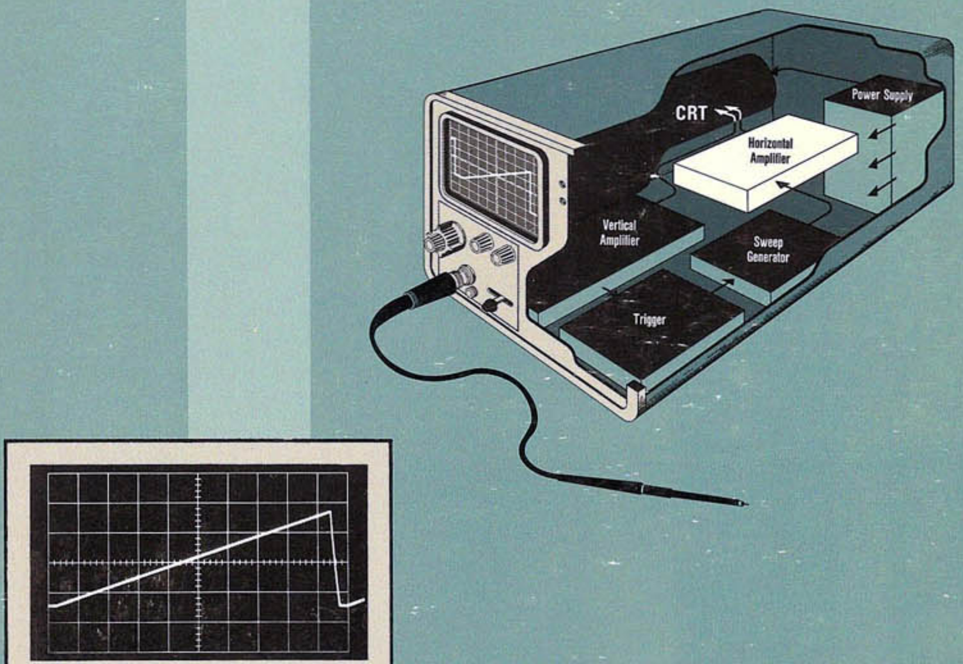
Horizontal Amplifier Circuits



Circuit Concepts Series

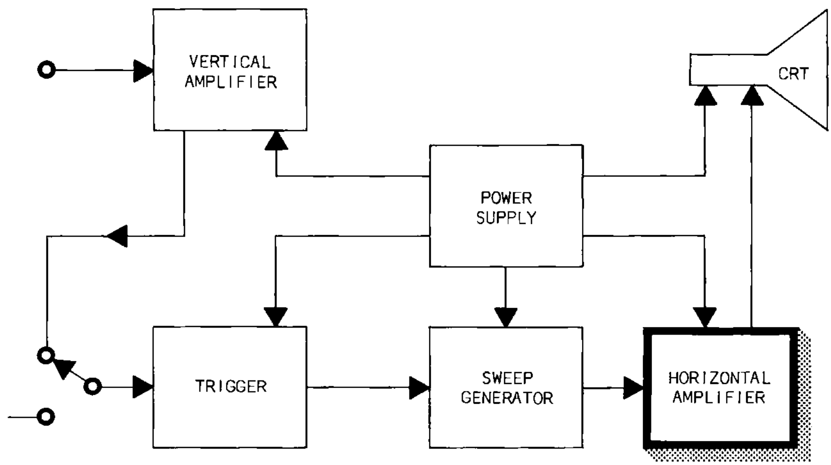
HORIZONTAL AMPLIFIER CIRCUITS

BY

KENNETH L. ARTHUR

TEKTRONIX®

CIRCUIT CONCEPTS


VERTICAL
AMPLIFIER
CRT
POWER
SUPPLY
TRIGGER
SWEEP
GENERATOR
HORIZONTAL
AMPLIFIER

INTRODUCTION

An oscilloscope is an electronic measuring instrument which displays an analogue of an electrical event. The conventional display is a plot of current or voltage amplitude as a function of time. Generally, the time base, or reference, is generated within the oscilloscope by a group of circuits collectively referred to as the *horizontal-deflection system*. This volume is concerned with one of the major units of the horizontal-deflection system, the *horizontal amplifier*.

The principal function of the horizontal amplifier is to convert the carefully developed time-base ramp to a driving voltage for the horizontal-deflection plates of the oscilloscope cathode-ray tube (CRT). Less frequently, the amplifier may be called upon to process an independent variable signal in much the same fashion as the vertical amplifier.

After a brief discussion of the factors which enter into horizontal-amplifier design, the reader is conducted through detailed examinations of eight different amplifiers. The selected circuits represent a cross-section of the horizontal amplifiers found in oscilloscopes currently produced at Tektronix, Inc. In analyzing the operation of these amplifiers, emphasis is placed on simple rule-of-thumb procedures rather than on rigorous mathematical treatment.

Although not a prerequisite to an understanding of this volume, *Vertical Amplifiers*, another book in this series, is highly recommended as a comprehensive introduction to oscilloscope amplifiers. There, the rule-of-thumb procedures mentioned above are developed and explained in detail. In addition, valuable information on circuit components, frequency and risetime characteristics and general amplifier theory is presented in terms which are easily understood -- information which is not repeated in this publication.

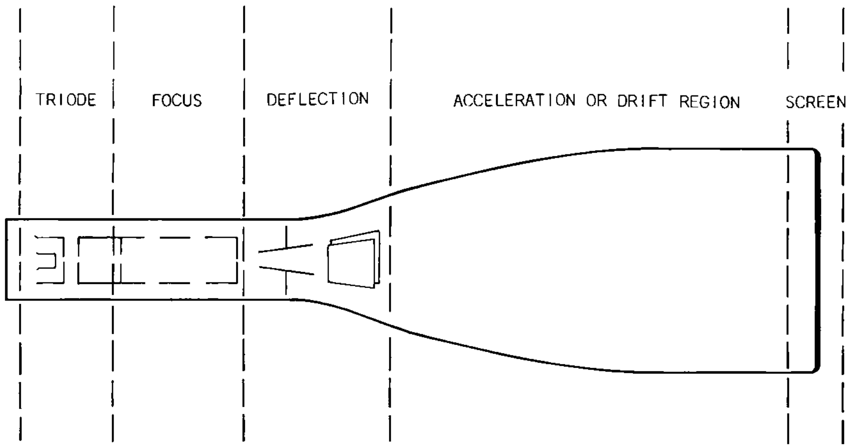


Fig. 1-1. CRT sections.

1

BASIC FUNCTION AND DESIGN

horizontal-amplifier function

The primary function of the oscilloscope horizontal amplifier is to convert the time-base (sweep-generator) ramp to an appropriate deflection-plate driving voltage for the cathode-ray tube (CRT). As a secondary function it provides a number of controls which contribute to the oscilloscope's measurement capabilities. The horizontal-amplifier configuration is therefore largely determined by the nature of the sweep-generator signal and the operating characteristics of the CRT, as well as the variety and scope of the control functions it must perform. It is the purpose of this chapter to examine these factors and their influence on the design and operation of the horizontal amplifier. To begin, let us consider the load which the amplifier must drive -- the CRT.

CRT

An oscilloscope CRT is a special-purpose vacuum tube consisting, essentially, of an evacuated glass or ceramic envelope, a heated cathode, a phosphor-coated anode (faceplate) and various control electrodes (see Fig. 1-1). In operation, a high potential, usually on the order of 10 kV, is established between cathode and anode. Electrons "boiled" off the heated cathode are accelerated by this potential, striking the anode at a high velocity. The phosphor molecules involved in the collision emit energy in the form of light, generating a spot of light on the CRT screen. Control electrodes focus the electrons into a narrow beam and determine the size, shape and brightness (intensity) of the spot. The position of the spot with respect to the x and y axes of the CRT graticule (coordinate grid) depends on the relative amplitude and polarity of the deflection-plate voltages. The deflection plates closest to the cathode have a greater effect (per unit of electrostatic charge) on the electron beam than the second pair, since the electrons at this point have less velocity. Because the vertical-deflection system must exhibit the greatest sensitivity, the output of the vertical amplifier is almost always applied to these more sensitive deflection plates.

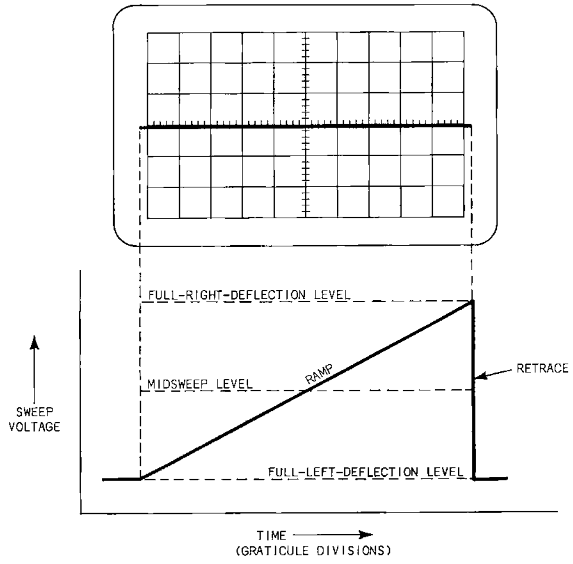


Fig. 1-2. Horizontal deflection – sweep versus trace.

When a sawtooth voltage is applied to the horizontal-deflection plates, the electron beam moves from left to right as the ramp rises (Fig. 1-2), then returns quickly to its starting position as the voltage falls to its starting level. For the sake of clarity and consistency, the sweep voltage will always be pictured as a positive-going sawtooth, rising from its DC starting level as the spot moves from left to right. This is the waveform that appears on the right-hand deflection plate. Sweep amplitude refers to total deflection voltage, whether single-ended or push-pull. The beam is cut off, or "blanked", during the retrace portion of the sawtooth so that no trace is generated during the retrace interval. The degree of beam displacement effected by a change in deflection-plate voltage depends on a number of variables, whose ratio is known as the (horizontal or vertical) *deflection factor* of the CRT. This characteristic is expressed in volts per division (or V/cm) and may be defined as the change in voltage required to produce one unit of CRT-beam deflection. (The term "deflection factor" may be used with respect to any point in the deflection system, but in the above instance refers to a fixed

sweep amplitude

deflection factor

parameter of the CRT itself.) Thus, if the horizontal deflection factor of the CRT is 10-V/div, the horizontal-deflection plates must be driven by a 100-V horizontal sweep to generate a 10-division trace.

DC level

Notice that we have only established the necessary sweep amplitude and not the *DC level* of the sweep. Here again, certain built-in characteristics of the CRT are the determining factors.

Most CRT's are designed so that in the absence of deflection forces (equal potential on both deflection plates) the spot appears in the center of the graticule (Fig. 1-3). This means that at the beginning of the trace the left-hand deflection

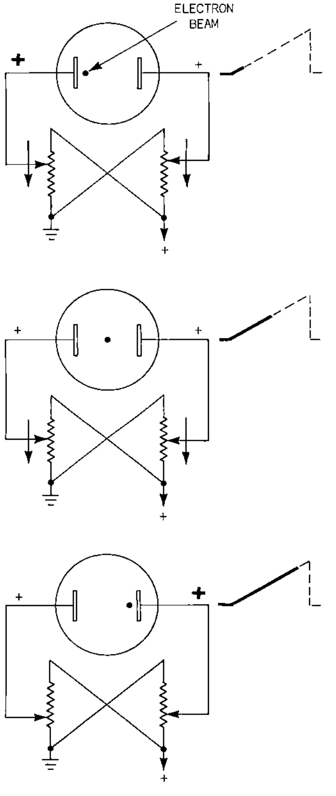


Fig. 1-3. Horizontal deflection – spot position versus deflection-plate voltage polarity.

plate must be positive with respect to the right-hand plate, while at the end of the trace the right-hand plate must be equally positive with respect to the left-hand plate. If these conditions are met, and sweep amplitude is appropriate to the CRT deflection factor, a 10-division trace should be obtained. It would seem, therefore, that the actual DC level of the midsweep voltage is unimportant. However, this is not true. The size, shape and focus of the spot created in the phosphor coating of the CRT depend, among other things, on the velocity of the electron stream at the anode. This velocity is a function of the acceleration imposed on the electrons by the electrostatic field between cathode and anode. The distribution of potential between these electrodes is called the *potential gradient*. (Fig. 1-4.) The gradient potential at the horizontal-deflection plates must be disturbed as little as possible by the deflection voltage. For this reason, the CRT cathode and anode voltages are selected so that the potential gradient at the deflection plates is neither so high nor so low as to require unrealistic plate or collector voltages in the output stage of the amplifier. It must be remembered, however, that even though the quiescent DC level at the deflection plates may be matched with the potential gradient of the CRT, both vertical- and horizontal-deflection signals will impose some change on the electrons's axial, as well as radial, velocity.

potential gradient

The most effective way to deal with this problem is to use *push-pull* rather than single-ended driving signals at the deflection plates. For example, if the sweep voltage is applied to only one of the deflection plates, the other plate must be connected to ground or some other reference (Fig. 1-5). Under these conditions, only one deflection plate changes potential during the sweep. The electrostatic field between the two plates and the preceding electrode (shown here as the accelerating anode for simplicity) is therefore distorted or unbalanced. At one extreme of the sweep, the electron encounters a negative equipotential surface before it is subjected to deflection. At the other end it encounters a positive equipotential surface. In the first case, the electron is decelerated and thus undergoes greater-than-average deflection (expansion). In the other case electrons are accelerated and undergo less-than-average deflection (compression).

push-pull deflection

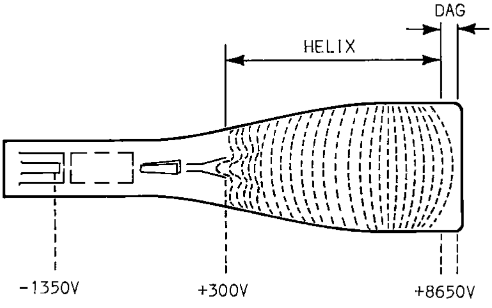


Fig. 1-4. Distribution of electrostatic potential in CRT.

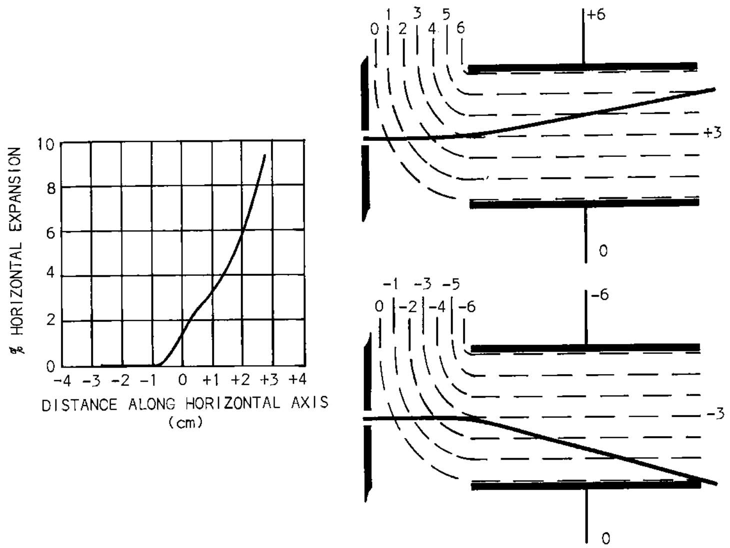


Fig. 1-5. Single-ended deflection.

symmetrical nonlinearity

Nonlinearity due to expansion causes positive measurement errors, while that due to compression causes negative errors. However, if *two* sweep voltages of opposite polarity and with half the amplitude of the single sweep are applied to the deflection plates, the field becomes balanced

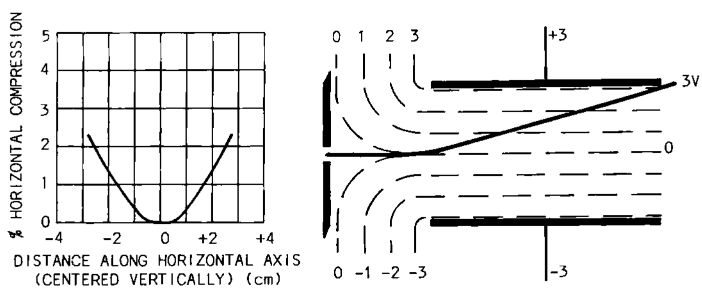


Fig. 1-6. Push-pull deflection.

(Fig. 1-6). Now all electrons enter the deflection-plate region at the same velocity, and therefore are deflected by the same factor at both ends of the sweep. Although, as stated, the deflection process itself tends to introduce small linearity errors, with *push-pull* deflection the same *type* of error will be produced on each side of the axis and, through appropriate engineering, can be confined to the outer ends of the trace. In addition, through deliberate introduction of *controlled* nonlinearity in the horizontal-amplifier circuits, the deflection-plate signal can be distorted in such a way that it compensates for the remaining small nonlinearities in the CRT. The advantages of push-pull deflection are so pronounced that single-ended deflection systems are seldom encountered in modern oscilloscopes.

trace position

Since the central portion of the trace is the most linear, it is also the best region for making accurate time measurements. In many cases, of course, the waveform under investigation will appear to the left or right of this region, depending on its time relation to the sweep trigger. However, the *position* of the *trace* with respect to the graticule can be adjusted by changing the DC level of the sweep deflection signal. (Fig. 1-7.) In this way, the waveform can be brought to the center-screen area. For example, assume that a sawtooth voltage, rising from zero to 100 volts, produces a 10-division trace centered on the graticule. A 50-V to 150-V sweep applied to the same CRT would then cause the trace to start at the center of the graticule and theoretically terminate

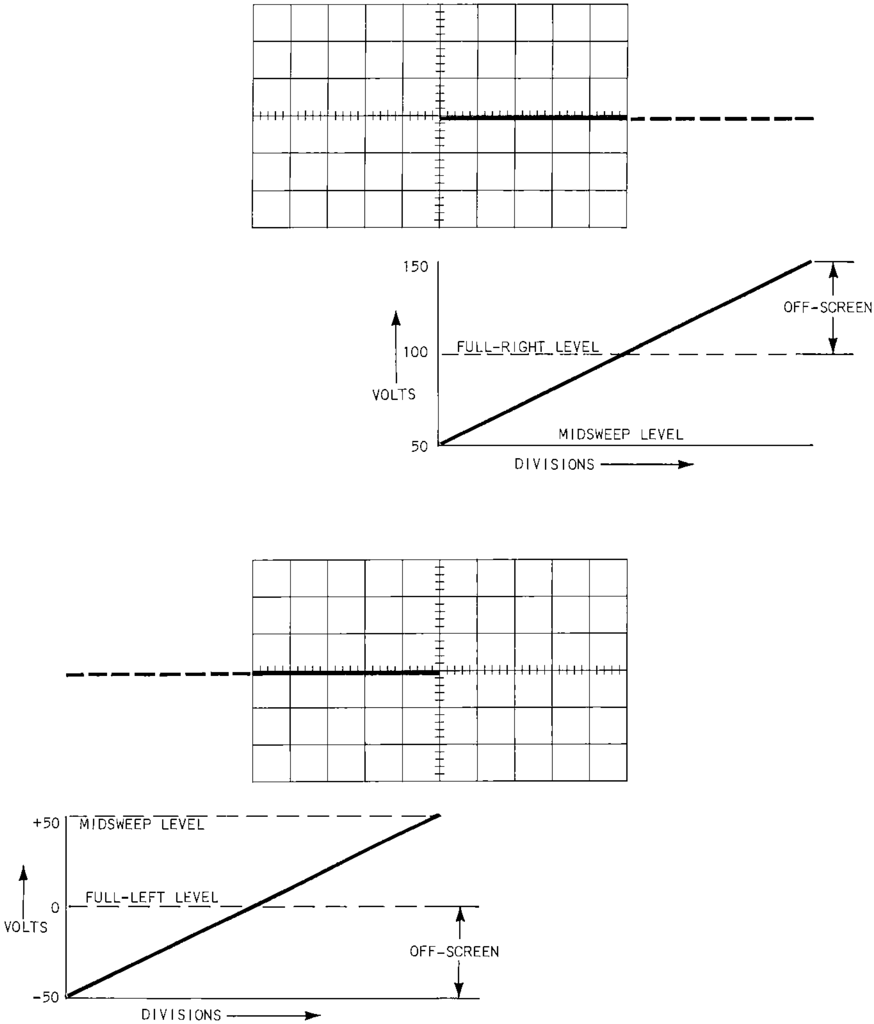


Fig. 1-7. Trace position.

about 5 divisions off-screen. (Actually, the deflection voltage is limited, as will be explained in the next chapter, to values only slightly in excess of the full-right and full-left deflection values.) Conversely, a -50 V to +50 V sweep would cause the trace to begin off-screen and end at the center of the graticule. Since the waveform under investigation maintains its relative position on the trace, it is clear that any waveform occurring within the trace limits can be positioned at the center of the graticule by adjusting the DC level of the horizontal-deflection voltage. For this reason, the DC-level control on the front panel of the oscilloscope is usually labeled HORIZONTAL POSITION.

deflection-plate capacitance

Another CRT characteristic which influences the configuration of the horizontal amplifier is deflection-plate capacitance. The two deflection plates themselves act as a capacitor. There also exists a small capacitance between the deflection plates and the other electrodes of the CRT. This capacitance, lumped together with the stray capacitance of the leads and output circuit and the plate or collector capacitance of the output amplifier, can easily reach a value of 20 pF.

constant current capability

To produce a linear trace, the electrostatic field between the plates must change at a linear rate. This in turn requires a linear change in deflection-plate voltage. At the same time, a linear change in voltage across the lumped capacitance described above requires a *constant* charging current, as shown by the familiar equation: $\frac{\Delta V}{\Delta t} = \frac{I}{C}$. Although the fastest sweep speed attained in the sweep generator is usually on the order of 50 ns/div, many instruments increase this value through use of a *magnified* sweep capability, as will be described later in this chapter. If this mode of operation is employed, the actual sweep speed at the deflection plates can be shortened to 5 ns/div and below. To calculate the current demand on the horizontal amplifier under these conditions we need only substitute the above values in the equation, using a typical horizontal-deflection factor of 25 V/div:

$$I = \frac{25 \cdot 20 \cdot 10^{-12}}{5 \cdot 10^{-9}} = 100 \text{ mA}$$

Current requirements on this order will obviously have a noticeable influence on the amplifier configuration.

Now let us consider the situation at the other end of the amplifier. The driving signal for the horizontal amplifier is, of course, the time-base ramp from the sweep generator (Fig. 1-8). This signal always has the same amplitude, but the risetime of the ramp gets shorter (faster) as sweep speed increases.

waveform fidelity

The range between the slowest and fastest risetimes in a conventional instrument is typically 5 seconds to 500 nanoseconds. Thus, compared to a typical oscilloscope vertical amplifier, the horizontal amplifier poses fewer problems in risetime response. Furthermore, the problems which do exist are simplified by the fact that, generally speaking, only the rising (ramp) portion of the sweep waveform need be faithfully reproduced -- that is, aberrations in the retrace portion of the waveform can be ignored, provided their duration does not exceed

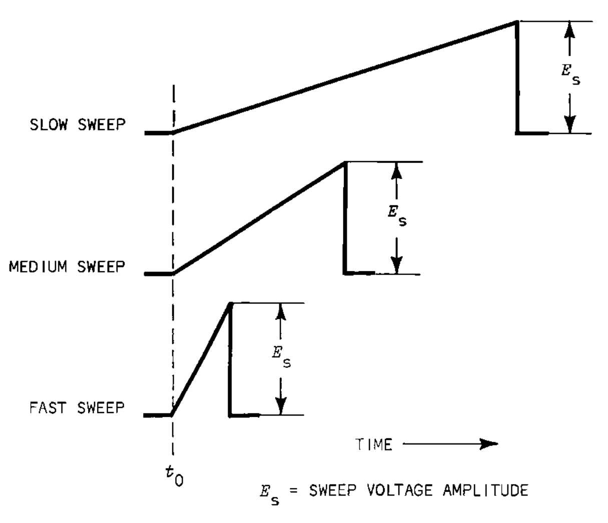


Fig. 1-8. Time-base ramp.

the normal hold-off period between sweeps (Fig. 1-9). The critical portion of the sweep waveform at high sweep speeds is, of course, the initial rise from sweep start. The maximum $\frac{dV}{dt}$ slope that can be attained in any circuit (expressed as a percentage) is

$$\% \text{ slope} = \frac{gm}{2.2\ C_o} \cdot 100,$$

where gm is the transconductance of the active device and C_o is the output capacitance of the processing circuit. For this reason, horizontal amplifiers that must process fast sweeps make generous use of high-gm devices, and amplifying circuits are designed to keep output capacitance low.

high-amplitude sweep signal

The time-base ramp developed in vacuum-tube run-up circuits may be as high as 150 volts in amplitude, while those generated in transistor circuits are usually on the order of 10 to 20 volts. Considering the difficulties involved in generating linear sweep signals of high amplitude, the casual observer is usually puzzled by the seemingly excessive dimensions of the sweep generator signal. His puzzlement is increased when he finds that these signals are attenuated considerably before being processed in the sweep amplifier.

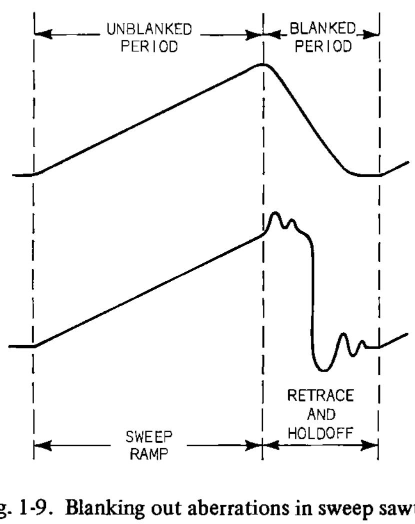


Fig. 1-9. Blanking out aberrations in sweep sawtooth.

The answer to this apparent anomaly is actually quite simple. Unavoidable power-supply ripple, present at the plates or collectors of the active devices in the sweep generator, adds to the sweep signal itself as a form of "noise." This noise, of course, degrades the linearity of the ramp and could cause inaccuracies in time measurements. However the ratio of signal-to-noise can be improved by increasing the amplitude of the sweep signal, since power-supply ripple has a constant amplitude. Thus, where a 5-volt sweep generator signal with 100-mV ripple would exhibit a 2% deviation from linearity; a 150-V signal from the same circuits would show a barely discernible deviation of 0.07%. In other words, the signal-to-noise ratio has been improved from 50 to 1500.

improving signal-to-noise ratio

The signal-to-noise ratio is not changed by attenuation, since both components of the signal are equally reduced. But what of the power-supply ripple in the horizontal amplifier? This question points up another advantage of push-pull deflection. Power-supply ripple appears to the push-pull stages of the horizontal amplifier as a common-mode signal and is characteristically rejected in the amplification process.

power-supply ripple

sweep-signal attenuation and amplification

Now that the high amplitude of the sweep generator signal has been accounted for, it remains only to explain why (in the case of the 150-V sweep generator signal) the signal must be attenuated *and then amplified* before it is applied to the deflection plates, rather than applied directly. In the first place, the horizontal amplifier provides a number of convenience and calibration controls which would otherwise have to be located in the sweep generator where they would severely complicate the problem of generating a linear sweep. Also, as we saw earlier, the CRT deflection plates act as a capacitance load, which at the higher sweep speeds would act as a low-impedance drain on the sweep generator. Finally, the sweep sawtooth must be converted from a single-ended to a push-pull driving signal for optimum linearity of the trace.

In view of these facts, it is easy to explain why the sweep generator signal is attenuated. If it were not, consider the situation presented by the HORIZONTAL POSITIONING control. This control will be examined in detail in the following paragraphs; for the present discussion it will simply be stated that to move the trace from the full-left to full-right position, a change of more than 240 volts is required across the deflection plates of a CRT (assuming a typical deflection factor of 24 V/div). Now, if a 150-V sweep generator signal were *not* attenuated before application to the push-pull amplifier, the amplifier's gain in the NORMAL mode of operation would necessarily be less than two. The sweep centering voltage would therefore be required to move through a range of about 150 volts. This 150 volts plus the sweep-generator signal amplitude totals 300 volts. This represents the range of voltage change which must be handled by the first stage of the amplifier alone. When one considers the power-supply voltages this arrangement would require and the effect of component and stray capacitance in the presence of rapid high-amplitude voltage changes, the disadvantages become obvious.

sweep-signal transmission

The time-base ramp is usually transmitted as a voltage signal; but when the sweep generator is located at a significant physical distance from the horizontal amplifier, as is often the case in dual-sweep or plug-in instruments, it is common practice to transmit the sweep generator signal in the form of current. This practice minimizes nonlinearities in the signal which might otherwise be introduced by stray capacitance in the transmission path.

impedance matching

Since the run-up circuits of the sweep generator (those which develop the time-base ramp) depend in principle on charging a particular value of capacitance through a known resistance, it is of prime importance that the horizontal amplifier not alter the overall impedance characteristics of the run-up circuit when coupled to the sweep generator. This requires careful attention to the input impedance of the attenuating circuits. Generally speaking, the input impedance offered by the attenuation circuits is very high.

To maintain a *constant* input impedance at all sweep speeds, the attenuator must be frequency-compensated according to the same principles that apply to attenuating circuits in the vertical amplifier.*

---

*See Tektronix Circuit-Concepts series, *Vertical Amplifiers*.

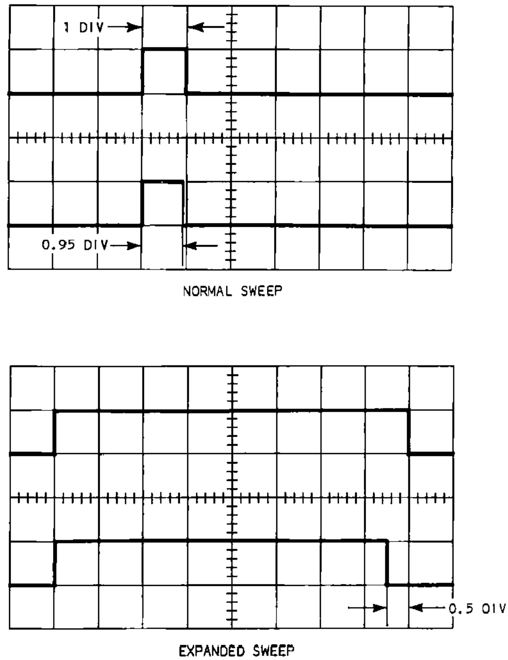


Fig. 1-10. Normal versus expanded sweeps.

Two other features of the horizontal amplifier remain to be considered. Neither is found in all sweep deflection systems, but each serves to extend the flexibility and versatility of an oscilloscope. The first of these is the *sweep magnifier*.

sweep magnifier

It is axiomatic that, other factors being equal, greater precision in time measurements can be attained if the horizontal dimensions of a displayed waveform are increased. This, of course, is the reason why high sweep speeds facilitate the measurement of shorter time intervals. If we accept the proposition that an upper limit to sweep speed is imposed by the characteristics of the sweep-generator circuits, it is clear that we must find another method to increase the horizontal dimensions of the signal under measurement. One rather obvious method is to increase the gain of the horizontal amplifier. For example, assume that at the sweep generator's fastest setting, two signals differ by only 0.05 of a horizontal division on the graticule (Fig. 1-10). Accurate measurement of such a small difference would be impossible with normal vision. If, however, the gain of the horizontal amplifier is increased tenfold, the signals will differ by 0.5 divisions and the difference can be measured with reasonable accuracy. Of course, the horizontal trace is now *theoretically* 100-divisions long, while only 10 divisions can appear on the graticule at one time. However, by adjusting the HORIZONTAL POSITIONING control, the desired portion of the trace can be brought to the center of the graticule for observation and measurement. The value of each graticule division is found by dividing the sweep time-per-division setting by the sweep magnifier setting.

external horizontal signal

A number of instruments are equipped to accept an *external* horizontal-deflection signal. This may be an external time-base signal, in which case the oscilloscope displays the familiar *Y-T* plot (signal amplitude versus time). In other applications the external signal is a dependent variable, and the display is referred to as an *X-Y* plot. A good example of this application is the generation of Lissajous figures for phase and frequency comparisons. To present a useful and meaningful display, this type of oscilloscope must preserve the relative amplitudes and phases of the signals under observation. This requires that the horizontal amplifier exhibit the same phase, frequency and delay characteristics over a specified bandwidth as does the vertical amplifier. Unfortunately these demands often conflict with the primary task of the horizontal amplifier -- processing the time-base ramp. Therefore, the upper frequency at which *X-Y* signals can usefully be displayed is usually established by the extent to which compromises can be made in the horizontal-amplifier design. However, oscilloscopes especially designed for *X-Y* mode applications feature matching amplifiers, equivalent in most respects to medium-bandwidth vertical amplifiers.

external signal preamplifier

One technique often employed to extend the range of useful X-Y performance is to add a preamplifier to the horizontal-amplifier section. Only *external* horizontal inputs are processed by the preamplifier. The output of the preamplifier is applied to the horizontal amplifier proper, where it is subject to the regular horizontal controls and processes. The added gain provided by the preamplifier brings its sensitivity closer to that of the vertical amplifier. Phasing controls may also be included to guarantee phase coincidence at the CRT deflection plates.

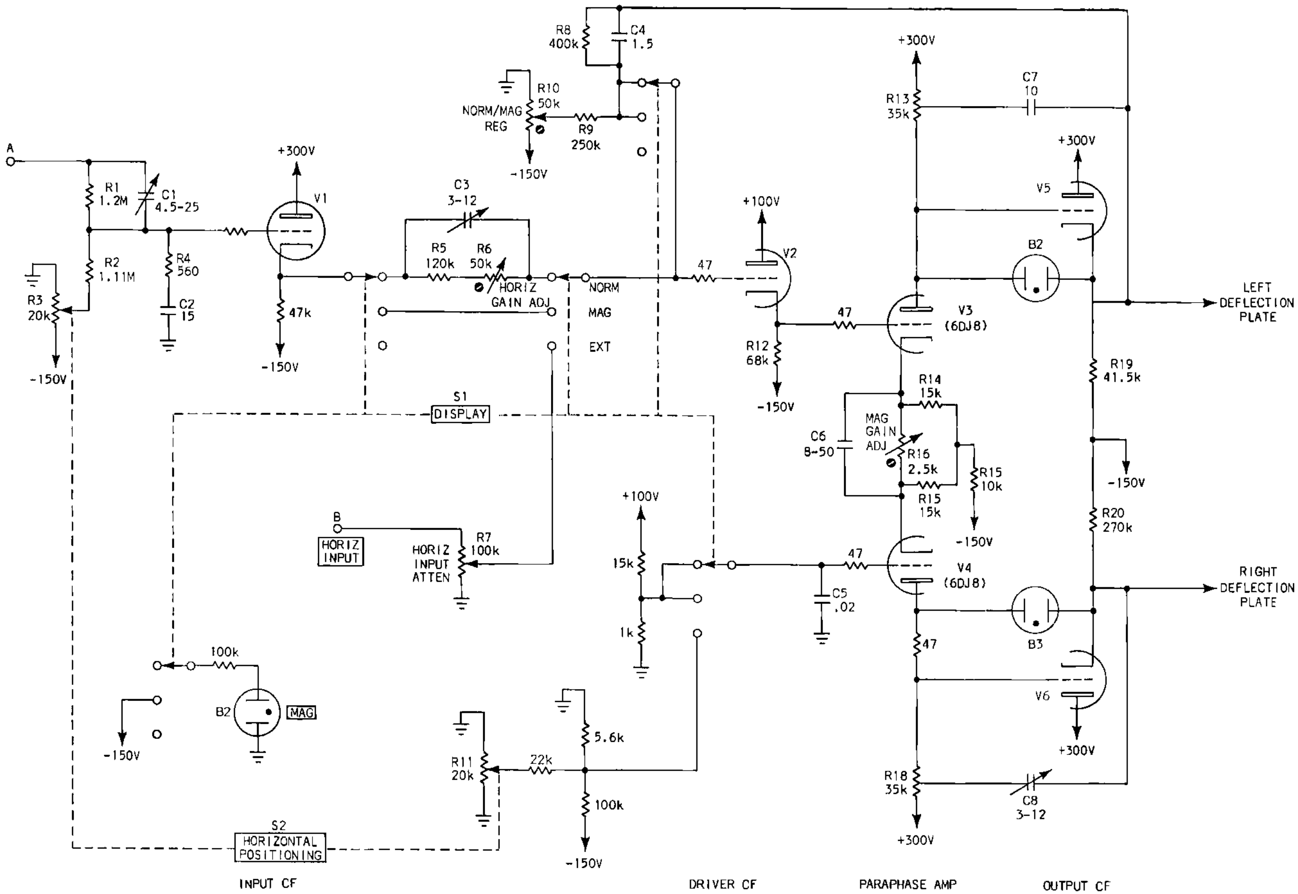


Fig. 2-1. Basic horizontal amplifier – schematic diagram.

2

HORIZONTAL AMPLIFIER

In this chapter we will examine in detail the operation, configuration and characteristics of a basic horizontal amplifier. Although simple in design, this amplifier carries out all the functions discussed in the previous chapter. A schematic diagram of the circuit is shown in Fig. 2-1.

method of analysis

It will be helpful at this time to adopt a systematic approach to our investigation into the behavior and characteristics of horizontal-amplifier circuits. Our first task, of course, is to examine the schematic diagram to determine what functional relationships exist between the various circuits and components which make up the whole. Next, we should make a block diagram of the amplifier to emphasize its basic operational characteristics and to facilitate calculation of the amplifier's gain. By comparing the results of these calculations against the input and CRT deflection factors, we can check our interpretation of the circuit's operation. Finally, we should return to the schematic diagram for a point-to-point examination of circuit details, noting similarities and differences between the circuit in question and those previously explored.

Let us begin with a preliminary examination of the schematic diagram (Fig. 2-1).

input circuits

The sweep generator (time-base) signal is applied to the amplifier at terminal A, where it is attenuated by voltage divider R1, R2 and R3 to about one-half its original amplitude. The attenuated signal drives the grid of Input Cathode-Follower V1. The cathode of V1 is directly coupled to one set of DISPLAY switch (S1) contacts. This switch has three positions. In the NORMAL position, the input signal is coupled through R5 and R6 to the grid of Driver Cathode-Follower V2. In the (X5) MAGNIFIED position, the signal is directly coupled to the same grid, bypassing R5 and R6. In the EXTERNAL position, the time-base signal is disconnected from the amplifier; in this mode, the amplifier accepts only the signal at terminal B, the HORIZ INPUT terminal.

feedback

In addition to the various input signals, the grid of V2 also receives a feedback signal from the amplifier's output in the NORMAL and MAGNIFIED modes of operation. In the EXTERNAL mode, the feedback circuit is disconnected.

paraphase amplifier

The cathode signal of V2 is applied to the grid of V3 which, together with V4 and their common-cathode circuit, makes up a paraphase amplifier. The grid of V4 is held at a fixed potential by one of two voltage-divider circuits, depending on the mode of operation.

output CF

The amplified plate signals of V3 and V4 are applied to the grids of Output Cathode-Followers V5 and V6 respectively. No common cathode connection exists between these tubes; however, since they are linked through the capacitance of the CRT deflection plates, they can be regarded as constituting a push-pull amplifier.

The terminology used in the schematic diagram to identify the stages of the horizontal amplifier is typical of that employed in Tektronix instrument instruction manuals. In labeling our block diagram however, we will employ the same terminology developed in Tektronix Circuit Concepts *Vertical Amplifiers*. To avoid confusion, the reader should frequently consult both the schematic and block diagrams to verify his identification of the circuit under discussion.

NORMAL mode

Fig. 2-2 is a block diagram of the same horizontal amplifier with the DISPLAY switch positioned for the NORMAL mode of operation. In this mode, the feedback circuit is connected, and resistor R5 plus the series resistance of HORIZ-GAIN-ADJ potentiometer R6 are switched into the grid circuit of V2. Thus we have a cathode follower driving a negative-feedback paraphase amplifier.

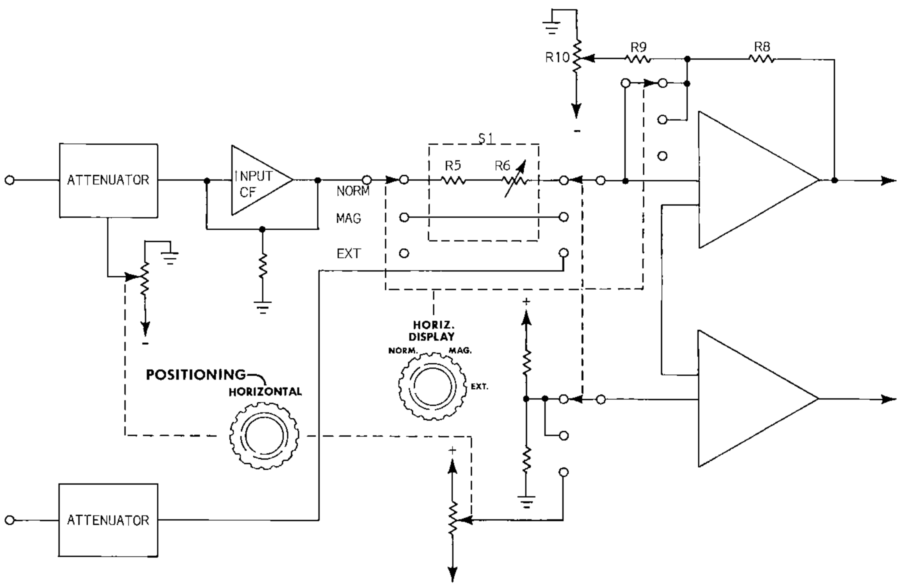


Fig. 2-2. Basic horizontal-amplifier block diagram – NORMAL mode.

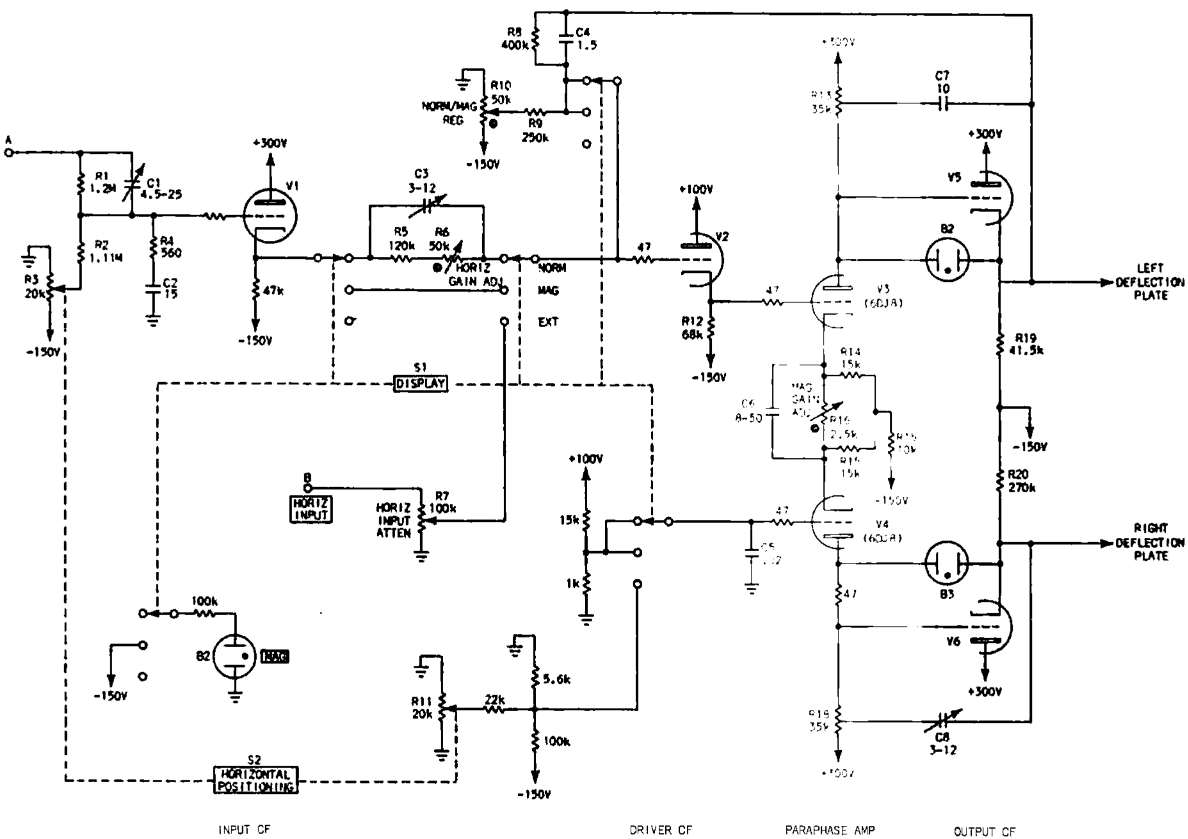

A
R1
1.2M
C1
4.5-25
R2
1.11M
R3
20k
-150V
R4
560
C2
15
+300V
V1
47k
-150V
C3
3-12
R5
120k
R6
50k
HORIZ
GAIN ADJ
NORM
MAG
EXT
S1
DISPLAY
R8
400k
C4
1.5
R10
50k
NORM/MAG
REG
-150V
R9
250k
+100V
47
V2
R12
68k
-150V
+300V
R13
35k
C7
10
+300V
V5
B2
V3
(6DJ8)
R14
15k
C6
8-50
MAG
GAIN
ADJ
R16
2.5k
R15
10k
R17
15k
-150V
V4
(6DJ8)
C5
47
B3
V6
+300V
R18
35k
C8
3-12
+100V
LEFT
DEFLECTION
PLATE
R19
41.5k
-150V
R20
270k
RIGHT
DEFLECTION
PLATE
B
HORIZ
INPUT
R7
100k
HORIZ
INPUT
ATTEN
+100V
15k
1k
100k
B2
MAG
-150V
R11
20k
22k
5.6k
100k
-150V
S2
HORIZONTAL
POSITIONING
INPUT CF
DRIVER CF
PARAPHASE AMP
OUTPUT CF

Fig. 2-3.

The attenuated time-base ramp is applied, together with a DC positioning voltage, to the input cathode follower whose output is then coupled through resistors R5 and R6 to the negative-feedback paraphase amplifier. Here the signal is converted to push-pull and amplified to the dimensions required by the horizontal-deflection factor of the CRT.

determining open-loop gain

Our first step in determining the NORMAL gain of the amplifier is to check the schematic diagram for an estimate of the amplifier's open-loop gain (Fig. 2-3). It is clear that only V3 and V4 have better than unity gain and therefore must provide the entire open-loop gain of the circuit. At 35 kΩ, R13 and R18 considerably exceed the probable value of internal plate resistance (r_p) in V3 and V4. We therefore cannot make our usual assumption that internal cathode resistance (r_k) is approximately 200 Ω, but must consult a tube manual to find the value of μ and r_p for these tubes (in this case 6DJ8's). The value of r_k can then be calculated from the equation:

$$r_k = \frac{r_p + R_p}{\mu + 1}$$

which takes into account the "reflected" impedance of R_p (R13 and R18) in the cathode circuits. With the plate and bias voltages shown in the schematic, V3 and V4 will exhibit a μ of about 30 and an r_p of approximately 4 kΩ. Thus for each of the tubes,

$$r_k = \frac{4 + 35}{31} = 1.25 \text{ k}\Omega.$$

Gain of this stage, with R16 at design center, is then:

$$A_V = \frac{2R_p}{2r_k + R_k} = \frac{70}{2.5 + 1.25} = \frac{70}{3.75} = \mathit{18.6}$$

Some cathode-follower degeneration will take place in V2, V3, and V4; so using a figure of 0.9 for the combined effect, open-loop gain appears to be about *16.8* for the amplifier as a whole.

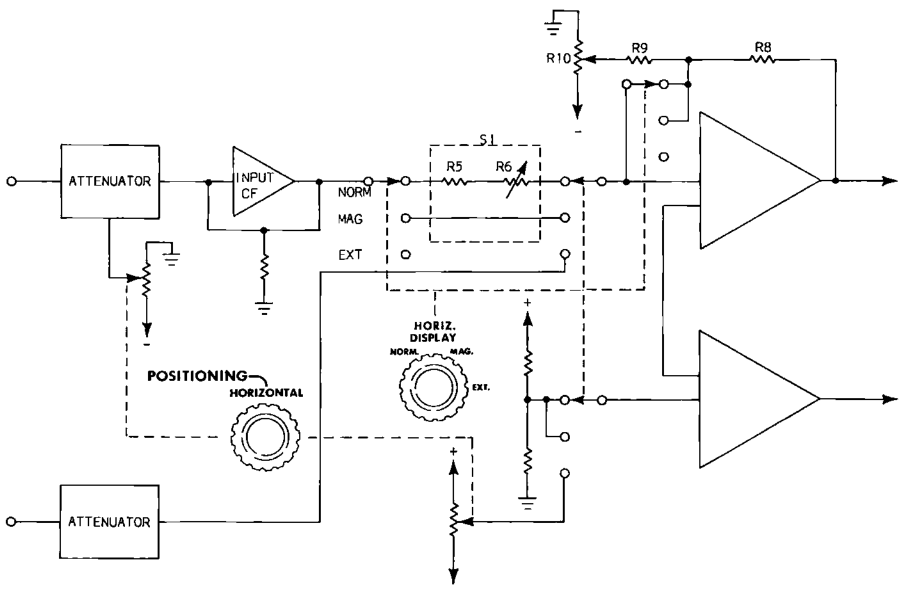


Fig. 2-4. Basic horizontal-amplifier block diagram – NORMAL mode.

closed-loop gain

The problem of calculating closed-loop gain is more complicated than it might appear at first glance. With a high open-loop gain to work with, we could ignore R9 and potentiometer R10 (Fig. 2-4), since they are connected to what would then be the null-point of the amplifier. However, the grid of V2 will respond by the amount $\frac{E_o}{1 - A_{o\ell}}$ to the input signal so a significant portion of the signal current flows through R9 and R10 and must be taken into account.

It will be remembered that the general equation for an operational amplifier is:

$$A_V = \frac{-Z_f}{Z_i}\left(\frac{A_{o\ell}}{A_{o\ell} - 1 - \frac{Z_f}{Z_i}}\right)$$

When modified to include the effect of an impedance in shunt with the negative input, and considering only the resistive impedance of the circuit, the equation becomes:

$$A_V = \frac{-R_f}{R_a}\left(\frac{A_{o\ell}}{A_{o\ell} - 1 - \frac{R_f}{R_a}\left(\frac{R_a + R_s}{R_s}\right)}\right)$$

where R_s = shunt resistance across the negative input.

Since feedback is applied only from one side of the amplifier, our open-loop-gain figure must be halved; also, it must be remembered that $A_{o\ell}$ is a negative number since the output is 180° out of phase with the input signal. Our calculation (all potentiometers at design center) thus becomes:

$$A_V = \frac{-\text{R8}}{\text{R5} + \frac{\text{R6}}{2}}\left(\frac{-8.4}{-8.4 - 1 - \left(\frac{\text{R8}}{\text{R5} + \frac{\text{R6}}{2}}\right)\left(\frac{\left(\text{R5} + \frac{\text{R6}}{2}\right) + \left(\text{R9} + \frac{\text{R10}}{4}\right)}{\left(\text{R9} + \frac{\text{R10}}{4}\right)}\right)}\right)$$

$$A_V = \frac{-400}{145}\left(\frac{-8.4}{-8.4 - 1 - \frac{400}{145}\left(\frac{145 + 262}{262}\right)}\right)$$

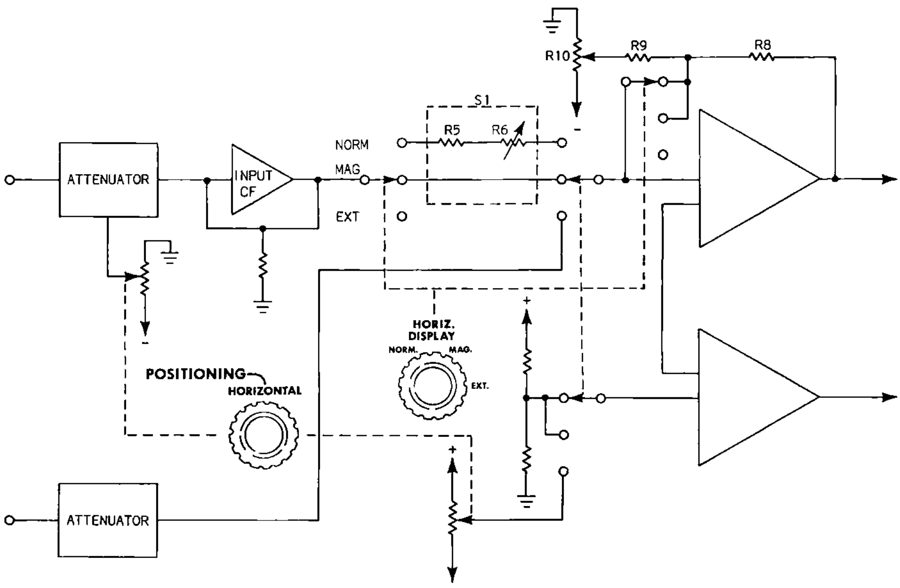


Fig. 2-5. Basic horizontal-amplifier block diagram – MAGNIFIED mode.

$$A_V = -2.76\left(\frac{-8.4}{-8.4 - 1 - (2.76 \cdot 1.55)}\right)$$

$$A_V = \frac{(-2.76)(-8.4)}{-9.4 - 4.27} = \frac{+22.3}{-13.67}$$

$$A_V = -1.63 \text{ per side (or } -3.26 \text{ push-pull)}$$

To check our interpretation of the circuit we must know the amplitude of the sweep generator signal and the deflection factor of the CRT. In the instrument for which this amplifier was designed, these values are 150 volts and 24 volts per division respectively. The sweep generator signal is reduced to 75 volts in the attenuator and again reduced to about 71 volts in the input cathode follower, yielding an input deflection factor of 7.1 V/div. To find the gain required of the amplifier we divide the CRT deflection factor by the input deflection factor, yielding the figure 3.38. Our calculated gain therefore appears to be confirmed by the amplifier's actual performance and we may assume that we have correctly analyzed the circuit's operation.

MAGNIFIED mode

Now let us examine the configuration taken by the circuit when S1 is positioned for the (X5) MAGNIFIED mode (Fig. 2-5). Resistor R5 and potentiometer R6 are bypassed, leaving only about 200 Ω (r_k of V1) as input resistance R_a. To all intents and purposes, the feedback signal is grounded so that the amplifier exhibits its full open-loop gain. When the 71-volt input signal is amplified by this factor (16.8), a theoretical deflection voltage of 1200 volts is generated. Note that at 24 V/div this represents 50 divisions of deflection, exactly what is required of the X5 MAGNIFIED sweep.

It appears, therefore, that our calculation of MAGNIFIED (open-loop) gain is also verified by the actual performance of the circuit.

EXTERNAL mode

In the EXTERNAL mode (Fig. 2-6), the feedback path is completely disconnected by the action of S1. The signal at V2 grid is therefore subjected to the full open-loop gain of the amplifier. Note that the action of S1 also disconnects the HORIZONTAL POSITIONING information applied to the grid of V1. However, potentiometer R11 is ganged with R3 and, in the EXTERNAL mode, determines the DC level of V4 grid through this same front-panel control. Potentiometer R7 acts as an uncalibrated variable attenuator for the external horizontal signal. The lack of frequency compensation at this input points up the amplifier's limited bandwidth when used in the *X-Y* mode of operation.

Norm/Mag Reg potentiometer R10 is another DC-level control. To explain its purpose we must digress a little.

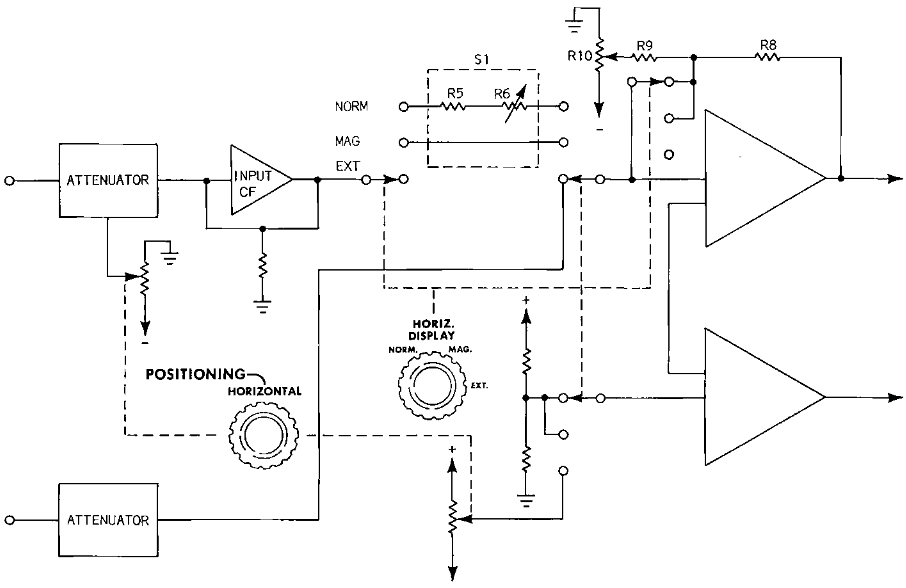


Fig. 2-6. Basic horizontal-amplifier block diagram – EXTERNAL mode.

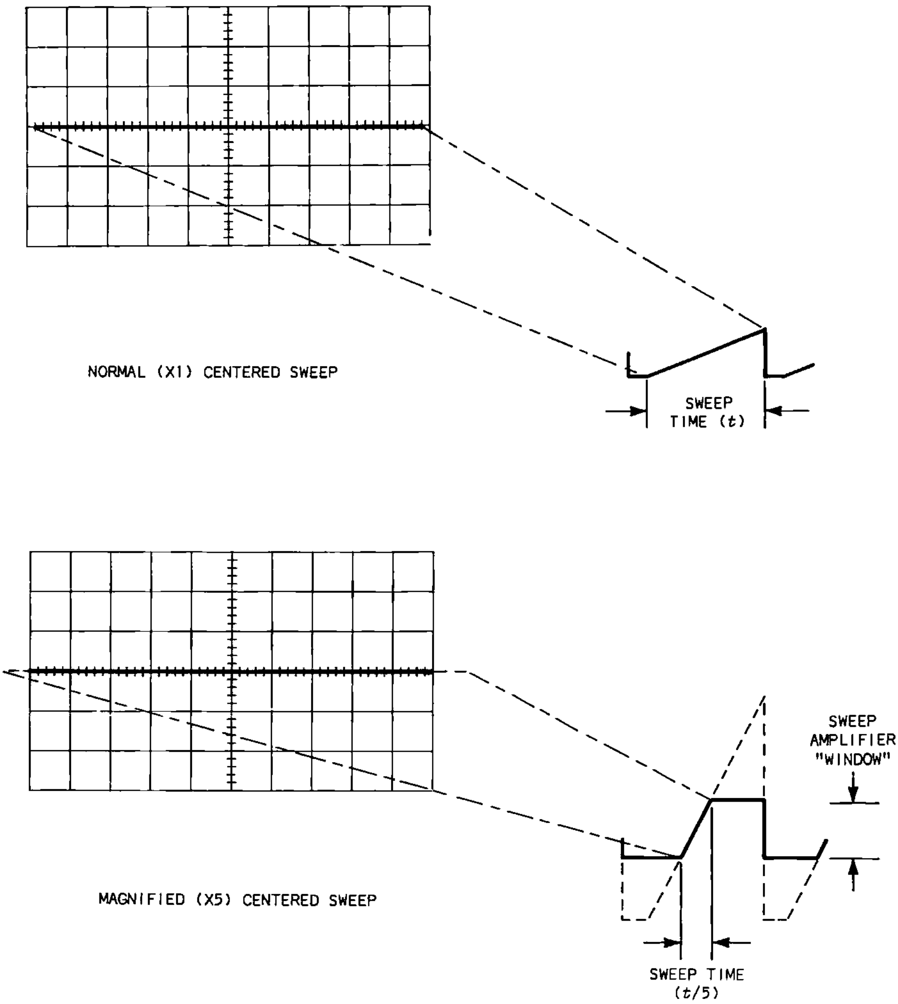


Fig. 2-7. Deflection-sweep limiting in MAGNIFIED mode.

As was stated earlier, when the amplifier is operated in the MAGNIFIED mode, or when the HORIZONTAL POSITIONING voltage is at other than its "centered" value, the CRT beam is driven off the CRT screen. In the X5 MAGNIFIED mode, for instance, there is room for only one-fifth of the trace theoretically generated by the magnified-sweep deflection voltage. In practice of course, the deflection plates do not generate such a uselessly large signal. What actually happens is that some stage of the amplifier goes into saturation or cutoff long before such a high-amplitude signal is produced. This action creates a sort of "window" at the amplifier output. The normal, centered, sweep signal appears wholly within this window. When a magnified or off-center sweep signal is generated, only that portion appearing in the window will cause CRT beam deflection. This effect is illustrated in Fig. 2-7.

sweep amplifier "window"

Since the real-time lapse between sweep trigger (sweep origin) and the appearance of a given signal at the vertical deflection plates does not change as the mode is switched, only those signals which happen to fall in the magnified portion of the normal trace would be visible.

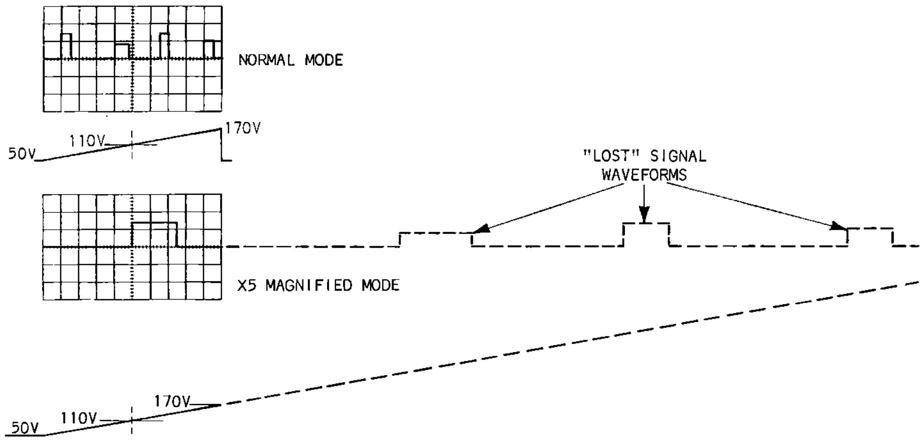


Fig. 2-8. Magnified sweep – registration at origin.

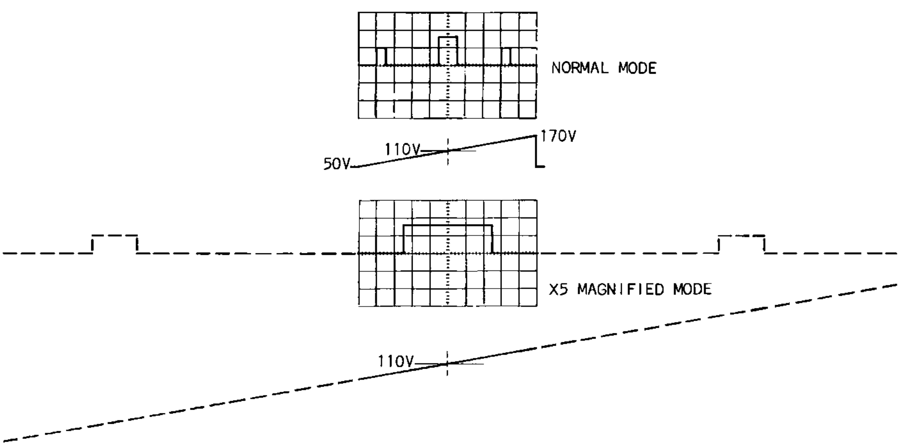
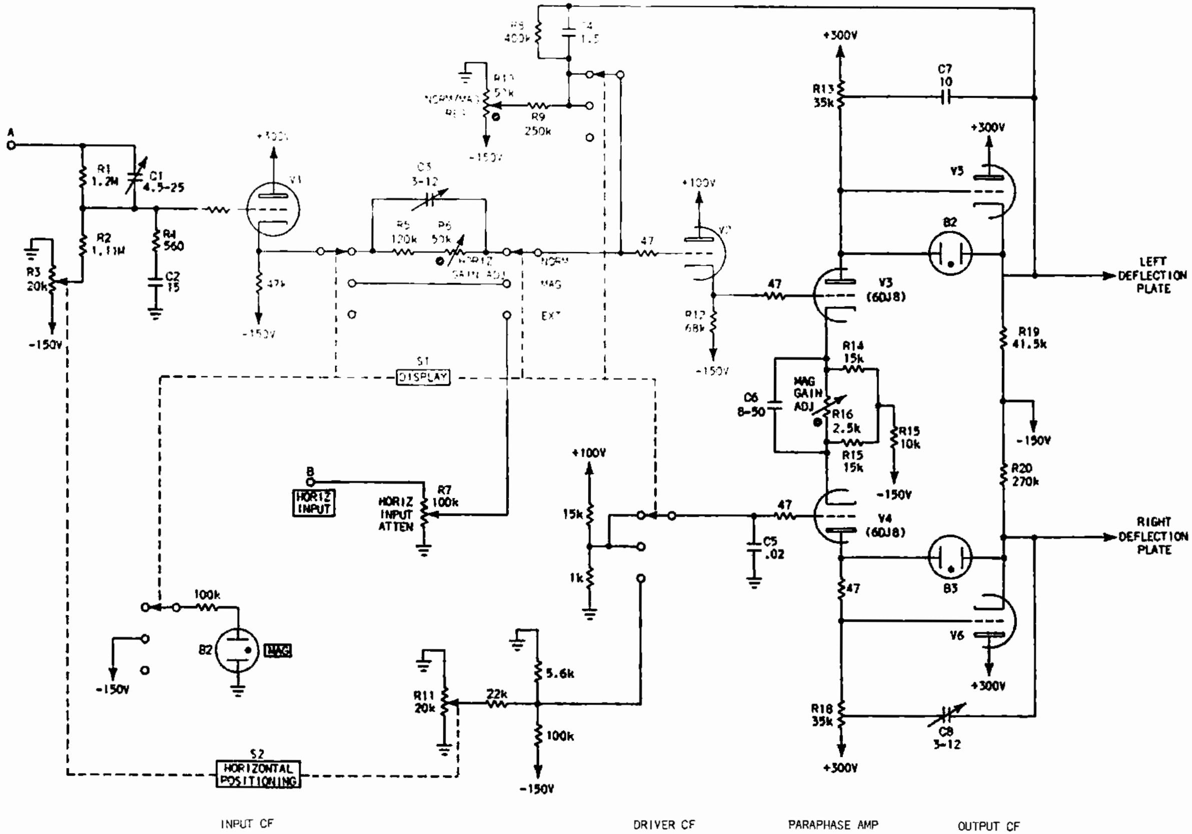


Fig. 2-9. Magnified sweep – center registration.

positioning magnified trace

Our problem here is to decide what portion of the normal trace we wish to see displayed when we switch from the NORMAL to the MAGNIFIED mode of operation. For example, if the X5 MAGNIFIED sweep were to start at the same DC level as the normal sweep, only the first two divisions of the normal trace would be displayed as a magnified trace (see Fig. 2-8). The "lost" signals, of course, could be brought back into view by adjustment of the HORIZONTAL POSITIONING control. This would lower the DC level at the grid of V1; and the time consumed by the sweep generator signal in bringing the grid back up to the trace-generating level would represent a controllable delay in the sweep. Therefore, in using the oscilloscope the operator would first adjust the HORIZONTAL POSITIONING control to bring the desired signal within the first two divisions at the left-hand edge of the graticule, then shift to the MAGNIFIED mode. However, since the central portion of the trace is the most accurate segment for taking measurements, it is more convenient to "register" the trace at the center of the CRT. To attain this condition, the DC level at the amplifier input must automatically shift to a lower level as the magnified sweep is switched into play so that the magnified ramp reaches the same voltage level at midsweep as does the normal ramp (Fig. 2-9).



A
R1 1.2M
C1 4.5-25
R2 1.11M
R3 20k
-150V
R4 560
C2 15
V1
+300V
47k
-150V
C3 3-12
R5 120k
R6 50k
HORIZ GAIN ADJ
NORM
MAG
EXT
R8 400k
C4 1.5
R10 50k
NORM/MAG REG
R9 250k
-150V
S1 DISPLAY
+100V
V2
47
R12 68k
-150V
+300V
R13 35k
C7 10
+300V
V5
B2
V3 (6DJ8)
LEFT DEFLECTION PLATE
R14 15k
MAG GAIN ADJ
C6 8-50
R16 2.5k
R15 10k
R15 15k
-150V
R19 41.5k
-150V
R20 270k
B HORIZ INPUT
HORIZ INPUT ATTEN
R7 100k
+100V
15k
1k
C5 .02
V4 (6DJ8)
RIGHT DEFLECTION PLATE
B3
47
V6
+300V
R18 35k
C8 3-12
+300V
100k
B2
MAG
-150V
R11 20k
22k
5.6k
100k
-150V
S2 HORIZONTAL POSITIONING
INPUT CF
DRIVER CF
PARAPHASE AMP
OUTPUT CF

Fig. 2-10.

Now we are ready to examine the Norm/Mag Reg adjustment (Fig. 2-10).

It has already been stated that when the CRT beam is horizontally centered on the graticule (midsweep condition) the deflection plates must be at the same potential. The grids of V3 and V4 must also be at the same potential, assuming that the feedback amplifier is properly balanced. The grid of V2 will therefore be slightly more positive than the grids of V3 and V4. This grid potential, whatever its value, we can call the "center-screen" voltage, E_{CS}. It follows that regardless of the mode of operation a signal will only appear at the center of the CRT graticule when the sum of the horizontal-positioning and sweep-sawtooth voltages are equal to E_{CS}. It should also be clear that if a signal is to remain at center screen as the DISPLAY mode is switched from NORM to MAG, the grid voltage of V2 and the cathode voltage of V1 must *both* be equal to E_{CS} at midsweep. The circuit is designed to create this set of conditions; however, to compensate for circuit and CRT variations, Norm/Mag Reg potentiometer R9 is included in the circuit. Since this adjustment affects the DC level of both modes, the display must be switched back and forth between the MAG and NORM modes as R9 is "tweaked" until no shift in position is apparent in a signal at center screen.

E_{CS} alignment

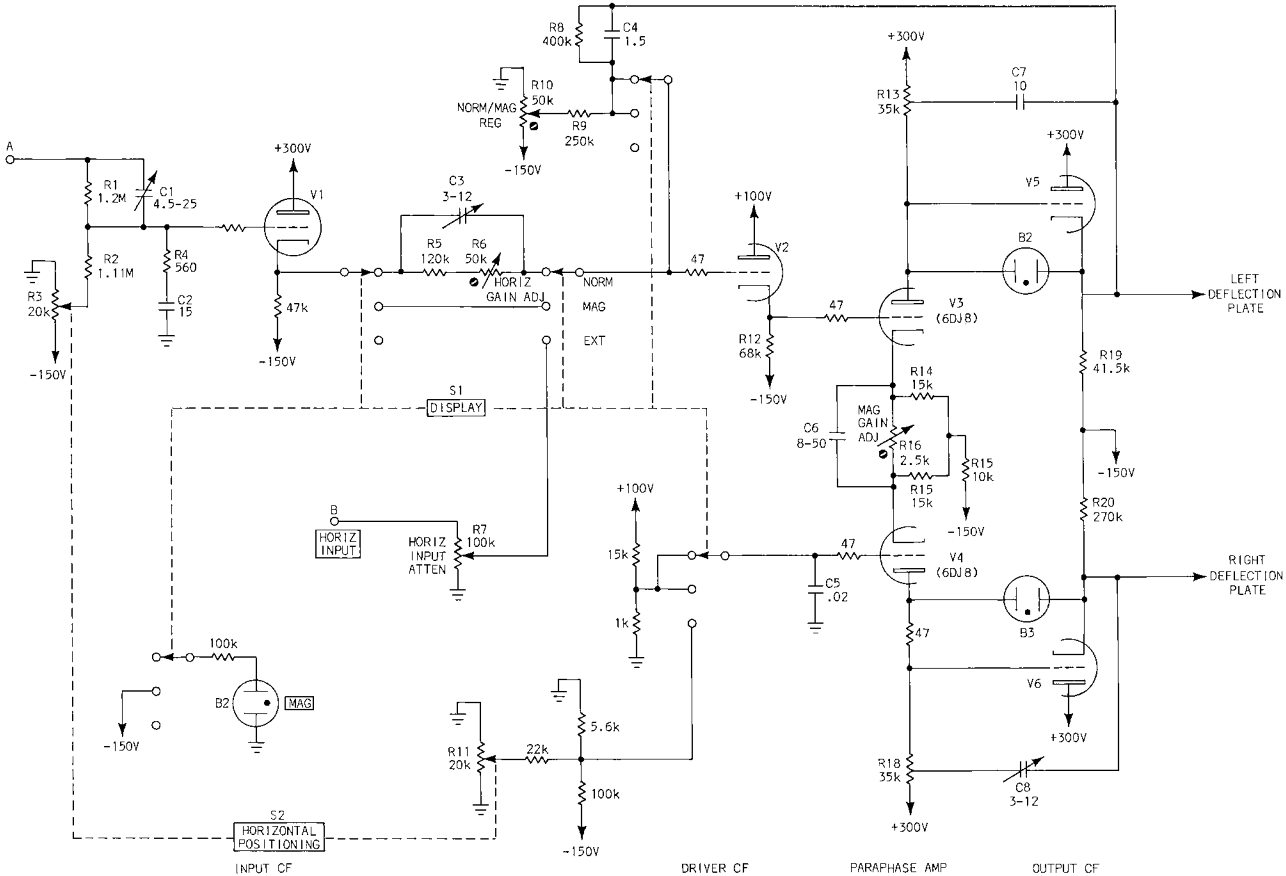


Fig. 2-11. Basic horizontal amplifier – schematic diagram.

Let us now review the schematic diagram for a few remaining details (Fig. 2-11).

As sweep speed increases, the deflection-plate driving voltage tends to fall off in amplitude and the sweep ramp tends to develop nonlinearities. The various factors which give rise to these tendencies are thoroughly examined in *Vertical Amplifiers* of the Tektronix Circuit-Concepts series and will not be discussed here. We should note, however, the presence of those circuits and devices which are installed to prevent degradation of the sweep signal at the faster sweep speeds. Our first example is found in the input attenuator. Capacitor C1 is installed to compensate for the effect of stray capacitance (and the input capacitance of V1), which can be regarded as being in parallel with R2. Its value is chosen so that the *product* of R1 and C1 is equal to the product of R2 and the input and stray capacitance. This assures that the input signal receives the same attenuation regardless of its risetime (or frequency component).

frequency compensation

offsetting negative resistance

Capacitor C2 and resistor R4 offset the negative resistance characteristic exhibited by V1 (and cathode followers in general) at high frequencies. Without this compensation the initial rise of the sweep ramp would suffer a degree of "rounding."

stabilizing amplifier gain

Capacitors C3 and C4, shunting R5, R6 and R8 respectively, assure that the feedback and input *impedances* of the feedback amplifier are maintained at a constant ratio at all sweep speeds, thereby stabilizing the amplifier gain. Capacitor C6 acts to reduce cathode degeneration at the faster sweep speeds. Bootstrap capacitors C7 and C8 also react to fast-risetime signals by supplying current from the cathode-follower output to the 7-k tap in R13 and R18. The overall effect is to increase the apparent plate impedance of these tubes, yielding increased gain at fast sweep speeds.

MAG warning

In spite of these measures, when the fastest sweep rates are used in conjunction with the MAGNIFIED mode the sweep suffers some distortion. To warn the operator that the display switch is in the MAGNIFIED position, neon tube B1 is energized by -150 volts when the instrument is switched from the NORMAL to MAGNIFIED mode of operation.

preventing damage to cold cathodes

Neon tubes B2 and B3 are installed to protect the cathodes of V5 and V6 when fast high-amplitude sweep signals are present on their grids. Under these signal conditions, the cathode voltage is unable to rise as quickly as that of the grid; a high grid-to-cathode voltage therefore tends to develop which could strip particles of electron-emitting coating from the cathode. However, B2 and B3 conduct before grid-to-cathode voltage reaches a damaging level.

NOTES

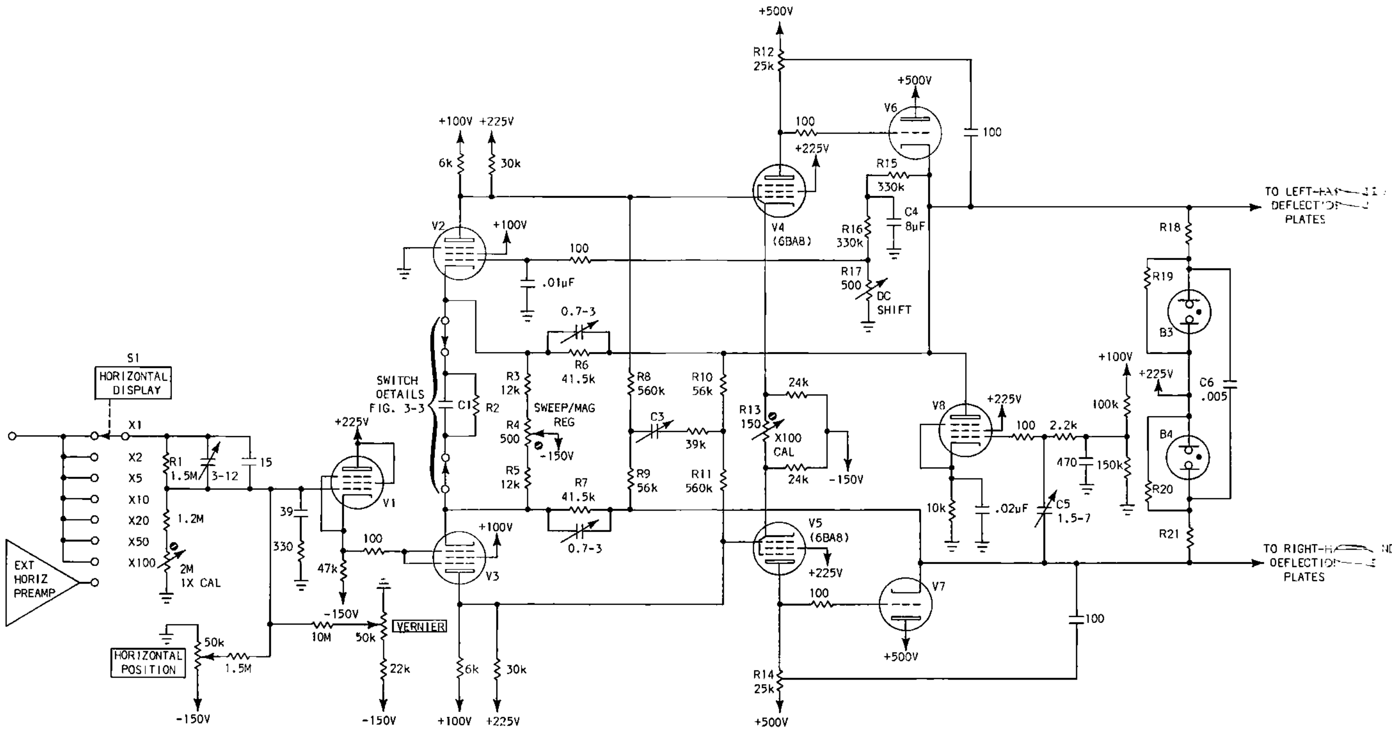


Fig. 3-1. Example-1 schematic diagram.

3

TYPICAL CIRCUIT CONFIGURATIONS

We are now ready to examine a selected cross section of horizontal amplifiers found in Tektronix instruments. As we proceed, the reader should keep in mind that our approach is a conceptual one and relies heavily on generalizations and approximations. Therefore, although the circuits discussed in this chapter come from actual production models, the material presented is not intended as a guide to instrument maintenance or repair.

EXAMPLE 1 -- VACUUM-TUBE AMPLIFIER WITH WIDE-RANGE MAGNIFICATION

HF capacitance driver

The two principal factors which contribute to greater complexity in horizontal-amplifier design are (1) demand for faster sweep speeds, and (2) solid-state component limitations. The influence of the first factor is illustrated in Fig. 3-1. This schematic closely resembles that of the basic amplifier discussed in Chapter 2; however, an additional active component, pentode V8 has been added. This tube is often labeled "HF Capacitance Driver" and is found in most sweep amplifiers of vacuum-tube construction, especially those which provide high-gain factors in the magnified modes.

A brief consideration of the dynamic conditions present in cathode-follower V6 at fast sweep speeds will reveal the purpose of the HF capacitance driver (Fig. 3-2).

deflection-plate capacitance

The deflection plates of the CRT, as stated earlier, can be regarded as constituting a low-value capacitor, C_{dp}. At slower sweep speeds (Fig. 3-2A), when the grid of V6 is driven negative by the sweep ramp, the cathode also moves in the same direction.

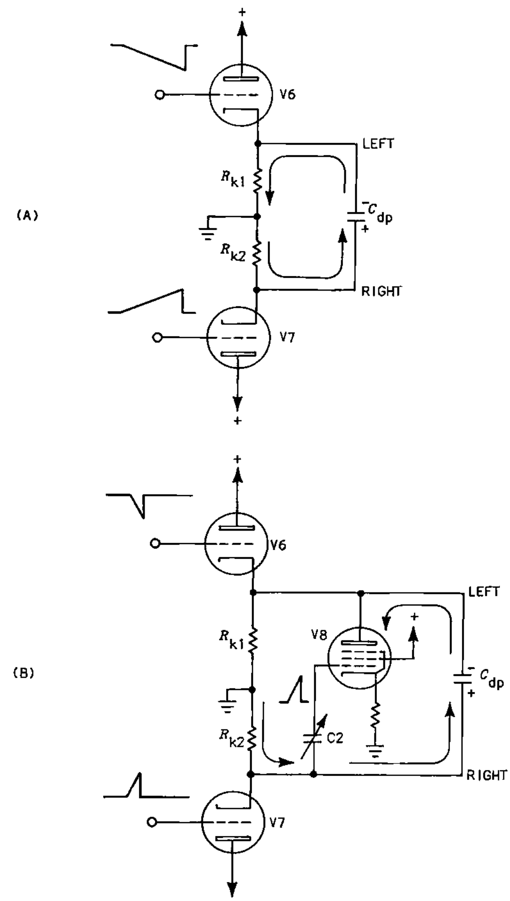


Fig. 3-2. High-frequency capacitance driver circuit.

Current to charge C_{dp} is supplied through cathode-resistor R_{k1} and V7. To generate a linear trace on the CRT this charging current must be relatively constant. No difficulty is encountered at the slower sweep speeds, since the RC time of the cathode-resistor deflection-plate capacitance circuit is short compared to the risetime of the sweep signal itself.

At higher sweep speeds, however, a problem is created in the left-hand side of the driving circuit (Fig. 3-2B). As the grid of V6 is driven rapidly negative, the cathode voltage lags behind because insufficient time has elapsed to charge C_{dp} to the voltage at the grid of V6. The tube therefore tends to go into cutoff. This has two undesirable effects: (1) the circuit no longer operates in push-pull so the charging rate is nonlinear; and (2) the deflection voltage fails to rise at the desired rate, resulting in slower beam deflection and error in sweep time-per-division measurements. HF capacitance driver, V8, eliminates this problem by providing additional charging current as sweep speed increases. It acts as a variable-resistance parallel leg in V6's cathode circuit. Under slow-sweep conditions it has little effect on the circuit as a whole, due to the very low value of capacitance represented by C2. However at fast sweep speeds, a positive-going signal is coupled to the grid of V8. This lowers the resistance of the pentode so that V6 is not driven into cutoff. Its cathode is therefore able to follow the constantly changing sweep voltage. Capacitor C2 is adjusted during calibration to a value which just offsets the risetime lag introduced by the dynamic behavior of V6. This technique makes it possible to generate relatively linear sweep deflection signals with typical amplitudes of 150 volts per side (300-V total) and extremely short risetimes.

calculating gain

In calculating the gain of this circuit, one must proceed warily. The use of pentodes in the negative-feedback amplifiers (V2, V4, V6 and V3, V5, V7) seems to indicate a high open-loop gain. On second glance, however, it will be noted that the gain of V2 and V3 is less than one. (Note that in the X1

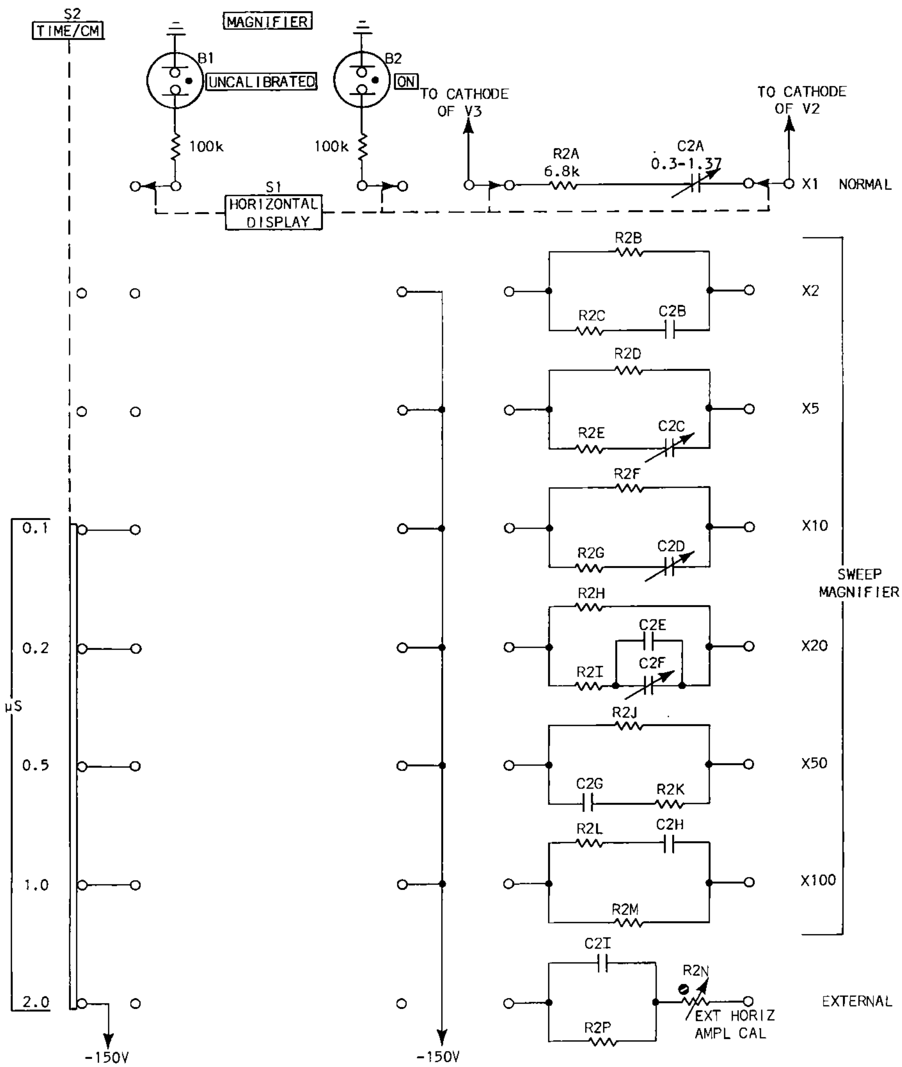


Fig. 3-3. Example-1 display-switch details.

position of the HORIZONTAL DISPLAY switch, Fig. 3-3, R2A and C2A are in series so that, except at high sweep speeds, no coupling exists between the cathode of V2 and V3. (See schematic, Fig. 3-1.) Using a value of 5 kΩ for the parallel plate-load resistors the gain calculation for the first stage becomes:

$$A_{o\ell 1} = \frac{R_{out}}{R_{in}} = \frac{5}{12.25} = 0.4$$

An analysis of the DC conditions governing the behavior of V4 and V5 will also yield some surprising results. From the schematic diagram it can be seen that the grids of V5 and V6 will quiescently ride at about 100 volts. The cathodes of these tubes will therefore show about the same potential. This places about 250 volts across the cathode resistors. Cathode current is therefore about 10 mA. Some of this current is drawn by the screen grid so that only about 7 mA flow through R12 and R14. From the tube manual it is found that with a plate current of only 7 mA, a 6BA8 tube exhibits a g_m of only 2500 micromhos. Its internal cathode resistance, r_k, is thus about 400 ohms. Gain of this second stage is therefore:

$$A_{o\ell 2} = \frac{R_{out}}{R_{in}} = \frac{25 \cdot 10^3}{\frac{75}{8} + 400} = \frac{25}{0.5} = 50$$

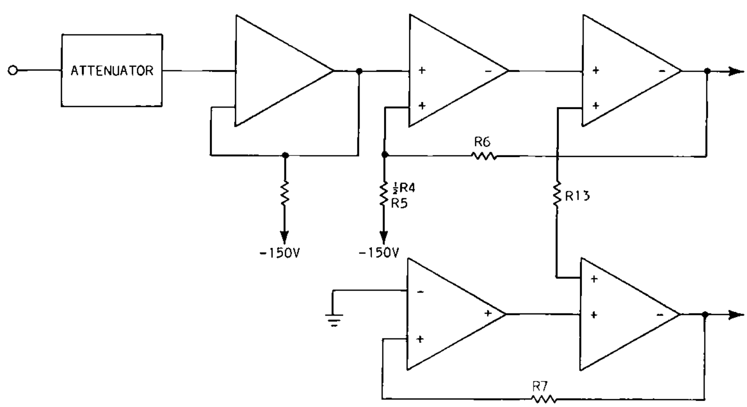


Fig. 3-4. Example-1 block diagram.

Open-loop gain of the two stages is therefore:

$$(A_{o\ell 1})(A_{o\ell 2}) = (0.4)(50) = 20$$

With this information we can draw our block diagram (Fig. 3-4) and calculate closed-loop gain for the *right* side of the amplifier using the general equation for negative-feedback amplifiers.

$$A_V = \frac{R_f - \dfrac{R_f}{A_{o\ell}+1}}{R_a + \dfrac{R_f}{A_{o\ell}+1}} = \frac{41.5 - \dfrac{41.5}{21}}{12.25 + \dfrac{41.5}{21}} = \frac{41.5 - 1.97}{12.25 + 1.97} = \frac{39.53}{14.22} = 2.8$$

After attenuation and coupling through V1, the sweep generator signal has an amplitude of about 50 volts. When amplified by the factor 2.8, this signal provides a 140-volt driving signal at the right-hand deflection plate of the CRT.

The left-hand side of the amplifier cannot be analyzed with the simple approximations at our disposal. Because no coupling takes place between V2 and V3 (in the X1 mode), the input signal for the left-hand side of the amplifier is at the cathode of V4. Negative feedback from the cathode of V6 is first dropped across R6 and then applied to the cathode of V2, which acts as a grounded-grid amplifier. The signal at V4's control grid is a negative-going sawtooth about 8 volts in amplitude. This, by itself, would tend to drive the left-hand deflection plates *positive* by approximately 480 volts. However, the signal at V4's cathode is approximately the equivalent of that present at the grid of V5.

Under closed-loop conditions the signal is a positive-going sawtooth about 10 volts in amplitude. The difference between these voltages (+2 volts), when amplified by the open-loop gain of V4, yields a negative-going deflection signal of 100 volts from the left side of the amplifier. The sum of left- and right-hand deflection voltages is thus 240 volts which, with a 24-V/div CRT deflection factor, provides ten divisions of horizontal deflection.

providing needed gain

When we consider the fact that the HORIZONTAL DISPLAY switch (S1) has a X100 position, we are faced with a problem. How can an amplifier whose *open-loop* gain is only 50 provide a closed-loop gain of 480? (This is the gain required to produce 1000 divisions of deflection at 24 volts per division from a 50-volt input signal.) The answer of course is that the open-loop gain of V2 and V3 increases enormously as the cathode resistance of V2 is lowered by the action of S1. In the X100 position, for instance, the open-loop gain of V3 is approximately 50-times greater than that exhibited in the X1 position giving the two stages a combined open-loop gain of about 750. This is more than enough to guarantee a gain of 480 in the closed-loop configuration.

Further improvement in the open-loop gain of V4 and V5 is effected by the positive-feedback resistors R8, R9 and R10, R11 which are cross-connected from the outputs of the amplifier to the grids of V4 and V5 respectively. Capacitor C3 is adjusted to guarantee that the plate impedances of V3 and V4 are properly matched.

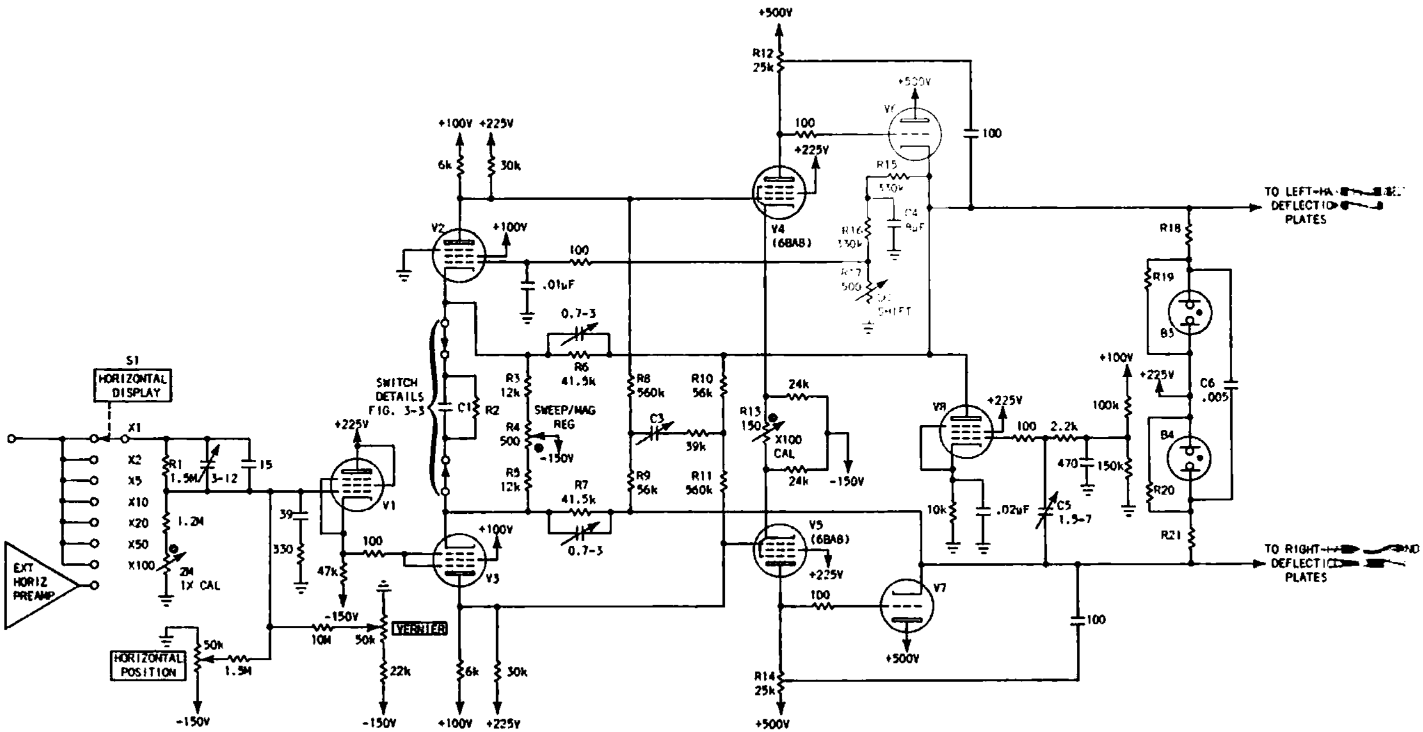

+500V
R12
25k
+500V
V6
100
100
+225V
R15
330k
C4
R16
330k
V4
(6BA8)
R17
500
+100V +225V
6k
30k
V2
+100V
100
.01µF
0.7-3
R6
41.5k
S1
HORIZONTAL
DISPLAY
X1
X2
X5
X10
X20
X50
X100
R1
1.5M
3-12
15
1.2M
2M
1X CAL
EXT
HORIZ
PREAMP
SWITCH
DETAILS
FIG. 3-3
C1
R2
R3
12k
R4
500
SWEEP/MAG
REG
-150V
R5
12k
R7
41.5k
0.7-3
R8
560k
R10
56k
C3
39k
R9
56k
R11
560k
+225V
V1
39
330
47k
100
+100V
V3
-150V
10M
50k
VERNIER
HORIZONTAL
POSITION
50k
1.5M
22k
-150V
-150V
6k
30k
+100V
+225V
R13
150
X100
CAL
24k
24k
-150V
V5
(6BA8)
+225V
100
V7
+500V
R14
25k
+500V
V8
+225V
100
2.2k
10k
.02µF
C5
1.5-7
470
150k
100k
+100V
+225V
C6
.005
R18
R19
B3
B4
R20
R21
100
TO LEFT-HAND
DEFLECTION
PLATES
TO RIGHT-HAND
DEFLECTION
PLATES

Fig. 3-5.

Another feature of interest in this amplifier is the *DC Shift* circuit, consisting of resistors R15, R16, potentiometer R17 and capacitor C4 (Fig. 3-5). The phenomenon called "DC shift" is actually a change in gain which takes place in a vacuum tube as it is shifted from AC to DC operation (Fig. 3-6). This effect is not thoroughly understood, but it is easily seen when a step voltage is applied to the grid. Because the fast rise of the step is equivalent to an AC signal, it is subject to normal AC gain. As the new level of grid voltage is sustained, however, gain falls off slightly. This causes the DC level at the plate to fall off. The new level represents the DC gain of the tube. The overall result of this shift on the horizontal amplifier would be a difference in gain between fast and slow sweeps.

DC-shift circuit

This effect is eliminated by the action of the DC Shift circuit. At the slowest sweep speeds a small positive feedback is applied to the control grid of V2 from the junction between R16 and R17. As sweep speed increases, more and more of the positive feedback signal is shunted to ground through the decreasing impedance of C4. When R17 is properly adjusted the circuit reduces gain just enough to offset DC shift as sweep speed increases.

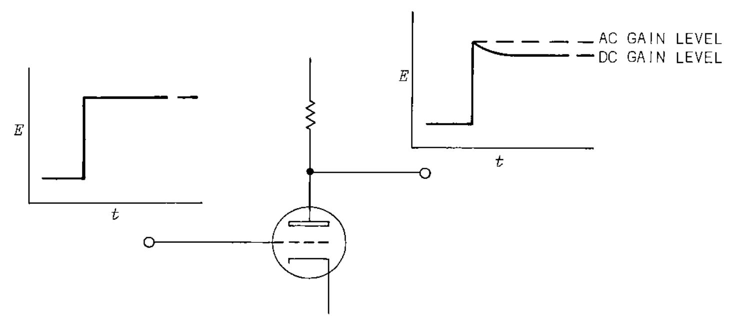
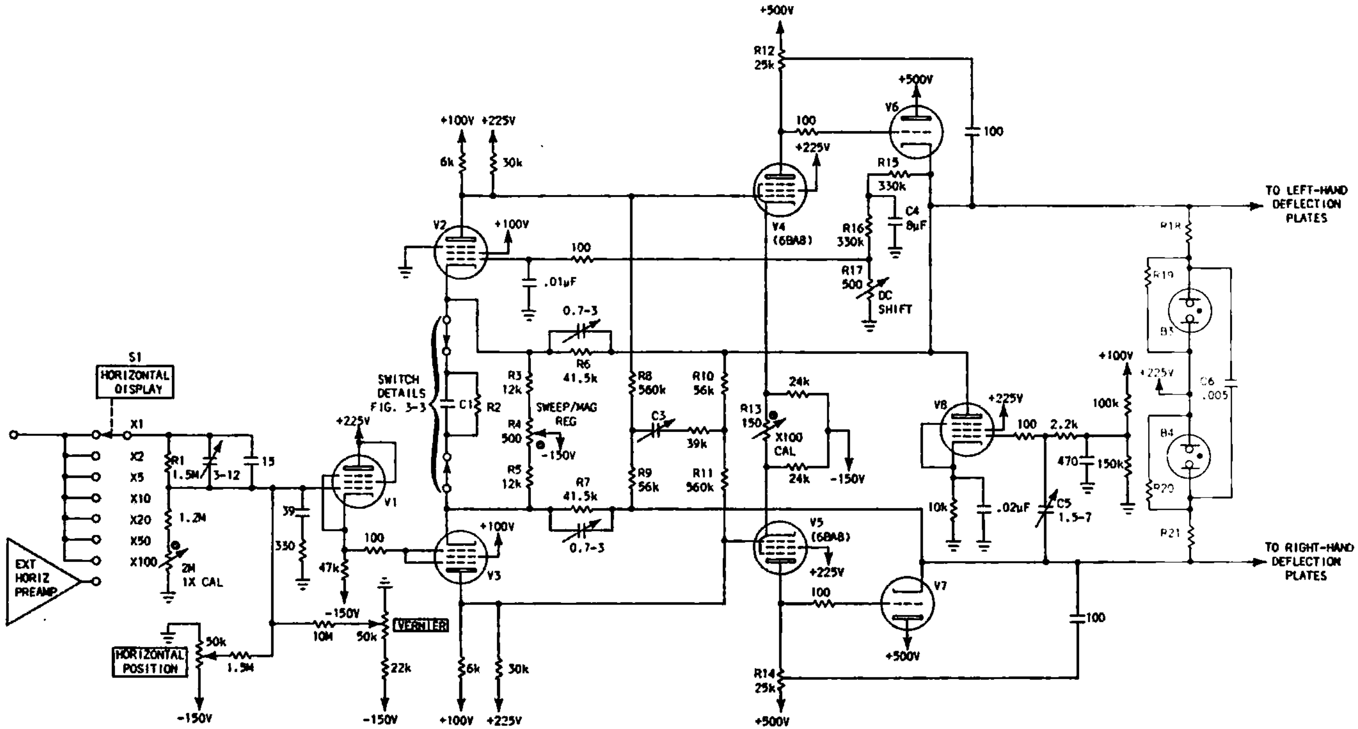


Fig. 3-6. DC-shift effect.



+500V
R12
25k
+500V
V6
100
100
+225V
R15
330k
TO LEFT-HAND
DEFLECTION
PLATES
+100V
+225V
6k
30k
V4
(6BA8)
R16
330k
C4
8µF
R17
500
DC
SHIFT
V2
+100V
100
.01µF
0.7-3
R6
41.5k
S1
HORIZONTAL
DISPLAY
SWITCH
DETAILS
FIG. 3-3
C1
R2
R3
12k
R4
500
SWEEP/MAG
REG
-150V
R5
12k
R8
560k
R10
56k
C3
39k
R9
56k
R11
560k
R13
150
24k
X100
CAL
24k
-150V
V8
+225V
+100V
100k
100
2.2k
470
150k
10k
.02µF
C5
1.5-7
C6
.005
R18
R19
B3
B4
R20
R21
TO RIGHT-HAND
DEFLECTION
PLATES
X1
X2
X5
X10
X20
X50
X100
R1
1.5M
3-12
15
1.2M
2M
1X CAL
EXT
HORIZ
PREAMP
+225V
V1
39
330
100
47k
R7
41.5k
0.7-3
V3
+100V
V5
(6BA8)
+225V
100
V7
+500V
100
R14
25k
+500V
-150V
10M
50k
VERNIER
22k
HORIZONTAL
POSITION
50k
1.5M
-150V
-150V
6k
30k
+100V
+225V

Fig. 3-7.

lights indicate location of beam

A convenience feature is provided by neon tubes B3 and B4 in the output of the amplifier (Fig. 3-7). One side of each tube is connected to 225 volts. The other sides are coupled through R18 and R21 to the deflection-plate leads. The tubes are also paralleled by R19 and R20. At midsweep each deflection plate receives approximately +300 volts and current through both sides of the output circuit is the same. The voltage drop across R19 at this time is $\frac{R19}{R18 + R19}(300 - 225) = 0.8(75) = 60$ V. (The same drop, of course, exists across R20.) This is slightly less than the ignition voltage of the neons, so neither will light up. However, if the beam moves to the left, the drop across R19 increases while the drop across R20 decreases. At the full-left screen position, the drop across R19 is 0.8(360 - 225) or 107 volts. This voltage is more than enough to cause B3 to fire. By the same reasoning, at the full-right position B4 lights up and B3 is extinguished. These indicator lights, located on the front panel of the oscilloscope, assist the operator in locating the beam when the trace itself does not appear on the CRT screen. Notice that capacitor C6 shunts both neon tubes, and as sweep speed increases, soon shorts them out of the circuit. This prevents annoying flicker of the tubes when the sweep is actually in operation.

UNCALIBRATED indicator lamp

The details of HORIZONTAL DISPLAY switch (S1) are shown again in Fig. 3-8. Included in the diagram is one wafer of the TIME/CM switch, This front-panel switch controls the risetime of the sweep generator signal. The two switches act together to light the UNCALIBRATED indicator lamp, B1, whenever the *equivalent* sweep speed is equal to or faster than 0.01 μs/cm. As shown in the diagram, the wiper of the TIME/CM switch is in the 0.1 μs/cm position and moves *down* as sweep-time/cm gets greater (sweep speed gets slower). For example, in the X1, X2 and X5 positions of the display switch B1 remains off. As soon as this switch is turned to the X10 position, a connection is made between the -150 V bus and the lamp, turning it on. At this time, and in any of the higher magnification modes, the combination of sweep speed and magnification factor produces an *equivalent* sweep-time/cm of 0.01 μs/cm or less. This tells the operator that his measurements no longer are calibrated. Now, if the TIME/CM switch is moved to the 0.2-μs/cm position (one step lower) the indicator will go out and will not come on again until sweep magnification is increased to X20 or more. Here again, sweep-time per centimeter divided by magnification factor (0.2/20) yields a sweep speed equal to or faster than 0.01 μs/cm.

EXTERNAL-mode gain

Resistor R2P, Ext Horiz Ampl Cal potentiometer R2N and the frequency-compensating capacitor C21 in the last (EXTERNAL) position of HORIZONTAL DISPLAY switch are inserted into the common-cathode circuit of V2 and V3. Gain of the horizontal amplifier in the EXTERNAL mode will thus fall somewhere between that of the X5 and X10 modes since the resistance of R2N and R2M is 3.25 kΩ at design center.

Explanation of the functions of the various adjustments, frequency-compensation circuits, bootstrapping circuits, etc. were made in the previous chapter and will not be repeated here. The external horizontal preamplifier will be examined in the last chapter.

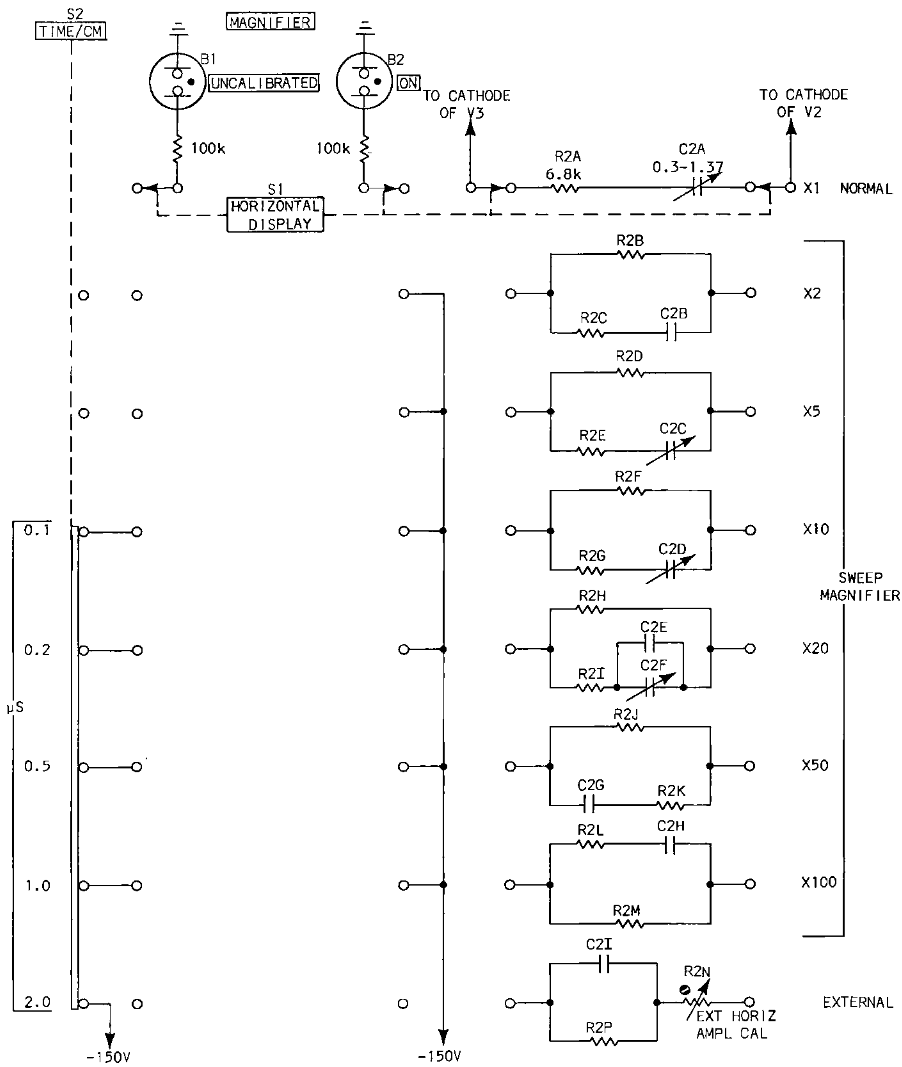


Fig. 3-8. Example-1 display-switch details.

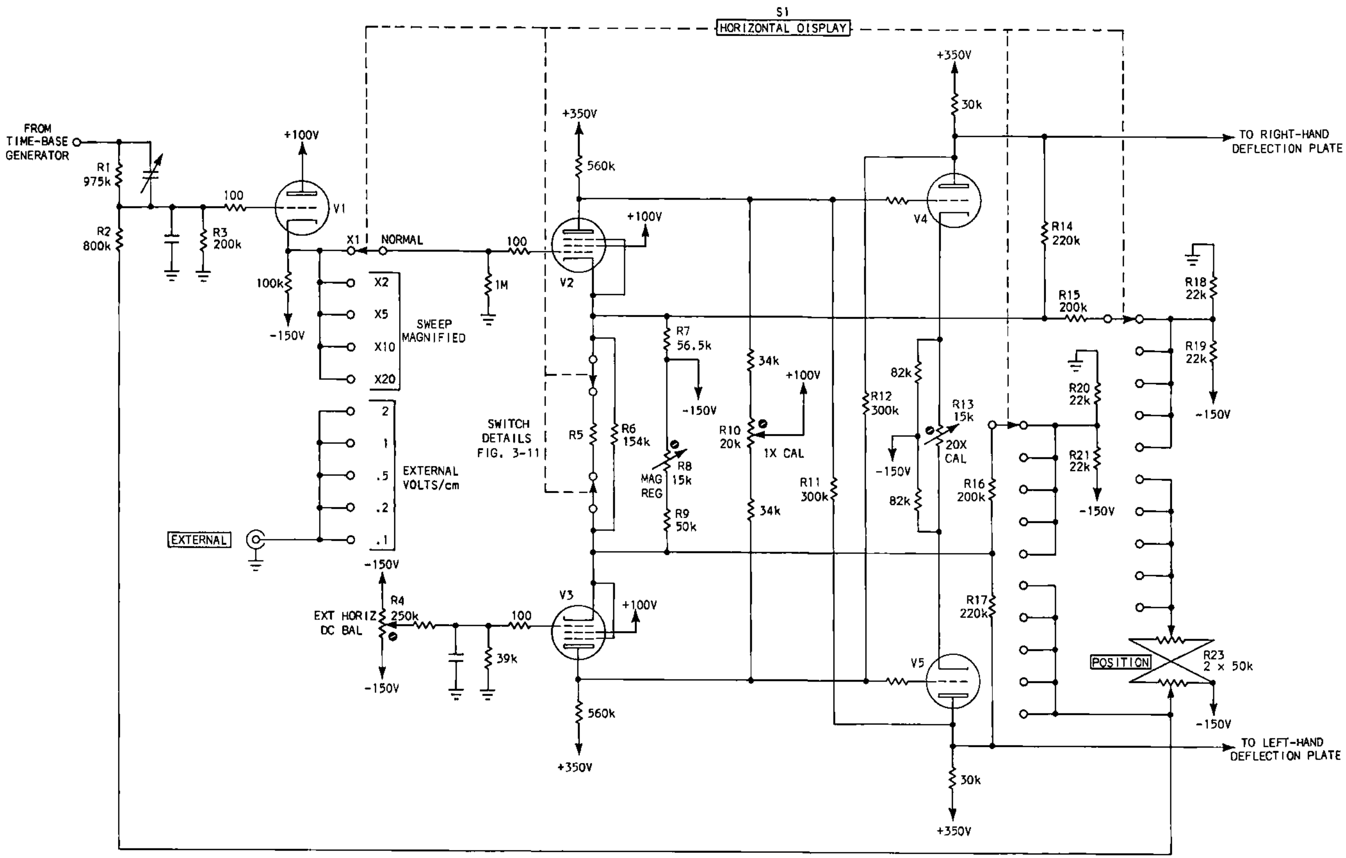


Fig. 3-9. Example-2 schematic diagram.

EXAMPLE 2 -- VACUUM-TUBE AMPLIFIER

Another interesting variation on the basic sweep-amplifier configuration is shown in Fig. 3-9. At first glance it does not seem to differ significantly from those already discussed. A close examination, however, will soon uncover circuit details whose purpose and function require additional consideration. First let us establish the general operating characteristics of the amplifier as an exercise in circuit analysis.

A block diagram of the sweep amplifier is shown in Fig. 3-10. It consists essentially of an attenuator, a cathode follower which isolates the sweep generator and the attenuator from the contact capacitance of HORIZONTAL DISPLAY switch S1, and a negative-feedback paraphase amplifier. To find the values of R_f and R_a we must turn to the schematic diagram. We see that feedback from the plates of triodes V4 and V5 is coupled to the cathodes of pentodes V3 and V2 through 220-kΩ resistors R14 and R17. Connecting these cathodes are four parallel resistive circuits: (1) R5, whose value depends on

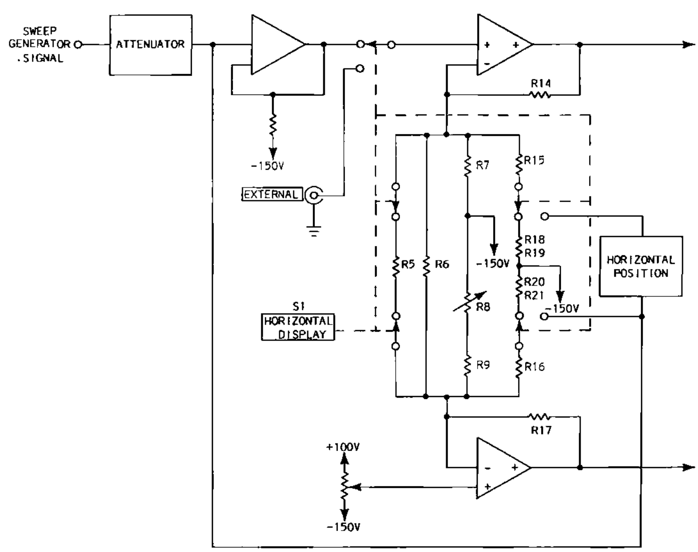


Fig. 3-10. Example-2 block diagram.

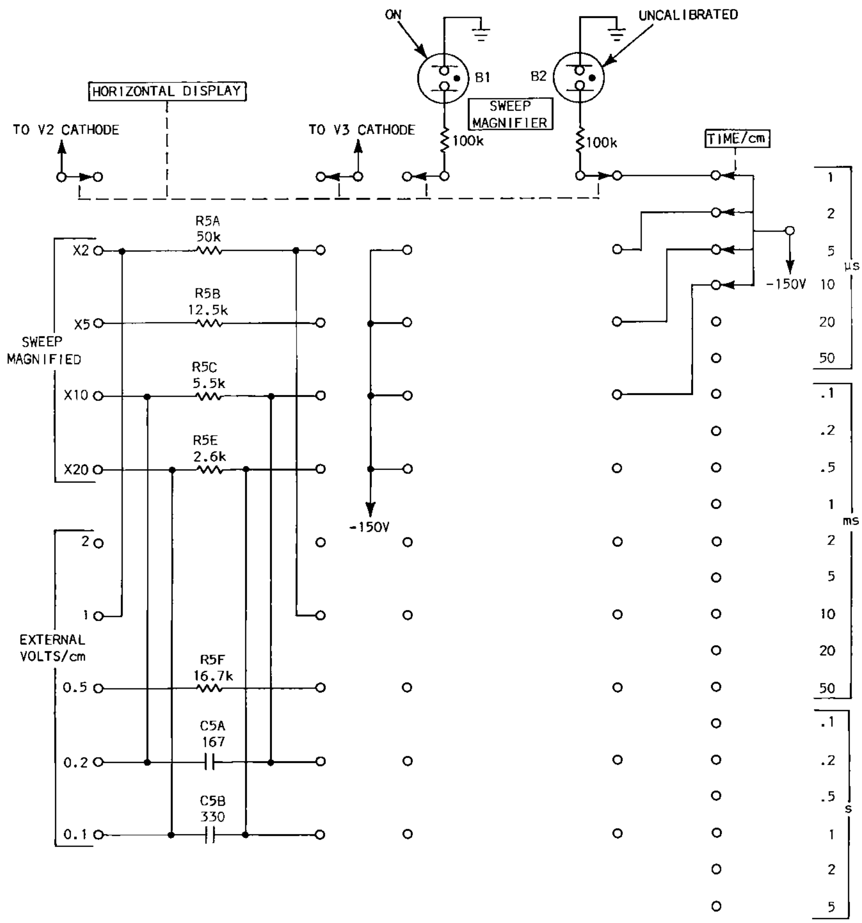


Fig. 3-11. Example-2 display-switch details.

the position of the HORIZONTAL DISPLAY switch, (2) resistor R6, (3) the series combination R7, R9 and MAG REG potentiometer R8 and (4) the series combination R15 and R16 plus the parallel values of resistors R18, R19 and R20 and R21. Gain again must be calculated separately for each side of the amplifier due to the inherent imbalance of the paraphase stage. In the X1 mode R5 equals infinity (see Fig. 3-11); that is, no resistor is inserted between the cathode contacts of S1. The value of R_a with all potentiometers at design center is thus the series equivalent of 154 kΩ, 114 kΩ and 452 kΩ in parallel, or 57 kΩ. Substituting values, the gain calculation for the right side is:

calculating gain

$$A_V = \frac{R_f + R_a}{R_a} = \frac{220 + 57}{57} \approx 4.8$$

and for the left side:

$$A_V = \frac{R_f}{R_a} = \frac{220}{57} \approx 3.8$$

for a total gain of approximately 8.6.

A 150-V sweep generator signal attenuated by the factor $\frac{R1}{R2 + R3}$ or 0.14, then amplified by a factor of 8.6, will produce a push-pull deflection drive of 180 volts or 18 V/div. These figures match the actual values of sweep generator amplitude and CRT deflection factor in the oscilloscope from which this example was taken. Our analysis of the circuit therefore appears to be confirmed.

Magnified sweeps are again generated by reducing the value of R_a. This is accomplished by switching lower-value resistors into the common-cathode circuit of V2 and V3.

providing needed gain

As was true in Example 1, positive feedback is employed to increase the open-loop gain of the circuit. However, a different technique is employed in this case. The plates of the output-amplifier tubes V4 and V5 are cross-coupled through resistors R11 and R12 to the opposite grids (Fig. 3-9). This arrangement results in a considerable increase in open-loop gain, but also creates a problem at the cathode of V3. Since its grid is held at a fixed DC potential by Ext Horiz DC Bal potentiometer R4, the voltage change at the cathode has the same polarity (although only a fraction of the amplitude) as the sweep signal applied to the grid of V2. However, because of the high-amplitude positive feedback signal present at the grid of V5, negative feedback to the cathode of V3 tends to offset the normal positive movement and drives the cathode in the negative direction. Although this effect is rather small it is made significant by the high gain of the pentode and tends to reduce the overall gain of this side of the amplifier. For this reason, a parallel leg of each plate circuit is connected to the 100-V power supply through potentiometer R10. Movement of the potentiometer wiper increases the plate load (and thus the gain) of one tube and decreases that of the other. Since V3 is responsible for amplifying an *undesired* effect, it is the gain of V3 which must be reduced. Therefore, when properly adjusted R10 actually *increases* the horizontal-amplifier gain by *reducing* the gain of V3, and is used as a calibration adjustment for the X1 or normal mode of operation.

As the amplifier is switched to higher magnification modes, the effect of R5 on gain decreases, since the amplifier is approaching its open-loop gain condition. Potentiometer R13 is therefore inserted as common-cathode resistance for V4 and V5. This provides a narrow-range gain control for the highest magnification modes.

Another departure from the basic configuration is found in the DC level (POSITION) controls. In the normal mode the DC level of the deflection-plate driving signals is controlled by POSITION potentiometer R23. Any change in DC level at the grid of V1 is amplified at the plates of V4 and V5, providing a trace-positioning capability. When the amplifier is switched to the external horizontal mode, V1 is bypassed and control of horizontal positioning is shifted to the cathodes of V2 and V3, the "minus" inputs of the operational amplifier. Potentiometer R4 at the grid of V3 is adjusted so that no change in cathode potential occurs at V2 and V3 as the amplifier is shifted between the internal and external modes of operation.

The HORIZONTAL DISPLAY switching circuits (Fig. 3-11) are much the same as those discussed in Example 1. Note, however, that there are five positions of the switch for the external mode and that in each position a different value of R5 is inserted in the cathode circuits of V1 and V2. Gain of the horizontal amplifier in the 2-V/cm position is seen to be the same as that in the X1 MAGNIFIED position; in the 1-V/cm position, the same as X2 MAGNIFIED and so forth. The HORIZONTAL DISPLAY switch again works in conjunction with the sweep TIME/CM switch. In this case the "Uncalibrated" neon, B2, lights up whenever the sweep TIME/CM and HORIZONTAL DISPLAY switch settings yield actual deflection-sweep times shorter than 1 μs/cm.

Other concepts exemplified by this amplifier have been considered in preceding discussions.

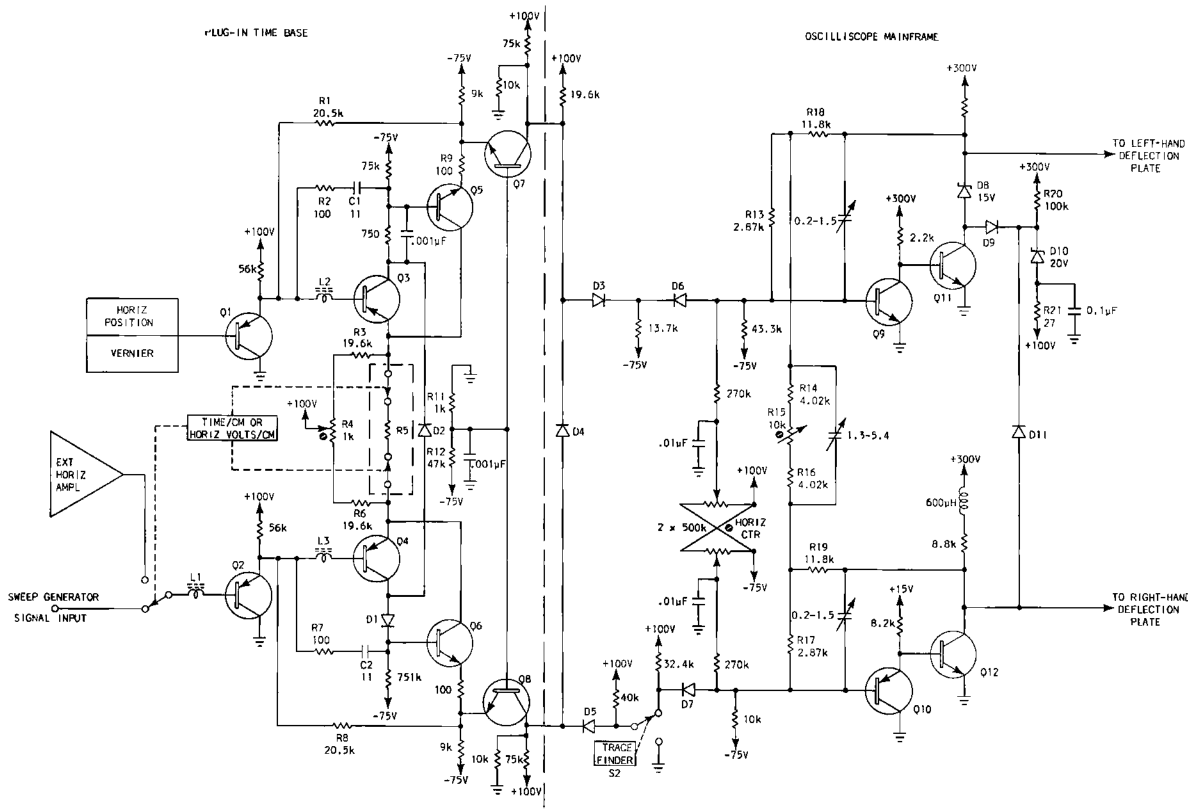


Fig. 3-12. Example-3 schematic diagram.

EXAMPLE 3 -- TRANSISTOR HORIZONTAL AMPLIFIER WITH PLUG-IN PREAMPLIFIER

Fig. 3-12 is a schematic diagram of one of the first transistor horizontal amplifiers to be developed for actual production at Tektronix. It introduces a new concept to our investigations in that it consists of two separated stages. Only the second stage (Horizontal Amplifier) is a permanent part of the mainframe. The first stage (Horizontal Preamplifier) is provided by a plug-in time-base unit, which also includes the sweep generator. (Although more than one type of plug-in time-base can be used in combination with the Horizontal Amplifier, those characteristics which affect the performance of the horizontal-amplifier stage are essentially the same.)

signal path

Development of a block diagram for this combination poses some new problems in circuit analysis. The simplest technique is to trace the complete path taken by signal current as the sweep sawtooth or external horizontal signal is applied to the input of the preamplifier.

First it should be noted that the bases of Q7 and Q8 are tied to a fixed potential by voltage divider R11-R12. This holds the emitters of these transistors at a fixed level. Therefore, the emitters of Q5 and Q6 are relatively immobile in regard to voltage changes (note the low resistance of R9 and R10). Since the base of Q3 is held at a fixed potential by emitter follower Q1 (whose base voltage is set by the horizontal-positioning circuits), no voltage change takes place at Q3's emitter in response to the sweep generator signal. In short, the only significant change in voltage level that takes place in the preamplifier (in response to the input signal) occurs at the emitter of Q4. This voltage change is reinforced by the action of Q6. When current increases or decreases in the collector of Q4, the voltage at Q6's collector moves in the same direction as the signal at the base and emitter of Q4. The small signal loss which normally takes place across the transistor's internal emitter resistance is thus effectively restored.

It follows, from this brief discussion, that the preamplifier acts as a transadmittance* amplifier whose output current is a function of (1) input signal voltage, (2) external emitter resistance of Q4 and (3) resistance in the preamplifier load.

The mainframe amplifier, consisting of Q9, Q10, Q11 and Q12 and their associated components, is a type of operational amplifier called a transimpedance amplifier. Feedback from the collectors of Q11 and Q12 creates a voltage null at the bases of Q9 and Q10. Therefore, all signal current passes through feedback resistors R13 and R18 on the left side and R17 and R19 on the right side. Since no resistance impedes the flow of current from the preamplifier, the output voltage of the horizontal (mainframe) amplifier may be expressed $E_{out} = I_{in}R_f$.

This is actually an oversimplification, since additional current will flow through the network R14, R15 and potentiometer R16, adding to the current through R18 and R19.

Sufficient current must be drawn through R13 to null the input signal current; therefore, the collector of Q11 must rise to a higher value than it would if R14 and R15 were not in the circuit. In other words, the presence of R14, R15 and R16 *increases* the expected gain of the amplifier.

---

*The terms *transimpedance* and *transadmittance* describe amplifiers in which the term "gain" is generalized to "transfer." Thus, in the common voltage-gain amplifier the symbol V_{ampl} stands for voltage transfer or "gain", $\frac{E_{out}}{E_{in}}$. An amplifier with low input and output impedances accepts a current signal and generates a voltage signal at its output terminal. Its "transfer" symbol is thus z_{ampl} or $\frac{E_{out}}{I_{in}}$ and is given in terms of resistance. That is, an amplifier with a z_{ampl} of one megohm will have a 10-volt output when a signal current of 0.01 mA is applied to its input terminal. Conversely, the symbol y_{ampl} describes the voltage-to-current "gain" $\frac{I_{out}}{E_{in}}$ of the transadmittance amplifier. Such an amplifier has high input and output impedances.

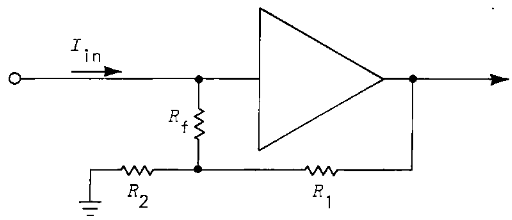


Fig. 3-13. Transimpedance amplifier with divided-feedback circuit.

gain

Now let us derive an equation which takes these facts into account using Fig. 3-13, a block diagram of a simple operational amplifier whose configuration corresponds to that of the circuit under consideration.

Since no current appears to flow into or out of an operational amplifier, the voltage drop across R_f is

$$(1) \quad E_{Rf} = I_{in}R_f$$

which also represents the voltage between the junction of R_f, R_1, and R_2 and signal ground. Current through R_2 is therefore

$$(2) \quad I_{R2} = \frac{I_{in}R_f}{R_2}$$

Since both I_{in} and I_{R2} flow through R_1, the voltage drop across R_1 is

$$(3) \quad E_{R1} = (I_{in} + I_{R2})R_1 = \left(I_{in} + \frac{I_{in}R_f}{R_2}\right)R_1$$

The output voltage of the amplifier is therefore $E_{R1} + E_{Rf}$ or, using equations (1) and (3),

$$(4) \quad E_{out} = I_{in}R_f + R_1I_{in} + \frac{R_1I_{in}R_f}{R_2}$$

$$= I_{in}\left[R_1 + R_f\left(\frac{R_1 + R_2}{R_2}\right)\right]$$

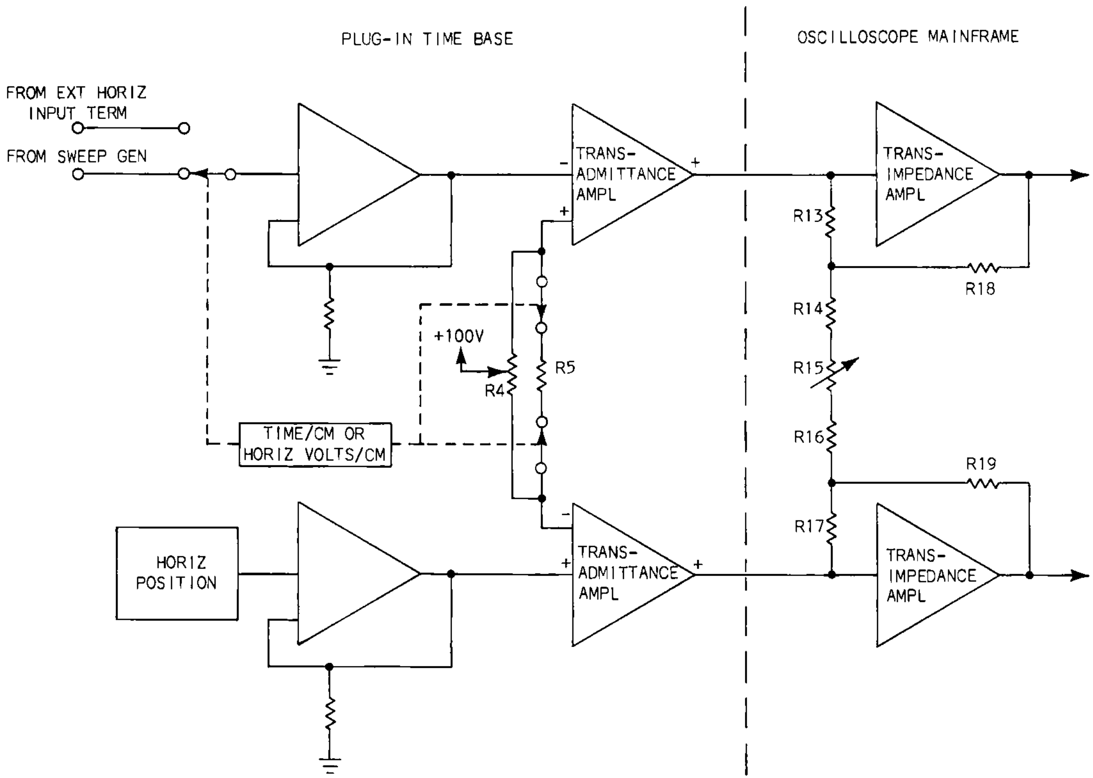


Fig. 3-14. Example-3 block diagram.

Now, since I_{in} in the above equation is the output current of the preceding stage, we can substitute the expression for this current in equation (4). Thus:

$$E_{out} = \frac{E_{in}}{R_a}\left[R_1 + R_f\left(\frac{R_1 + R_2}{R_2}\right)\right]$$

and

$$\frac{E_{out}}{E_{in}} = \frac{R_1 + R_f\left(\frac{R_1 + R_2}{R_2}\right)}{R_a}$$

which we recognize as the gain equation for the combined preamplifier and mainframe amplifier.

We can now apply this equation to the complete horizontal amplifier whose block diagram is shown in Fig. 3-14. Assuming all potentiometers to be at design center, and with switch S1 in the X1 position, we obtain the following values:

R_1 = R18 = 11.8 kΩ

R_2 = R14 + $\frac{\text{R15 (at midsetting)}}{2}$

= 4.02 + 2.5 = 6.52 kΩ

R_f = R13 = 2.87 kΩ

R_a = R5 = 2.67 kΩ (This multiple value resistor is located in the TIME/CM or HORIZ VOLTS/CM switch, not shown in the schematic.)

When we substitute these values in the modified equation, the calculation of gain (per side) is:

$$A_V = \frac{2.87\left(\frac{11.8 + 6.52}{6.52}\right) + 11.8}{2.67}$$

$$= \frac{2.87(2.82) + 11.8}{2.67} = \frac{8.2 + 11.8}{2.67}$$

= 7.5 per side, or 15-volts push-pull

In the actual instrument the sweep generator signal at the amplifier input is 10 volts in amplitude so that the input deflection factor is 1 volt per division. The CRT deflection factor is 15 volts per division. The result of the foregoing calculation is therefore quite consistent with the actual performance of the amplifier and we are justified in assuming that our analysis is correct.

Now that the gain and basic operation of the amplifier have been established we may return to Fig. 3-12 for an examination of a few significant details.

In most respects the circuit configuration can be directly related to the vacuum-tube configurations discussed earlier in this chapter. Emitter followers Q1 and Q2 perform the same function as their correspondingly placed cathode followers. Sweep-positioning and magnification circuits introduce no new concepts.

Paraphase amplifier Q3-Q4 also operates on the same principles as its vacuum-tube counterpart. However, certain modifications lend it a greater complexity and require explanation.

effects on linearity

We have seen that in vacuum-tube amplifiers the magnified sweep voltage or an extreme left or right positioning voltage is allowed to drive the amplifying stage of the horizontal-deflection circuit into saturation and/or cutoff. If this same practice were followed in transistorized amplifiers the linearity of the deflection signal would be adversely affected.

When a transistor is driven into saturation a "minority carrier charge" develops across the base-collector junction. To bring the transistor out of saturation this charge must first be drained off. The time required to do so represents a delay in the collector's response to the signal at the base. At fast sweep speeds, particularly in the magnified sweep modes, this delay would result in nonlinear and, therefore, inaccurate trace generation.

The cutoff region of a transistor's collector-current curve is also nonlinear. It is true that the nonlinear region (which occurs between about zero and 0.6 volts of forward bias in a silicon transistor) is quite limited, but it is sufficient to cause nonlinear trace generation at the start of the sweep.

The critical period, of course, is the sweep-starting period. Aberrations in the sweep voltage which occur after the beam has completed its left-to-right excursion are of little or no consequence. Therefore, our concern is that neither side of the amplifier output stage be in saturation or cutoff at sweep start. We will find that the same precautions do not apply to stages closer to the amplifier input since the resulting delays and nonlinearities occur either before or after the trace-generation period.

In the normal (X1) mode of operation, with the horizontal trace centered, there is no tendency toward saturation or cutoff in the transistors which make up the amplifier under discussion. In the magnified modes, however, or when the horizontal positioning control is adjusted to the extreme left position, Q3 will be driven into cutoff, either by the positive sweep-centering voltage at Q2's base or by a reduction in the emitter-coupling resistance.

The collector of Q3 will thus move toward the -75 V power-supply level, tending to cut off Q5. However, diode D2 conducts as soon as the collector of Q3 becomes more than about 0.6 V more negative than the collector of Q4. The potential at the base of Q5 is therefore held above the cutoff level. As the sweep generator signal rises, Q4's collector also begins to go negative, D2 cuts off and Q6 follows the subsequent fall in Q4's collector voltage at a linear rate. No similar measures are taken to prevent the cutoff of Q3 at the other end of the sweep, since the trace at this time has been driven off the right side of the CRT screen.

Diode D1 acts as a voltage-dividing resistor and also temperature compensates D2.

Resistors R1 and R8 are called "feed-forward" rather than feedback resistors. No feedback current flows in these resistors; however they do serve to absorb the small current that flows in R6 and one half of R4, which parallel R5. For example, if the base of transistor Q4 rises one-volt positive, its emitter will rise by an almost equal amount, causing a discrete current change in R6 and R4 which adds to the signal current change in R5. Since the base (and therefore the emitter) of Q3 is held at a fixed level, no corresponding current flows through R3. However, since the collector of Q6 is tied to the low input impedance of the mainframe amplifier, it too is held at a fixed level. Therefore the current which flows through R8 when the base of Q4 goes positive subtracts from the collector current. Note that the value of R8 is almost exactly equal to the value of R6 plus one half that of R4. Thus the "imbalance" current normally present in a paraphase amplifier is almost completely cancelled. What little imbalance remains is removed by the common-mode rejection action of the following push-pull amplifier.

Although it is not uncommon to find both negative and positive feedback connections in the same amplifier, it is unusual to find one negative feedback loop within another, as is seen in this amplifier. R2 and C1 as well as R7 and C2 provide a negative feedback path whose impedance decreases as sweep speed increases. Their purpose is to compensate for stray (switch contact) capacitance (across R5) which provides too much HF peaking of the sweep signal. The RC feedback network decreases gain at these higher frequencies to offset the peaking effect.

Ferrite-bead inductors L1, L2 and L3 are inserted to suppress parasitic oscillations in the respective transistors.

Transistors Q5 and Q6 are compound-connected for high gain and to assure that gain is dependent almost entirely on the value of R5. For example, a positive signal at the base of Q3 produces a negative signal at the base of Q5. This in turn causes the collector of Q5 to go positive, pulling up the emitter of Q3. As a result of this action, almost no change in voltage occurs across the transistor's internal-emitter resistance so that only the resistance of R5 affects the gain of the paraphase-amplifier section.

Transistors Q7 and Q8 are connected as grounded-base amplifiers to provide a high-impedance (current) source for the following transimpedance amplifier.

Moving to the second stage of the horizontal amplifier, we find another limiting circuit. Under any conditions which produce an on-screen trace, diodes D3, D5, D6 and D7 are all forward biased, while diode D4 is cut off. However, as a horizontal-positioning voltage or a magnified sweep signal drives the trace offscreen to the left, diodes D3 and D5 disconnect before the left side of the amplifier is cut off or the right side saturated. Further increases in left-positioning current are channeled through diode D4. Fast recovery of the amplifier is thus guaranteed as the sweep rises to an on-screen value. In the offscreen-to-the-right condition, diodes D6 and D7 disconnect before the right side of the amplifier saturates or the left side is cut off.

Zener diode D10 together with resistors R20 and R21 form a voltage divider which sets the upper limit on collector excursions for transistors Q11 and Q12.

Should some accidental occurrence tend to increase the collector voltage of either transistor above about 120 volts, diode D9 and D11 would conduct to prevent any further increase and resulting damage to the transistors.

Zener diode D8 increases the voltage level at the left deflection plate of the CRT so that at midsweep the deflection plates are at identical voltage levels.

locating the beam

A new convenience feature is introduced in this circuit -- the TRACE FINDER switch. This is a spring-loaded switch which, when manually depressed, closes the contacts which conduct the right-hand deflection signal to ground, causing the CRT to no longer have push-pull deflection. The left-hand side of the amplifier has insufficient gain to drive the CRT beam off-screen, so that regardless of sweep mode or horizontal positioning the beam can be easily located.

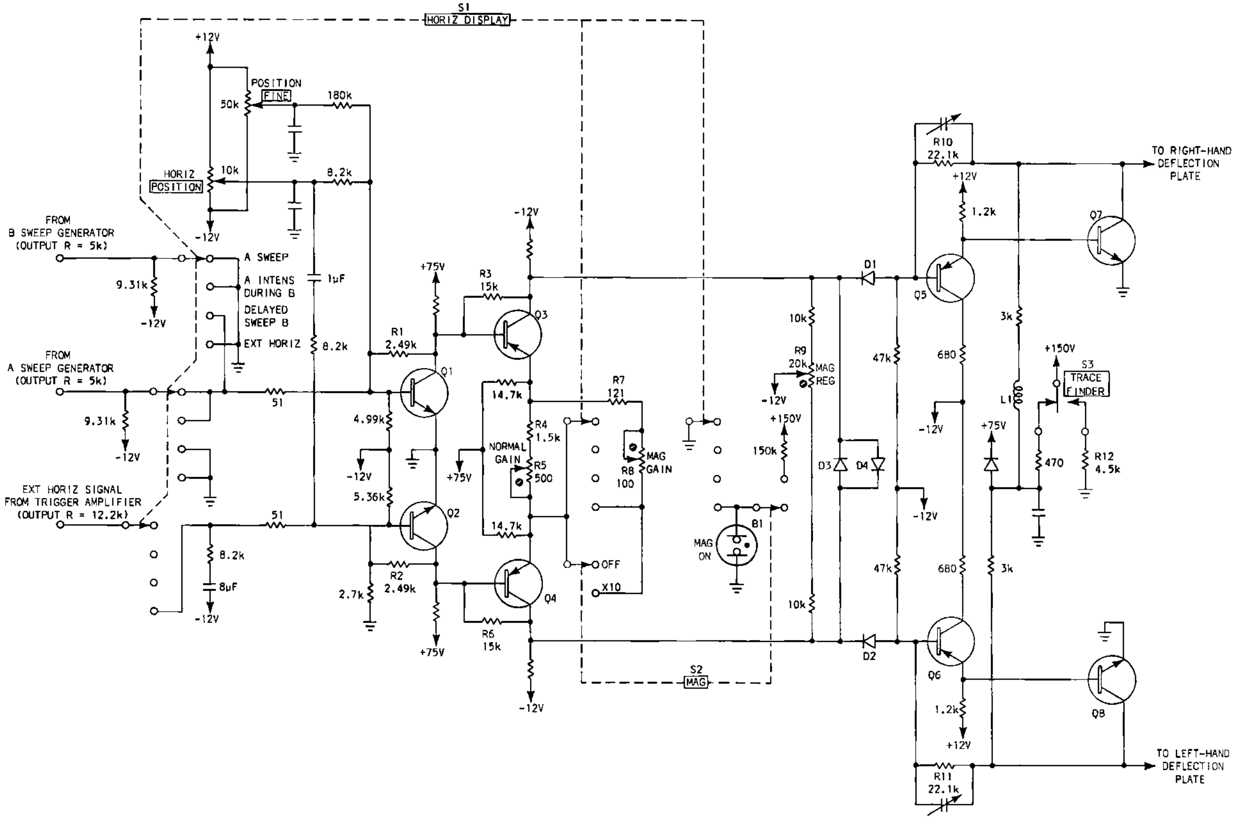


Fig. 3-15. Example-4 schematic diagram.

EXAMPLE 4 -- MODERN TRANSISTOR HORIZONTAL AMPLIFIER

A transistor horizontal amplifier of more recent design is shown schematically in Fig. 3-15. It will be of considerable advantage to our analysis of this amplifier if we note its similarities to the amplifier we have just examined.

gain

In the first place, Q5 and Q7 as well as Q6 and Q8 form independent negative-feedback amplifiers and provide sufficient open-loop gain to qualify as operational amplifiers. Since the collectors of Q3 and Q4 are connected to the null point of the operational amplifiers, no change in signal voltage occurs at the collectors. This means that no signal current flows in R3 and R6. Thus we have almost the same arrangement as we found in the immediately preceding case -- a transadmittance amplifier driving a transimpedance amplifier. Gain of the combined stages is therefore equal to the ratio of R_f in the second stage to R_a of the first stage. Also, it is clear that R3 and R6 serve the same current-balancing function as their counterparts in the previously described circuit. A block diagram of the complete amplifier is shown in Fig. 3-16.

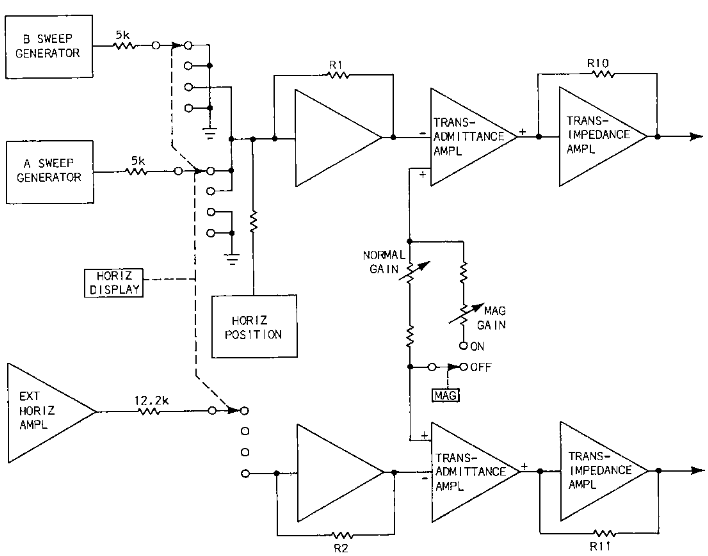


Fig. 3-16. Example-4 block diagram.

The input signal is taken from one of two sweep generators and applied to the base of transistor Q1 (Fig. 3-15), or from the EXT HORIZ input terminal and applied to the base of Q2. No cathode coupling exists between the two transistors, so they operate independently, although horizontal positioning information is always present at the base of Q1. Thus the output of this pair of transistors is always single-ended and is applied to either the base of Q3 or Q4, depending on the mode of operation. Resistors R1 and R2 conduct negative feedback to the base of the transistors. Gain of these transistors is sufficient to justify their treatment as operational amplifiers. Input resistance of the amplifiers is provided by the output resistance of their respective signal sources. In the instrument for which this horizontal amplifier was designed, the output impedance of the A and B sweep generators is 5 kΩ each. Gain of the first stage is therefore

$$\frac{R_f}{R_a} = \frac{2.49}{5} \approx 0.5 \text{ k}\Omega$$

for the A and B sweeps. The external horizontal signal is supplied from a voltage divider in the Trigger Amplifier circuit of the oscilloscope. From the horizontal amplifier's input terminal the impedance of the voltage divider is about 12.2 kΩ. Gain impressed on the external horizontal signal is thus

$$\frac{R_f}{R_a} = \frac{2.49}{12.2} = 0.22$$

Push-pull gain of the combined paraphase and output amplifier stages is also quite easily calculated. In the normal (X1) mode, with all potentiometers at design center;

$$A_V = \frac{R_f(\text{left}) + R_f(\text{right})}{R_a} = \frac{\text{R10} + \text{R11}}{1.5 + 0.25} = \frac{44.2}{1.75} = 25$$

Front-to-back gain of the horizontal amplifier in the normal A and B sweep modes is thus

$$(A_{V1})(A_{V2}) = (0.5)(25) = 12.5$$

Since both sweep generator signals are approximately 10 V/div in amplitude, and the deflection factor of the CRT is about 12 V/div, the calculations appear to be verified.

When the MAG switch is set to the X10 position, resistor R7 and Mag Gain potentiometer R8 are placed in shunt with R4 and R5. This reduces the value of R_a to $\frac{(1.75)(0.171)}{1.75 + 0.171}$ or 0.156 kΩ, roughly one-tenth its value in the unmagnified mode of operation, thus raising the gain of the amplifier by a factor of ten. R5 is adjusted to compensate for circuit variations and CRT differences so that the magnified sweep has exactly ten times the deflection factor of the unmagnified sweep.

When S1 is in the EXT HORIZ position, the amplifier is automatically placed in the X10 MAG position. Gain imposed on the external horizontal signal is therefore:

$$A_V = (0.2)(25)(10) = 50$$

However, note that MAG ON neon tube B1 only lights up when MAG switch S2 is in the X10 position.

The Mag Reg control (potentiometer R9) operates in the same manner described in earlier discussions. The Normal Gain control performs the same function as those formerly labeled "Sweep Cal" -- that is, it is used in calibration procedures to adjust *normal* gain to the required factor compensating for variations in CRT deflection factor, transistor characteristics, and passive-component values.

preventing saturation and cutoff

As we should expect from our experience with the first transistorized amplifier, measures have been taken to prevent saturation and cutoff of the output-amplifier transistors. Diodes D1 and D2 together with D3 and D4 accomplish this task. At the right deflection plate Q7 would tend to saturate as the magnified deflection signal tried to drive the CRT beam off-screen to the left. However, as the collector of Q3 drops to about 0.5 volts, D1 becomes back-biased and disconnects. This prevents any further change at the base of Q5 and Q7 so that Q7 does not go into saturation.

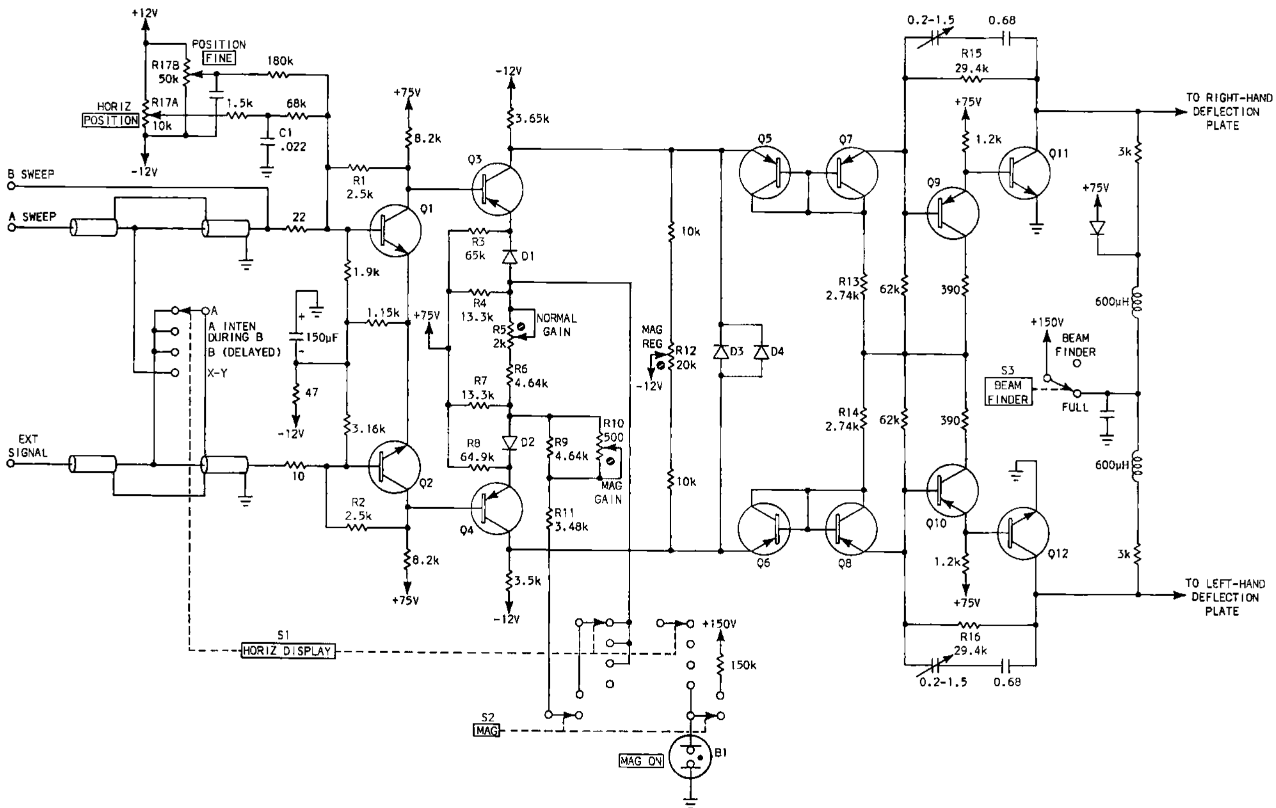


Fig. 3-17. Example-5 schematic diagram.

When the sweep rises to a level which allows D1 to turn on again, Q7 follows immediately and the ramp remains undelayed and linear. The same effect can be observed in diode D2 and transistors Q6 and Q8 on the other side of the amplifier. When diode D1 or D2 disconnect, Q3 and Q4 lose their low-impedance load and would thus tend to go into saturation at this time. However, when D1 or D2 cut off, either D3 or D4 goes into conduction to maintain the required low impedance.

Even though both transistors conduct throughout the application of a normal ramp, the collector currents of Q7 and Q8 cannot always add to a constant value. That is, as the ramp rises Q7 draws less current than required to maintain a constant deflection-plate charging current. This additional current is supplied by the collapsing magnetic field in inductor L1.

TRACE FINDER

In the collector circuit of Q7 is a TRACE FINDER switch whose function is the same as that of the TRACE FINDER switch discussed in Example 3. When Q7 and Q8 are disconnected from the 150-V power supply, the sudden drop in current demand would cause a sharp momentary rise in the unregulated supply voltage. This would probably disturb the operation of other oscilloscope circuits supplied from the same source. However when the spring-loaded switch is depressed another set of contacts connect the power supply to "dummy-load" resistor R12, maintaining the current demand at a reasonably constant level.

EXAMPLE 5 -- TRANSISTOR HIGH-PERFORMANCE HORIZONTAL AMPLIFIER

A more sophisticated version of the horizontal amplifier which we have just examined has been developed for a solid-state wide-bandwidth (150 MHz) oscilloscope. This instrument offers *X-Y* display capabilities up to 2 MHz, which far exceed those of all but special-purpose oscilloscopes. A schematic diagram of this amplifier is shown in Fig. 3-17.

The general arrangement of the amplifier is similar to that of the two preceding examples. The input signal is first processed by a low-gain negative-feedback paraphase amplifier, then is applied to a transadmittance-transimpedance combination that functions as a push-pull operational amplifier.

The time-base ramp is applied from either the A or B sweep generator to the upper half of the horizontal amplifier, while the external horizontal signal is applied to the lower half, depending on the position of the HORIZ DISPLAY switch. In this instrument sweep-mode selection is performed in the sweep generators by other contacts of the HORIZ DISPLAY switch. Note that one set of contacts on this switch grounds the shield of the lead which provides the input signal and, at the same time, grounds the inner conductor of the other coaxial input. The input resistance of the first amplifier depends on the display mode selected. In the A or B sweep mode, the output impedance of the selected-sweep generator supplies the input impedance of the first stage while in the *X*-*Y* (external) mode, resistors in the trigger preamplifier perform the same function.

calculating gain

Turning to a block diagram of this circuit (Fig. 3-18) let us first calculate the gain of the paraphase amplifier. In either of the internal sweep modes the gain is found by dividing the feedback resistance (R1 or R2) by the input resistance (sweep-generator output impedance). This yields a gain of about 0.5 per side, or a push-pull gain of 1. In the *X*-*Y* mode, R_{out} (XY) can be varied between about 1.75 and 1.2 kΩ for a first stage gain between approximately 2.8 and 4.2.

No special compensation is provided for the inherent imbalance of the paraphase amplifier. In the first place, since the emitters of Q1 and Q2 are tied directly together, signal current in the opposite collectors is only slightly out of balance. This slight imbalance is corrected by the common-mode rejection characteristic of the following push-pull amplifier.

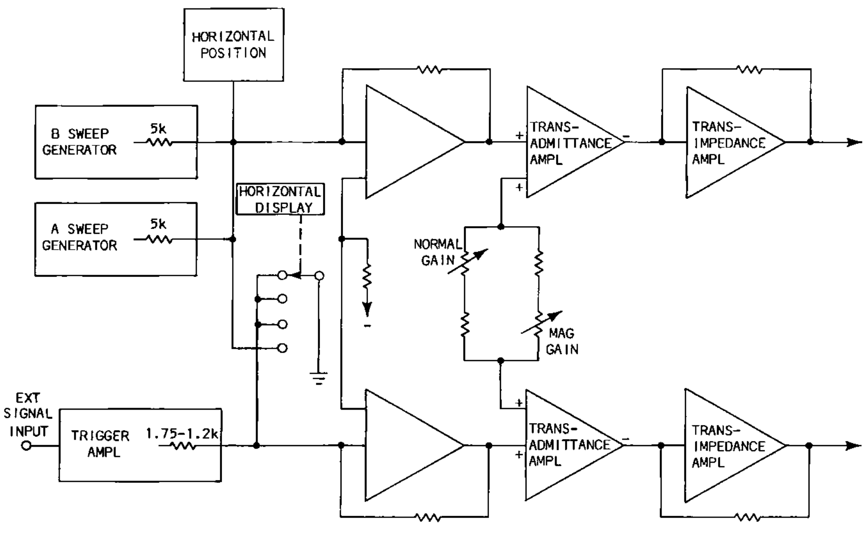


Fig. 3-18. Example-5 block diagram.

Horizontal position (DC level) of the sweep is established by ganged potentiometer R17, as explained in previous discussions. C1 forms a simple decoupling network with the resistance of the potentiometer to remove any AC ripple from the input of the amplifier.

The paraphase amplifier output, a fairly well-balanced push-pull signal, is applied to the push-pull amplifier. Gain of this stage is determined by feedback resistors R15 and R16 and the emitter resistance of Q3 and Q4. With all potentiometers at design center, the calculation becomes

$$A_V = \frac{\mathrm{R15} + \mathrm{R16}}{\mathrm{R5} + \mathrm{R6}} = \frac{58.8}{4.64 + 1} \simeq 12$$

With the nominal 10-V time-base ramp, this gain provides the 120 volts of deflection signal required by the CRT deflection factor of 12 volts per division. In the magnified mode of operation the cathode-coupling resistance of Q3 and Q4 is reduced to about one-tenth its former value, by the insertion of R11 and potentiometer R10 with R9 in parallel, increasing gain by a factor of ten. Mag Gain potentiometer R10 is adjusted *after* Normal Gain potentiometer R5 has been set, to compensate for circuit and CRT deflection-factor variations. The Mag Reg potentiometer R12 is adjusted, as explained earlier, so that the normal and magnified deflection signals reach the same level at midsweep, thus assuring equal expansion on both sides of center screen.

preventing cutoff and saturation

The adverse effects of saturation or cutoff of the amplifying transistors have already been explained. Cutoff of Q3 and Q4 is prevented by diodes D1 and D2. At the midscreen signal level, both Q3 and Q4 conduct at the same level and no current flows through R5 and R6. At any other time, however, current in the two emitter circuits is unbalanced. For example, as Q3 conducts less current, its emitter becomes increasingly positive while Q4's emitter moves in the negative direction by an equal amount.

Current therefore flows through R5 and R6 which adds to that already flowing through D1 from R4. This means that D1's anode becomes increasingly negative while its cathode goes positive. Before Q3 goes into cutoff and about the time the CRT beam has been driven slightly off-screen, diode D1 is disconnected by reverse bias, increasing Q3's emitter resistance by a factor of 5 or more. This degrades the gain of Q3 and at the same time increases its ability to follow the base signal without cutting off.

Saturation of Q3 and Q4 is prevented by the diode-connected transistor pairs, Q5-Q7 and Q6-Q8. Each transistor pair acts as a pair of diodes connected cathode-to-cathode. Keep-alive current is supplied through resistors R13 and R14 so that small changes in the collector currents of Q3 and Q4 are transmitted to the bases of Q9 and Q10. However, should this current rise or fall beyond a specific limit, one of the diode-transistors will cut off. For example, suppose current rises in Q4. Current through Q6 will also increase due to the increase in forward bias across its junction. However, the resulting increase in current through resistor R14 will also drive the base of Q8 positive and eventually cause it to disconnect. When this takes place diode D3 will go into conduction, maintaining the load impedance of Q4 at its former low value.

The reason for using paired transistors in diode configuration, rather than conventional diodes, is to attain a more linear transition between the conducting and nonconducting states of the diodes.

inductor for C_{dp}

The remaining portion of this amplifier is similar to that explained in Example 4 with two minor exceptions. In this amplifier both right *and* left collector circuits include an inductor to supply added charging current for deflection-plate capacitance at fast sweep speeds. Also note that no dummy load is switched into the 150-V power supply when the spring-loaded BEAM FINDER switch is depressed. This indicates that the load imposed by the horizontal amplifier is only a small part of the total load on the power supply, so no disturbance is created when it is suddenly disconnected.

BEAM FINDER

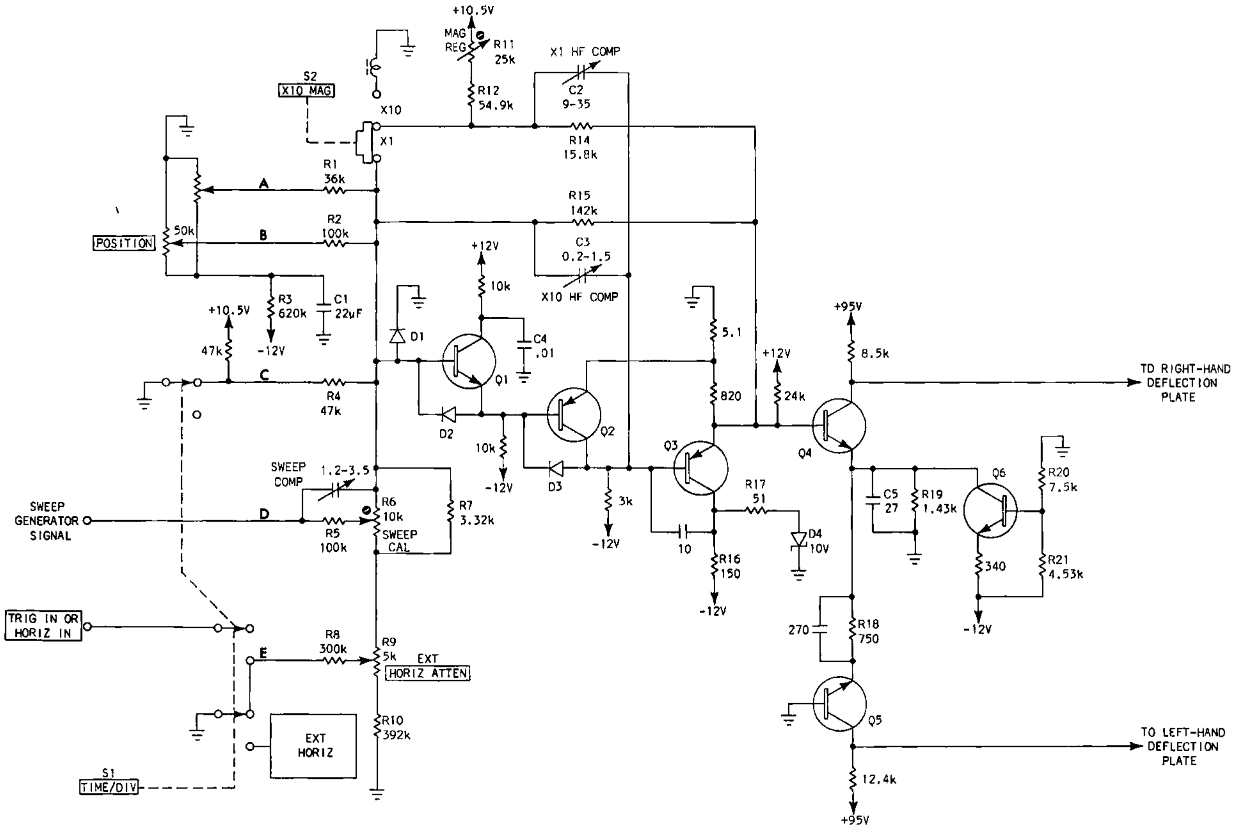


Fig. 3-19. Example-6 schematic diagram.

Unlike the previously described amplifier, this one does not automatically switch to magnified gain when used in the external, or *X-Y*, mode of operation; the switching circuit is so arranged that only normal gain is available in this mode. Remember, however, that the gain of the amplifier in the *X-Y* mode is variable between 3 to 4 times that in the NORMAL mode, due to the lower output resistance of the *X* signal source.

X-Y mode variable gain

EXAMPLE 6 -- TRANSISTOR HORIZONTAL AMPLIFIER FOR BATTERY-POWERED OSCILLOSCOPE

When practical semiconductors became commercially available it was inevitable that battery-powered portable oscilloscopes would be developed for the market. The advantages of a lightweight, compact instrument that requires no external power source are as obvious as they are numerous. However, conventional circuit configurations often had to be modified or totally discarded to meet the more stringent demands of these new instruments.

The chief requirement of battery-powered instruments is *power economy*. New power supplies were therefore developed which exhibited much higher efficiency and versatility but could regulate properly only with a nearly constant load. This restriction added further to the problems already facing the solid-state-circuit designer. The horizontal amplifier represented by the schematic diagram in Fig. 3-19 is one of the configurations in which this challenge has been met successfully.

power economy

constant load

The complete amplifier consists of an operational amplifier, Q1, Q2 and Q3, followed by a conventional paraphase inverter, Q4 and Q5. (See block diagram Fig. 3-20.) Switch S2 in the feedback path changes the R_f/R_a ratio to provide a X10 magnified sweep. Inputs to the operational amplifier are (1) DC horizontal positioning (terminals A, B and C), (2) the time-base ramp from the A or B sweep generator (terminal D), and (3) an external horizontal signal (terminal E). (The A and B sweep generators remain connected at all times, but generate no signal in the EXTERNAL mode of operation.) All amplification takes place in the paraphase inverter since the gain of the operational amplifier is considerably less than unity. The input of the operational amplifier is utilized as a low-impedance point into which the switch capacitance and various control circuits may be safely inserted.

The base of Q1 is the input to the operational amplifier. The DC feedback signal is supplied from the low impedance at the emitter of Q3, while the AC feedback comes from the high impedance at the collector of Q2. This arrangement lowers the propagation time of the AC feedback signal, yet provides the necessary current drive for the DC feedback. Resistor R13 and the series combination R12 and Mag Reg potentiometer R11 are a part of the current-summing positioning network and may be viewed simply as additional DC inputs to the operational amplifier. R14 and R15 are the feedback resistors, compensated for high sweep speeds by C2 and C3.

R3 and C1 constitute a decoupling network for the -12 V supply to the positioning network. R5 provides the principal input resistance for the sweep signal, while R8 performs the same function for the EXT HORIZ signal.

calculating gain

The natural open-loop gain of the operational amplifier is augmented by positive feedback supplied from the emitter of Q3 to the emitter of Q2. A point-to-point calculation would yield an open-loop gain of about 1000. Since the nominal deflection factor of the CRT in this instrument is 11.2 V/div, requiring a 10-division deflection voltage of 112 volts, it is quite apparent that the base voltage of Q1 is required to change only by about 100 mV for full-screen deflection. It is important to keep this fact in mind as we examine the operation of the amplifier.

The A and B sweep amplitudes are idential at about 31 volts, starting at +3 volts and rising to +34 volts. The voltage-dividing effect of R9, R10 and sweep-calibrating potentiometer R6 determines what portion of the signal current is applied to the base of Q1. We will treat this effect as a determinant of the operational-amplifier input resistance. In this way we can calculate gain without resorting to the lengthly shunt impedance equation used on page 25.

If the pickoff wiper of R6 is in its uppermost position, no signal current will flow through the potentiometer, so it is effectively removed as input resistance. If, on the other hand, the pickoff is in the lowest position, part of the signal current is shunted to ground and the full value of R6 (in parallel with R7) must be added to the input resistance. Because of the high open-loop gain of the amplifier, the shunting effect of R1, R2 and R4 can, as usual, be ignored.

Assuming all potentiometers to be set at design center, let us calculate the gain of the operational amplifier (Fig. 3-20). In the normal mode R_f consists of the parallel resistance of R14 and R15, about 14.2 kΩ. R_a will be equal to R5 modified by the signal-current-division ratio of R9, R10 and the parallel resistance of potentiometer R6 and R7.

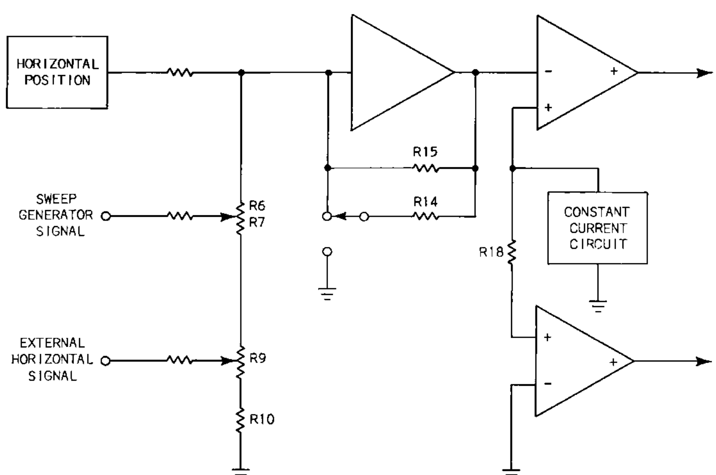


Fig. 3-20. Example-6 block diagram.

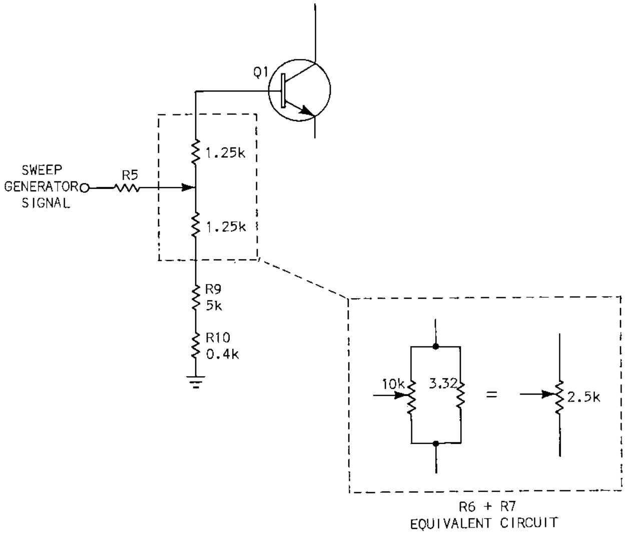


Fig. 3-21. Input-signal current divider.

When reduced to a series equivalent circuit and with R6 at design center, the voltage divider appears as shown in Fig. 3-21. Signal current is thus attenuated by the factor $\frac{6.65}{7.9}$ or 0.84. This current division has the same effect as an *increase* in the value of R5 by the factor 1.18 (the reciprocal of 0.84). The gain calculation for the operational amplifier is therefore $A_V = \frac{R_f}{R_a} = \frac{14.2}{100 \cdot 1.18} = 0.12$ for the normal mode of operation. When the X10 MAG switch is actuated one leg of the feedback path is grounded, leaving only R14 to pass the feedback current. At 142 kΩ the feedback resistance is now 10 times its former value, increasing the operational-amplifier gain by the same factor.

Calculation of the paraphase-inverter gain must take into account the presence of the longtailed current-supply transistor Q6 and its associated components. The fixed forward bias developed across voltage divider R20 and R21 assures that Q6 generates a constant current for the paraphase inverter at a higher level than could be attained with a longtail resistor. R19 supplies required additional current, and together with R17, R16 and zener diode D4, forms a current-balancing network which overcomes the inherent imbalance of the paraphase inverter.

Since the base of Q5 is grounded, and all signal current at its emitter must pass through R18, gain through the left half of the amplifier is simply $\frac{R_{out}}{R_{in}} = \frac{12.4}{750} = 16.5$. Input impedance for the other half is found by calculating the parallel resistance offered by Q6, R19 and R18. The collector impedance of Q6 is so high it has little influence on the problem, so R_a becomes $\frac{1.43 \cdot 0.75}{1.43 + 0.75} = \frac{1.07}{2.18} = 0.493\ \text{k}\Omega$. Gain of this half thus equals $\frac{R_{out}}{R_a} = 8.5 \cdot 493 = 17$. Note how the inherent imbalance of the paraphase inverter has been almost totally eliminated by this arrangement.

Overall gain of the inverter is, of course, 17 + 16.5 or 33.5. Combined with that of the operational amplifier this yields a total gain of 0.12 · 33.5 or about 4 for the complete amplifier. With a 30-volt time-base ramp at the input, the deflection voltage applied to the CRT is about 120 volts. This is slightly more than enough for 10.5 divisions of deflection in the normal mode of operation.

normal mode operation

Let us return now to the operational amplifier and examine its response to the time-base ramp in both the normal and X10 MAG configuration. At quiescence, in the normal mode with the horizontal-position controls centered, the CRT beam is positioned a little more than 5 divisions to the left of center screen. All transistors are in conduction at this time. The currents of all the input circuits add algebraically to a value very close to zero volts at the base of Q1. The emitter of Q1 and the base of Q2 are therefore about -0.6 V. This places the emitter of Q2 at about zero volts so that it passes only a small trickle of current. D3 is therefore back-biased. Q3, on the other hand, is forward-biased by the negative collector voltage of Q2 and is therefore well into conduction in the class A mode. As the time-base ramp begins to rise, feedback current tends to nullify any voltage change at the base of Q1. However, because the amplifier lacks infinite gain, a slight rise in Q1 base voltage takes place. Most of this rise is coupled through Q1 emitter to Q2 base where it tends to reduce the current in that transistor. The result is an amplified negative signal at Q2 collector which is coupled to Q3 base. The output from Q3 emitter is also negative-going and is applied to the paraphase inverter.

It must be remembered that the base voltages of Q1 and Q2 must change by only a few millivolts to produce about 3-volts change in the emitter of Q3 and all the transistors remain in a conducting state throughout the sweep.

X10 MAG mode

When the circuit is shifted to the X10 MAG mode, a number of changes take place in the behavior of the operational amplifier. When the upper feedback circuit is grounded, one of the positive inputs to the current-summing network is removed from the circuit, causing a considerable negative shift in DC level at the base of Q1. This change is inverted in Q1 and transmitted to the base of Q2 causing current to increase in Q2. As Q2's collector moves in the positive direction, however, D3 becomes forward biased and goes into conduction, tying the anode of D2 to some positive value so that it also turns on. This limits the developing reverse bias on Q1's base-emitter junction to a safe value. Q1, of course, cuts off. Only a slight change is noticed at the output of the amplifier -- just enough, in fact, to deflect the CRT beam slightly off-screen to the left.

With Q1 in cutoff, the feedback loop is open and the operational amplifier is disabled. No change takes place until the sweep voltage rises sufficiently to bring the base of Q1 to the normal sweep-starting level -- about zero volts. At this time diodes D2 and D3 cut off, Q1 goes into conduction and the feedback loop is reestablished. When the ramp rises a few millivolts farther, the positive signal at Q1's emitter cuts off Q2 and the feedback loop is again opened, preventing any further change at the collector of Q3. At this time the beam is slightly off-screen to the right, having traversed the X axis during the very short period in which the operational amplifier functioned as such. When the base of Q1 rises more than about 0.6 volts above ground, D1 goes into conduction to prevent damage to Q1.

It should be apparent that the operational amplifier will respond in exactly the same fashion when the horizontal-positioning voltage is shifted in either direction. Only that portion of the normal (or magnified) sweep, appearing in the "window" established by the limiting action described above, will cause a trace to appear on the CRT.

One important feature of the operational-amplifier action is that no transistor is allowed to saturate in the magnified sweep mode.

preventing saturation

As stated earlier, by preventing saturation of the amplifying transistors it is possible to preserve the linearity of the magnified sweep. Preventing saturation also helps to eliminate large current swings in the power supply which, as we have stated, must be avoided in this type of circuit.

Another feature which also tends to prevent large current changes is the choice of resistive elements in the collectors and emitters of the transistors. As the ramp rises Q1 and Q4 draw more current, while Q2 and Q3 draw less. The drop in current through one pair of transistors almost equals the rise through the other pair.

Frequency compensation in this horizontal amplifier is similar to that in previously described circuits with a few minor additions. Because the collector resistors of Q4 and Q5 are not of the same value, the RC charging time for the deflection-plate capacitance differs from one side of the paraphase inverter to the other. C5 is inserted in the emitter circuit of Q4 to bring the charging times back into balance. C4 reduces the collector impedance of Q1 at high sweep speeds, improving the gain at the emitter by reducing the usual Miller effect exhibited by plate-loaded amplifiers.

Note that in the magnified mode of operation the Mag Reg adjustment affects the DC level at the base of the paraphase amplifier rather than at the input to the operational amplifier.

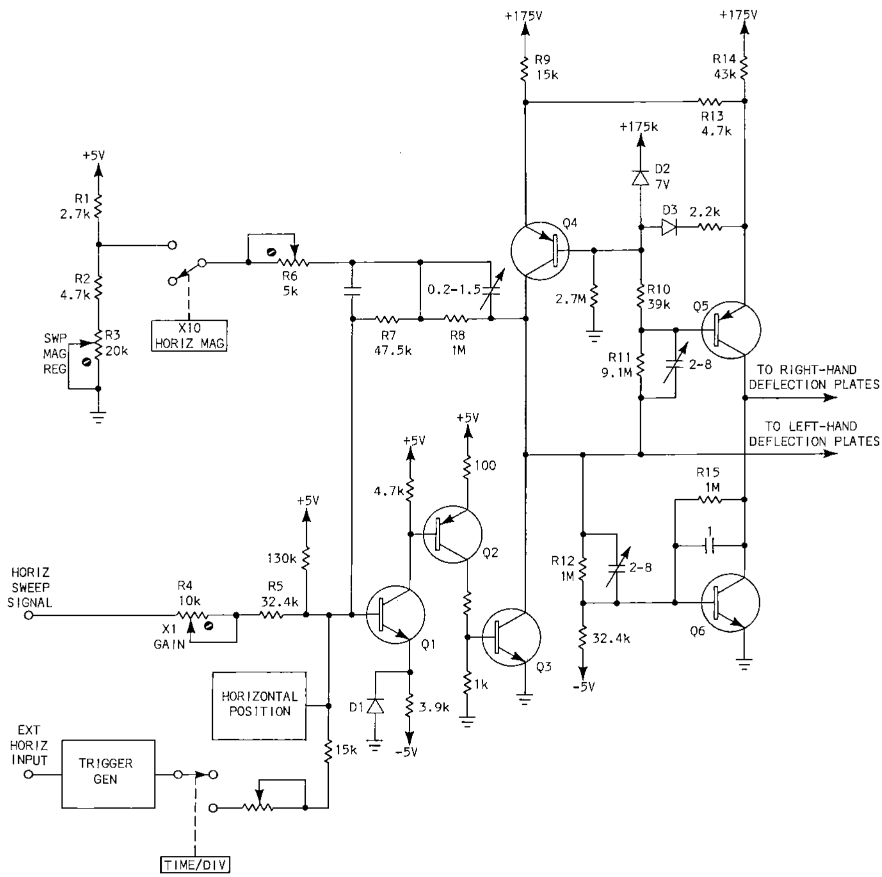


Fig. 3-22. Example-7 schematic diagram.

EXAMPLE 7 -- TRANSISTOR HORIZONTAL AMPLIFIER WITH CURRENT-CONTROL CIRCUIT

Another example of power-saving techniques is provided by the sweep amplifier whose schematic diagram is shown in Fig. 3-22. In this amplifier, the most notable departure from configurations already encountered is the lack of a paraphase inverter. Here the output of the operational amplifier, Q1, Q2, and Q3, is applied directly to the left deflection plate. The right-deflection-plate signal is developed by unity-gain signal-inverting operational amplifier, Q6, using the left-deflection-plate signal as its input. A current-control circuit, Q4 and Q5, acts to minimize the power dissipation of the circuit. A X10 magnified sweep mode is provided through control of the feedback resistance in the first operational amplifier. Like the immediately preceding example, this horizontal amplifier is designed for a battery-powered instrument, where reduced power consumption and a constant load on the power supply are of primary importance. A block diagram of the amplifier is shown in Fig. 3-23.

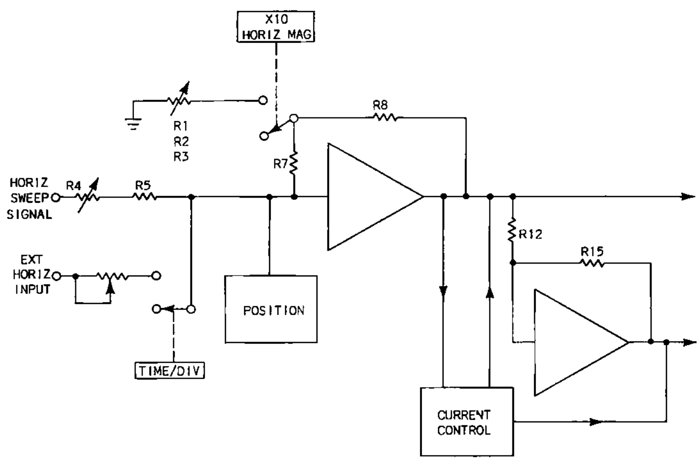


Fig. 3-23. Example-7 block diagram.

gain

The first or "output" operational amplifier has a different configuration from those encountered in previous discussions. When the X10 HORIZ MAG switch is open, feedback resistance R_f is simply the combined resistance of R7 and R8. When closed, however, R8 and the series equivalent of R6, R1, R2 and R3 act as a voltage divider, so a greater change in output voltage is required to maintain a null at the amplifier's input. (See page 63 for another example of this same situation.) (To simplify the equation we will include the series equivalent of R1, R2 and R3 in the value assigned to R6.) Therefore the value of R7 must be modified by the ratio of $\frac{R8 + R6}{R6}$. The equation for the gain of the magnified configuration thus becomes:

$$A_V = \frac{R7\left(\frac{R8 + R6}{R6}\right) + R8}{R_a}$$

Input resistance R_a is supplied by R5 and X1 GAIN potentiometer R4. Gain in the X1 mode (assuming all potentiometers at design center) is therefore:

$$\frac{R_f}{R_a} = \frac{R7 + R8}{\frac{R4}{2} + R5}.$$

$A_V = \frac{1.0475 \text{ M}}{0.0374 \text{ M}} = 28$. Since the sweep generator signal is about 5 volts in amplitude, a 140-volt deflection signal is applied to the left-hand side of the CRT.

The 140-volt deflection signal at the collector of Q3 is applied through R12 to the base of Q6. Feedback from the collector is returned to the base through another 1-MΩ resistor, R15. The R_f/R_a ratio is therefore unity, so although the input signal is inverted it suffers no change in amplitude.

The deflection factor of the CRT in this instrument is a nominal 27 volts per division, so about 10.5 divisions of deflection will be generated by the total 280-volt change at the deflection plates.

Substituting appropriate values in the equation for the magnified sweep configuration, the calculation is

$$A_V = \frac{0.0475\left(\frac{1 + 0.00475}{0.00475}\right) + 1}{0.0374} \approx 280$$

which satisfies the reqired X10 increase in gain factor.

current control

The current-control circuit exemplifies one of the new techniques developed for power conservation in battery-powered instruments.

At quiescence, current through feedback resistors R7 and R8 is maximum, while that through R15 is minimum. The base of Q4 is held at a fixed potential by zener diode D2. The high positive potential at the base of Q4 limits current in Q4 to a very low value. The bases of Q4 and Q5 are essentially at the same potential because of the high voltage-divider ratio of resistor R11 to R10. The emitters of these transistors are therefore also at about the same potential; almost no current flows through emitter-tying resistor R13. As the sweep at the base of Q3 rises, its collector goes negative, increasing current through Q5. As the emitter of Q5 goes negative, current begins to flow through R13, while current through R9 changes very little, due to the much higher resistance of R14. Thus, the loss of current supplied to R9 through R7 and R8 is balanced by the increased current through R15 and R13. The overall effect, therefore, is to maintain an almost constant current through both R9 and R14 in spite of the changing current in the feedback resistors. R9, in effect, acts as the load resistor for both operational amplifiers.

This amplifier has the added advantage of consuming little power under no-sweep conditions. With no sweep present at the input, neither side of the amplifier draws much current, due to the very small forward bias on both Q3 and Q6. Only when the sweep is actually underway, therefore, does the circuit dissipate operating power of any significance.

In the magnified mode of operation Q1 and Q3 cut off during most of the lower portion of the sweep, while Q5 saturates during most of the higher portion. No ill effect is produced by saturation of Q2 in this case however, since the delaying and distorting effects mentioned earlier are imposed only on the retrace portion of the sweep.

Both the saturation and cutoff periods occur when the beam is off-screen. The 10% portion of the sweep which actually produces a trace on the CRT has the same upper and lower voltage level as the unmagnified sweep. Swp Mag Reg potentiometer R3 is adjusted to assure that the midsweep levels of both normal and magnified sweep are the same. This, of course, causes the trace to expand equally on both sides of the graticule center.

NOTES

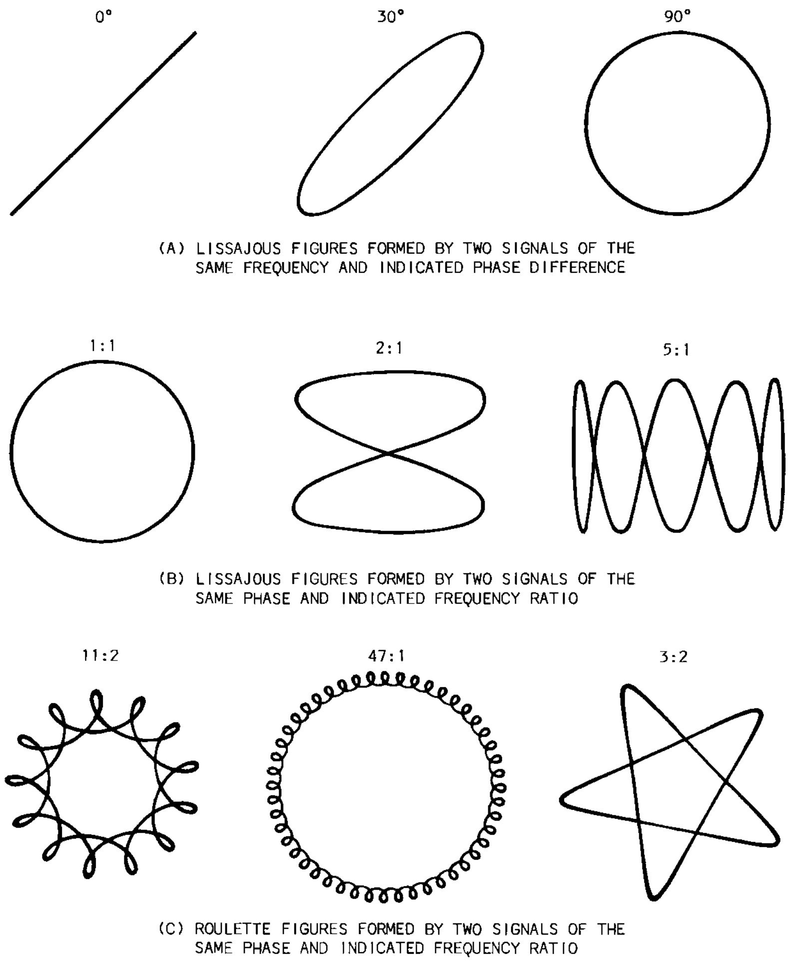

0°
30°
90°
(A) LISSAJOUS FIGURES FORMED BY TWO SIGNALS OF THE SAME FREQUENCY AND INDICATED PHASE DIFFERENCE
1:1
2:1
5:1
(B) LISSAJOUS FIGURES FORMED BY TWO SIGNALS OF THE SAME PHASE AND INDICATED FREQUENCY RATIO
11:2
47:1
3:2
(C) ROULETTE FIGURES FORMED BY TWO SIGNALS OF THE SAME PHASE AND INDICATED FREQUENCY RATIO

Fig. 4-1.

4

EXTERNAL HORIZONTAL PREAMPLIFIERS

There are a number of useful applications for the oscilloscope in which two independent variables are compared by applying one of them to the vertical deflection system and the other to the horizontal system. When used in this way, the oscilloscope is said to be operated in the *X-Y* (rather than *Y-T*) mode. In most cases, the signal characteristics of interest are phase, frequency and amplitude.

It is not the purpose of this chapter to explore the *X-Y* technique nor list its many applications; however, a few principles of the technique must be mentioned in order to establish the requirements it imposes on the horizontal amplifier. When using the normal (*Y-T*) mode of operation, the observer is usually interested in the dimensions and shape of a single waveform, or in a comparison of two or more waveforms as they occur in respect to a given point in time. When using the *X-Y* mode, the phase and frequency of two signals are of primary concern. By the application of one signal to the horizontal and another to the vertical deflection system of the oscilloscope, characteristic patterns are formed which can be interpreted in terms of the relative frequency, phase and amplitude of the two signals (Fig. 4-1).

phase, frequency and amplitude

It is clear that such displays will be meaningful only if the same phase delay, deflection factor and amplifier bandpass are presented to the input signals. Otherwise the display would represent not the true relationship between the signals, but their relationship at the deflection plates of the CRT, which would include the errors introduced by the difference in amplifier parameters.

A very few instruments are designed especially for *X-Y* displays. These offer matched vertical and horizontal amplifiers as well as CRT's with similar horizontal and vertical deflection factors. Most instruments, however, include the *X-Y* capability as an auxiliary mode of operation. These instruments must include circuits which adapt the instrument for the *X-Y* mode of operation.

X-Y auxiliary mode

The first adaptations which must be made pertain to the relative gain of the vertical and horizontal deflection systems. In a conventional oscilloscope, the vertical amplifier is designed to amplify very small, fast-risetime (high-frequency) signals. The horizontal amplifier, on the other hand, usually processes a fairly high-amplitude and much slower-risetime signal. Furthermore, to produce the same degree of beam deflection from a given signal amplitude, the horizontal amplifier must have a gain several times that of the vertical amplifier, since the horizontal deflection factor of the CRT is several times larger (requires more volts per division) than the vertical deflection factor. In terms of gain and bandwidth, the vertical amplifier is superior to the horizontal amplifier by several orders of magnitude.

relative-gain deflection systems

deflection-system preamplifier

This situation can be improved to some extent by the addition of a preamplifier to the horizontal deflection system. When the instrument is switched to the *X-Y* mode, the external horizontal (*X*) signal is applied to the preamplifier, whose output is connected to the main amplifier input circuits (Fig. 4-2). Horizontal-gain potentiometer R1 makes it possible to bring the vertical and horizontal input deflection factors into correspondence.

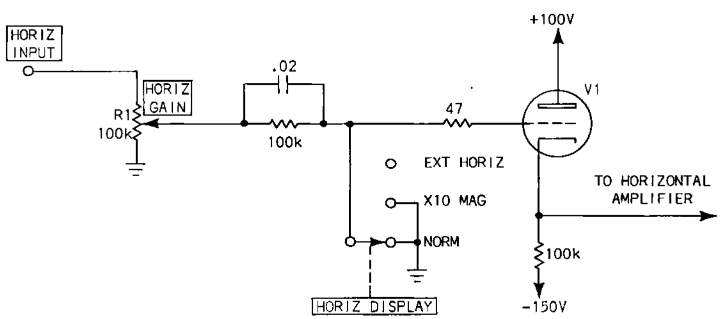


Fig. 4-2. Simple external horizontal preamplifier.

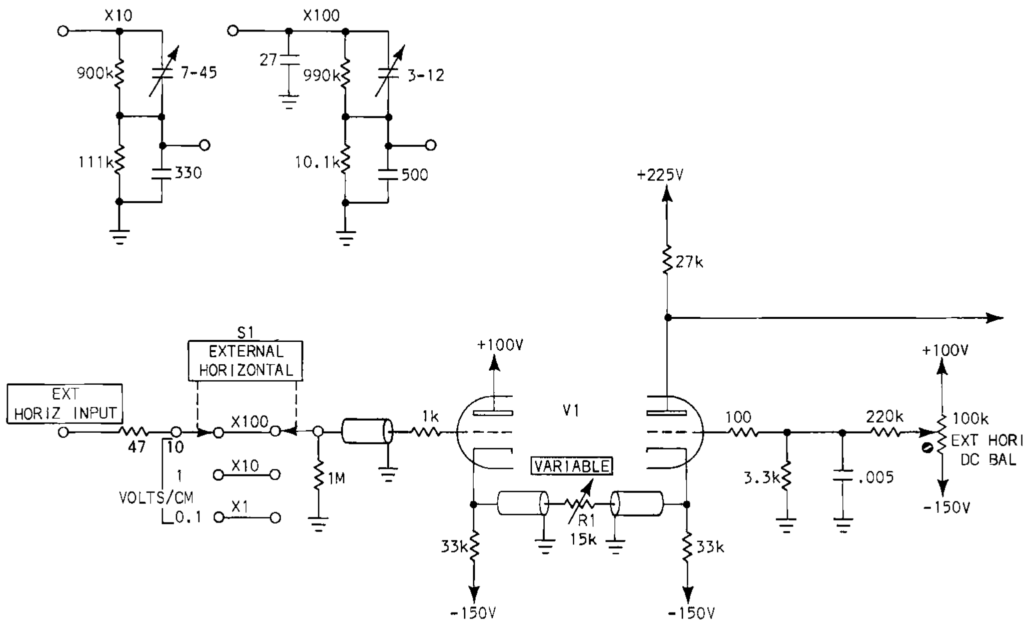


Fig. 4-3. External horizontal preamplifier with calibrated volts/division.

A signal generator or calibrating fixture is necessary for this adjustment, since the gain control itself is not calibrated. However, the preamplifier shown in Fig. 4-3 is more convenient. Two frequency-compensated input attenuators, providing X100 and X10 factors, together with a straight-through connection for X1 operation are selectable by means of the EXTERNAL HORIZONTAL switch (S1). The circuit is calibrated so that when the VARIABLE potentiometer (R1) is in its detent position, the deflection factor may be read directly in volts per division. Thus the vertical and horizontal amplifiers can be adjusted for the same deflection factor without repeated calibration.

In a few cases, especially in later instruments, preamplification of the external horizontal signal is accomplished in the trigger circuits. Fig. 4-4 is a schematic diagram of a trigger preamplifier (part of the trigger generating system) whose primary function is to amplify the output of the trigger pickoff circuit before it is applied to the A and B trigger generators. Note that TRIGGER switch S1 has two positions: Normal and *X-Y*. In the NORMAL mode, the input signal for the preamplifier comes from the trigger pickoff circuit. In the CH 1 ONLY or *X-Y* position, the input comes from the channel 1 amplifier. When used in the *X-Y* mode, the external horizontal signal is simply connected to the channel-1 vertical-input terminal. Gain of the preamplifier is about 10. Gain-adjust potentiometer R1 is a part of the input resistance for the following operational (horizontal) amplifier when connected to it by the horizontal display switch (see page 76). The combined gain of the two amplifiers is thus variable and may be calibrated to match the selected deflection factor of the vertical amplifier.

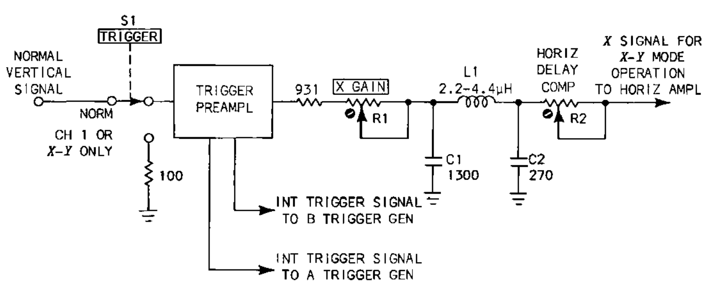


Fig. 4-4. Trigger preamplifier used as an external horizontal preamplifier.

This circuit also introduces an adjustment to compensate for differences in signal-propagation delay between the horizontal and vertical amplifiers. Like most wide-bandwidth instruments, the instrument from which this example was taken incorporates a

delay

delay line in the vertical amplifier. The delay in signal propagation is deliberately introduced in the vertical amplifier to give time for the horizontal deflection system to initiate a sweep. In *X-Y* operation however, as described earlier, it is very important that both vertical and horizontal deflection systems impose the same delay on the *X* and *Y* signals, so that their phase relationship be preserved. The

delay compensation

delay compensation network, consisting of C1, C2, L1 and R2, is actually a π-section filter whose response approximates a fixed delay. This delay (about 110 nanoseconds) corresponds to the delay between the vertical and horizontal amplifier in normal (*Y-T*) operation. R2 and L1 are adjusted at middle and high frequencies respectively for minimum phase shift at those frequencies. The effect of this compensation network is to raise the maximum frequency of *X-Y* operation from the hundred-kilohertz range of ordinary instruments to about 2 MHz. Beyond this frequency the π-section filter no longer approximates a fixed delay.

frequency range

The third problem which is encountered in *X-Y* operation, that of the relative frequency response of the vertical and horizontal amplifiers, cannot be solved by preamplifiers or other auxiliary circuits. Because the horizontal-deflection plates of the CRT have a much higher deflection factor (poorer sensitivity), the horizontal amplifier would be required to exhibit several times the gain required of the vertical amplifier for signals of equal amplitude. Since high gain and wide bandwidth are mutually antagonistic, the upper frequency limit of meaningful *X-Y* operation in conventional oscilloscopes is set by the bandwidth of the horizontal amplifier. As mentioned earlier, this limit is typically on the order of a hundred kilohertz. Thus the only way to increase the frequency range of *X-Y* mode operation is to build a horizontal amplifier especially for this purpose. In most cases the extra expense and increased circuit complexity required for this effort are not justified, since the applicability of *X-Y* techniques of measurement is quite limited.

NOTES

NOTES

INDEX

BOOKS IN THIS SERIES:

CIRCUIT CONCEPTS

title	*part number*
Digital Concepts	062-1030-00
Horizontal Amplifier Circuits	062-1144-00
Oscilloscope Cathode-Ray Tubes	062-0852-01
Oscilloscope Probe Circuits	062-1146-00
Oscilloscope Trigger Circuits	062-1056-00
Power Supply Circuits	062-0888-01
Sprectrum Analyzer Circuits	062-1055-00
Storage Cathode-Ray Tubes and Circuits	062-0861-01
Sweep Generator Circuits	062-1098-00
Television Waveform Processing Circuits	062-0955-00
Vertical Amplifier Circuits	062-1145-00

MEASUREMENT CONCEPTS

Automated Testing Systems	062-1106-00
Engine Analysis	062-1074-00
Information Display Concepts	062-1005-00
Probe Measurements	062-1120-00
Semiconductor Devices	062-1009-00
Spectrum Analyzer Measurements	062-1070-00
Television System Measurements	062-1064-00