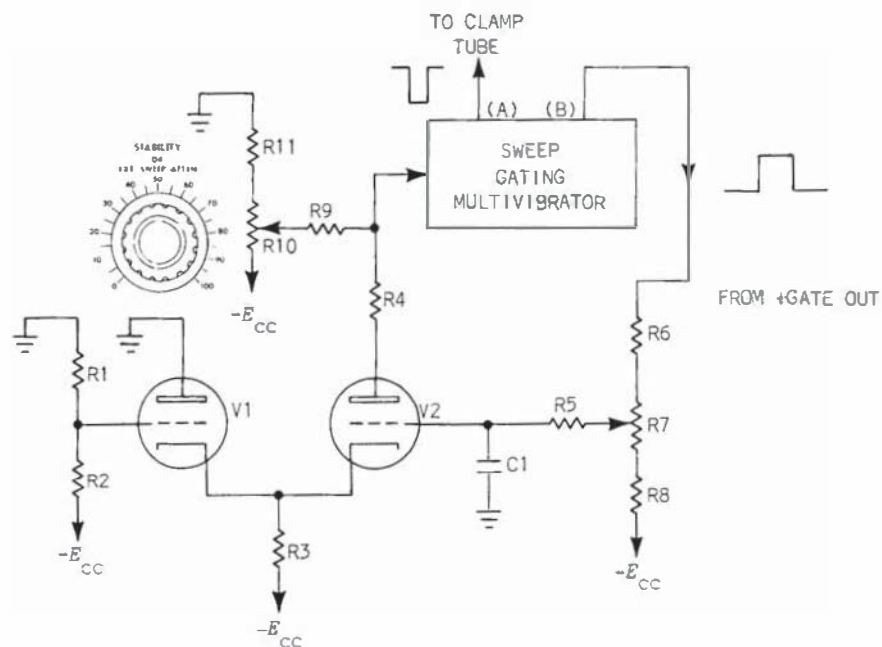

Fig. 3-7. Duty-cycle limiter.

TRIGGER SIGNAL TO GATE MULTI

SAWTOOTH OUTPUT

DUTY-CYCLE LIMITER ACTION

$T_1 - T_0$ = C1•R5 CHARGE TIME
$T_2 - T_1$ = LIMITING ACTIVITY
$T_3 - T_2$ = C1•R5 DISCHARGE TIME

GATE MULTI OUTPUT (A)

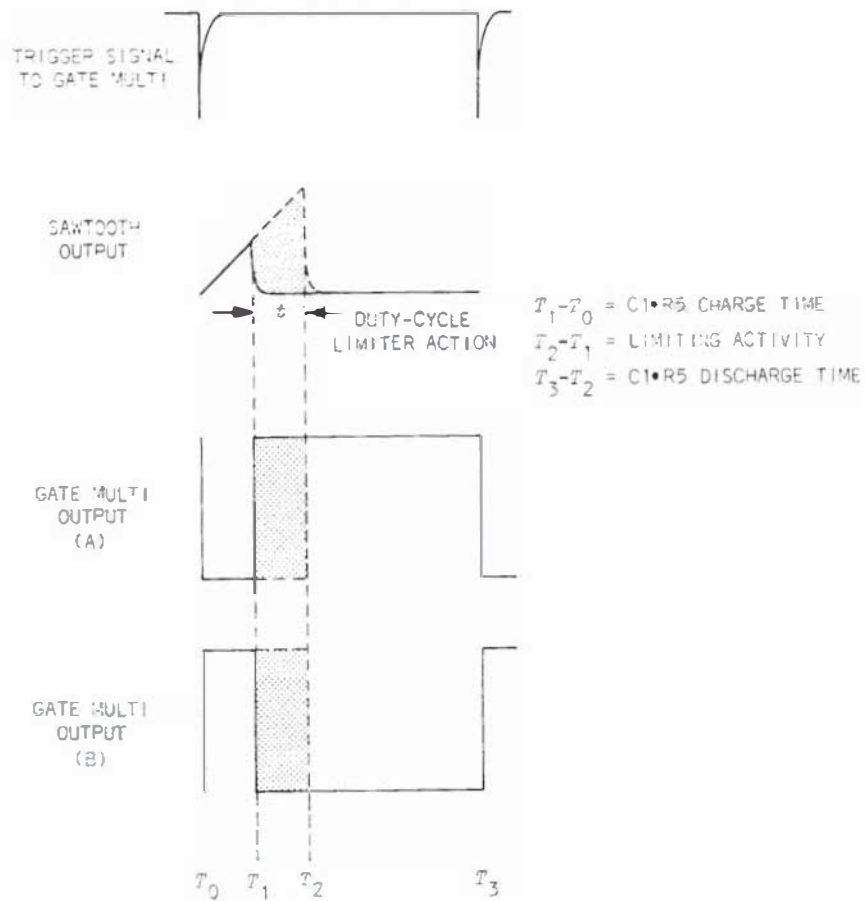GATE MULTI OUTPUT (B)

T_0 T_1 T_2 T_3

Fig. 3-8. Effects of duty-cycle limiting.

The ladder diagram in Fig. 3-8 illustrates the effect of limiting the sweep duty cycle. Without the limiter, the sweep period would be $T_2 - T_0$, the period established by the RC timing components of the gating multi. The addition of the duty-cycle limiter automatically reduces the sweep period to $T_1 - T_0$. In short, at relatively low trigger-repetition rates the RC-timing components of the gating multi establish the sweep duty cycle. However, if the trigger repetition rate increases to a point where C1 does not have sufficient time to discharge, the period of the sweep will be reduced. This effect is revealed by a reduction in the trace length on the CRT. If the trigger rep-rate continues to increase, the duty-cycle limiter can disable the sweep generator completely.

As wider bandpass vertical systems were developed, faster sweep rates were required. If we require the sweep generator to rise to a given voltage in a given period of time, we must supply the timing capacitor with a given amount of current. The equation $\Delta I_C = C\frac{\Delta V}{\Delta T}$, expresses the current demand of the sweep generator for a given ramp rate-of-rise. It is evident that if we wish to increase sweep rate, we must supply a specified change in voltage to the horizontal deflection plates in a shorter period of time. If we assume the sweep generator timing capacitor to be as small as possible, we realize that reducing time (ΔT), keeping ΔV and C constant, the current demand of the sweep generator increases. The power supply must be capable of delivering this current. Power handling capabilities of circuit components must also be increased to insure reliability of the system. Risetime and falltime periods become shorter, requiring additional frequency compensation of the circuits. Special attention must be given to the transmission of various signals within the system to insure proper circuit events that are time related. For example, the CRT must be unblanked prior to sweep generation, if the *entire* sweep is to be displayed. We must be aware of these considerations when we introduce practical techniques of high-speed sweep generation.

high-speed sweep generation

When a signal is applied to the vertical system of an oscilloscope, it may arrive at the CRT deflection plates before the horizontal sweep is started. The vertical signal must therefore be *delayed* to allow the horizontal deflection system sufficient time to start the sweep. (See Vertical Amplifier Circuit Concepts Book discussing delay lines.) One of the primary considerations in the design of the sweep generator system then, would be to start the sweep display as soon as possible after the application of signal to the vertical system, thereby reducing the amount of delay required by the vertical system.

signal delay

When a trigger signal is applied to the horizontal deflection system, a finite period of time elapses between the occurrence of the trigger and the start of the sweep display. Under certain conditions, this period may be too long. A practical circuit used to reduce this transition time, using only

passive devices, is shown in Fig. 3-9. The circuit provides gating and unblanking pulses of proper duration, while preserving fast-rise characteristics of the applied trigger signal.

clip
lines

The following is a discussion of gate generation using the *Clip Line*.

Upon application of a positive trigger pulse, through a lumped-constant delay line, to the clip line, the pulse will travel down the delay line in a predetermined period of time. The pulse is felt by the unblanking amplifier prior to its arrival at the gate amplifier. Initiating the unblanking circuits prior to starting the sweep is necessary if we are to see the entire sweep display. As the pulse travels down the line, the sweep-gating amplifier is activated, which in turn starts the sweep generator. These circuits will continue in an *on* state until the pulse has had sufficient time to travel down the clip line and be reflected. Due to the short-circuit at the end of the clip line, 100% of the pulse is inverted and reflected back to the source. When the reflected pulse arrives at points B and A, in that order, the associated circuits are turned off. The duration of the sweep-gating pulse, or sweep display, is determined by the *length* of the clip line. The clip line technique of gate generation is useful only at very fast sweep speeds, typically 80 ns or faster.

The gate duration is limited by several factors. First, it cannot be longer than the duration of the driving pulse. Secondly, the physical length of the clip line is limited by the available space in the instrument.

gate
extender

At slower sweep speeds, gate generation is provided by the familiar multivibrator. We must retain ability to start the sweep as quickly as possible. The risetime of the trigger pulse must be preserved while providing longer-gate periods. The gate extender is a device that responds to a fast-rise trigger pulse and assures a suitable *gate duration* for sweep gating and unblanking at slower sweep rates.
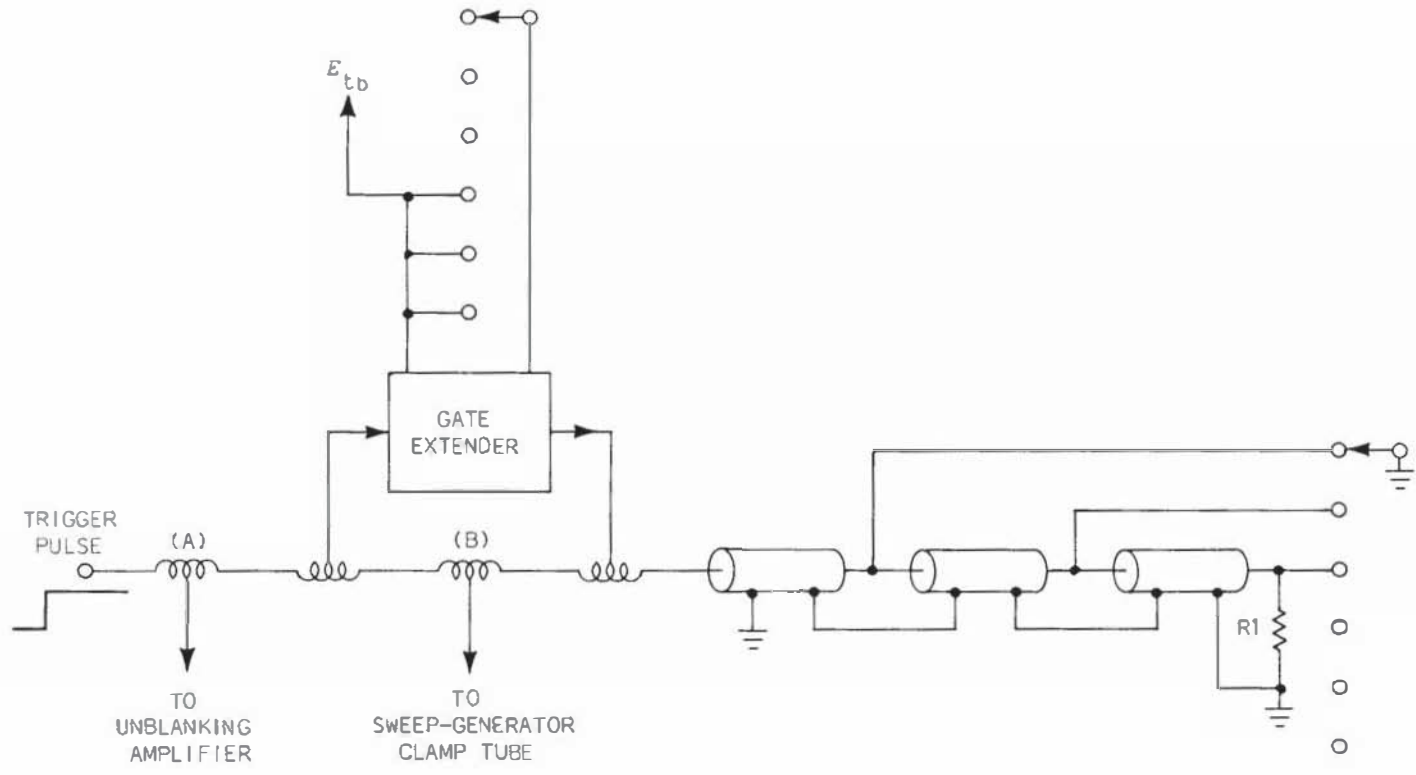
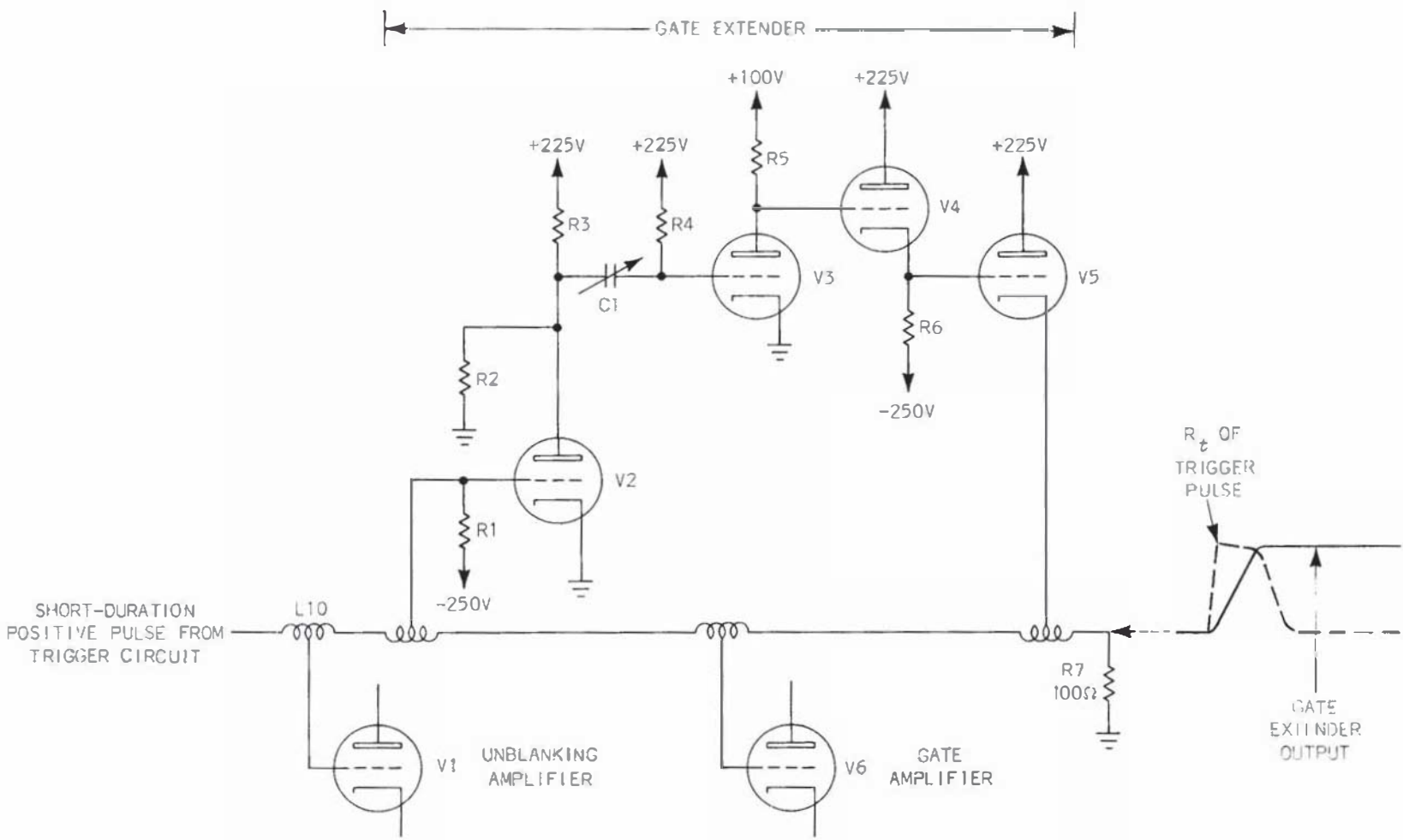Fig. 3-9. Clip line and gate extender.

Fig. 3-10. Simplified gate extender.

gate
extender
operation

Fig. 3-10 represents a simplified gate-extender circuit. A positive trigger pulse, of very short risetime and pulse duration, is applied to delay line L10. The pulse is first felt by the unblanking amplifier V1, to get the CRT unblanked before sweep start. Later, the gate extender action is initiated. Finally, the sweep-gate amplifier is activated, which in turn drives the clamp tube in the sweep generator to start ramp generation. The trigger pulse is then dissipated in R7 to prevent reflections along the transmission line.

As it travels down the delay line, the positive trigger pulse brings V2 out of cut-off. The voltage at the plate of V2 falls, driving the grid of V3 negative through C1. The voltage at the plate of V3 steps positive toward the supply, carrying the grid of V4 with it. The cathode of V4 follows the grid, thereby increasing the cathode voltage of V5. The regenerative action created by coupling the positive signal at the cathode of V5 to the control grid of V2 insures the quasi-stable state of the gate extender for a period determined by the R4 · C1 time constant.

The output of the gate extender alone cannot start a sweep display fast enough for our requirements. It *is* fast enough however, to maintain the peak voltage level attained by the trigger pulse even after the pulse has been terminated (see composite signal at R7, Fig. 3-10). This peak-voltage-level period (gate-pulse duration) is made adjustable by providing a variable capacitor (C1). The adjustment of C1 then, determines the ramp duration of the sweep generator.

The gate extender operates as a monostable, or one-shot multivibrator, regenerating the input signal to prescribed gate pulses.

DC coupling the sawtooth waveform through some device to the horizontal deflection plates eliminates the need for DC restoration, a practice that also reduces the load capacitance that must be driven by the sweep generator. Without DC restoration, the need for ramp-start-reference assurance again becomes apparent. (See Chapter 1.)

Fig. 3-11 is a block diagram of the bootstrap
generator and its associated control circuits.
Particular emphasis has been placed on the block
labeled Clamp-Tube Plate Regulator, which will be
referred to simply as the plate regulator. The plate
regulator provides a method of feeding back a
specific occurrence from the output of the generator
point (A), to the generator input, point (B). The
feedback signal establishes a specific DC level
(ramp-start reference), to which the sweep generator
will recover after termination of the sawtooth
leading edge. The term feedback "signal" is
deliberately employed to indicate that the plate
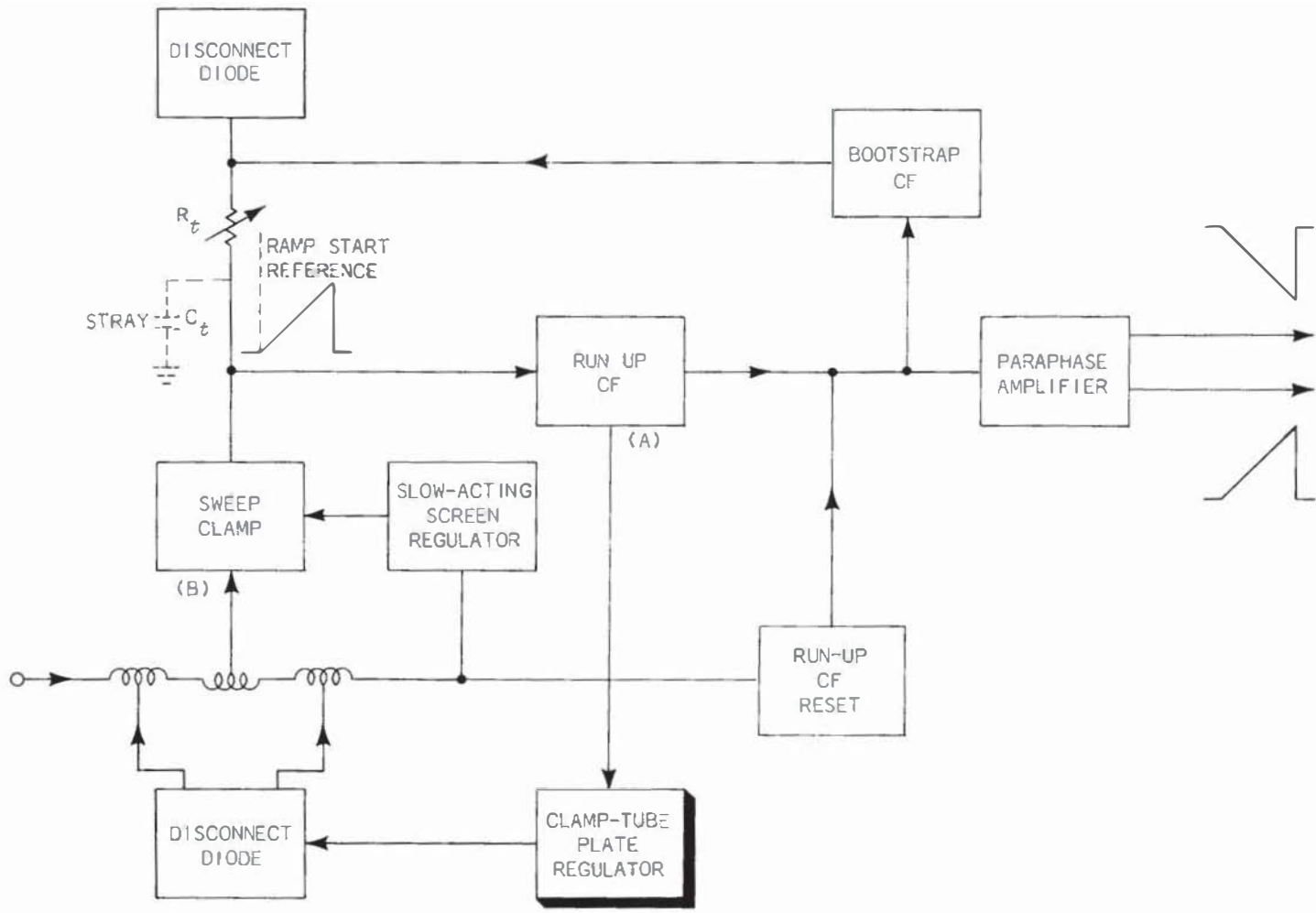regulator responds to fast voltage changes.
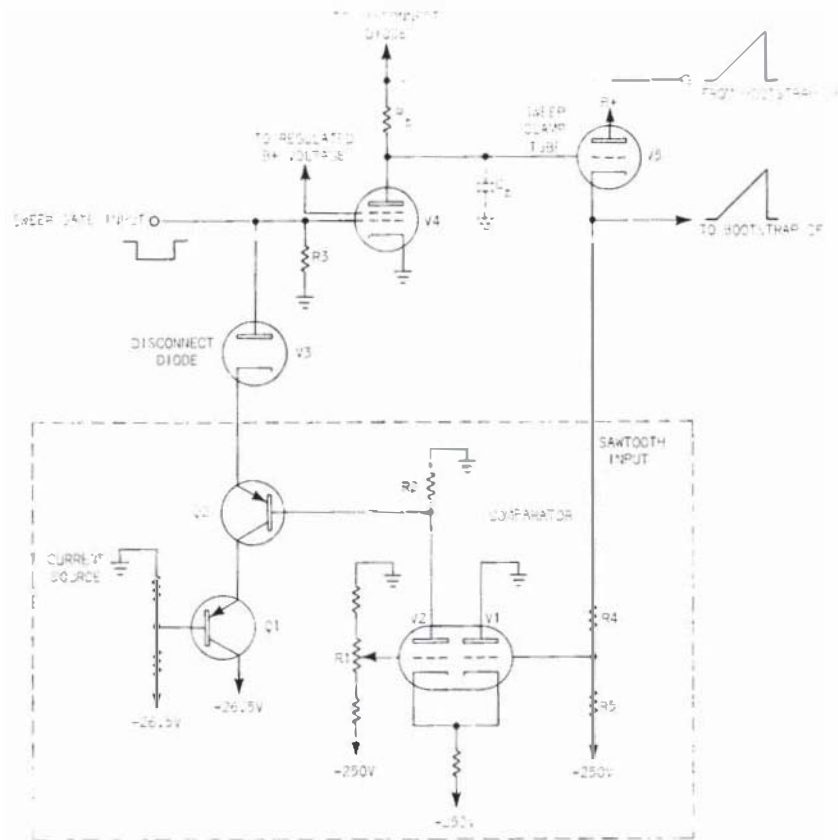
Fig. 3-11. Bootstrap-sweep-generator block diagram.

Fig. 3-12. Clamp-tube-plate regulator.

The plate regulator consists of a comparator
amplifier and a controlled-current source. Fig. 3-12
is a simplified diagram of the plate regulator
circuit. V1 and V2 make up the comparator amplifier.
Q1 and Q2 provide current to the disconnect diode V3.

For this discussion, we will establish the desired
ramp-start-reference voltage (plate voltage of clamp
tube V4) at 150 volts.

plate
regulator
operation

Assume for now that the plate voltage of V4 is
something other than 150 volts. Potentiometer R1,
located in the comparator circuit, is the means by
which the plate voltage of V4 is adjusted to a
desired value. For example, if the plate voltage
at V4 is *higher* than 150 volts, R1 must be adjusted
to provide a more negative voltage to the control
grid of V2. The common cathode bus of the
comparator will follow the negative grid voltage
change, reducing the plate current in V2 and
increasing plate current in V1. The comparator
becomes unbalanced for the moment. The reduced
plate current through R2 reduces the bias current
delivered to Q2, resulting in less current drive
to V3 and R3. Reducing the current through R3
reduces the negative-grid voltage developed across
R3, increasing plate current in V4. Increasing
the V4 plate current *lowers* the plate voltage. The
voltage change developed at the plate of V4 is
coupled through the cathode follower V5, and applied
through the divider R4 and R5, to the control grid
of V1. The voltage at the grid of V1 will fall
until it approaches the voltage at the grid of V2.
When the voltage at the grid of V1 becomes equal to
that of V2, the comparator achieves a state of
balance and all activity within the feedback loop
ceases. A *decrease* in V4 plate voltage is
compensated by the same mechanism.

Upon receiving a sufficiently large negative gate
pulse, the sweep generator begins to generate the
ramp voltage. The plate regulator is disconnected
by reverse biasing the disconnect diode V3. The
ramp portion of the sawtooth will exist for the
duration of the negative gate pulse. During the
ramp period, (or sweep display period), the action
of the plate regulator holds V3 in cut-off.

Upon termination of the ramp, the sweep generator
enters into the recovery period. The sawtooth
trailing edge is fed back to the grid of V1. The
V1 grid is carried negative until it reaches a
predetermined voltage level (set by the adjustment
of R1) at which time V3 conducts, clamping the
control grid of V4 to that bias voltage that will
establish the V4 plate voltage at 150 volts, the
level from which the ramp started initially.

4

THE MILLER INTEGRATOR SWEEP GENERATOR

Miller
effect

The sweep generators discussed in this chapter differ from those of the preceding chapters, in that their operation is based on a phenomenon called "Miller Effect."*

Briefly, Miller effect is an *effective* increase in input capacitance brought about by feedback through plate-to-grid interelectrode capacitance in a plate-loaded vacuum-tube amplifier. The "Miller capacitance" in a given amplifier can be calculated from the equation, $C_{Miller} = (A + 1)C_{pg}$, where A represents the stage gain (without feedback), and C_{pg} represents the static interelectrode capacitance between plate and grid.

The Miller effect was first noted as a phenomenon of vacuum-tube operation, but the concept can be employed in the analysis of other circuits utilizing multi-terminal active devices.

The following circuit discussion will serve to illustrate how Miller effect influences the input characteristics of an amplifier.

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*J. M. Miller: "Dependence of the Input Impedance of a Three-Electrode Vacuum Tube upon the Load in the Plate Circuit." Sci. Papers of the Bureau of Standards, C1919, 15; page 367.

In Fig. 4-1, the plate-loaded vacuum-tube amplifier V1 is in its quiescent state. C_{pg} is charged to the plate voltage of V1. When a voltage source, with internal impedance R_{in}, is connected to the input of V1, at the first instant, the entire voltage is impressed across R_{in} due to the inability of the input capacitance to change its charge in zero time. After some period the capacitors $C_{gk} + C_{gp}$ begin to charge through R_{in}. If we assume that V1 has a voltage gain of 10, $C_{gp} = 5$ pF, $C_{gk} = 55$ pF, and $R_{in} = 1$ K, the generator would "see" $C_{gk} + C_{Miller}$, or an *effective capacitance*, C_{eff}, being equal to $C_{gp} (A + 1) + C_{gk}$ or 110 pF. If the generator voltage is to be developed across the grid-to-cathode circuit, some time must elapse while C_{eff} charges.

effective
capacitance

We will assume the fall period of the generator to be much shorter than the RC time constant of the input circuit. The period required for the grid to reach the generator voltage will be determined by $5(R_{in} \cdot C_{eff})$ or $5(1$ K $\cdot$ 110 pF$) = .55$ μs. If we increase the gain of the stage by a factor of 10, we find that C_{eff} seen by the generator is $C_{eff} = (100 + 1)5$ pF $+ 55$ pF, or a total of 560 pF. The period required for the grid to reach the generator voltage will now be $5(1$ K $\cdot$ 560 pF$) = 2.8$ μs.

We can conclude from the previous examples, that Miller effect was more pronounced when amplifier gain was increased. By increasing the stage gain ten times, the voltage developed at the grid (during the same period) was less (approximately 63%) than that developed in the first case. Therefore the Miller effect can also be observed as providing negative feedback that tends to oppose any instantaneous change in the input circuit.

A. T. Blumlein obtained a British patent resulting from his development of an integrator circuit, based upon Miller's discovery. The circuit introduced a large degree of negative feedback from the plate to control grid of a pentode vacuum tube through a *deliberately added* external capacitance. Sec Fig. 4-2.

In the discussion to follow, we will see how Miller effect is utilized in the development of extremely accurate integrating circuits.
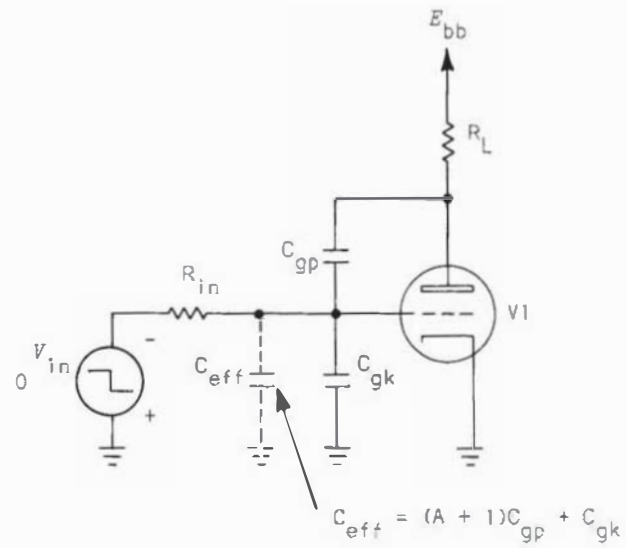
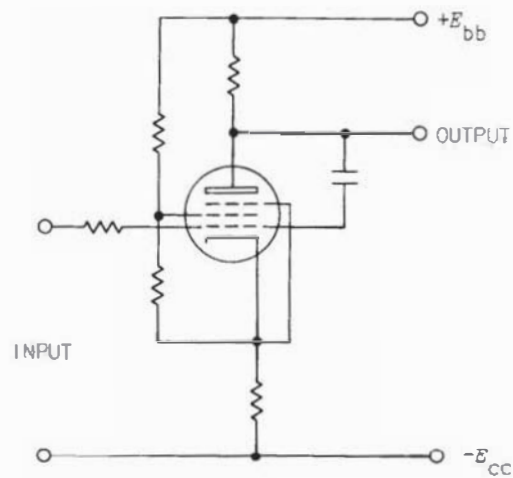Fig. 4-1.   Effective input capacitance due to "Miller effect."

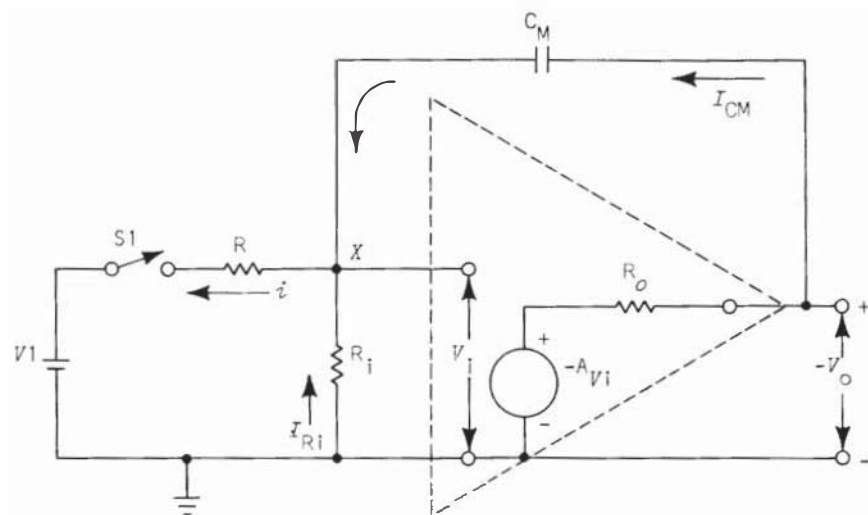Fig. 4-2.   Blumlein's Miller-capacitance integrator.

Fig. 4-3.

THE OPERATIONAL INTEGRATOR

actual
ground

It is common practice, when designing and building electronic circuits, to establish a common point of reference, to which all other points within those circuits are referred. The reference point is usually connected to a metallic chassis on which the circuitry has been constructed. The chassis then, is referred to as *ground*, and customarily all voltages within the circuitry are referred to that chassis or *actual ground*.

virtual
ground

In the discussion of the operational integrator, the term *virtual ground* will be used. For our discussion we will define virtual ground as; that point, driven by two signals, being equal in amplitude and frequency, but exactly opposite in phase. This virtual ground may be referred to as a null point.

Recalling the apparent multiplication of input capacitance (Miller effect), we can also conclude that the input impedance *decreases* with stage gain. At first glance, this phenomena would appear to be objectionable. It would appear that the effect of reducing the input impedance would reduce the stage gain, and so it does. This loss of gain may or may not be objectionable, depending on the purpose of the circuit.

Let's examine the circuit in Fig. 4-3. R_i represents the circuit input impedance. R represents the internal impedance of the source ($V1$). The base amplifier (triangle) is described by its Thevenin's equivalent: R_o = the amplifier output impedance; $-A_{Vi}$ = open-circuit voltage gain.

Assume that Sl is closed at T_1. (See Fig. 4-4B.) Current through R_i develops a voltage across R_i. A voltage change will occur at the output terminals, $(-A \cdot V_i)$. [The (-) sign before A indicates polarity inversion.] The voltage $(-V_o)$ will cause a current through capacitor C_M. The capacitor current (I_{CM}) will flow through R in a direction that cancels I_{Ri}. The I_{CM} current *opposes* the voltage change initiated by the current I_{Ri}. This feedback action causes an apparent reduction in the input impedance of the circuit. That is, with sufficient gain, the base amplifier will provide enough current to cancel I_{Ri}. In that case V_i will be zero. Point X therefore, appears as a very small impedance to any signal voltage, behaving as a *virtual ground* or a null point.

The amplifier, using negative feedback, functions as a self-balancing bridge, providing through the feedback element, whatever current is necessary to maintain point X at a null. The amplitude of the output signal is a function of this current and the impedance of the feedback element C_M.

In Fig. 4-4A the circuit in Fig. 4-3 has been replaced by its Thevenin's equivalent, and,

$$V' = \frac{V_1 \cdot R_i}{R_i + R} \qquad \text{and} \qquad R' = \frac{R_i \cdot R}{R_i + R}$$

To continue the explanation of the operational integrator, we will assume the following (See Fig. 4-4B.):

$C_M = 1 \ \mu F.$

$R' = 1 \ M.$

$V' = 1$ volt for a period of one second.

When switch Sl closes, the base amplifier will seek to deliver a balancing current through C_M to produce a null at point X. To provide a *steady*, or *constant current* of 1 μA through 1 μF, the base amplifier must provide a *continually changing* output voltage of 1 volt per second. The output voltage will continue to change at this rate until the input is changed, or until the base amplifier saturates.

The mathematical expression for the output voltage level reached during the period $T_2 - T_1$ is:

$$E_{out} = \frac{-1}{R' \, C_M} \int_{T_1}^{T_2} e_{in} \, dt$$

It is important to remember that the output of the integrator, at the end of some time interval, is proportional to the *net number* of *volt seconds* applied to the input.
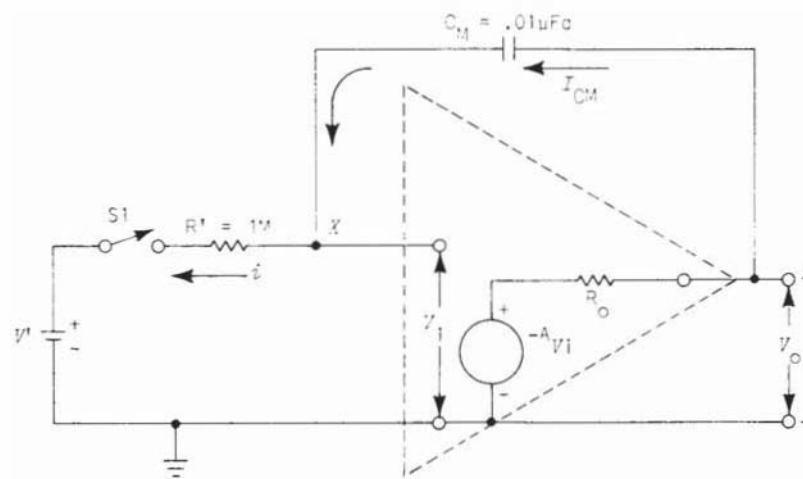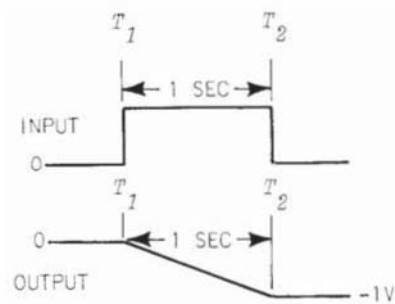

Fig. 4-4A.   An operational integrator.

Fig. 4-4B.

If we assume the magnitude of the generator voltage
V' to be equal to the voltage across the capacitor
C_M, the voltage at the input (point X) of the
amplifier must be zero. Ideally, *all of the current*
flow in R' will flow through capacitor C_M. The
ramp voltage *rate-of-change* appearing at the output
will be determined by the values of R', C_M and the
input voltage (V').

$$\text{Sweep speed} = \frac{V'}{R' \, C_M}$$

Also, in order to obtain an output equal to a true
integral of the input, the gain (A) of the amplifier
should be infinite.

In practice, it is not possible to build an infinite-
gain amplifier. We can however, achieve open-loop
gains on the order of many thousands. The operational
integrator can achieve a very high degree of ramp
linearity, because of its ability to deliver
reasonably constant current to a capacitor through
the provision of a large degree of negative feedback.

The effect of the finite gain on the operational
integrator is an *integrating error*, caused by an
error voltage (V_i) which causes a voltage to be
developed in the finite output impedance of the
amplifier. This small output voltage (E_{Ro}) is
proportional to the input signal itself, rather than
to its integral. However, this undesirable effect
is insignificant if the loop gain is sufficiently
high. The effect of integration error can be
assessed by considering the time constant of
integration. C looks like (1 + A)C at the input
terminal, so the time constant (τ) = (1 + A)RC.
Integration is good only when integration time is
short compared to this time constant.

When the Miller effect principle has been
successfully applied to the development of highly
linear sweep generators, the following
characteristics are evident:

    1. A high degree of negative feedback of
       the type just described. This feedback
       keeps the grid potential (in vacuum-
       tube terminology) within the grid base
       of the vacuum tube, in addition to
       keeping grid voltage changes small with

respect to plate voltage changes. The higher the *open-loop-gain* (gain without feedback) of the circuit, the smaller the grid voltage changes during the run-up or run-down period of the sweep generator.

2. A high constant plate current in the base amplifier resulting in extreme ramp linearity.

3. A high degree of amplification with very low output impedance in the complete system.

One consideration which influences linearity is input impedance. For best linearity, it is essential that the input impedance of the base amplifier be very *high*. Therefore, in the following discussion, we will use the vacuum tube as the base amplifier of the Miller integrator. Later, we will examine the input impedance characteristics of the transistor Miller integrator.

Miller integrator sweep generator

The block diagram in Fig. 4-5 represents a basic sweep generator system, incorporating a Miller integrator and its control circuits. The following discussion will deal primarily with the Miller integrator and disconnect-diode blocks, which include the base amplifier, the output cathode follower, and the disconnect-diode circuits.
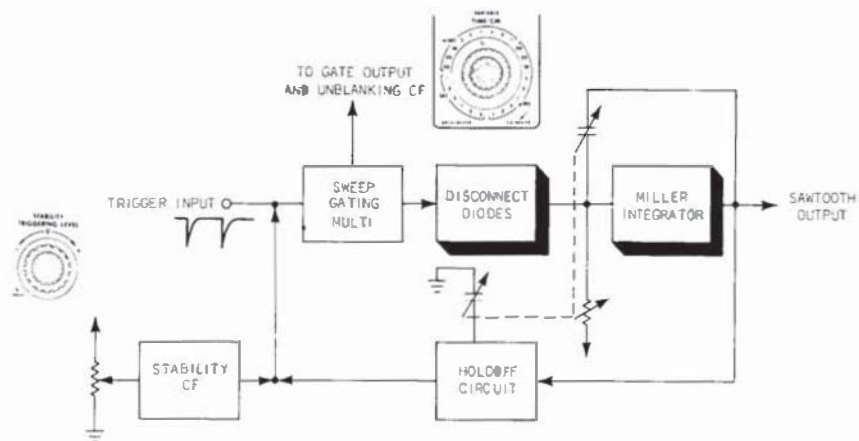

Fig. 4-5. Miller integrator sweep generator block diagram.

Fig. 4-6. Practical Miller-runup sweep generator
described in the text.

A schematic diagram of a practical Miller integrator sweep generator is shown in Fig. 4-6A. The ladder diagram of Fig. 4-6A illustrates events occurring at different points within the circuit during given time periods.

The basic sweep generator consists of base amplifier V1, resistors RL and RT, and capacitor C_T. Diodes V2 and V3 function as *disconnect diodes,* or voltage driven switches, and will be discussed later in the text. V4 isolates the plate circuit of V1 from whatever load may be placed upon the generator, in addition to providing a low-impedance charging path for I_{Ct}.

Prior to sweep start, the control grid of the base amplifier is clamped to a -4 V source through the conducting diode V2. The voltage at the grid of V1 establishes a plate voltage of approximately 40 volts DC. The voltage at the control grid of V4 is approximately -15 V, set by the divider action of B1 and R2 connected between the -150 V supply and the plate of V1. The cathode of V4 is held at -4.5 V when V3 is conducting. At the time the grid of V4 moves to -15 volts, the cathode follows this change. However, because the cathode of V4 is connected directly to the cathode of V3, when the V3 cathode reaches some voltage (-4.5 V), V3 will come into conduction, clamping the cathode of V4. This level is the *ramp-start reference.* V3 is conducting most of the current through R_K, therefore the upper plate of the timing capacitor (C_T) is disconnected from the plate of the base amplifier. The timing capacitor has virtually no charge at this time.

The command for ramp start is initiated by changing the condition of the V1 control grid. Since the current (I_{Rt}) was flowing from the -150 volt supply, through R_t and the conducting diode V2 to a -4 V source, none of the (I_{Rt}) current flows into the timing capacitor. To begin generation of the ramp, I_{Rt} must be diverted to the timing capacitor. This current diversion is accomplished by reverse-biasing diode V2, or gating the diode *off*. A gating multivibrator usually provides this gating signal. (The gating multi will be discussed in the control circuit description.)

When the gating source changes from -4 V to -8 V,
diode V2 disconnects (recall the description
"disconnect diode") the gating source from the
control grid of V1. The V1 control grid immediately
seeks the -150 volt supply level. The moment the
grid of V1 moves negative, the plate voltage moves
positive "gain" times as much. As the plate of V1
starts to rise, the control grid of V4 is carried
with it. V4, in its conducting state, will carry
all of the current through R_K. This *is* the charging
current for the timing capacitor during ramp run-up
time. Diode V3 is cut-off when its plate is driven
negative and its cathode is driven positive by the
plate voltage of V1. The positive-going voltage at
the top end of the timing capacitor forces a charging
current through the timing resistor. This current
opposes the initial tendency of the V1 control grid
to seek the -150 supply level. As was stated
previously, if the voltage at the V1 grid were held
constant for the period $R_t \cdot C_t$, the output voltage
change would be 150 volts. Example: R_t = 1 M;
C_t = .01 μF; E_{in} = -150 volts.

$$\Delta E_{out} = \frac{E_{in}}{R_t C_t} = \frac{150}{.01 \text{ s}} \text{ or } 15 \text{ volts/ms}$$

Notice that for a period of 10 ms, the current I_{Rt} =
150 μA. This is the charging current for the timing
capacitor.

Recall that the charging current will remain constant
only if the base amplifier has an *infinite gain* so
that a *virtual ground* is maintained. These ideal
conditions cannot be achieved in practice; therefore,
we must recognize that the output circuit will have
a definite exponential characteristic. The total
applied voltage as seen by the timing capacitor
will be: (A)E_{in}.

Referring to Fig. 4-6B(d), the voltage at the grid
of V1 is not -2 V, but approximately -2.8 V after
period $T_2 - T_1$. The plate voltage change of V1 is
approximately 150 volts at time T_2. The gain,
therefore, is equal to: $\frac{E_{out}}{E_{in}}$ or $\frac{150}{.8}$ = 175. The
total applied voltage, as seen by the timing
capacitor, would be: V = (150)(175) = 26,250 volts.
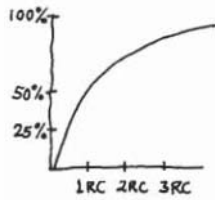
slope error

Referring to the universal time constant chart in the margin, it is evident that the sweep will be extremely linear. Less than 1% of the total exponential will be used for the sweep function voltage. The slope error will be approximately:

$$e_s = \frac{V_s}{V_{in}(A)} = \frac{150}{26,250} = .57\%$$

Where: V_s = output voltage.

deviation

The percentage of deviation from an ideal ramp will be approximately: $e_d = 1/8 \; e_s \approx .07\%$.

Other factors that may contribute to slope error and displacement error within the sweep generator are: tolerances of R_t and C_t, degree of grid current in V1, and leakage current drawn by V2 when in the nonconducting state.

In Fig. 4-6A, the path for the charging current during ramp-time extends from the -150 volt supply, through the timing resistor R_t, to the lower plate of C_t, from the upper plate of C_t and through the conducting tube V4.

retrace

Upon completion of the ramp run-up period, the gating source reverts to its previous state [time T_2, Fig. 4-6B(a)]. The gating source will raise the plate of V2 sufficiently to cause V2 to conduct. [See point X in Fig. 4-6B(a).] At time T_2, V2 comes into conduction, diverting current from C_t to the gating source. This current through the timing resistor, V2, and the gating source raises the voltage at the control grid of V1 toward its initial level. This voltage change occupies a finite period. This period is determined by the time constant C_t, the forward resistance of V2, the gate generator source impedance, and R_K. The period $T_3 - T_2$ then, is the generator *recovery* time, or the *retrace* time of the sweep.

Meanwhile, at the instant the V1 grid moves positive, the plate voltage falls rapidly. During the fall period the timing capacitor is discharging. The control grid of V4 follows the negative plate voltage change at V1. The cathode will follow this grid voltage change, but only to a point [point Y in

Fig. 4-6B(a)]. Notice that at instant T_3, in
Fig. 4-6B(c), the voltage at the cathode of V4 is at
the *ramp-start reference* level. When that level is
reached, V3 comes into conduction [point Z,
Fig. 4-6B(a)]. The voltage level at the plates of
V2 and V3 is established, for the period T_4 - T_3,
by the current through the two diodes into the
gating source. The voltage change at the V1 grid
ceases; therefore, the V1 plate comes to rest at
the original quiescent level.

Since the timing capacitor was discharging during
the generator run-down period, at time T_3 (Fig. 4-6B)
C_t has completely discharged. The circuit is once
again in its quiescent state, and will remain so
until another gate signal is applied.

The neon lamp, B1, provides DC coupling of the
sawtooth waveform to the control grid of V4 without
attenuating the signal. Resistor R1 damps out any
oscillations that B1 might tend to generate as a
neon-tube relaxation oscillator. Capacitor C1 acts
as a "speed-up" cap, bypassing rapidly changing
signal components around B1 and R1 to the V4 grid.
B2 acts as a protective device for V4. If, for
example, the V4 grid-to-cathode potential should
exceed the ignition potential of the neon, (this
could occur prior to the V1 and V4 filaments
reaching proper operating temperature), the neon
would ignite, maintaining a safe 50-volt potential
between the grid and cathode of V4. The neon will
extinguish after the circuit stabilizes. Capacitor
C2 provides bootstrapping, i.e., positive feedback
to the upper plate of the timing capacitor at fast
(10 to 1 µs) sweep speeds.

THE TRANSISTOR MILLER INTEGRATOR

forward-
biased
junction

When transistors are used in a Miller integrator,
their relatively *low* input resistance becomes
evident when we consider that the base-to-emitter
junction of a transistor is regarded as a
semiconductor diode. (See Fig. 4-7.) As we know,
when a forward voltage is applied to the base-to-
emitter junction, the junction will exhibit a
characteristic similar to that shown in Fig. 4-8.

Line (1) on the graph (Fig. 4-8) is a plot of base-
to-emitter voltage versus base current; line (2) is

a plot of base-to-emitter voltage versus emitter
current. Any point along these two lines may
represent some value of resistance (within the
dynamic operating range).

The resistance defined at any point along line (1)
is approximately $\beta + 1$ times that shown at line (2)
at any common point of intersection (point A). It
follows then, if by some means we can find the
resistance at one point on a line, and if we know
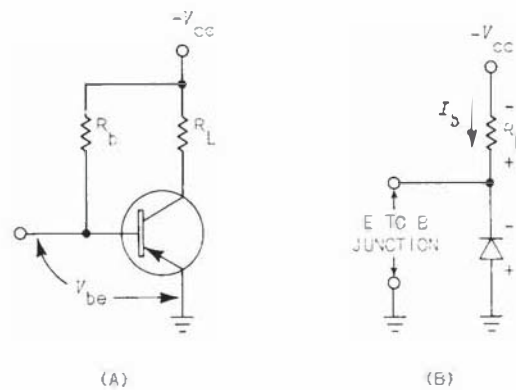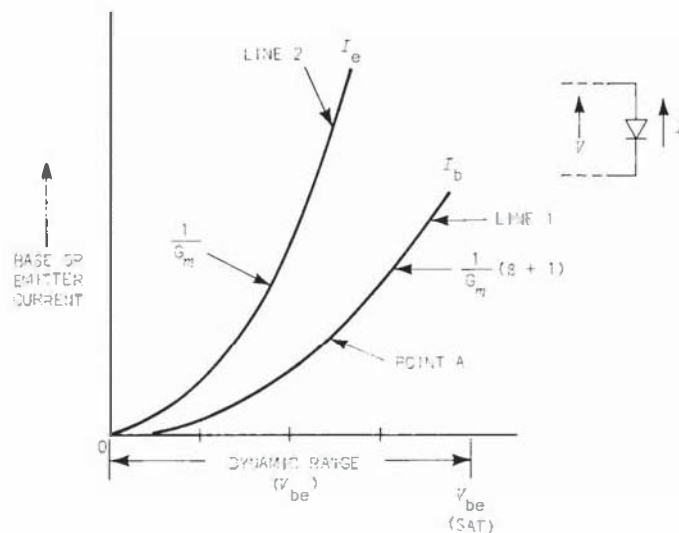Beta, we can find a corresponding value of resistance
on the other line.

Fig. 4-7.

Fig. 4-8. Forward biased junction.

Fortunately, there is a "rule of thumb" equation useful in determining the resistance along either line. $\dfrac{0.026 \text{ V}}{I_e}$ defines any resistance along line (2) and $\dfrac{0.026 \text{ V}}{I_b}$ defines any resistance along line (1), where I_e or I_b is the DC base or emitter current at the point selected. The value $0.026/I_e$ is justified by the basic physics of the device and will not be discussed here.

Now let's consider what this has to do with transistor circuits. Note that any portion of line (2) may be used to find $\dfrac{\Delta V_{be}}{\Delta I_e}$. Assuming that the transistor has a very high Beta, a change in input current (ΔI_e) will very nearly equal a change in collector current (ΔI_c). In vacuum-tube theory, the expression, $\dfrac{E_{in}}{I_{out}}$ is called $\dfrac{1}{G_m}$. So let's call the resistance calculated from line (2) by the same name $\dfrac{1}{G_m}$.

All we've said so far, is that the impedance looking into the emitter of a transistor is approximately equal to $1/G_m$ of the device, and can be calculated by:

$$\frac{1}{G_m} = \frac{26 \text{ mV}}{\text{DC value of emitter current (mA)}}$$

In series with $\dfrac{1}{G_m}$ is a small resistance (R_{eb}) that is made up of the ohmic resistance of the leads and the semiconductor material. R_{eb} usually amounts to about 2 Ω to 5 Ω. Therefore, the expression,

$R_{eb} + \dfrac{1}{G_m}$ is an operating characteristic of the

trans-
resistance

device which we shall call "transresistance." The notation for this expression is r_{tr}, therefore,

$$r_{tr} = R_{eb} + \frac{26 \text{ mV}}{I_e \text{(mA)}}.$$ To find the input resistance

(R_{in}) to the device in Fig. 4-7(A) we can use the equation:

$$R_{in} = r_{tr} \, (\beta + 1)$$

If we assume a Beta of 100, and an r_{tr} of 8 Ω, the
input resistance of the transistor in Fig. 4-7(A)
is equal to 808 Ω. When we compare this to a vacuum
tube with a very high input impedance, it is apparent

Miller
transistor
input
resistance

that if the transistor is to be used as the base
amplifier for the Miller integrator, methods must be
found to overcome the low input resistance
characteristic of the transistor. Recall that one
of the prerequisites for good integration was the
high input impedance amplifier.

The low input resistance of the transistor in
Fig. 4-9 provides a current path for base current
that will flow in the timing resistor and not in the
timing capacitor. Recall that for good integration,
the current in the timing resistor must equal that
in the timing capacitor during the ramp period. The
presence of base current in the timing resistor
would contribute to a slope error at the sweep
generator output.

One method of increasing the integrator's input
resistance would be to use a high Beta transistor
as the base amplifier. Recalling the equation:

$$R_{in} = r_{tr} (\beta + 1)$$
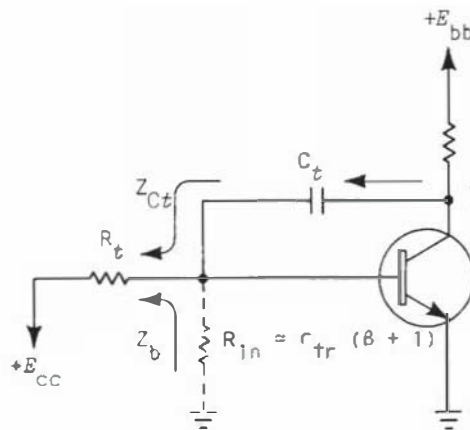
Fig. 4-9.
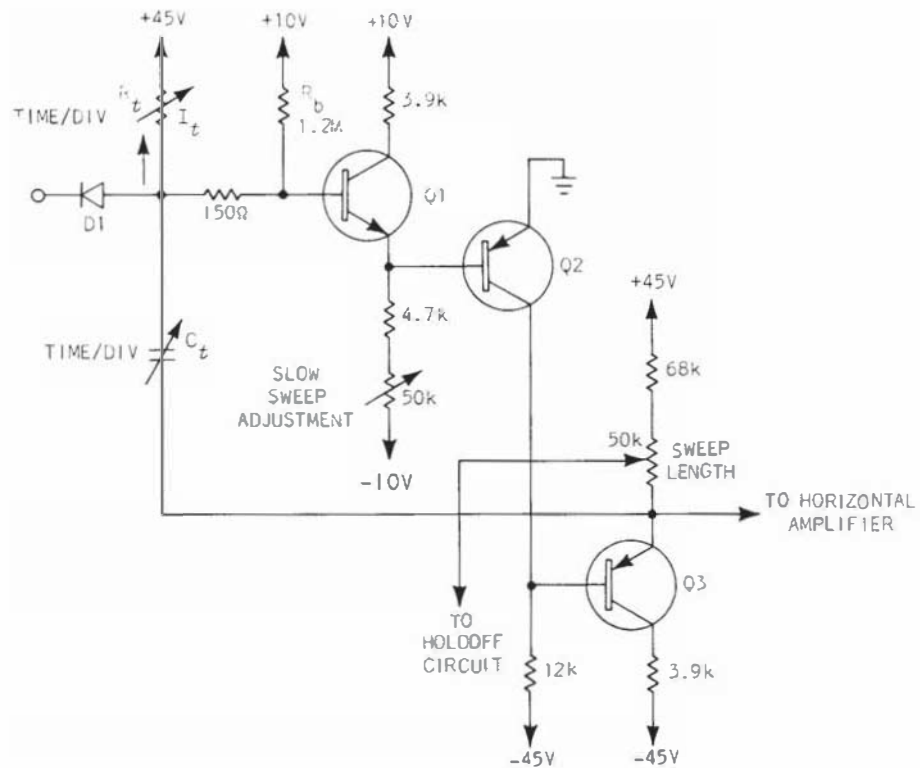
Fig. 4-10. Transistor Miller integrator.

Another method of improvement isolates the "null-
seeking" point from the base amplifier input (See
Fig. 4-10). This configuration increases the
generator input impedance *considerably*, since the
input impedance of the generator now becomes the
input impedance of Q2, times the Beta of Q1 plus one.

isolating
timing
circuit

The bias resistor R_b, establishes a fixed base current
for both transistors that will not flow in the timing
circuit. The input resistance of the integrator
base-to-emitter thus remains relatively fixed in
spite of input signal current variations.

The best method of isolating the timing circuit from
the base amplifier input is to use a cathode follower
or a source follower. Both of these devices exhibit
a very high input impedance.

HOLD-OFF

In the discussion of "Sawtooth Characteristics" (Chapter 1), the subject of "hold-off" was introduced briefly. We will now discuss the hold-off function and the circuits represented by the hold-off block in Fig. 4-11.

Let us return for a moment to Fig. 4-6B(c). Notice that throughout period $T_4 - T_3$ the sweep generator is in its quiescent state. This period is referred to as *hold-off*. The term is derived from the function of specific circuits incorporated in most modern sweep generators. The hold-off function serves two purposes:

1. To allow the output voltage of the sawtooth generator to fall to its quiescent level (ramp-start reference), corresponding to the CRT electron beam position at the left side of the CRT.

2. To provide sufficient time for any transients that may develop, during ramp or retrace, to disappear.

The hold-off circuits thus provide the necessary *waiting period* or hold-off time, by rendering the gating source insensitive to any trigger signals that may arrive prior to complete sweep recovery.
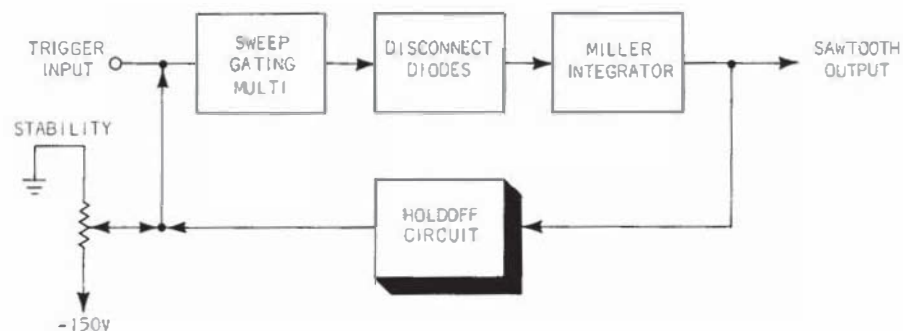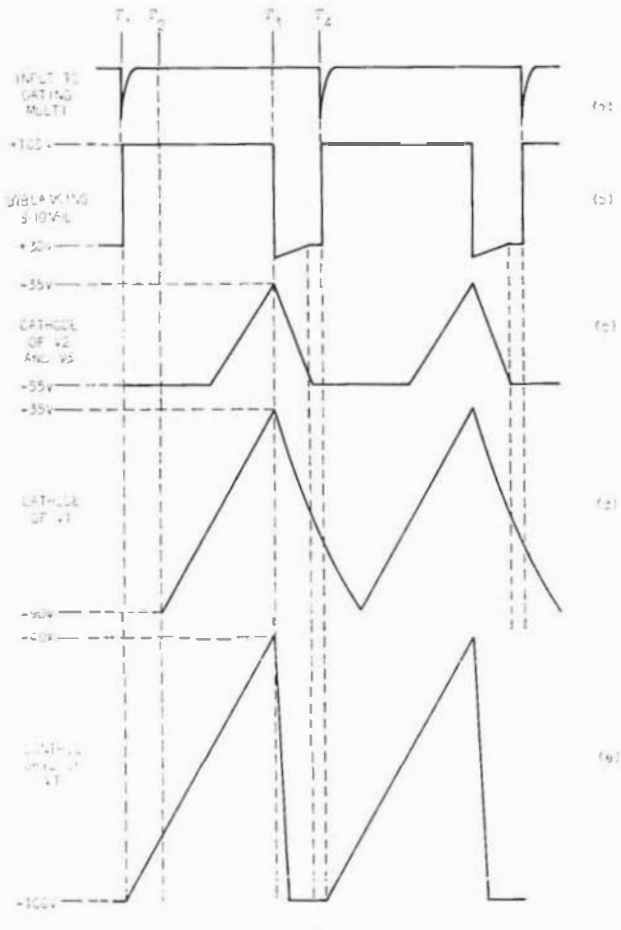

Fig. 4-11. Miller integrator sweep generator.
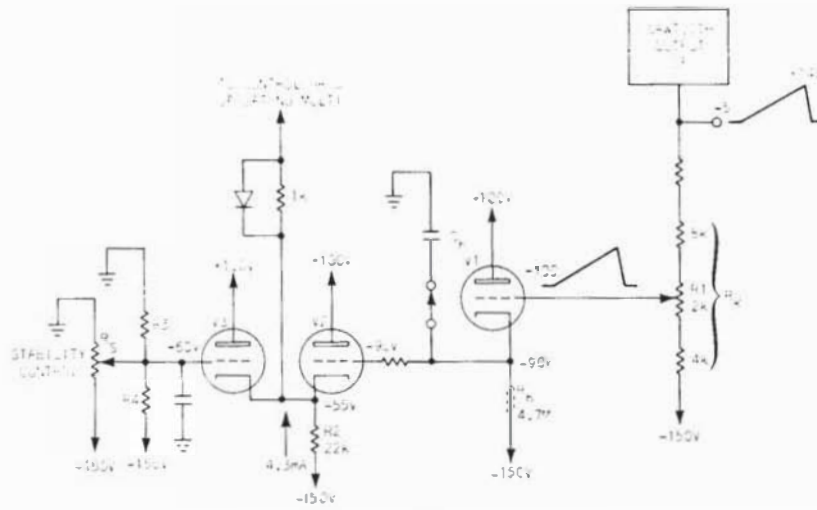
Fig. 4-12. Practical holdoff circuit.

Fig. 4-12A is a schematic diagram of a practical circuit used to provide the hold-off period for the sweep generator. The significant components of the circuits are: hold-off capacitor C_h, cathode resistor R_h, and cathode follower V1. C_h is one of a number of capacitors which are switched into the circuit as the TIME/CM switch is adjusted for different sweep rates. This arrangement insures a maximum sweep duty cycle.

bistable
Schmitt
multivibrator

Before we examine the operation of the hold-off circuit itself, we must investigate certain characteristics of the sweep-gating multivibrator. A good example of this type of circuit is commonly known as the bistable Schmitt multivibrator. The term "bistable," indicates that the multi has two stable states. In one state, which we'll call the "ready" state, the multi generates an output only when provided with a specified input trigger signal. In the other state, the multi is insensitive to the input triggers and must be "reset" to its *ready* state before it can provide another output.

sweep length
adjustment

Referring to Fig. 4-12A, quiescently, a DC voltage of -100 volts is established at the control grid of V1 by the adjustment of the Sweep Length potentiometer R1. (The reason this particular voltage was selected will be explained later in the text.) The cathode of V1 operates at about -90 volts, biasing V1 near cut-off. ($I_{Rh} = 13$ µA.) The direct coupling between the V1 cathode and the control grid of V2 also establishes the latter at about -90 volts. The cathode of V2 rests at approximately -55 volts. Due to the large negative bias on V2, current through R2 flows only through V3.

three-state
multi

Potentiometer R5, labeled STABILITY, indirectly determines the DC level at the input of the sweep-gating multi. Quiescently, the DC level at the cathodes of V2 and V3 determines which of the following three states the gating multi will be in:

1. Not triggerable -- not responsive to trigger signals.

2. Triggerable -- will provide an output upon receipt of a trigger.

3. Freerunning -- but capable of being
   synchronized. (Within certain
   frequency limits. See Fig. 4-13.)

For the purpose of this discussion, assume that the
STABILITY control is set to a DC level that will set
the gating multi in its triggerable, or "ready" state.
(See Fig. 4-14.) When a negative trigger signal,
large enough to "punch through" the "trip" level,
is applied to the gating multi, the multi will change
its state and provide an output that will start ramp
generation. The gating multi, having been switched
to its other stable state, will not respond to
trigger signals during sweep time.

Referring again to Fig. 4-12A, notice that a portion
of the positive-going ramp developed across R_K, is
applied to the control grid of V1. As the grid of
V1 moves positive, the V1 cathode and the V2 control
grid move with it. Hold-off capacitor C_h, previously
charged to -90 volts, discharges through the low
impedance of (now) conducting V1. This low impedance
discharge path allows the voltage on C_h to follow the
rising ramp very closely. As the control grid of V2
moves positive, the V2 bias is reduced. As the grid
of V2 approaches -55 V, the current through R2
switches to V2 and the signal voltage at the V2 grid
will be developed across R2. [See Fig. 4-12B(c).]
As the grid of V2 takes control of the (V2 - V3)
common cathode bus, the bus rises toward the "reset"
level of the gating multi.

When the cathode bus reaches the "reset" level, the
gating multi switches states, terminating the ramp.
Notice in Fig. 4-14 that the multi cannot yet be
triggered, since its grid voltage is high above the
"trip" level during most of the retrace period.

During recovery, a portion of the trailing-edge
voltage is applied to the control grid of V1
(Fig. 4-12A). The cathode of V1 would follow the
grid, at the same rate, were it not for the presence
of C_h and R_h. Remember that during the sawtooth
run-up period, C_h discharged from its quiescent level
of -90 volts to about -35 volts. The rate-of-change
(voltage) across the hold-off capacitor could keep
pace with the voltage change at the V1 grid because
of the low impedance path provided by conducting
tube V1. During the recovery period however, the

rate-of-change at the V1 grid is about 20 times
faster than at run-up time. The grid of V1
therefore, loses control of the cathode. The voltage
at the cathode can then fall only at a rate
determined by C_hR_h. The C_hR_h charge time constant
is *much greater* than the discharge time constant
(C_h times the Z_k of V1). The *time* required for the
hold-off capacitor to charge to the initial -90
volts is the *hold-off period*.

When the voltage at the V1 cathode falls (charging
C_h) to its quiescent level of -90 volts, the control
grid of V1 will regain control of the cathode. The
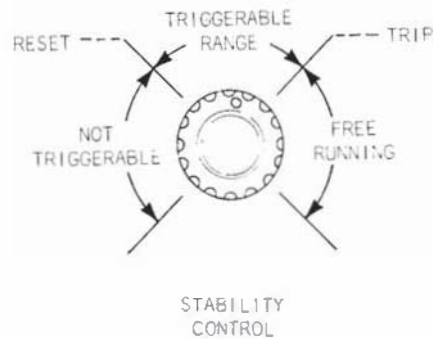
Fig. 4-13. General example of the range of
            stability control coverage.
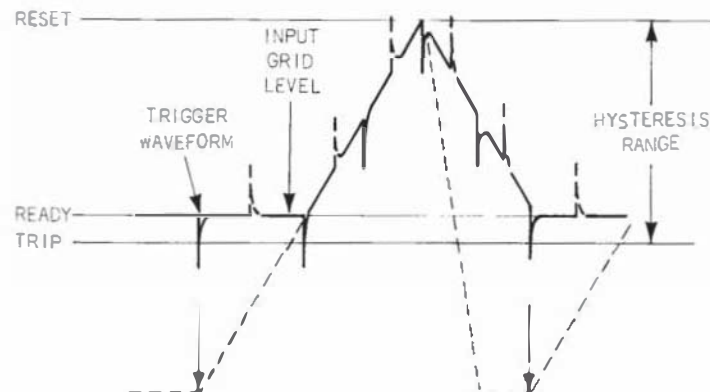
Fig. 4-14. Composite illustration of input
            conditions occurring at sweep gate multi.

common cathode bus of V2 and V3 will fall at the
C_hR_h rate, but only to the "ready" level of the
gating multi. That level, you will recall, was
previously set by the STABILITY control. At this
level, the gating multi is again ready to receive
an incoming trigger signal, which will initiate
another ramp.

sweep
duration

Adjustment of potentiometer R1 (Sweep Length control)
determines the duration of the sweep (sweep length).
We shall refer to the illustration in Fig. 4-15
during the discussion.

The sweep duration is determined by the duration of
the negative gate pulse from the gating
multivibrator. Fig. 4-15 shows the relationship
between the sawtooth feedback voltage, the gate
duration, and the duration of the sweep. Notice the
ramp intersecting point (B) in Fig. 4-15(c). The
amplitude of the waveform is such that the ramp
intersects the "reset" level of the gating multi at
point (B), or after the period, $T_3 - T_1$. This point
of intersection determines the duration of the sweep-
gating pulse, which in turn determines sweep length.
(Notice however, that sweep rate is unaffected by
this adjustment.) In the foregoing example, the CRT
sweep display covers 10 graticule divisions.
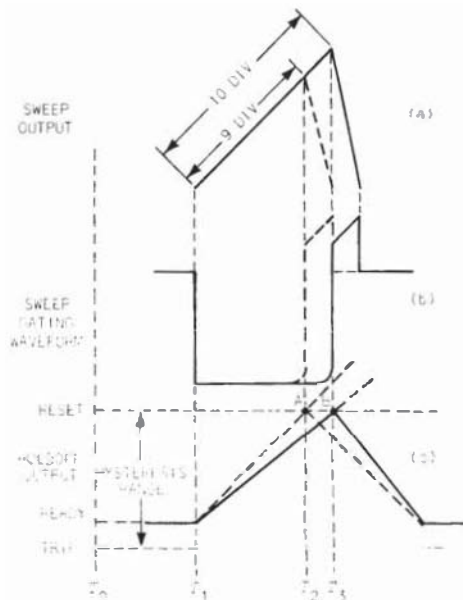
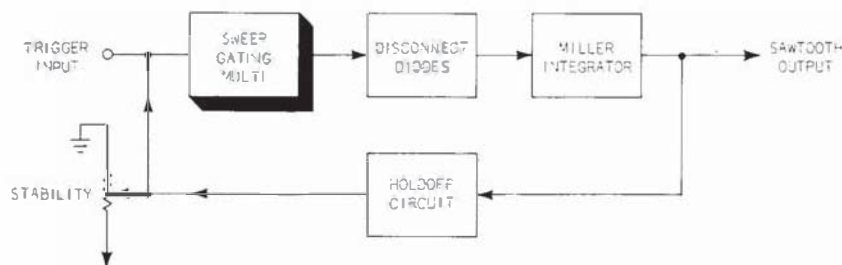Fig. 4-15. Adjusting sweep length.

Fig. 4-16. Miller integrator sweep generator.

If the Sweep Length control were adjusted to provide a larger-amplitude feedback signal, the leading edge of the signal would intersect the "reset" level at point (A). Resetting the multi at point (A) terminates the ramp sooner. Therefore, as indicated, the sweep length will be 9 graticule divisions.

Summing up then, the amplitude of the feedback signal, as determined by the Sweep Length control setting, determines the duration of the gate and consequently, the sweep length.

SWEEP GATING MULTI

The block labeled "Sweep-Gating Multivibrator" (Fig. 4-16), represents a circuit commonly used to control the Miller sweep generator. The circuit functions as an electronic switch, responding only to signals possessing certain predetermined characteristics. The output of the gating multi exists at only one of two possible levels. One level allows the sweep generator to run up (or run down) while the other level holds the generator in a quiescent state (no sweep). The sweep-gating device described below is a cathode-coupled bistable multivibrator, commonly called a "Schmitt trigger," after the inventor of the vacuum-tube version. Some advantages of this type of multi are listed below:

two-level output

1. The input terminal is not affected by the regenerative switching. When the circuit switches states, the voltage at the input terminal is not altered. This feature is desirable if we wish to maneuver the input terminal characteristics.

2. The output load resistance is not
    necessary to the bistable operation.
    This resistance may therefore vary
    over a wide range, depending upon the
    output requirement.

3. Capacitive loading at the output will
    not slow the regenerative action
    (switching time) although such loading
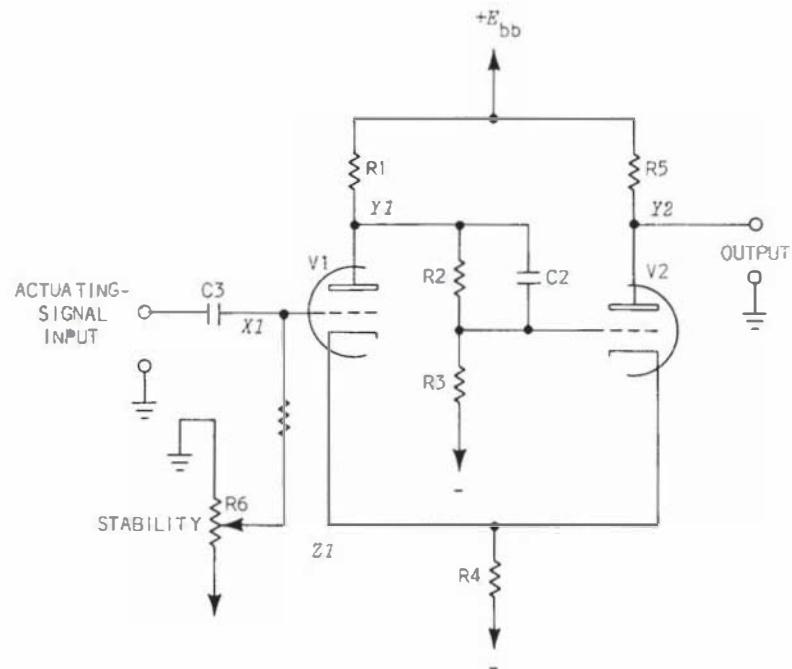   will increase the rise and fall time
    of the output signal.

Fig. 4-17. Simplified sweep gating multivibrator
"Schmitt trigger".

Fig. 4-17 is a simplified diagram of a Schmitt
trigger, designed to provide sweep *gating* to the
Miller-sweep generator. We will refer to this
figure while discussing the basic operation of the
sweep-gating multi.

sweep
gating

Let us assume V1 is conducting plate current. A
voltage developed across R4 appears at Z1. The
voltage at *X1* therefore is near that voltage at Z1.
The multi is in one of its two possible states. V2

is cut off and the output level will be $+E_{bb}$. If we wish to switch the multi to its opposite state, we could increase the negative grid voltage at V1. By adjusting the potentiometer (R6) toward the negative supply, we cause the voltage at $X1$ to move negative. V1 will continue conducting until the V1 grid voltage reaches its cut-off or the "trip" level (described in the hold-off discussion). When V1 cuts off the voltage at $Y1$ increases toward $+E_{bb}$ rapidly. The voltage change at the plate of V1 is: $(\Delta e_{in})(-A)$, indicating amplification in V1. A portion of this positive signal at $Y1$ is applied to the grid of V2 through the divider R2 and R3. The V2 cathode follows the positive change at the grid, rapidly increasing the negative bias on V1. This action is regenerative and results in a very rapid transition of the multi. During this transition, the current to R4 was diverted from V1 to V2. The control grid of V2 will continue to rise until V1 cuts off. During this transition the voltage at $Y2$ changes from $+E_{bb}$ to $E_{bb} - I_2R5$.

To restore the multi to its initial, or "ready" state, we must change the voltage at point $X1$ in a positive direction until the "reset" level is reached. This could be done by adjusting the STABILITY control. At this time, the regenerative action again occurs, and results in a rapid output voltage change from $E_{bb} -I_2R5$ to E_{bb}.

To complete one cycle of multi operation, the grid of V1 must be driven sufficiently negative to start the regenerative action. It must then be driven in the opposite direction to "reset" the multi to its "ready" state. Note (Fig. 4-14) that the voltage difference between the "reset" level and the "trip" level is defined as the "hysteresis" of the multi.* The hysteresis range is determined by the "loop gain" of the multi, which in turn, is primarily determined by the value of R1.

hysteresis

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*For a more detailed discussion of Schmitt trigger "hysteresis," refer to Sect. 10-11, p. 389-394, Millman & Taub, "Pulse, Digital, and Switching Waveforms."
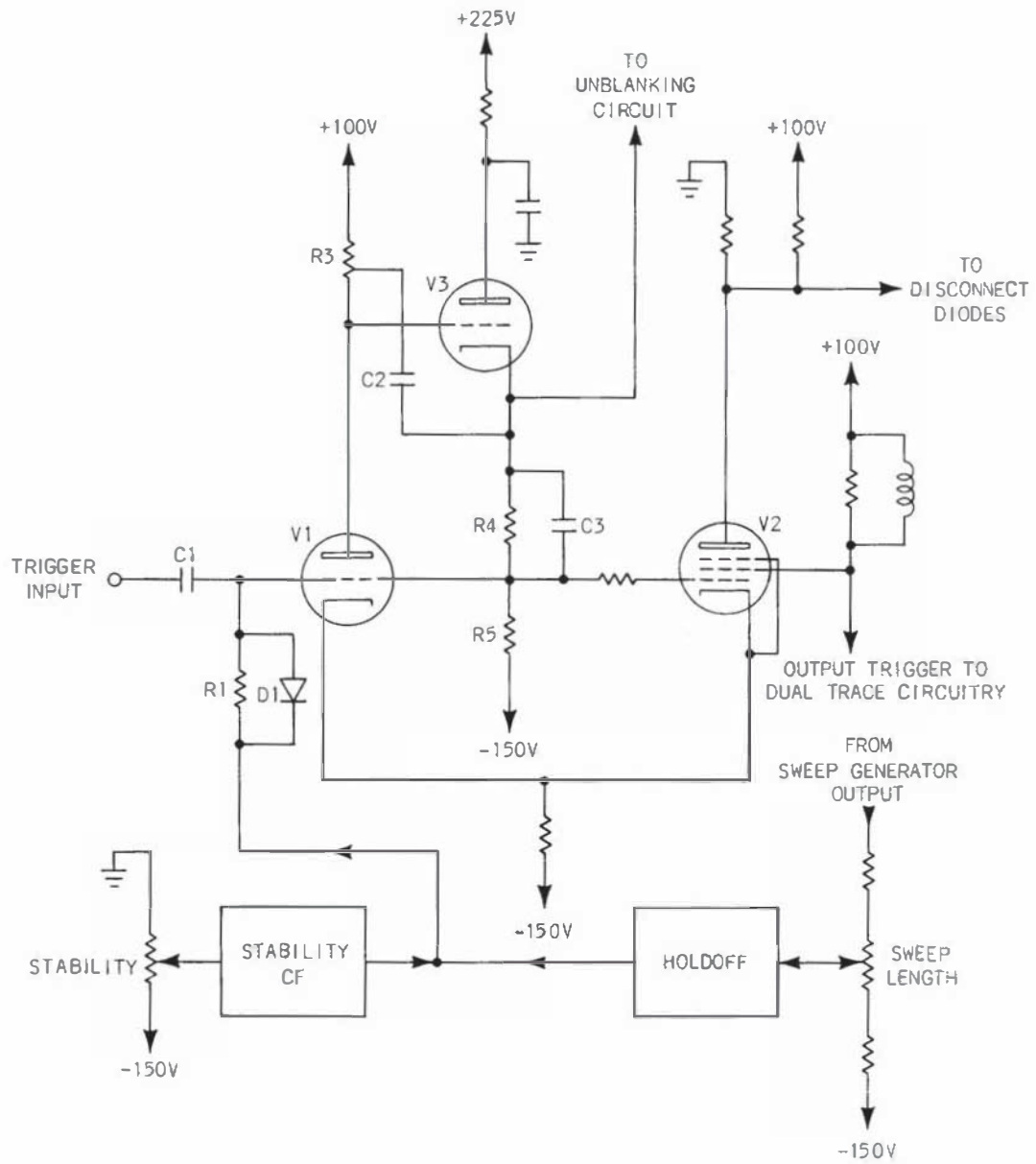
Fig. 4-18.

The vacuum-tube version of the Schmitt trigger requires the hysteresis region to be fairly large, for reasons of reliability. As the tubes age their amplification factor decreases; consequently the loop gain falls off. If the hysteresis region were originally narrow, a small reduction in loop gain would result in unstable sweep triggering, eventually causing the sweep to free-run. On the other hand, loop-gain is limited by transition time requirements. When we recall that an increase in the value of R1, (Fig. 4-17) increases the loop gain, it follows that the risetime of V1 would increase and this would increase the multi-transition time. As we learned earlier, the elapsed time from signal insertion to sweep start *must* be kept to a minimum.

A practical sweep-gating multi is represented by the schematic in Fig. 4-18. Having just described the basic operation, we shall now look at some of the special features that have been provided for improved performance of the device.

cathode-follower V3

Notice the cathode-follower V3. The output of V1 is coupled, via the cathode follower, to the grid of V2. We include the cathode follower to reduce the effect of shunt capacitance loading on the plate of V1. This improves the risetime of V1 considerably, which improves the multi-transition time.

diode D1

Diode D1 is added to the input differentiator to shunt positive-going trigger pulses around R1. Since the gating multi, in its ready state, will respond to negative trigger pulses, the positive pulses are unnecessary, and in fact impose undesirable effects on sweep length. Referring to Fig. 4-14, notice that the positive pulses (dotted lines) would "punch through" the "reset" level before the ramp itself could reach that level. If this occurred, the sweep length would be *shorter* for a triggered sweep than for a free-running sweep. The sweep length would also vary with trigger repetition rate variations. This possibility is eliminated however, since the *positive* differentiated pulses do not see the resistance of R1, but the small resistance of the forward-biased diode D1. Therefore, the negative trigger pulses are the predominate signals appearing at the input to the gate multi.

bootstrap
capacitor C2

Another method of decreasing the gating multi transition period is the addition of a bootstrap capacitor in the plate circuit of V1. This capacitor (C2) connected between the cathode of V3 and a tap of R3, couples the fast-changing voltages, appearing at the V1 plate, to the tap at R3, thereby providing an additional short-term current source that will assure faster transitions at the V1 plate.

frequency
compensation

Capacitor C3, in conjunction with the input (grid-to-cathode) capacitance of V2, frequency-compensates the divider R4 and R5 to further improve multi-transition time.

The Schmitt trigger proved to be very valuable when incorporated in triggered sweep generator systems. In many current applications of the Schmitt trigger, the transistor "flip flop"* has replaced the vacuum-tube binary. The basic operation of this device is much the same as that of the vacuum-tube Schmitt trigger. For a detailed discussion of the transistor bistable multi, see the "Oscilloscope Trigger Circuits" book, one of the series of Tektronix Circuit Concepts.

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*Common term for the bistable multivibrator.

5

SWEEP GENERATOR CONTROL CIRCUITS

In this chapter we will describe some of the circuits incorporated in modern oscilloscopes that provide a means of sweep generator control. The main topics of discussion will be:

1. Hold-off techniques.

2. Sweep gating.

3. Sweep lock-out.

4. Automatic sweep generation.

5. Sweep switching.

hold-off
using the
disconnect
diode

In Chapter 4 we discussed the hold-off method which utilized a cathode follower to drive the hold-off capacitor. Another method is illustrated in Fig. 5-1. In this circuit the hold-off capacitor is driven by a disconnect diode.
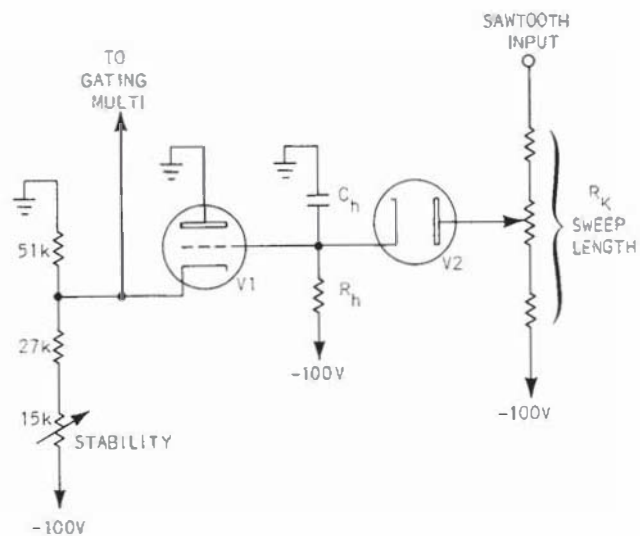
Fig. 5-1. Holdoff (using a disconnect diode).

Quiescently, C_h will charge toward -100 volts until the cathode of V2 reaches the approximate voltage level at the plate. Cathode follower V1 is cut off due to a large negative bias. The setting of the STABILITY control determines the DC level at the input of the gating multi.

When a sweep is initiated, a portion of the sawtooth is fed back to the plate of V2 through the divider R_K. As the plate of V2 rises with the ramp, C_h discharges toward ground. The grid of V1 rises with the V2 cathode until V1 comes into conduction. When this occurs, the positive-going ramp is coupled to the gate multi input, eventually resetting the multi. Resetting the gating multi terminates the ramp which the hold-off causes the generator to recover. The rapidly falling period sawtooth trailing edge, now present at the V2 plate, back-biases V2. The cathode of V1 however, cannot fall at the sawtooth recovery rate, but at a rate determined by the values of R_h and C_h. This slower rate allows the generator to completely recover before the gating multi is set "ready." The hold-off capacitor will continue to charge until the disconnect diode conducts. At this point the circuit is in its quiescent state and the gating multi is ready to receive another trigger pulse.

The disconnect diode hold-off technique is used in slower-sweep-speed generators, usually those whose sweep speeds are no faster than 1 div/µs. If this method were used at faster sweeps, the hold-off circuit would present an increasing capacitive load to the sweep generator. In this event, a follower of some type would be necessary to isolate the hold-off circuit from the generator.

TUNNEL-DIODE GATING MULTIVIBRATOR

The introduction of the tunnel diode led to the development of an almost ideal means of sweep gating. The tunnel diode is a single active device that performs all the functions of the Schmitt trigger, and is capable of transition periods of subnanosecond durations.

If the tunnel diode is to serve as a bistable switch or multi, the load resistance in the circuit must be greater than the tunnel diode's negative resistance ($R_L > -R_d$). Of course, proper biasing and circuit construction are also necessary.
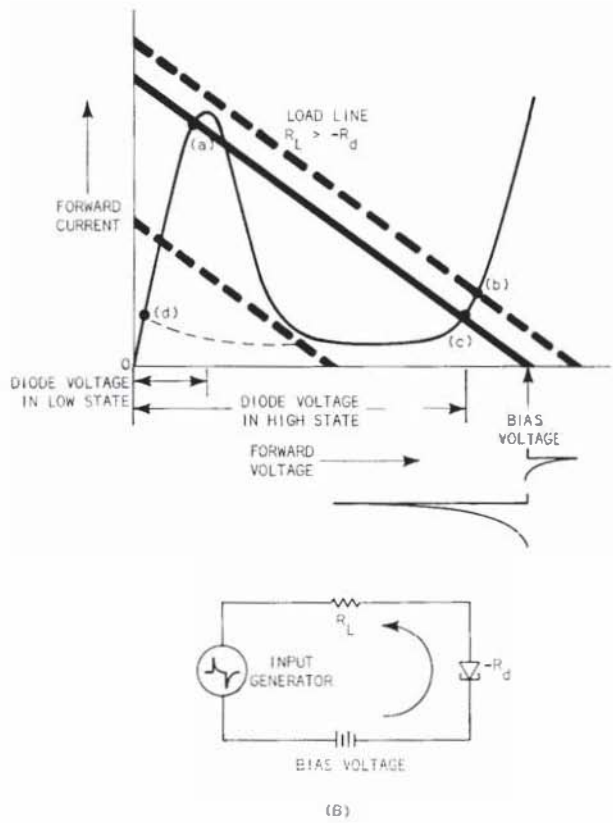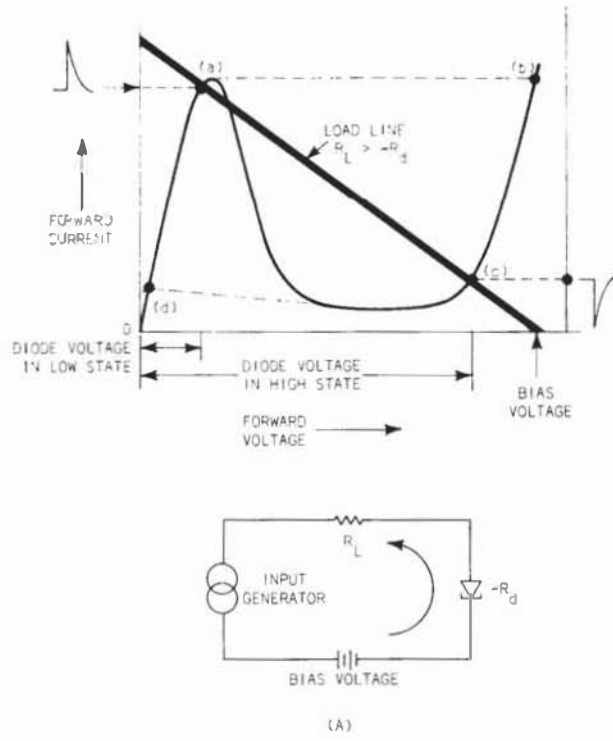
Fig. 5-2. The bistable tunnel-diode switch.

The tunnel diode shown in Fig. 5-2A is biased quiescently at point (a) on its characteristic curve. If the input current is increased above the peak current point of the tunnel diode, the voltage across the TD will switch rapidly to point (b). The transition time from point (a) to (b) may be as fast as .1 to .3 ns. Once the tunnel diode voltage is resting at point (c) (determined by R_L) it will remain there until the input current is reduced below the *valley* current level. When this occurs, the TD voltage will switch rapidly from point (c) to point (d) and come to rest at (a) once again. The circuit is bistable, in that it takes a trigger to switch it from its low (forward bias) state to its high state and must also be triggered back to its low state.

In Fig. 5-2B we see a similar action, in that the TD behaves as a bistable switch; but in this case the *applied voltage* is varied to cause the TD to switch. If the same conditions are established, with the TD voltage resting at point (a), an increase in forward bias *voltage* by the generator will move the load line beyond the first positive slope. The TD voltage will switch to point (b), fall to point (c) and remain there. To restore the TD, the forward bias must be reduced below the *valley voltage*. When this is done, the TD will switch back to point (c) and again come to rest at point (a).

If a germanium TD is used, point (a) exists at about 50 mV, and point (b) exists at about 0.35 to 0.5 V.

If the tunnel diode functions as a switch and is biased at point (a), the TD would be extremely sensitive to any external circuit activity; that is, the TD could switch to its high-voltage state accidentally. We may however, reduce the sensitivity of the device by decreasing the forward bias applied to it.

reverse
biasing
the TD

In practical circuitry, it often becomes necessary to bias a TD in the *reverse* direction. This will insure that the device will not accidentally change states. To place the TD in this reverse-biased

state, we simply apply a reverse-bias voltage, producing a reverse current. [See point (*c*) in Fig. 5-3.]

In the discussion to follow, when we talk about the TD drawing reverse current, the operating point will exist somewhere in the reverse current region of the TD's characteristic.

In any countdown circuit (a sweep-gating multi may be considered such a device), the device used to generate the gate must have a transition time significantly shorter than the risetime of the input signal. A fast transition may occur only during one half cycle of the device. As long as *this* transition occurs during trigger interrogation, the device will provide jitter-free triggering. Therefore, the ability of the gating multi to interrogate very fast-risetime or closely spaced pulses, is of prime importance. As we know, the tunnel diode is capable of subnanosecond transition times.

practical
tunnel-diode
gate
generator

A schematic diagram of a practical tunnel diode sweep gating circuit, used to gate a Miller sweep generator, is shown in Fig. 5-4. The circuit consists of a tunnel diode D2, rated at approximately 4.7 mA, a PNP germanium transistor (Q1) used as a gate amplifier, and a disconnect diode (D1), used to disconnect the TD from incoming trigger pulses.

With the circuit in its quiescent state, the sweep-reset multi provides about 3 mA of current for the TD through R1 and R2. The current through the TD sets the diode to its low forward voltage, or "ready" state. With a base-to-emitter voltage of -.05 V, Q1 is cut off.

triggering
the TD gate
generator

To initiate a sweep, the tunnel-diode must be provided with an additional 1.7 mA of forward current. A current trigger is applied through the disconnect diode to the TD. The trigger supplies about 5 to 8 mA of current, which is more than sufficient to switch the TD to its high-voltage state. The TD cathode falls to about -500 mV, back-biasing the disconnect diode. The back-biased diode (D1) will "lock out" any triggers that may arrive during sweep time. The negative pulse generated by the TD is sufficient to drive the gate amplifier (Q1) into

saturation. The collector of Q1 rises rapidly toward the grounded emitter. The positive voltage step at the collector of Q1 is sufficient to back-bias the generator disconnect diode (D3) initiating a ramp.

The saturated gate amplifier draws about .5 mA of base current through R1, R2 and R3, holding the TD in its high state during sweep time.
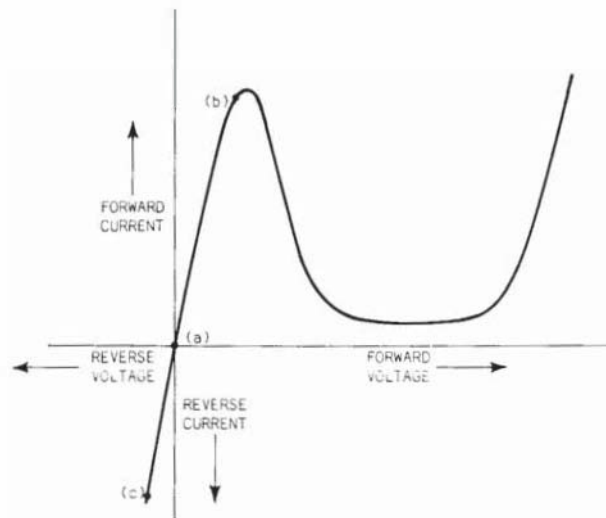

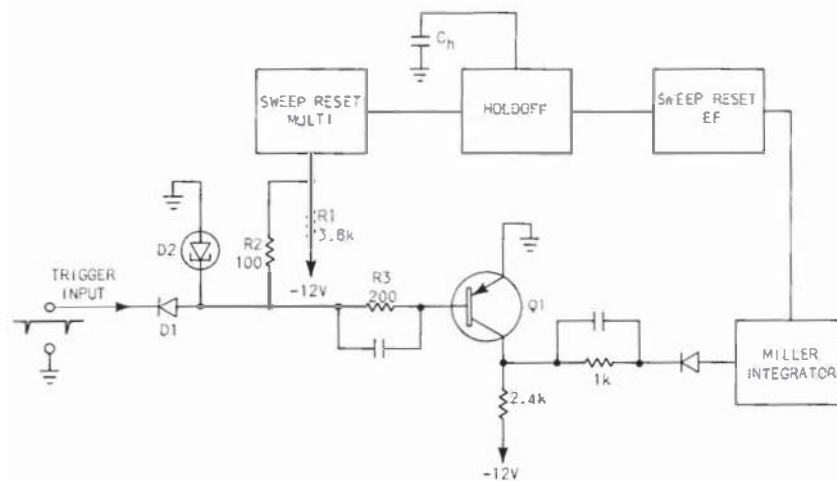Fig. 5-3.   Reverse biasing the tunnel diode.
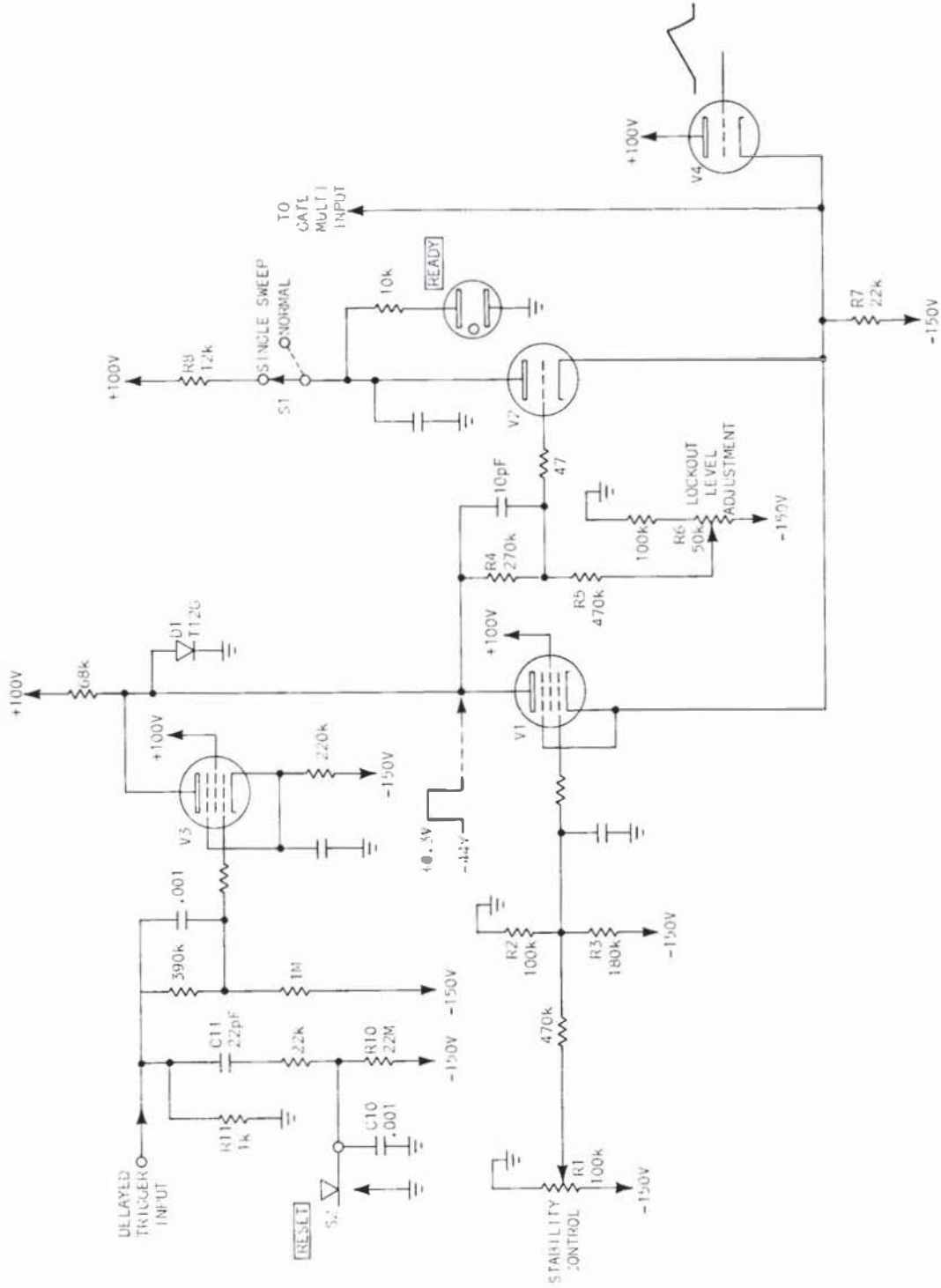

Fig. 5-4.   Practical tunnel-diode gate generator.

Fig. 5-5. Sweep lockout.

When the sweep generator reaches a predetermined
level, the sweep-reset multi switches, pulling the
junction of R1 and R2 positive. The positive voltage
step at the R1 - R2 junction *reverse biases* D2,
switching the TD to a locked-out condition (not
triggerable). The switching action of the TD back
biases the E-to-B junction of the gate amplifier,
cutting off the Q1 collector current. This event
forward biases the sweep disconnect diode (D3),
terminating the ramp. The sweep generator now enters
its recovery period.

restoring
the TD

After the hold-off period the sweep-reset multi
switches back to its quiescent state, again forward
biasing the TD. The TD is now in its low (forward
voltage) state. Diode D1 is again forward biased
and the circuit is "ready" to receive another
trigger.

SWEEP LOCK-OUT

Another type of control device, incorporated in
many sweep generator circuits, is usually referred
to as a *lock-out* circuit. The lock-out circuit
provides a method of disabling the sweep generator
for an indefinite period.

Upon command, the lock-out circuit will set the
input of the sweep generator to a "ready" state.
In this state, if a trigger is applied, a ramp will
be generated. Upon completion of a ramp, the lock-
out circuit will hold the sweep generator
insensitive to any subsequent trigger signals. The
sweep generator cannot be triggered until a second
command signal is applied to the lock-out circuit
to "ready" the sweep generator.

A schematic diagram of a practical lock-out circuit
is shown in Fig. 5-5. The position of S1 activates
or deactivates the lock-out circuit. (The reason
for the SINGLE SWEEP label will become obvious
after the discussion.) With switch S1 in the
NORMAL position, V2 cannot conduct. V1 functions
as a cathode follower, following the activity of the
STABILITY control. Of the total 5 mA of cathode
current through R7 only 1.5 mA flows in the V1 plate
circuit. The balance is drawn by the screen grid.
This configuration allows changes in grid voltage,
without appreciable plate voltage changes. The
common-cathode voltage of V1, V2, and V4 can be

adjusted by the STABILITY control, thus allowing the gate multi to be set to its triggerable, or "ready" state, during normal operation.

As we recall from the "hold-off" discussion, as the sweep runs up, a portion is fed back to the common cathode bus. The rising ramp cuts off V1 and continues to rise, *resetting* the gating multi. When the sawtooth starts negative, this negative voltage is fed back to the gate input, pulling the gate input voltage down until the stability CF (V1) turns on. This establishes the gate multi "ready" level.

single sweep
lock-out

When S1 is switched to the SINGLE SWEEP position the +100 V supply is applied to V2. V2 will not conduct, due to the large negative bias present (about -20 V). The READY light (located on the front panel of the instrument) will be lighted, indicating to the operator that the sweep is "ready" to receive a trigger. Now we'll trigger the sweep and describe the activity within the lock-out circuit.

After applying the input trigger, the ramp begins. The positive-going ramp is fed back to the common cathode bus, eventually cutting off the stability CF (V1). When this happens, the plate voltage of V1 rises toward the +100 volt supply, but is clamped at approximately +.3 volts. Clamp diode D1 establishes a reference point to provide stability to the Lock-Out Level adjustment. The positive voltage step at the V1 plate raises the control grid of V2 from about -70 to -40 volts, turning on V2. The cathode bus is now at about -40 volt level. The rising ramp *resets* the gate multi, and the sawtooth voltage starts falling. The negative-going ramp is coupled to the grid of V4, eventually cutting it off. *Normally* the cathode of V4 would follow the negative-going ramp, lowering the cathode bus to the "ready" level. Because V2 is conducting, the negative-going voltage at the V4 grid will not affect the common cathode bus. V2 holds the input level of the gate multi sufficiently high to insure that the multi cannot be triggered. The sweep generator is *locked-out* and will remain so, until some method of restoring the gate multi to its "ready" state is implemented.

resetting
the sweep
generator

One method of resetting the sweep generator to its "ready" state is to apply a pulse to the grid of V2 sufficient in amplitude to cut off plate current. The V2 cathode would then fall to the gate input "ready" level (set by the STABILITY control).

The "reset" or "enabling" pulse, may be generated in a circuit consisting of S2, R10 and C10, or by the Delayed Trigger generator, incorporated in some oscilloscopes (See discussion of Delay Pick-off in "Oscilloscope Trigger Circuits" book.)

enabling
pulse

To understand the generation of the "enabling" pulse, using the RESET switch S2, we should recognize that a charge of -150 volts appears across C10 when S2 is open. When S2 is closed, C10 discharges through the switch, creating a pulse that starts at a -150 V level and rises to 0 volts (almost instantaneously). This positive pulse is differentiated by R11 and C11 and is applied to the control grid of V3. The pulse is amplified (about 10X), inverted, and applied to the control grid of V2 through the divider R4 and R5. The pulse applied to the grid of V2 is approximately -30 V. The V2 cathode will follow the negative-going grid until the common cathode bus reaches the V1 grid voltage. At that time, V1 will conduct, clamping the falling cathode bus to the *preset* "ready" level. The V1 plate will fall to about -40 volts, insuring the cut-off state of V2. When V2 cuts off, its plate voltage increases toward the +100 volt source, providing sufficient ignition potential to the READY neon. The READY neon fires, indicating a "ready" sweep condition at the front panel.

Another method of sweep lock-out will be described while referring to Fig. 5-6. In following this description, keep in mind the requirements of the lock-out circuit previously described.

Fig. 5-6. Sweep generator reset circuit.

attained, the neon fires. After ignition, the voltage across B1 *stabilizes* at about 60 V. The negative voltage step created by the neons changing from its ignition state to its stable state (-20 V), changes the charge across C2. The resulting negative current change is applied to the base of Q1, driving it into saturation. The positive voltage change at the collector of Q1 is coupled through C3 to the base of Q2, turning it off. When Q2 cuts off, the current through R9 is switched to Q3, raising the collector voltage of Q3 from -9 V to about -2 V. Sweep-reset multi switching action has two results:

1. The RESET lamp lights, due to the change in collector voltage of Q3, which turns on Q4.

2. The TD is forward biased by the negative collector voltage change at Q2, setting the TD to its low (forward voltage), or "ready" state.

So far, we have described the activity that returns the sweep generator from a "locked-out" to a "ready" condition. Now, we will apply a trigger pulse and describe the subsequent operation of the circuit.

triggering
When a negative trigger pulse is applied through the forward-biased coupling diode (D1) the TD will switch to its high (voltage) state if the trigger contains sufficient energy. This activity back-biases the trigger-coupling diode, and also saturates the gate amplifier. This saturated gate amplifier reverse biases the sweep disconnect diode (D3) allowing the sweep generator to generate a negative-going ramp. The negative-going ramp is coupled through the sweep reset EF to the back-biased diode (D13). As the ramp falls from the +13 volt level to the level at the base of Q2 (about 4.5 V, established when Q3 conducts), D13 will conduct, allowing the base of Q2 to follow. When the ramp drives the Q2 base to about +2.3 V (.6 V more negative than the common emitter level), the multi will switch to the condition: Q2 *on*, Q3 *off*. This sets the base of Q2 to about +.5 V. The conduction of Q2 increases current through R6, causing the TD to switch to its low (reverse voltage) state, drawing about 9 mA of *reverse current* from the TD. The TD, in its reverse-biased state, will not respond to input triggers. Therefore,

the sweep generator will be *locked-out* until the TD can be biased to its low *forward voltage* (high forward current), or "ready" state. As Q3 switches from its *on* to *off* state, the E-to-B junction of Q4 is back-biased. This extinguishes the RESET lamp, indicating at the front panel a "locked-out" sweep.

Meanwhile, as the TD became reverse biased, the gate amplifier was cut off, terminating the sweep. The generator entered its recovery period, rising positive to the ramp-start reference. The positive-going output of the sweep generator, coupled through the sweep reset EF, back-biases D13. The base of Q2 is thus set at about +.5 V and will remain there until a *reset pulse* is applied from the RESET circuit.

Normally (when the hold-off circuit is connected to the Q2 base) the base would follow the positive excursion of the hold-off capacitor until the voltage was sufficient to cut off Q2. This would set the input circuit to its "ready" state.

In the foregoing discussion we have studied two methods of lock-out control. These were selected as typical examples of techniques used to incorporate lock-out into sweep generator systems.

AUTOMATIC SWEEP TRIGGERING

The ability to generate a sweep automatically is a desirable feature incorporated in many more sophisticated oscilloscopes. This feature provides a reference sweep to the operator for beam positioning prior to application of a vertical signal. Even though the sweep is running, it retains the ability to be triggered. That is, if a trigger is applied, the sweep will function as a triggered sweep (within limits).

One of the earliest automatic-sweep techniques incorporated a free-running multivibrator, used as a sweep-trigger source. (See discussion of Automatic Operation in the "Oscilloscope Trigger Circuits" book.

automatic
sweep
generation

A more recent automatic-sweep technique incorporates a circuit which automatically sets the *state* of the sweep generator. Recalling that in earlier instruments, the sweep generator could be set to one

of three states by the STABILITY control (Locked-out, Ready, or Free-running), we will now discuss a circuit that *automatically* sets the sweep generator to either the "free-run," or "ready" state, depending on trigger circuit activity. The term "AUTO-STABILITY" will be used to describe this circuit.

bright-line auto

Sometimes the term "Bright-Line-Auto" is used when referring to the Auto-Stability technique. This term originated mainly to emphasize one advantage of the later method of automatic sweep generation. In the early instruments, a free-running multi functioned as a pulse generator. The output of the multi was processed and used as a trigger source for the sweep generator. This technique provided a triggered sweep, even in the absence of externally applied trigger signals. A frequency of approximately 40 to 50 Hz was chosen for the multi, mainly to facilitate measurement of audio frequencies and 50- or 60-Hz line frequencies. Therefore, in the absence of an external trigger, the sweep *repetition rate* was limited to the 40-Hz trigger-multi frequency. This low rep-rate sweep resulted in a gradual decrease in trace brightness as sweep speed was increased. The Auto-Stability technique prevents trace dimming considerably, by setting the sweep generator in a free-running condition in the absence of an external trigger.

The repetition rate of the free-running sweep generator is usually much higher than 40 Hz (typically in excess of 100 kHz). Consequently, the term "Bright-Line-Auto" is sometimes used when referring to an instrument which incorporates the Auto-Stability circuitry.

A schematic diagram of an Auto-Stability circuit is shown in Fig. 5-7A. The integral parts of the circuit are:

auto-stability circuit

    A. Trigger pulse amplifier and shaper (Q1).

    B. Monostable multi (Q2 and Q3).

    C. Transistor switch (Q4).

    D. Diode gate (D4 and D5).

Fig. 5-7. Automatic stability control circuit.

The main purpose of the circuit is to set the collector level of Q4 to one of two possible conditions, either +12 V or -1 V. When the voltage at the Q4 collector is +12 V, the input to the sweep generator is set to a "ready" condition. When the Q4 collector level is -1 V, the sweep generator free-runs. The output level of Q4 is maintained at +12 V only if triggers are present at the input of Q1, and the period between triggers is not greater than 50 ms.

Quiescently, the output voltage (Q4 collector voltage) is -1 V. At this level, D4 is reverse biased. The sweep-reset multi provides about 3 mA of forward current to the TD, setting the TD to its "ready" state. The level at the junction of D5 and R10 forward biases D5, providing an additional 2 mA of current from R6 to the gating tunnel diode (TD6). The additional 2 mA of current into TD6 causes it to switch to its high (voltage) state, which provides a sweep start. As long as the collector of Q4 remains at -1 V, the TD will oscillate at a rate determined by the switching rate of the sweep reset multi; i.e., the sweep will free-run.

While referring to Fig. 5-7, we will examine the circuit activity initiated by an applied trigger pulse.

When a trigger pulse is applied to the trigger-shaper amplifier (Q1), the pulse is inverted, then differentiated by the collector load L1. The positive output pulse is applied through the forward biased diode (D1), to the base of the normally *off* transistor Q2. The pulse forward biases the E-to-B junction of Q2, bringing it into conduction. The negative-going voltage step at the Q2 collector is coupled through C3 to the base of Q3, cutting it off. The transition (Q2 conducting and Q3 cutting off), is accompanied by a regenerative action in the common-emitter circuit, causing the multi to switch to its quasi-stable state. The negative-going collector voltage of Q2 disconnects diodes D2 and D3, allowing the base of Q4 to fall toward ground. The base of Q4 falls toward ground until the E-to-B junction becomes forward biased, at which time Q4 saturates. When Q4 saturates, its collector rises to the +12 V level, forward-biasing diode D4. When D4 turns on, the quiescent 2 mA of bias current formerly flowing into the TD gate is switched from D5 to D4. This action reverse-biases D5, eliminating

the current path to the TD gate. Consequently, the gate generator receives only 3 mA from the sweep-reset multi. This current reduction terminates the free-running activity, and sets the gate generator to its "ready" state.

The Auto-Stability circuit has a recovery time of approximately 85 ms. If, after one trigger is applied, the multi is allowed to recover, the output will switch back to its -1 V level, setting the sweep generator to its "free-run" state once again.

SWEEP SWITCHING

A number of Tektronix oscilloscopes incorporate two separate sweep generators. One sweep may be employed as a delay circuit for the other, permitting a variety of useful display modes not available in instruments employing only one sweep generator. This feature may be combined with a system of mechanical and electronic switches, gating circuits and interconnections to provide sweeps that are *alternately* applied to the horizontal deflection system of the oscilloscope. Many advantages, otherwise found only in dual-beam instruments, are gained by this "time sharing" of a single-beam CRT.

When a dual-trace vertical plug-in unit is incorporated in the instrument, the sweep-switching feature provides the user with the equivalent of two complete oscilloscope systems in one package.

Before we examine the actual mechanics of sweep switching, we should become familiar with the principal modes of display made available by the dual-sweep feature itself.

display modes

The individual sweeps displayed by a dual-sweep instrument may be labeled "A" and "B." Either "A" or "B" sweep may be displayed separately, or a mode ("A" ALT "B") may be selected to display "A" alternating with "B" sweep.

Another mode, "B" INTENS BY "A," may be selected.
In this mode, "B" sweep is displayed, a portion of
which is intensified during "A" sweep's ramp duration.
The time between the start of "B" sweep and the
intensified portion is determined by a comparator
circuit within the instrument. Therefore, "A" sweep
is *delayed* by "B" sweep.*

The feature "A" sweep delayed by "B" introduces
another mode of display available to the user. The
mode, "A" DLY'D, will provide a display of "A"
sweep after it has been delayed by the "B" sweep.

In addition, one more mode is made available by the
alternate-sweep switching feature, the mode "B"
INTENS BY "A" - ALT - "A" DLY'D. In this mode, "B"
sweep is displayed while being intensified by "A,"
after which, "A" sweep is displayed after being
delayed by "B"; i.e., "B" INTENS BY "A" is alternated
with "A" DLY'D.

sweep-
switching
circuitry

Fig. 5-8 is a simplified logic diagram of the sweep-
switching circuitry. Diode gates are located between
the sweep generator outputs and a common horizontal
amplifier input. The symbols used to represent these
gates imply an *AND* function, that is, two
predetermined signals must appear simultaneously
at each input of the gate if a sweep is to be
displayed.

The gates are controlled by a sweep-switching
multivibrator, the outputs of which (being at
opposite levels) hold one gate open and the other
closed. For example, in the "A" ALT "B" mode, if
the switching multi is in a state that opens "A"
gate and "A" sweep runs, it will be displayed, and
"B" gate will be closed. When "A" sweep terminates,
the multi will switch states, closing "A" gate and
opening "B," allowing "B" sweep to be displayed.

"A" ALT "B"

When operating in the "A" ALT "B" mode (Fig. 5-9),
a provision is made to insure that while one sweep
is being displayed, the other is disabled. A signal,
coincidental with that which holds one of the gates
closed, is applied to the hold-off multi of the sweep
not being displayed. This "lock-out" signal insures
that the undisplayed sweep will be disabled for the
duration of the displayed sweep.

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*See Delaying Sweep in "Oscilloscope Trigger Circuits"
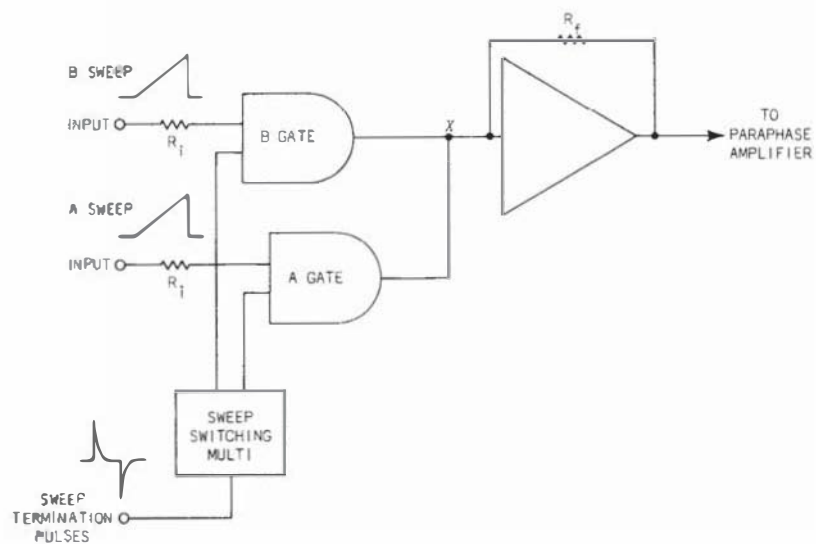book.

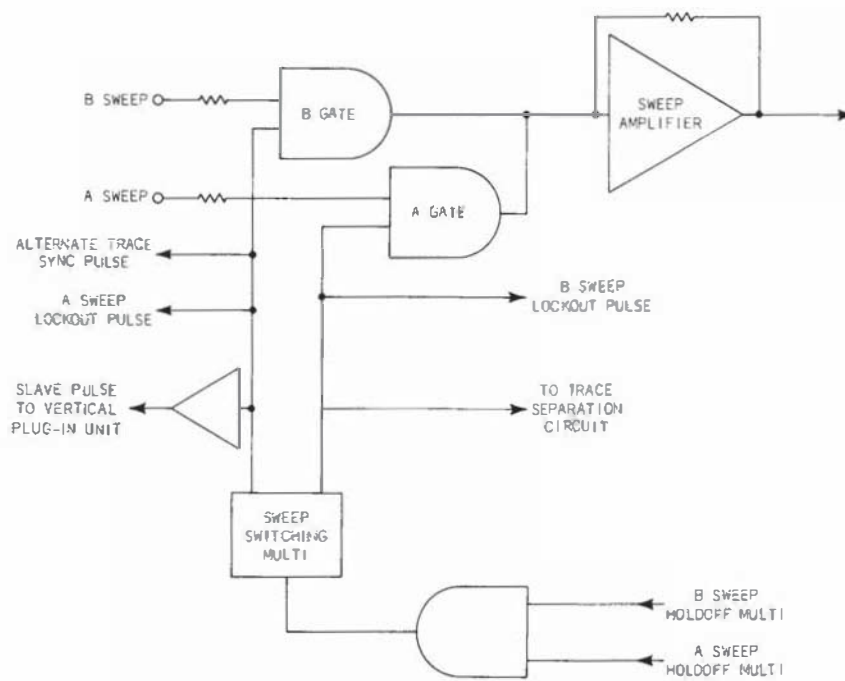Fig. 5-8.   Sweep switching schematic.

Fig. 5-9.

slave pulse

In addition to providing lock-out pulses, the sweep-switching multi generates a pulse used to "slave" a dual-trace vertical plug-in unit. This slave pulse is applied to a trace-switching multi (within the plug-in), to *insure* that "A" sweep will be *locked* to channel 1 and "B" sweep to channel 2. In addition, "alternate-trace sync pulses" are applied to the trace-switching multi, which allows the vertical channels to be "time-shared" with either "A" or "B" sweep or both (depending on the horizontal-display mode).

The sweep-switching circuit must perform another function if the alternating sweeps are to be distinguishable on the CRT display. Without this provision, alternating sweeps would share a common base-line. This would be confusing to the viewer. Therefore, a provision is made to vertically separate the traces. An output is provided by the sweep-switching multi to a "trace-separation" circuit. The output of this circuit provides a push-pull DC-offset voltage to the main vertical amplifier whenever "A" sweep is displayed, which offsets the "A" trace from the "B" trace. Signal fidelity is preserved, of course, since we simply change the vertical position of the trace *automatically*. The magnitude and

trace
separation

polarity of the "offset voltage" is set by the TRACE SEPARATION control. The sweep-switching multi turns the trace-separation circuit *on* whenever "A" sweep is displayed, whether the plug-in is a multiple or single-channel type.

In the "A" ALT "B" mode of operation then, it is possible to make two individual measurements on the CRT and have many of the advantages of a dual-beam instrument.

sweep rates

It must be remembered, however, that because the channels *time-share* the CRT, if the repetition rates of the two sweeps are widely separated, the channel driven by the faster sweep will not appear often enough to create the desired trace intensity. This occurs because the *slowest sweep* limits the sweep-switching repetition rate. The oscilloscope has a separate BRIGHTNESS control to compensate for this effect (within limits). When the sweep speeds are widely separated, the "A" sweep should be operated at the faster rate, and the INTENSITY control adjusted for a nominal "A" trace brightness. The BRIGHTNESS control should then be adjusted to bring the "B" trace brightness to the level of "A." This procedure is necessary because the BRIGHTNESS control simply attenuates the "B" unblanking pulse; therefore, the "B" trace cannot be brighter than the "A" trace.

single
sweep

Another important characteristic of the sweep-switching feature relates to "single sweep" presentations. It must be recalled that both sweeps cannot run at the same time. Consequently, if both sweeps are triggered from a common signal, time coincidence measurements cannot be made using both sweeps simultaneously.

"B" INTENS
BY "A"

When the instrument is operated in the "B" INTENS BY "A" mode, if "A" sweep were to continue after "B" sweep terminates, the intensified zone could extend back into the start of the next "B" sweep display. This would appear as a confusing double brightened area on the display. To prevent this, a pulse is taken from the "B" sweep hold-off multi (when the multi transition terminates "B" sweep) and applied to the hold-off multi in the "A" sweep generator circuit. This pulse terminates "A" sweep coincidentally with the completion of "B," a technique sometimes referred to as "anti-wrap around."

In the discussion to follow we will investigate the most significant subcircuits, which when combined, go together to make up the complete sweep-switching circuit. We will confine our discussion mainly to the "A" ALT "B" mode of operation, with occasional references to the other modes as the need arises.

sweep-
switching
multi

The first circuits to be considered will be the sweep-switching multi, made up of Q1, Q2 and their associated components. In addition, we will examine switching-pulse amplifier Q3, its associated components, and its relationship to the sweep-switching multi. See Fig. 5-10.

The sweep-switching multi is a bistable device. When operated in the alternate mode, either Q1 or Q2 will conduct until a negative input pulse switches the multi to its opposite state. Diodes D4 and D5 are "catching diodes," used to limit the collector swing of the switching transistors. This is done for two reasons: First, the diodes insure that the C-to-E breakdown voltage of the switching transistors will not be exceeded, and second, the clamping action improves the multi transition time. This is true because the transistor collector voltage must change only 27 volts of a potential 100-volt change.
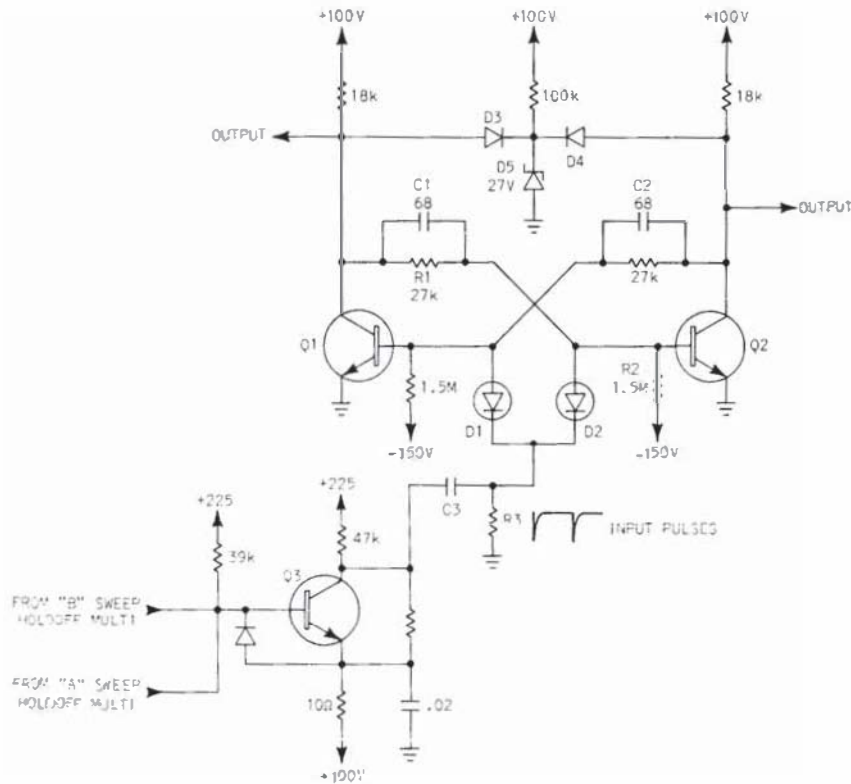
Fig. 5-10. Sweep switching multivibrator.

Let us assume that Q2 is conducting (saturated).
The voltage at the base of Q2 will be more positive
than that at the base of Q1. A negative pulse,
applied to the common cathode of D1 and D2, will
forward bias D2, and reverse bias the E-to-B
junction of Q2, cutting it off. A portion of the
positive voltage change at the Q2 collector is
coupled to the base of Q1, bringing Q1 into
conduction. When Q1 conducts, its collector voltage
falls. This negative voltage change is coupled to
the base of Q2, insuring the cut-off state of Q2.
Thus, the first trigger pulse causes the multi to
switch states, resulting in an output voltage change
from 0 to 27 volts at the Q2 collector, and a
complementary voltage change from 27 to 0 volts at
the Q1 collector. If a second trigger were applied,
the multi would revert to its initial state. Now
that we realize that the multi is "switched" by
the applied trigger pulses, it should be apparent
steering   that the diodes (D1 and D2) "steer" these pulses
diodes   to their respective inputs. Diodes used in this
manner are sometimes referred to as "steering
diodes."

The amplifier (Q3) is designed to respond to a
specific occurrence at its base-to-emitter junction.
Two input leads are connected to this junction.
Each of these input leads is connected to the plate
circuit of one side of its respective hold-off multi.
When both input levels go *high* (voltage), Q3 will
saturate. When Q3 saturates, a negative step occurs
and is differentiated by R3 and C3. The differentiated
pulse drives the steering diodes, one of which will
steer the negative pulse to the appropriate base
circuit of the sweep-switching multi, causing the
multi to switch states.
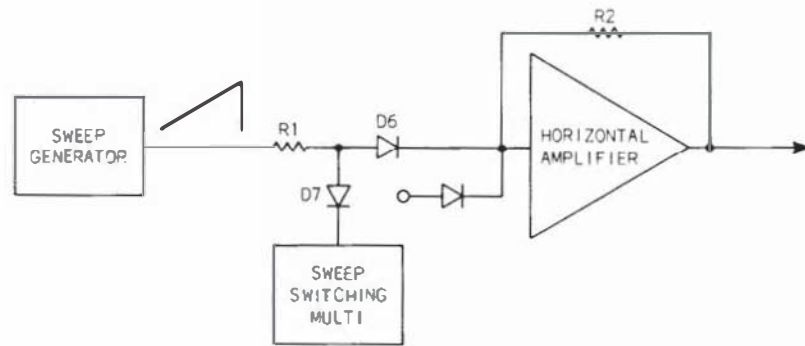
Fig. 5-11. Sweep gate.

The next significant circuit incorporated in the sweep-switching circuit is the sweep gate. Fig. 5-11 shows a schematic diagram of one of the gates. Since the two gates are identical, the following discussion will serve for both.

The output of the sweep-switching multi, as it appears at D7, will exist at one of two levels (27 or 0 volts). When the output level is 27 volts, D7 is reverse biased and D6 conducts. If a sawtooth appears at the gate input, D6 will couple the sawtooth current to the input of the horizontal amplifier. When the sweep-switching multi changes states, D7 becomes forward biased and D6 is reverse biased. Any sawtooth current appearing at the gate will be shunted to ground through D7 and the saturated transistor within the multi. At the same time, D6 (being reverse biased) will prevent any sawtooth current from appearing at the horizontal input.

Referring to Fig. 5-8 once again, we can easily understand the symbol used to represent the diode gates. To obtain an output from the gate, both a sawtooth *and* a positive pulse from the switching multi must simultaneously appear at the inputs to the gate.

Fig. 5-12 shows a schematic of the trace-separation circuit. If we assume that time-base "B" is being displayed, the output level of the switching multi will be 0 volts. This level will forward-bias D1

and D2, setting their anodes equally at about 0.5 volts, clamping the control grids of the amplifier to that level. Under these conditions, the TRACE SEPARATION control *will not* affect the vertical position of the trace. However, when time base "A" is displayed, the voltage at the common cathode bus of D1 and D2 will be 27 volts. This level back-biases the diodes. Under these conditions, if the TRACE SEPARATION control is adjusted, a voltage change will appear at the control grids of the amplifier. A push-pull output voltage change will appear at the plates of the amplifier. This *offset* voltage positions the "A" trace in accordance with the setting of the TRACE SEPARATION control. When time base "B" is again displayed, D1 and D2 become forward biased, clamping the amplifier control grids once again. The trace returns to its previous vertical position.
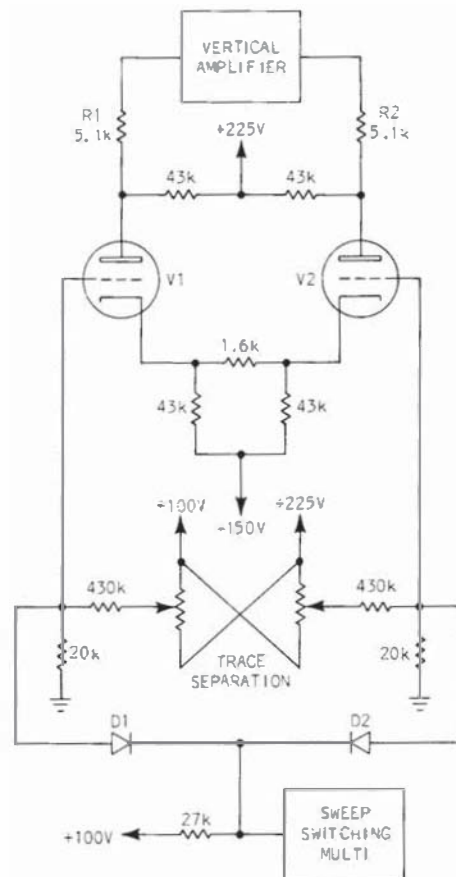
Fig. 5-12. Trace separation amplifier.

Fig. 5-13. Alternate sweep switching.

In those modes not utilizing sweep switching, the
D1 - D2 common cathode bus is grounded directly,
disabling the trace separation amplifier.

Fig. 5-13 is a schematic diagram incorporating those
circuits which go together to make up the complete
sweep-switching control circuit. The previous
discussion of the integral parts of this circuit
should contribute to a clear concept of sweep
switching.

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