



# Oscilloscope Probe Circuits



Circuit Concepts Series

# OSCILLOSCOPE PROBE CIRCUITS

BY  
JOE WEBER



CIRCUIT CONCEPTS

FIRST EDITION FIRST PRINTING NOVEMBER 1969

062-1146-00



udu 621-317/18-1  
621-317-75



Inv. # 3274/73-D

© TEKTRONIX, INC.; 1969  
BEAVERTON, OREGON 97005  
ALL RIGHTS RESERVED

## CONTENTS

	PREFACE	1	
1	INTRODUCTION	3	
2	PASSIVE VOLTAGE PROBES	9	
3	$Z_0$ PROBES	29	
4	ACTIVE VOLTAGE PROBES	39	
5	CURRENT PROBES	55	
6	DIFFERENTIAL PROBES	91	
7	HIGH VOLTAGE PROBES	107	
	INDEX	113	

## PREFACE

The oscilloscope is an invaluable tool to workers in many fields of science. This tool requires proper use and maintenance for trouble-free operation. How much knowledge is necessary to obtain trouble-free operation? Routine use and maintenance do not require an intimate knowledge of the working mechanism of any machine. Many drivers have no idea of how an automobile works. A mechanic can learn when and how to change spark plugs without ever understanding the principles that determine the operation of a spark plug. However, to obtain the *maximum* benefit from a machine, we must have a thorough knowledge of its operation. The specification sheet is our guide to efficient use and maintenance.

The purpose of this book is to convey the understanding necessary to obtain the maximum benefits from one particular portion of the oscilloscope, the probe. We will discuss the mechanical and electrical make-up of oscilloscope probes so that the reader will understand how the specification sheet is derived.

Before you read this book, we expect you to understand vertical amplifier circuitry, especially attenuator and input active device circuits. Other areas of vertical amplifiers depend on your interest. That is, if you are interested in differential measurements, a knowledge of differential amplifiers is desired to better appreciate differential probes. We include a discussion of sampling probes which is limited to those probes that are designed to couple the signal of interest to the sampling head circuits. We will not discuss probes that contain sampling head circuitry as these are covered in a separate book. A knowledge of sampling principles is not necessary to understand the probes covered in this book.

We hope the reader will not only understand probe circuits, but also why these circuits are needed. The reader should also be able to relate the calibration adjustments to the specification sheet and know when and how to calibrate the probe. The reader should be capable of interpreting the specification sheet and data curves to determine when the measurement conditions are marginal and when the probe may be damaged. While this is primarily a circuit concepts book, some measurement information is given. Our intent is to show how measurement information is derived from probe circuitry. A companion volume *Probe Measurements*, explains how to use this information for measurement results.

1

## INTRODUCTION

"Is the probe compensated?"

"Of course," I replied, "I've been using this bench set-up for the past six months and my probe is ALWAYS compensated."

"That's nice," replied the staff engineer. "Your bench scope was taken for recal this morning and you have a new one. Did you check probe compensation?"

"No, sir," I replied and added lamely, "but I will."

After the probe was properly compensated to the oscilloscope, that strange, unexpected waveform was transformed to the regular signal I had expected to find.

One of the first lessons usually given to an oscilloscope operator is the need for proper probe compensation and how it is accomplished. But, it is easily overlooked and nearly everyone is embarrassed at least once because he forgot to compensate the probe. With the sophisticated test equipment used today, compensation is only the first step in obtaining accurate measurements. Since the early 1950's, the oscilloscope has evolved from a crude tool of limited use to a sophisticated machine with wide performance capability. The oscilloscope probe has evolved in similar fashion. Thus today, Tektronix customers can purchase probes whose performance range matches the performance range of the oscilloscope.

evolution

As we would expect, Tektronix produces a wide variety of probes to meet the requirements of the many general purpose and special application oscilloscopes manufactured by Tektronix. And probe circuitry is becoming more sophisticated to insure that the probe does not degrade oscilloscope performance. This more sophisticated circuitry requires calibration just as the oscilloscope circuitry requires calibration. In many cases, the probe-oscilloscope combination must be calibrated as a measurement system to insure measurement accuracy.

Tektronix probes can be divided into three general classes:

- (1) *Voltage sensing.* These probes are designed to provide a high impedance in shunt with the circuit under test to monitor circuit voltages as shown in Fig. 1-1. Voltage measurements are the most frequent because they are usually the easiest to make. Thus, Tektronix produces more voltage probes than any other type of probe.

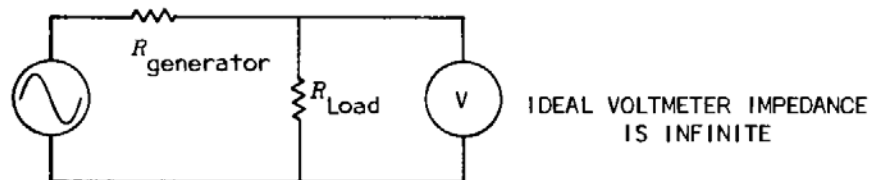


Fig. 1-1. Voltage measurement.

- (2) *Current sensing.* These probes are designed to provide a low impedance in series with the circuit under test to monitor circuit current as shown in Fig. 1-2. Tektronix current probes are designed to monitor circuit current without a direct electrical connection to the circuit under test. The development of current driven devices has increased the need for current measurements although these measurements are not as easily accomplished as voltage measurements.



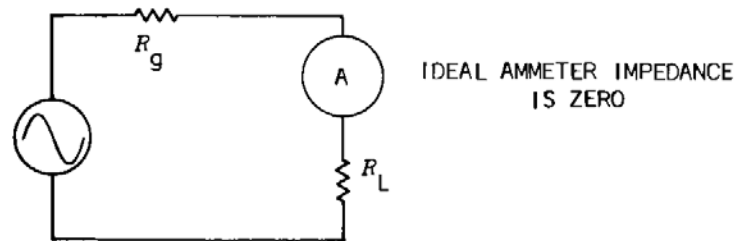


Fig. 1-2. Current measurement.

- (3) *Power transfer.* These probes are designed to provide an impedance which matches the impedance of the circuit under test as shown in Fig. 1-3. Impedance matching techniques are commonly used for processing high frequency signals with minimum reflections. We will not discuss  $Z_0$  cables as signal transfer mediums because an abundance of this information exists. We will discuss probes that transfer the signal from high impedance environments to low impedance environments, say from 1 k $\Omega$  to 100  $\Omega$ .

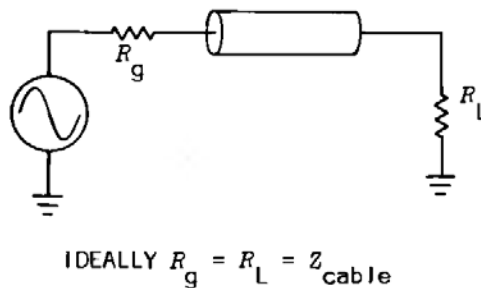


Fig. 1-3. Power transfer — ideal voltmeter or ammeter monitors  $R_L$ .

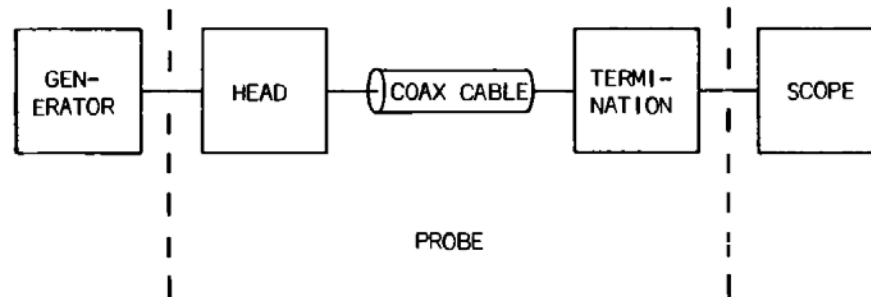


Fig. 1-4. General probe block diagram.

Fig. 1-4 is a general block diagram which is applicable to all Tektronix probes. The probe head contains the signal sensing circuitry. This circuitry may be passive, such as a  $10\text{ M}\Omega$  resistor shunted by  $7\text{ pF}$  or a 125 turn transformer secondary or it may be active, such as a Source Follower or a Hall Generator. A coaxial cable is used to couple the output of the probe head to the termination. The type of coax cable used depends on the type of probe.

termination

The termination has two functions: (1) terminate the cable in its characteristic impedance and (2) provide the source impedance required by the oscilloscope. The termination may be passive or active circuitry. The generator is any circuit to which the probe is connected. Thus we use generator for the circuit under test or investigation, for a specific signal generator used during calibration, or what have you. The output impedance of the generator is usually complex, that is, it contains resistance, capacitance and inductance. For the most part, we will ignore the affects of the generator output impedance. This subject matter is discussed in *Probe Measurements*.

$t_r$  and  
frequency  
response

We will refer to some portions of the calibration procedure that are performed with signal generators whose output impedance is a specified resistance. The Tektronix Type 106 Square-Wave Generator and the Type 191 Constant Amplitude Signal Generator have a specified 50  $\Omega$  output resistance. These generators terminated in their output resistance provide the 25  $\Omega$  source impedance required for checking risetime and frequency response of Tektronix instruments.

We will use the terms signal, input signal or output signal to refer to that "thing" which the probe couples to the oscilloscope for processing and display. We will also use the term, common-mode signal, to describe the identical signals that are applied to both input terminals of a differential system. Any other "signals" that we discuss will have some descriptive modifier attached to signal, such as switching signal and so on.

The three general classes of probes are too bulky for use as subject headings. Ergo, we will divide the subject matter as follows:

In Chapter 2 we discuss passive voltage probes, the most popular type of probe. We will pay particular attention to the resistive center conductor coaxial cables used with these probes and the requirements for terminating such cables.

A special form of passive voltage probe, the  $Z_0$  probe, is discussed in Chapter 3. These probes are essentially impedance transforming probes and are designed to have a low input capacitance. Probes of this type are primarily used in high frequency measurements.

Chapter 4 is devoted to active voltage probes. These probes are designed to present a higher input impedance than the passive voltage probe with less signal attenuation. Active termination circuits are introduced here.

Current probes, active and passive, are discussed in Chapter 5. The active and passive termination associated with these are included here. A brief discussion of Hall Effect precedes the explanation of the DC current probe.

In Chapter 6 we discuss differential probes, active and passive. We will note the differences between these probes and the probes designed for single-ended signals.

High voltage probes are discussed in Chapter 7. Here we examine the special techniques required to couple kilovolt signals to conventional oscilloscopes.

## 2

## PASSIVE VOLTAGE PROBES

advantages

The passive voltage probe is the most popular device for coupling the signal of interest to the oscilloscope. The passive voltage probe is convenient to use, maintains the accuracy of the oscilloscope ( $\pm 3\%$ ), has a wide dynamic range and is inexpensive. The attenuating passive voltage probes present a high impedance to the circuit under test which minimizes probe loading. Passive voltage probes are fairly rugged and will tolerate more abuse, mechanical and electrical, than other types of probes. (We are not endorsing probe abuse.) Another advantage of the passive voltage probe is the circuit simplicity and ease of calibration and maintenance. Admitted, later model probes have more complex termination circuitry but the passive voltage probe is still less complicated than its counterpart in any other type of probe.

disadvantages

The passive voltage probe is not without its disadvantages. The input impedance of the passive voltage probe will decrease as the applied frequency is increased. We can get a higher input impedance if we are willing to accept greater signal attenuation. Thus to minimize source loading, we must tolerate a smaller signal.

The passive voltage probe is so labeled because:

- (1) active devices are not contained in the probe head or termination circuitry and
- (2) it is normally connected where the probe input impedance is greater than the source impedance. Thus the circuit under test will appear as a voltage source to the oscilloscope.

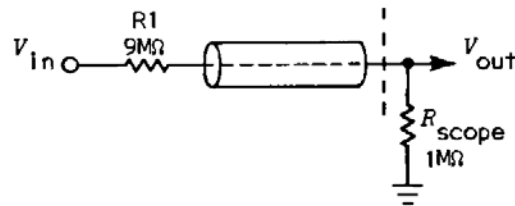
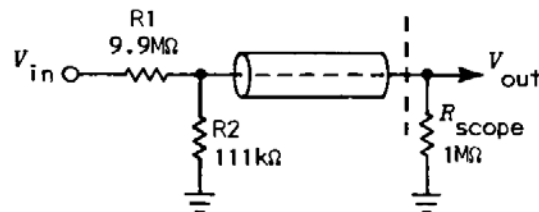


Fig. 2-1. 10X DC probe.

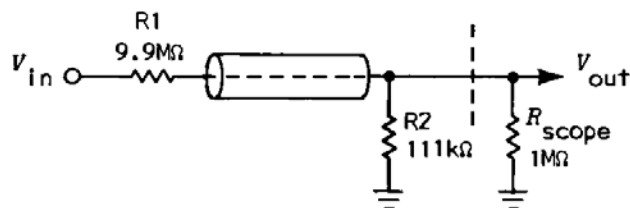
DC  
requirements

We will begin the analysis of passive probes by considering what we need for DC signals. We assume the oscilloscope input is 1 MΩ shunted by 20 pF unless otherwise stated. A 1X probe is trivial at DC, thus we begin with the 10X probe of Fig. 2-1. The probe is essentially a 9 MΩ resistor, R1, and a length of insulated wire for the probe cable. The probe-oscilloscope combination is a DC voltage divider whose transfer characteristic is given by:

$$V_{out} = \frac{(V_{in}) R_{scope}}{R1 + R_{scope}}$$



A



B

Fig. 2-2. 100X DC probe.

Two versions of a 100X probe are shown in Fig. 2-2A and B. In Fig. 2-2A, the probe head consists of a 9.9 M $\Omega$  resistor, R1, and a 111 k $\Omega$  resistor, R2. Since R2 parallels the oscilloscope input resistance,  $R_{scope}$ , the probe-oscilloscope combination is again a DC voltage divider. The transfer characteristic is given by:

$$V_{out} = V_{in} \frac{\frac{R2 \cdot R_{scope}}{R2 + R_{scope}}}{R1 + \frac{R2 \cdot R_{scope}}{R2 + R_{scope}}}$$

In Fig. 2-2B, R2 is located in a termination box which is between the probe cable and the oscilloscope input. The DC input resistance of the 10X and 100X passive probes is 10 M $\Omega$ . The passive probe considerations for DC signals are straightforward and shall not be belabored.

The probe-oscilloscope combination should be capable of processing a wide range of signals equally well. Thus, a 10X divider at DC should also be a 10X divider at the 30% down frequency of the oscilloscope and should be a 10X divider to a pulse with a fast risetime. We need a compensated voltage divider similar to the vertical amplifier attenuator. The probe cable must be shielded to isolate the probe from any radiated energy and to prevent the probe cable from acting like an antenna.

We proceed to convert the 10X DC probe to a 10X probe which is suitable for processing a wide range of signals. As a first step, a thin, flexible 50  $\Omega$  coax is selected to replace the insulated wire for the probe cable. (A 75  $\Omega$  or a 100  $\Omega$   $Z_0$  coax could have been selected without altering the results.)

converting  
to wide  
range

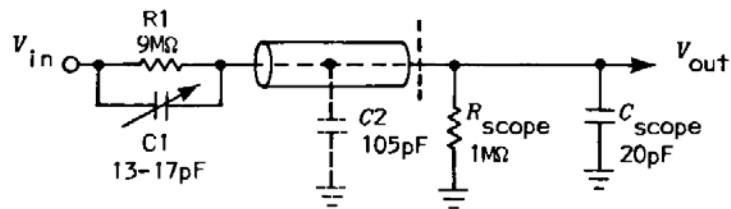


Fig. 2-3. Compensated 10X probe.

compensated  
divider

Now the resistor in the probe head is shunted by an adjustable capacitor to compensate the voltage divider. The situation is shown in Fig. 2-3. To obtain a properly compensated voltage divider, the time constant of the input network,  $R_1C_1$ , must equal the time constant of the output network,  $R_{scope}(C_2 + C_{scope})$ . (Compensated voltage divider theory is thoroughly discussed in *Oscilloscope Vertical Amplifiers*.)  $C_{scope}$  is the oscilloscope input capacitance and  $C_2$  is the capacitance of the cable. Typically, the capacitance per foot for 50  $\Omega$  coax is 30 pF. For a 3.5 foot cable, add 105 pF to the oscilloscope input capacitance!

The time constant of the output network is:

$$TC_{out} = 1 \text{ M}\Omega (105 \text{ pF} + 20 \text{ pF}) = 125 \text{ }\mu\text{s}.$$

For  $R_1 = 9 \text{ M}\Omega$ ,  $C_1$  must be:

$$C_1 = TC_{out}/R_1 = 125 \text{ }\mu\text{s}/9 \text{ M}\Omega = 13.88 \text{ pF},$$

since  $TC_{in}$  must equal  $TC_{out}$  to have a compensated voltage divider.

The input capacitance of an oscilloscope can be from 15 pF to about 50 pF depending on the individual instrument. Thus  $C_1$  must be adjustable from about 13 pF to 17 pF. The total capacitance shunting  $R_1$  is greater than  $C_1$ . Environmental capacitance must also be included. This environmental capacity will depend on the type of probe tip (add 1 pF to 2 pF for a pincer tip) and the type of probe head construction.

The circuit of Fig. 2-3 is nominally 10  $\text{M}\Omega$  shunted by 14 pF. As the frequency of the input signal increases, source loading is increased because  $C_1$  shunts  $R_1$ . If the probe of the circuit in Fig. 2-3 were increased to 6 feet,  $C_1$  must be 7.5 pF larger



to compensate the voltage divider. The capacitance would be greater than 20 pF, the input capacity of the oscilloscope!

The capacitive loading can be reduced by using a 100X probe as shown in Fig. 2-4. The output time constant of the 100X probe with 3.5 foot cable is given by:

$$TC_{out} = R_{eq} (C_2 + C_{scope}),$$

where  $R_{eq}$  is the shunt combination of  $R_2$  and  $R_{scope}$ .

$$TC_{out} = 100 \text{ k}\Omega (20 \text{ pF} + 105 \text{ pF}) = 12.5 \text{ }\mu\text{s}.$$

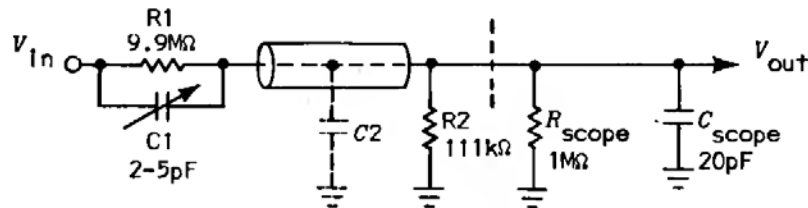


Fig. 2-4. Compensated 100X probe.

The time constant is reduced by a factor of 10 which means  $C_1$  can be reduced by 10.

$$C_1 = \frac{TC_{out}}{R_1} = \frac{12.5 \text{ }\mu\text{s}}{9.9 \text{ M}\Omega} = 1.26 \text{ pF}.$$

Again the environmental capacity of the probe head must be added to  $C_1$  to obtain the capacitance which shunts  $R_1$ , say 3 pF. By using a 100X probe we markedly reduce loading, however, the signal is smaller because of the increased attenuation. For any given measurement, we must weigh the relative advantages-disadvantages of the 10X and 100X probes.

The non-attenuating 1X probe with a 50  $\Omega$  coax cable is essentially a large shunting capacitance with an input terminal that is located 3.5 to 9 feet from the oscilloscope input. The 1X probe presents an extremely large load to high speed signals and therefore, is normally limited to such measurements as power supply ripple with the oscilloscope input AC coupled.

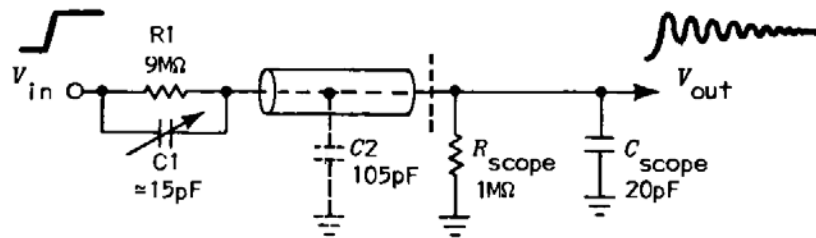


Fig. 2-5. Response to fast rise pulse.

I have ignored one aspect of lossless  $50\ \Omega$  cables. A lossless cable, as its name implies, is a circuit with a high  $Q$ . A high  $Q$  circuit will oscillate when it is excited by a burst of energy such as a fast rise pulse. The situation is shown in Fig. 2-5. The output is a series of sine waves with a frequency around 60 MHz for a 3.5 foot  $50\ \Omega$  cable. These oscillations, often called ringing, will eventually damp if no further energy is applied to the cable. If the bandwidth of the oscilloscope is relatively low, say 30% down at 5 MHz or less, the ringing will not be seen on the CRT display. If the 30% down frequency is 10 MHz or more, the oscilloscope is capable of processing the ringing. Then the CRT display will include the distortion introduced by probe cable.

We find the circuit of Fig. 2-3 is unacceptable for measurement systems with 30% down frequencies greater than 10 MHz. About a generation ago, 10 MHz was the state-of-the-art for high speed oscilloscopes. The design of passive probes had to change to provide an acceptable device for coupling signals to these oscilloscopes. We can suppress the oscillations by adding resistance to the circuit which lowers circuit  $Q$ . Since we want good transient response, the resistance must provide critical damping. Ergo, let us place an unbypassed resistor in series with the cable. The location of the unbypassed resistor will affect transient response. Placing the resistor at the oscilloscope end of the cable results in a slow transient response. Placing the resistor at the probe end of the cable or using equal value resistors at both ends of the cable gives some improvement in transient response.

We can obtain the resistance required for critical damping in another manner. Let us replace the high conductivity center conductor of the 50  $\Omega$  coax with a resistance wire such as a nickel-chromium. Assume the resistance required for critical damping is 350  $\Omega$ . Then a 100  $\Omega$ /foot wire will satisfy the resistance requirement of a 3.5 foot cable. We now label this coax as an  $R$  cable to differentiate with a 50  $\Omega$  or  $Z_0$  cable.<sup>1</sup> When a  $Z_0$  cable is replaced with an  $R$  cable, critical damping is realized and we will have a faster risetime than is obtained with series damping resistors.

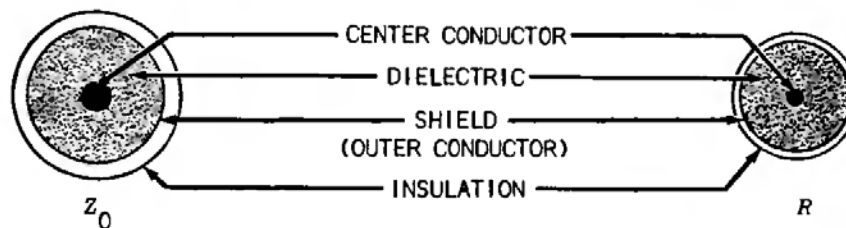


Fig. 2-6. Comparison of  $Z_0$  and  $R$  cable.

Let us compare  $R$  cable with  $Z_0$  cable. Fig. 2-6 shows an end view of both cables. The center conductor of the  $R$  cable has the smaller diameter. A strand of resistance wire is stronger than an equal size strand of conductance wire. Thus we can reduce the center conductor diameter and retain mechanical strength. Since the center conductor diameter is decreased, we can also reduce the diameter of the outer conductor. Thus we can make a thinner, more flexible cable with resistance wire.

Another advantage is obtained when we reduce the size of the conductors. The capacitance of the cable decreases. In Tektronix probe cables capacitance is from 6 pF/ft to 9 pF/ft. We will use 8 pF/ft for ease of computation and appropriately change the circuit of Fig. 2-3 to that shown in Fig. 2-7.

1.  $R$  cable is also called lossy cable to differentiate from lossless or  $Z_0$  cable.

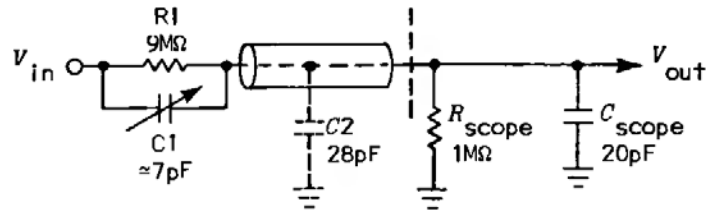


Fig. 2-7. Compensated 10X probe with 8 pF/ft  $R$  cable.

Let us determine the required value of  $C_1$ . For a 3.5 ft cable with 8 pF/ft,  $C_2$  is 28 pF.  $C_{\text{scope}}$  and  $C_2$  shunt  $R_{\text{scope}}$  by 48 pF, however, about 2 pF should be added for connector capacitance.  $C_1$  can reasonably be approximated at 5.5 pF. Again add about 1.5 pF for probe tip capacity to obtain 7 pF at the probe tip. The capacity shunting  $R_1$  has been reduced by a factor of 2 (from 14 pF to 7 pF) by the use of  $R$  cable!

We must resort to transmission line theory to analyze the response of this passive circuit to high speed signals. To obtain the best results from any transmission line, we should terminate the line in its characteristic impedance. The characteristic impedance of a coaxial cable is given by:

$$Z_0 = \sqrt{\frac{R_{\text{series}} + j \omega L}{G_{\text{shunt}} + j \omega C}}$$

In a lossless cable,  $G_{\text{shunt}}$ , the dielectric loss, is insignificant and the series resistance of the high conductance wire is also insignificant. Letting  $R$  and  $G$  go to zero, we obtain the familiar formula for the characteristic impedance of a lossless line.

$$Z_0 = \sqrt{L/C}$$

In a lossy line, the dielectric losses can also be ignored but the series resistance is significant. Thus, letting  $G_{\text{shunt}}$  go to zero and rearranging the terms we have:

$$Z_0 = \sqrt{\frac{j \omega L + R_{\text{series}}}{j \omega C}}$$

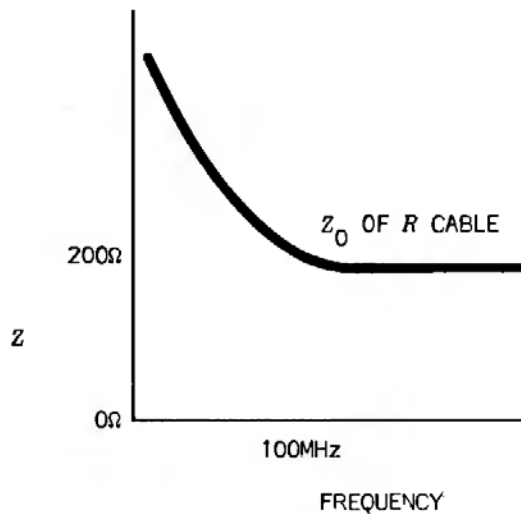


Fig. 2-8.  $Z_0$  of  $R$  cable vs frequency.

The term on the left reduces to the familiar  $\sqrt{L/C}$  which is independent of frequency. The term on the right varies with frequency. As frequency increases,  $j\omega C$  increases. Thus we state that  $Z_0$  goes to  $\sqrt{L/C}$  as frequency goes to infinity. In practice,  $Z_0$  approximates  $\sqrt{L/C}$  when the applied frequency is greater than 100 MHz. See Fig. 2-8. For Tektronix probe cables, the  $\sqrt{L/C}$  ranges from 175  $\Omega$  to 200  $\Omega$ .

The response of  $R$  cable to a step function input is shown in Fig. 2-9. The output signal is an attenuated fast rise followed by a slow "dribble up" to the amplitude of the input step signal. The amplitude of the fast rise portion is inversely proportional to the resistance of the center conductor. We note some ringing exists on the dribble-up portion of the output signal because the cable is not terminated. For oscilloscopes with 30% down frequencies of 30 MHz or less, an unbypassed resistor in series with the cable at the probe head will meet the requirements of a termination. See Fig. 2-10B.

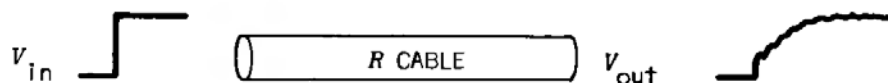
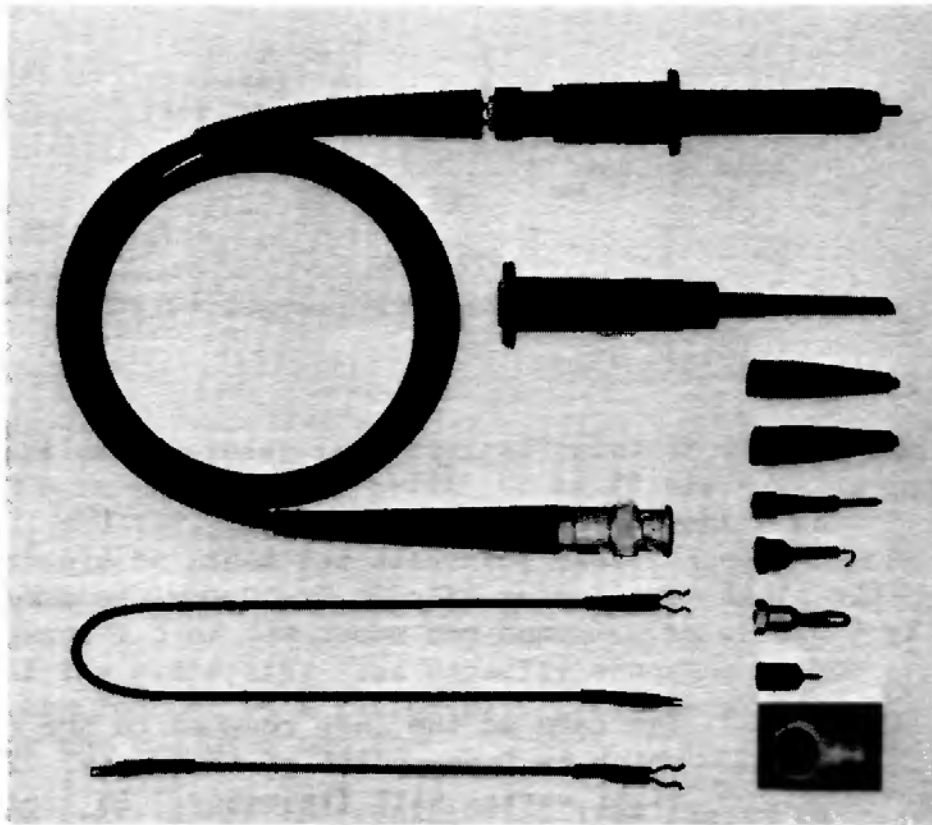
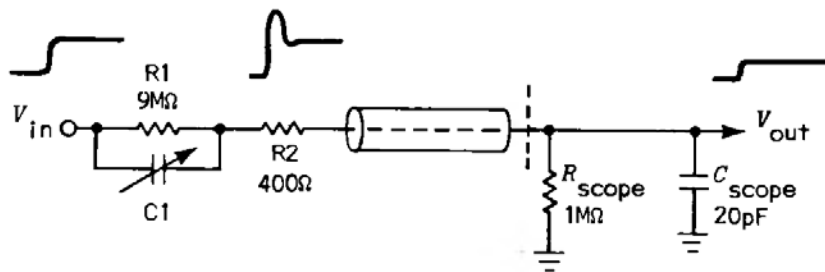


Fig. 2-9. Response of  $R$  cable to fast rise pulse.



A



B

Fig. 2-10. 10X probe with  $t_r \approx 6.6$  ns.

The only adjustment is the compensation capacitor, C1. C1 is adjusted for flat response to a 1 kHz calibrator signal as shown in Fig. 2-11A and B. The unbypassed resistor causes some "spiking" at the front corner which is attenuated in the lossy R cable. The signal at the oscilloscope input is an attenuated pulse.

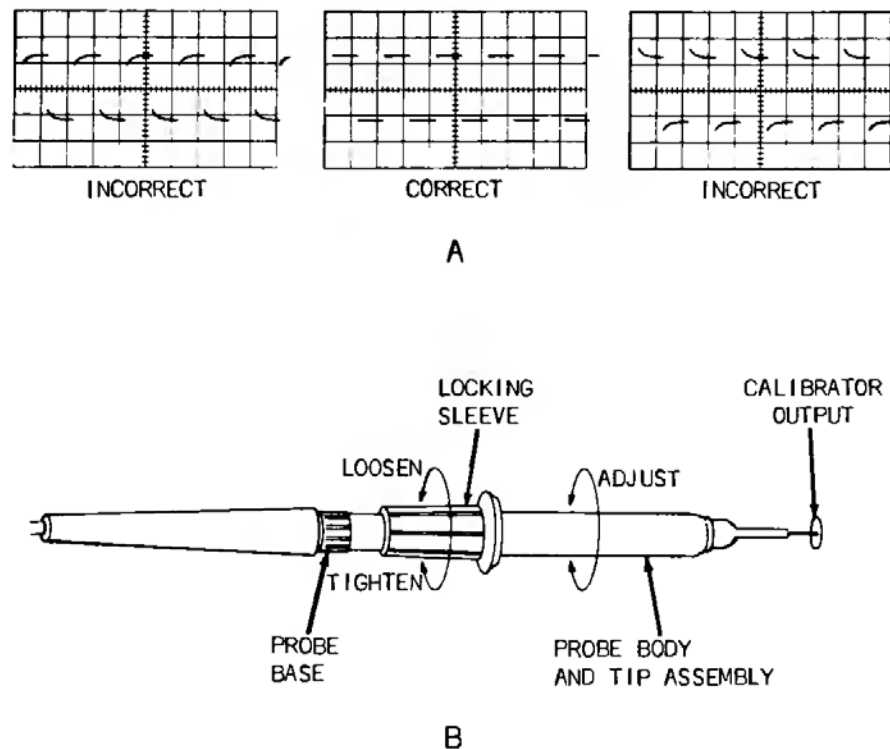
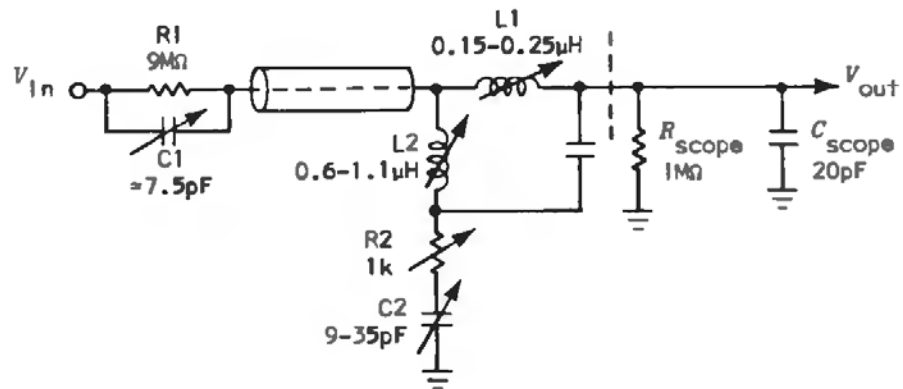


Fig. 2-11.

The probe circuit of Fig. 2-10B is not acceptable for use with oscilloscopes that have 30% down frequencies higher than 30 MHz. A more sophisticated termination is required. A 10X probe with risetime faster than 3 ns is shown in Fig. 2-12A and B. The termination circuitry is contained in the termination box at oscilloscope end of the probe cable. The compensation capacitor is in the probe head and is adjusted by rotating the probe base as shown in Fig. 2-11B. The termination resistor,  $R_2$ , is adjusted for minimum ringing when a fast rise pulse ( $t_r < 1$  ns) is applied.  $C_2$  blocks DC signals to maintain a high impedance and, with  $L_1$ ,  $L_2$  and  $C_3$ , terminates the reactive portion of the cable impedance throughout the performance range of the probe.



A

10X PROBE WITH INDUCTIVE PEAKING  $t_r \leq 3$  ns

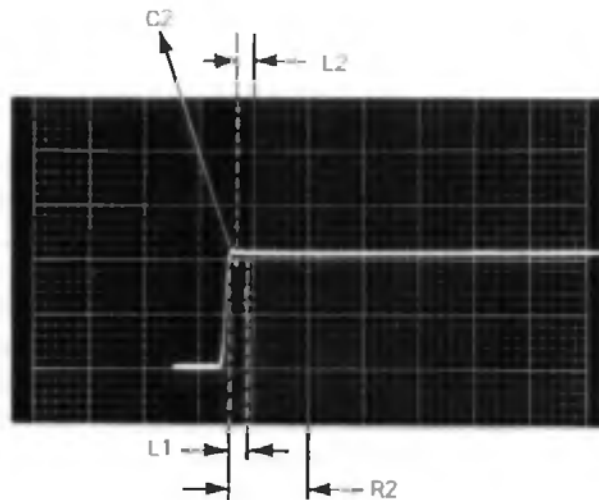
B

Fig. 2-12.

C2 introduces some rolloff at the front corner (see Fig. 2-13) which is cancelled by adjusting L1 for the best leading edge and L2 for the portion of waveform following leading edge. After the



transient response is calibrated, compensation should be rechecked because an adjustment of C2 changes the capacitance shunting  $R_{\text{scope}}$ , the oscilloscope input resistance. If C1 requires adjustment, the transient response must be checked again.



L1 AND L2 AFFECT PORTION OF WAVEFORM TO LEFT OF INDICATED GRATICULE LINE.

R2 -- DAMP RINGING IN FIRST TWO MAJOR DIVISIONS

C2 -- SPIKES LEADING EDGE AND AFFECTS PULSE AMPLITUDE

Fig. 2-13. Response of P6008-580 to pulse.

Probe risetime is less than 3 ns and this probe can be used with a 50 MHz oscilloscope with little performance degradation. If the probe is used with an 80 MHz oscilloscope, probe-oscilloscope bandwidth is on the order of 66 MHz. A 100X version of this probe with 9 ft cable is shown in Fig. 2-14.

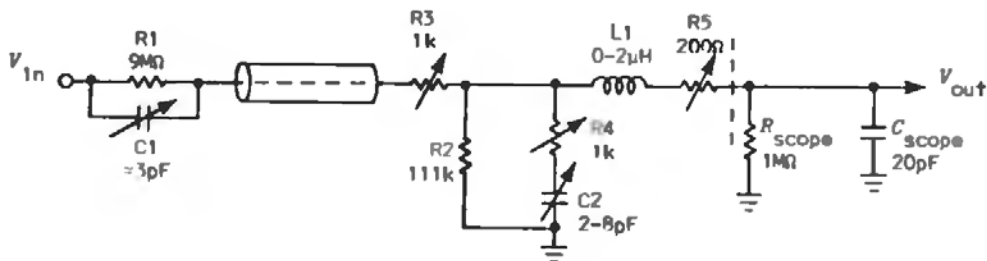


Fig. 2-14. 100X probe with  $t_r \leq 2$  ns.

R5 is adjusted for minimum ringing and R2 shunts the oscilloscope input to get the required DC voltage division. The need for inductive peaking is minimized because C1 is small. L1 corrects for the dip following the leading edge introduced by blocking capacitor C2. R3 is adjusted for best leading edge and R4 is adjusted for overall flattop of the pulse waveform. The 100X probe risetime is about 2 ns and the probe can be used with 100 MHz oscilloscopes without performance degradation.

slim body  
probes

The circuits discussed previously are used in medium size probe heads. These probes can be compensated for use with oscilloscopes whose input capacitance is from 15 pF to 47 pF. The medium body probes can be awkward in high density transistor circuits. Slim body probes were introduced to provide a more convenient tool. The small size of the slim body probe made it difficult to place an adjustable capacitor in the probe head. The compensation capacitor was placed in the termination box. A schematic and photograph are shown in Fig. 2-15.

This probe has a series termination for high speed signals and a shunt compensation capacitor, C3. R2 is the termination resistance which in conjunction with C3 tends to roll off the front corner.

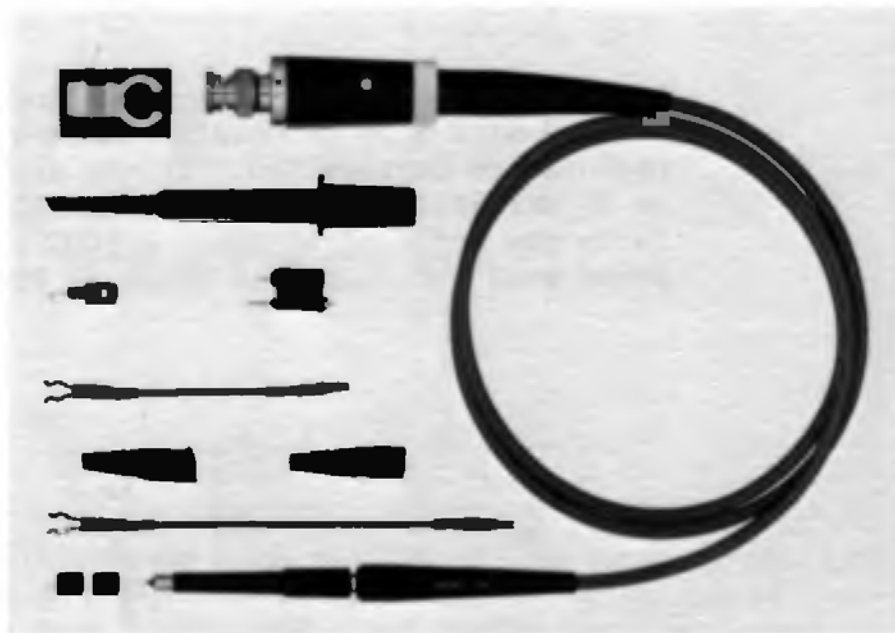


Fig. 2-15A.

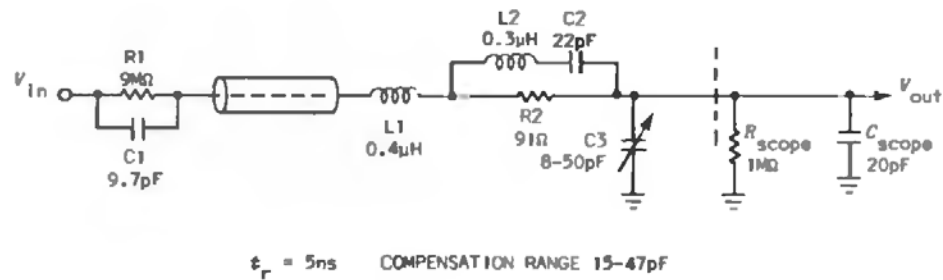


Fig. 2-15B.

L1, L2 and C2 provide peaking to cancel this rolloff. The primary disadvantage of this circuit is the need for a larger probe head capacitor to maintain a compensated voltage divider. The total capacity in shunt with the oscilloscope input resistance (the cable capacitance, the oscilloscope input capacitance and the termination capacitance) is increased by the compensation capacitor. Note the value of C1 has increased to 9.7 pF to maintain compensation. This probe circuit does have a risetime faster than 5 ns without requiring any high frequency calibration. The only adjustment is the voltage divider compensation, C3. The increase in capacitance at the oscilloscope end of the probe does limit the best risetime we can obtain. Some improvement in risetime can be obtained by using a more complex termination which requires calibration. In general, high speed oscilloscopes, 50 MHz or greater, have input capacitance of 15 or 20 pF. Thus a smaller compensating capacitor can be placed in the termination box.

A schematic of a 10X probe with risetime faster than 2 ns is shown in Fig. 2-16A. R3 is the termination resistance and L1 with blocking capacitor,

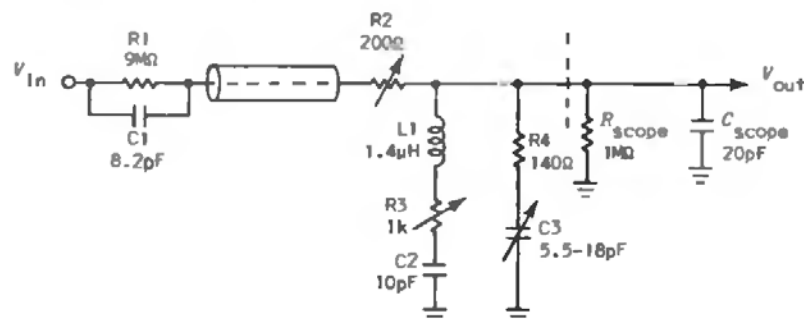
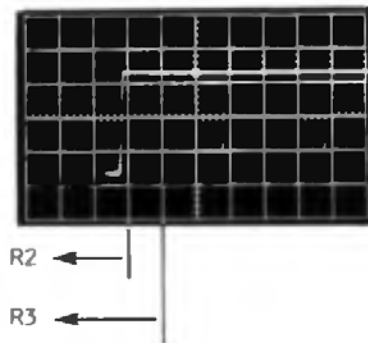


Fig. 2-16A. 10X probe with compensation range 15-20 pF.



R2 -- AFFECTS AREA OF WAVEFORM TO LEFT OF  
GRATICULE LINE  
R3 -- FIRST MAJOR DIVISION OF WAVEFORM

Fig. 2-16B. Transient response.

C2, terminate the reactive component of the cable impedance. The high speed shunting effect of compensation capacitor C3 is minimized by R4. R2 is adjusted for best leading edge. After the probe is compensated and R2 and R3 are adjusted for damping and leading edge in that order, some readjustment of the resistors may be required to obtain best overall pulse response. See Fig. 2-16B.

Even faster risetimes can be realized with the use of a T-coil type termination for the probe cable. A schematic is shown in Fig. 2-17. The cable is terminated by the T-coil, C5 and L5. The T-coil is terminated by R7 and blocking capacitor, C7. After R7 is adjusted for minimum ringing, C5 is adjusted for minimum ringing at the front corner consistent with smooth high-frequency rolloff. R4 controls front corner peaking and R9 is adjusted for an even flat top waveform. L1 and L2 are ferrite beads which reduce overshoot at the cable input.

An interesting feature of the passive voltage probe is that the input capacitance tends to decrease as applied frequency increases. At 1 MHz, the circuit of Fig. 2-17 has an input capacitance of 10 pF. At 100 MHz, the input capacitance is 7 pF. This decrease of input capacity is inherent to all passive voltage probes. A graph of input impedance versus frequency is shown in Fig. 2-18. The graph is called an  $R_p$ - $X_p$  curve, where  $R_p$  is the input resistance of the probe and  $X_p$  is the input capacitive reactance of the probe. A graph of  $C_p$ , input capacitance versus frequency is shown in the lower right corner.

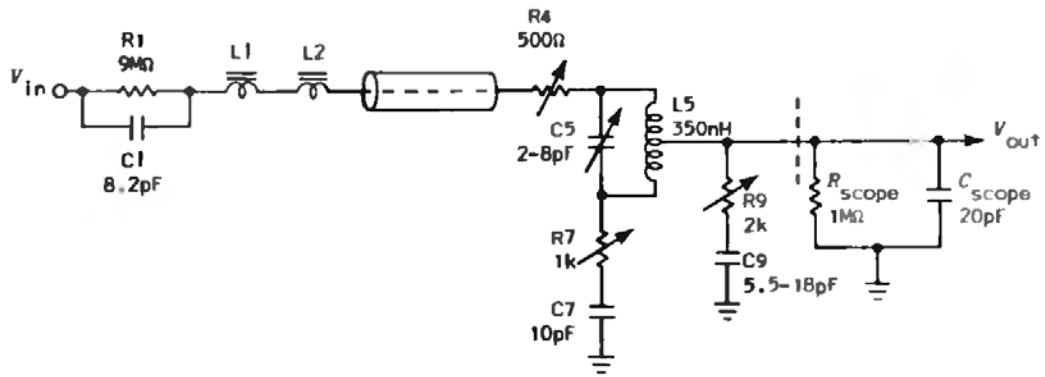


Fig. 2-17. 10X probe with T-coil type termination.

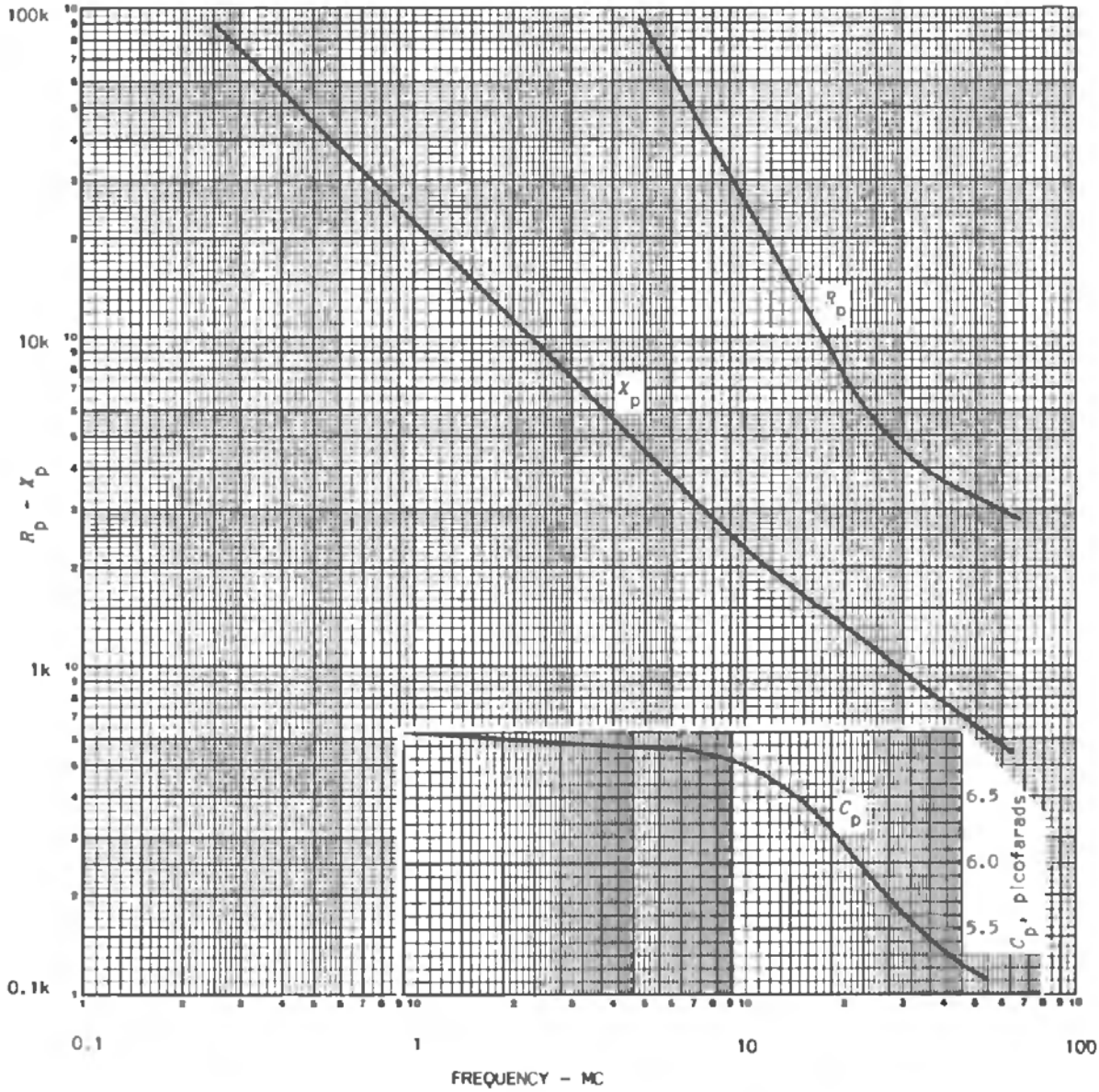


Fig. 2-18. Input resistance and reactance vs frequency curves.

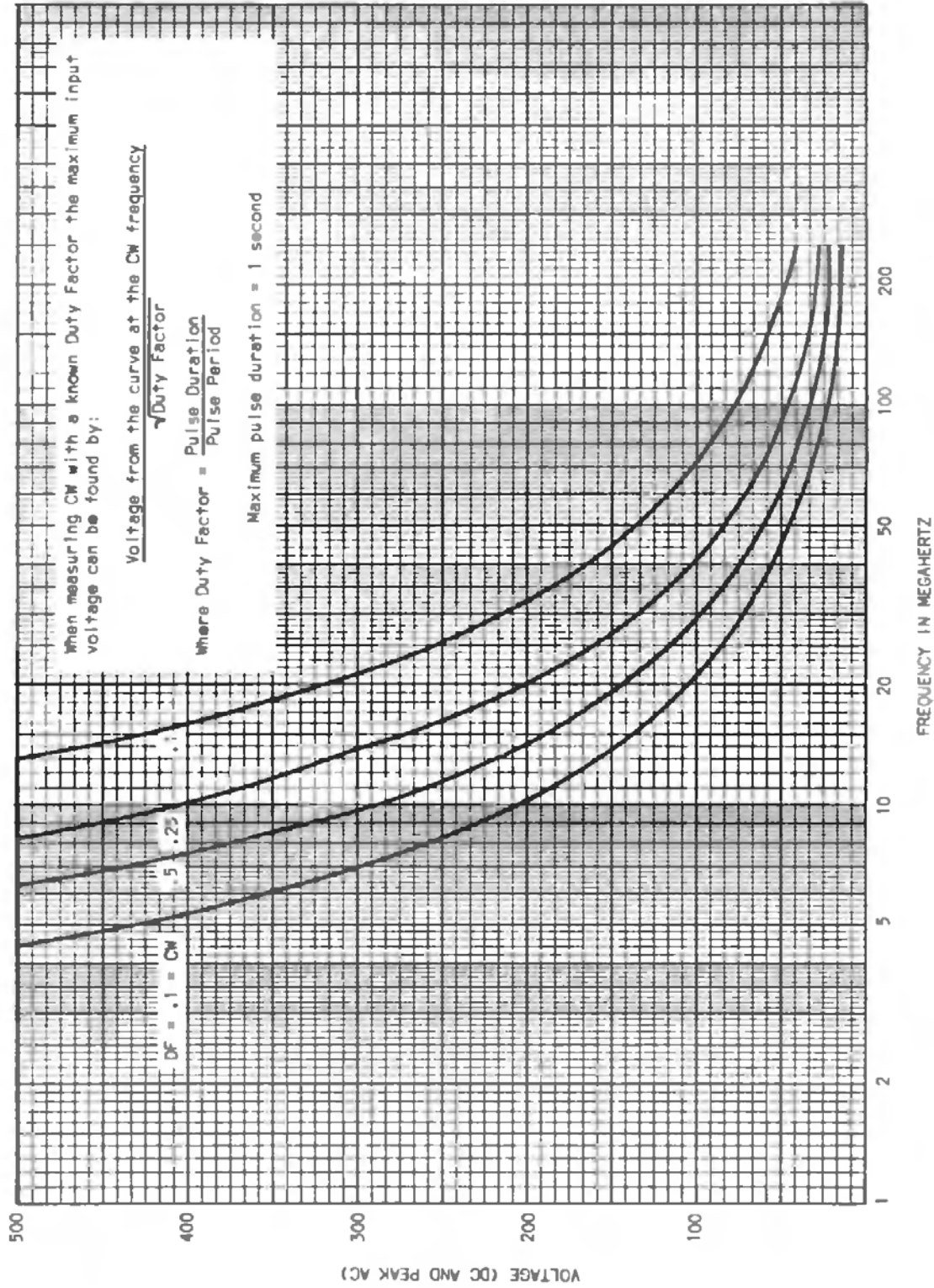


Fig. 2-19. Maximum applied voltage at specific duty factors.

Since probe input impedance decreases as applied frequency increases, it is important to choose low impedance points in the circuit under test for high speed measurements. At Tektronix, both bandwidth and risetime specifications are checked with signal generators whose source impedance is 25  $\Omega$ .

In the probe manual, the risetime of the *probe only* is specified. With this information and the risetime of the oscilloscope, the risetime of the probe-scope combination can be determined. The formula is:

$$t_r = \sqrt{(t_{rprobe})^2 + (t_{rscope})^2}$$

When checking probe-scope risetime, the risetime of the signal source should be considered. The risetime of the signal at the probe tip is a function of the source resistance and capacitance and the input capacitance of the probe. The display risetime is the vector sum of source, probe and scope risetimes.

$$t_r = \sqrt{(t_{rsource})^2 + (t_{rprobe})^2 + (t_{rscope})^2}$$

The maximum input signal amplitude for any probe changes with frequency. A high voltage versus frequency graph is shown in Fig. 2-19. Since probe input impedance decreases as applied frequency increases, probe current increases with frequency. The maximum input voltage rating is decreased so that the power rating of the probe components is not exceeded. In probes that have a reverse termination resistor in the probe head, this 1/2 watt resistor will determine the derating curve. In probes that have the termination circuitry at the oscilloscope end of the cable, the  $I^2R$  characteristic of the cable determines the derating curve.

power  
rating

The passive voltage probe is an inexpensive, versatile device for coupling the signal of interest to the oscilloscope. Presently, passive probe-conventional oscilloscope combinations can provide risetimes faster than 1.2 ns and 30% down frequencies greater than 150 MHz. Passive voltage probes are available with 1X, 10X or 100X attenuation and in cable lengths from 3.5 ft to 12 ft.

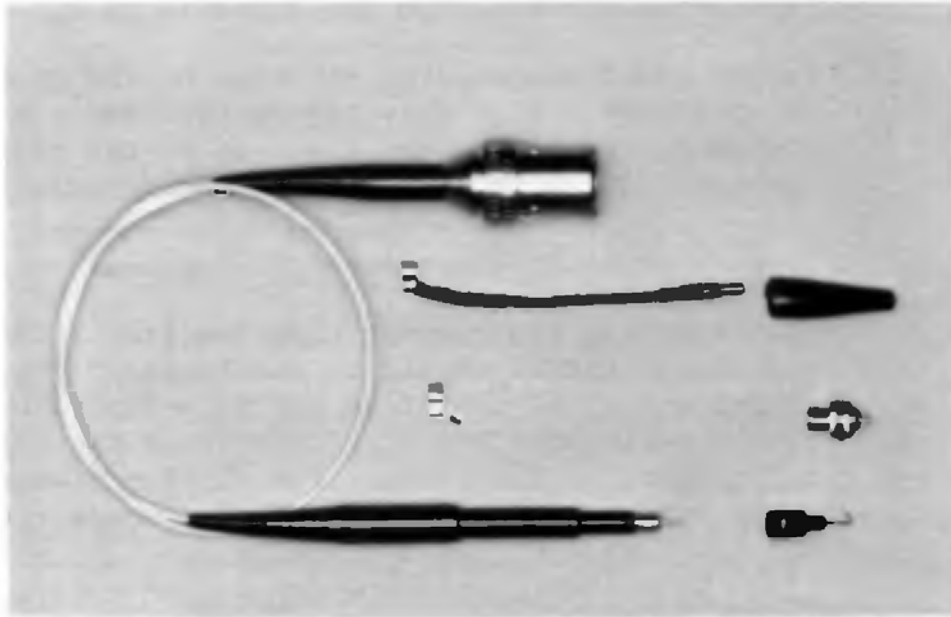
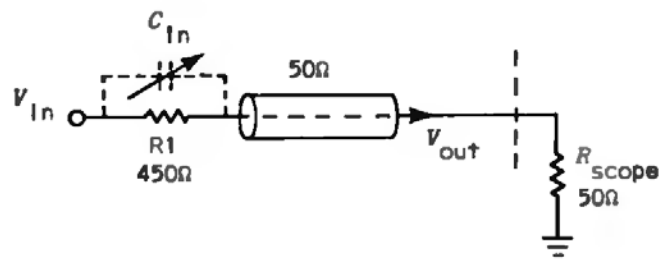


Fig. 3-1.

Fig. 3-2. 10X  $Z_0$  probe.



## 3

 $Z_0$  PROBES

$Z_0$  probes are a special type of passive voltage probe.  $Z_0$  probes were originally designed to provide sampling systems with a relatively high impedance, attenuating, passive probe. The use of  $Z_0$  probes has been extended to include conventional oscilloscopes. Hence, there are two classes of  $Z_0$  probes: (1) those probes designed to drive a  $50\ \Omega$  load and (2) the probe designed to drive a load of  $1\ \text{M}\Omega$  paralleled by 15 or 20 pF.

The  $Z_0$  probe has a low input capacitance (less than 1 pF) which is desirable for high speed signals. The input resistance is relatively low to all frequencies,  $5\ \text{k}\Omega$  or less depending on the particular probe. An AC coupling technique is available to minimize DC loading, however the maximum allowable DC input voltage is less than that of the high impedance passive probes.

Tektronix  $Z_0$  probes designed to drive  $50\ \Omega$  loads have a slim body assembly and a GR\* connector at the oscilloscope end of the cable. A photograph is shown in Fig. 3-1. A thin  $50\ \Omega$  cable is used to couple the probe head to the GR connector. From the photograph the probe looks like a delicate device that will not stand abuse. It is and it will not.

A schematic of a 10X  $Z_0$  probe is shown in Fig. 3-2.  $R_1$  in series with the  $50\ \Omega$  cable and the  $50\ \Omega$  input of the oscilloscope provide 10:1 attenuation.  $C_{in}$ , shown as an adjustable capacitor, is the environmental capacitance of the probe head that shunts  $R_1$ . This capacity can be changed by

---

\*General Radio Corp., West Concord, Massachusetts.

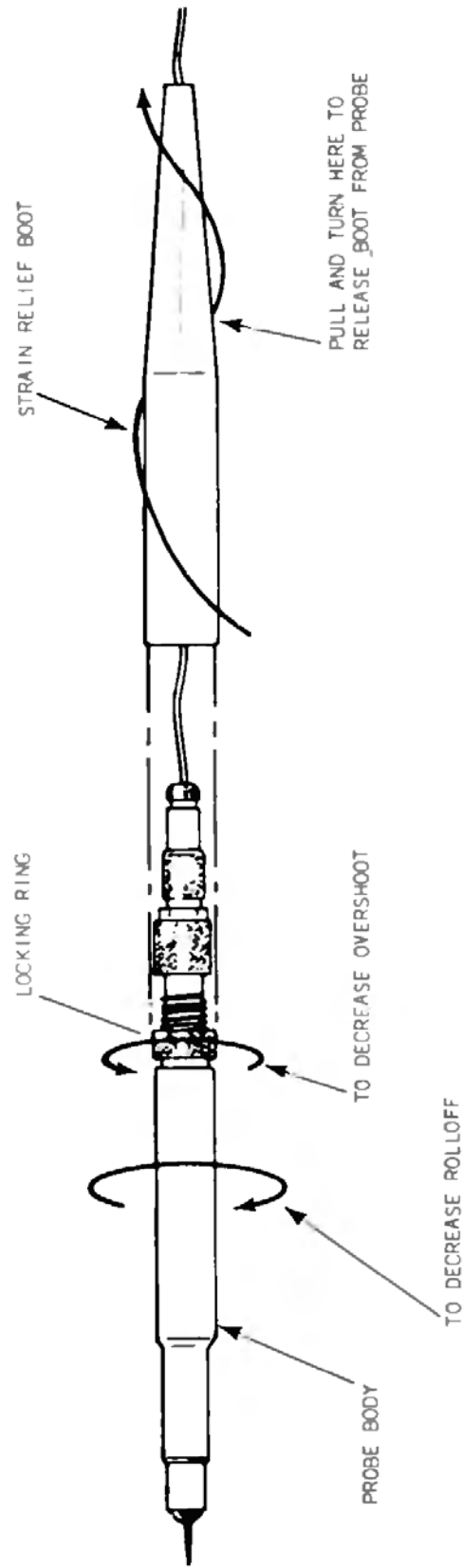


Fig. 3-3. Probe compensation.

positioning the probe tip with respect to the base of the probe head assembly. See Fig. 3-3.

Adjustment of  $C_{in}$  is termed "probe compensation" which is a misnomer. The probe is calibrated by (1) checking the output of a fast pulse generator ( $t_r \leq 250$  ps) with a sampling oscilloscope, Fig. 3-4A, and (2) adjusting  $C_{in}$  for proper pulse response, Fig. 3-4.. It is difficult to discriminate between an under-compensated probe (C) and proper compensation (D). The rolloff at the front corner is due to cable losses.

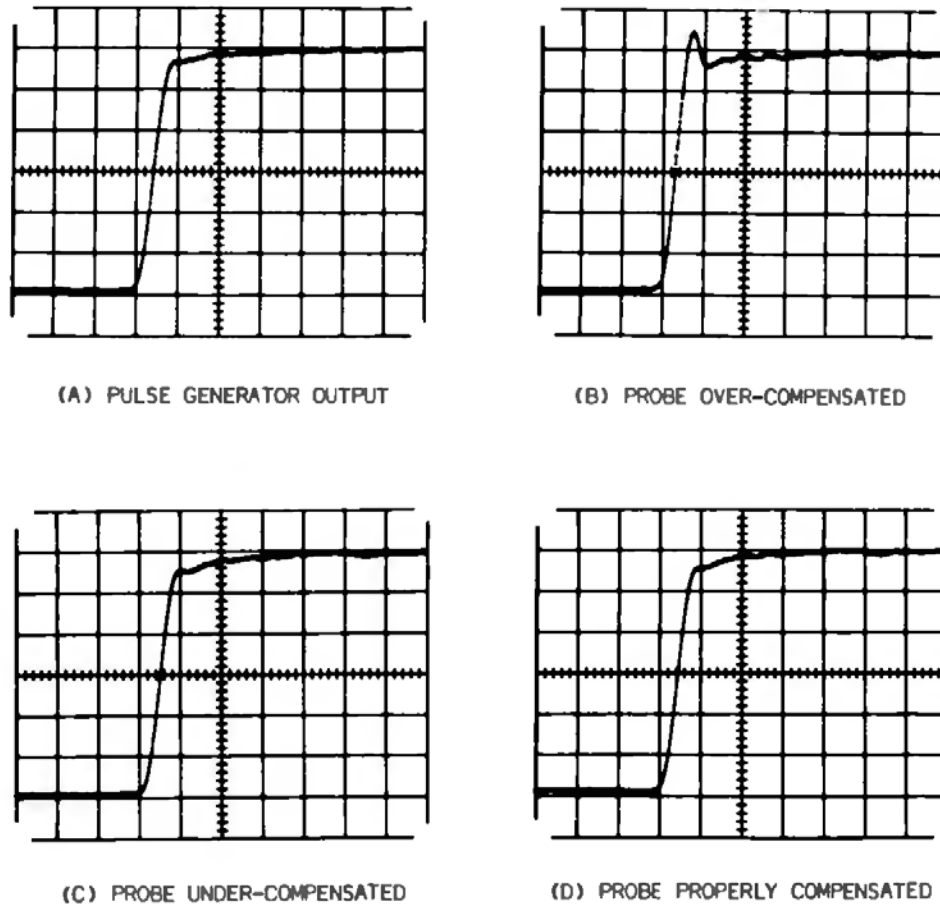


Fig. 3-4. Probe compensation waveforms.

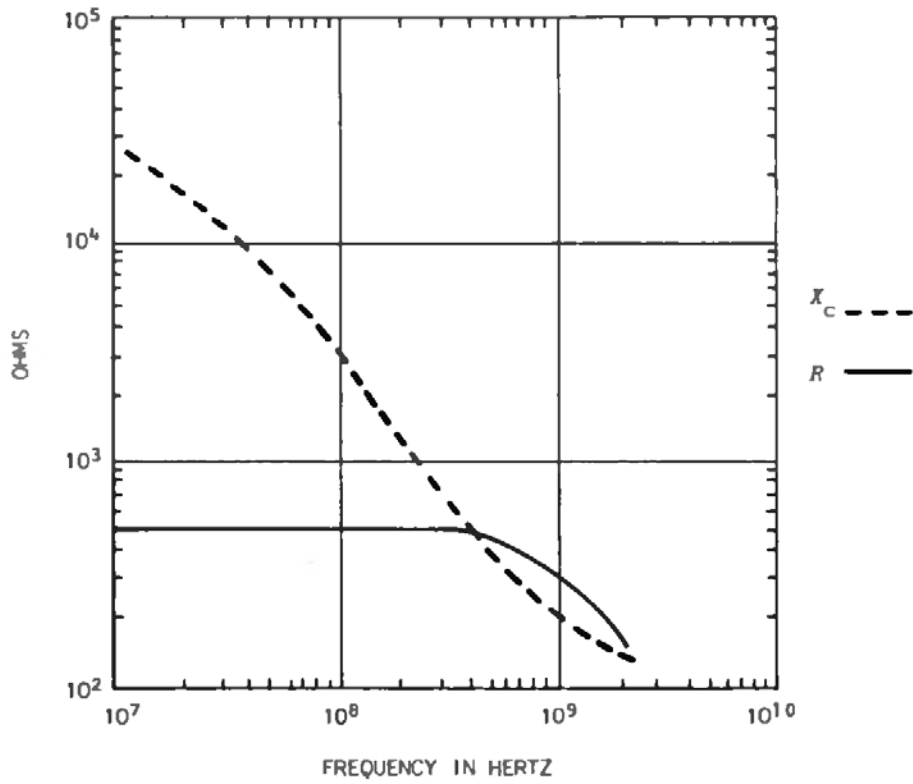


Fig. 3-5. Input resistance and capacitive reactance vs frequency curves.

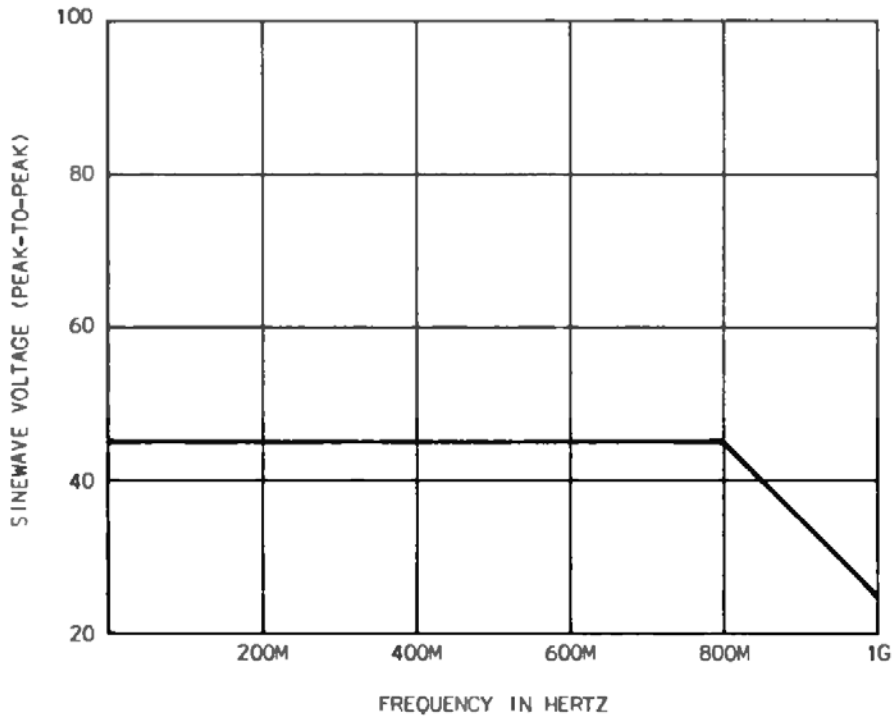


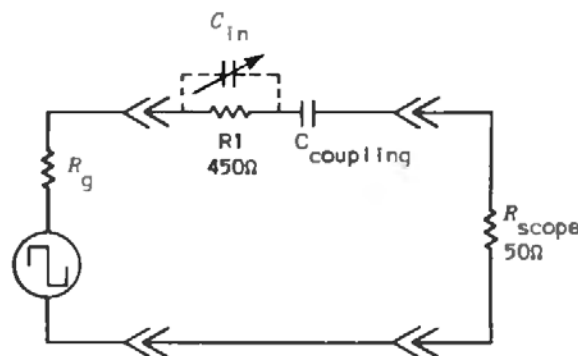
Fig. 3-6. Sine wave voltage derating.

The input impedance of the  $Z_0$  probe will decrease as applied frequency is increased like the passive voltage probe. An  $R_p$ - $X_p$  curve is shown in Fig. 3-5. At 1 GHz, the input resistance has fallen to about 300  $\Omega$ . Probe risetime is 100 ps or faster and the 30% down frequency is 3.5 GHz as calculated from the risetime specification.

Since R1 is a 1/2 watt resistor, the maximum allowable DC input voltage is 16 V. Allowable sine-wave voltage is 45 peak voltage above ground or less. A sine-wave voltage versus frequency derating curve is shown in Fig. 3-6. A GR 874-K Coaxial Coupling Capacitor can be used to block DC voltages up to 500 V. The coupling capacitor is typically 4700 pF. The schematic arrangement is shown in Fig. 3-7. The use of the coupling capacitor limits low frequency response. The 30% down low frequency can be determined by the following formula:

$$f = \frac{1}{2\pi C (R_p + R_s)} ,$$

where  $R_p$  is the probe input resistance,  $R_s$  is the source resistance, and C is the 4700 pF coupling capacitor.



$R_g$  = TEST POINT IMPEDANCE

$R_1 + R_{scope} = 500\Omega$

$C_{coupling}$  = TYPICALLY 4700pF WITH GR 874-K COAXIAL COUPLING CAPACITOR

LOW FREQUENCY  
-3dB POINT =  $\frac{1}{2\pi RC} = \frac{0.159}{C_{coupling} (500\Omega + R_g)}$

Fig. 3-7. AC-coupled probe low frequency response.

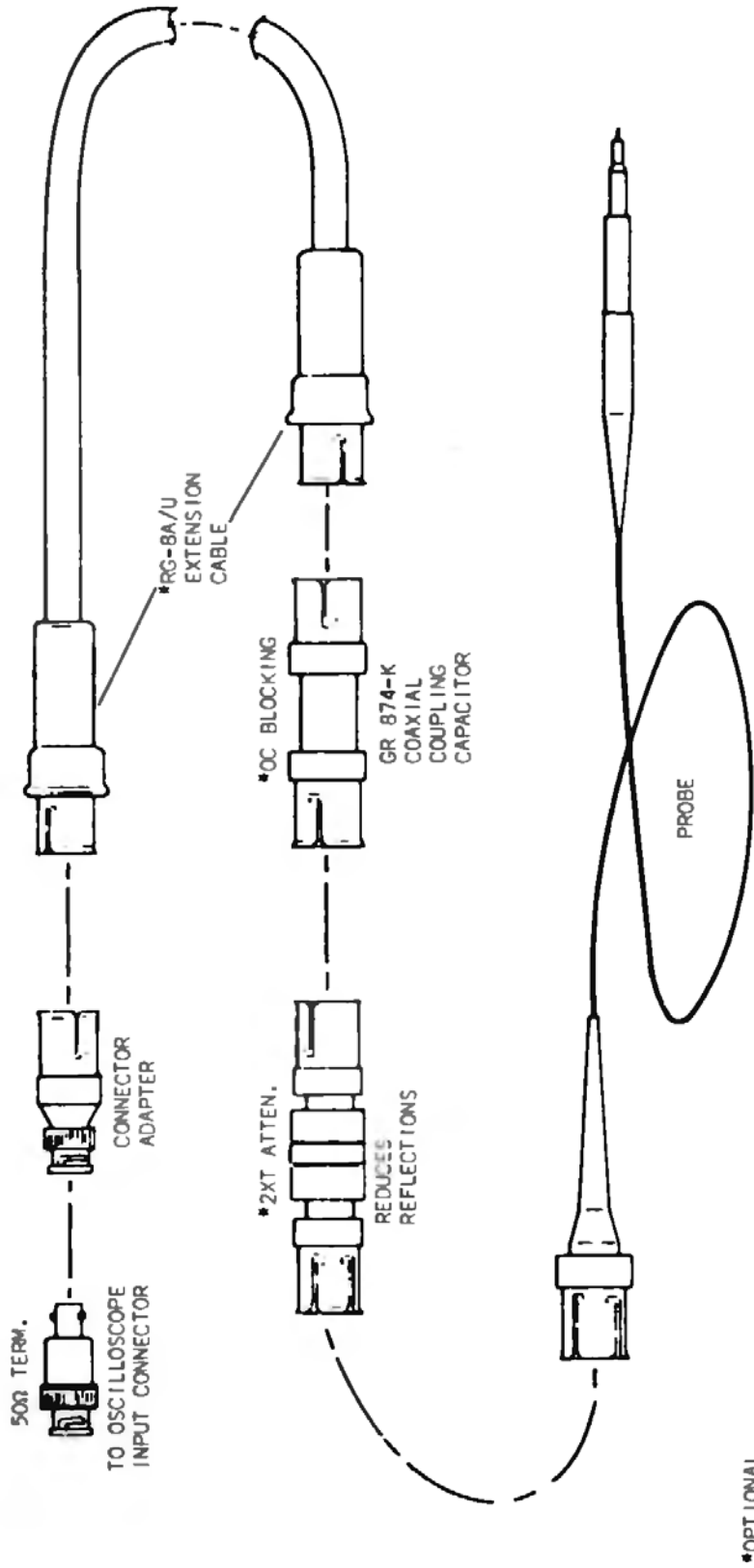


Fig. 3-8. Component sequence in non-sampling oscilloscope applications.

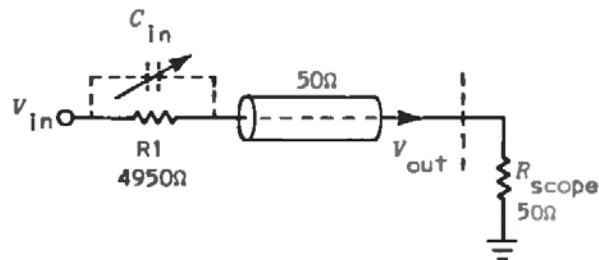
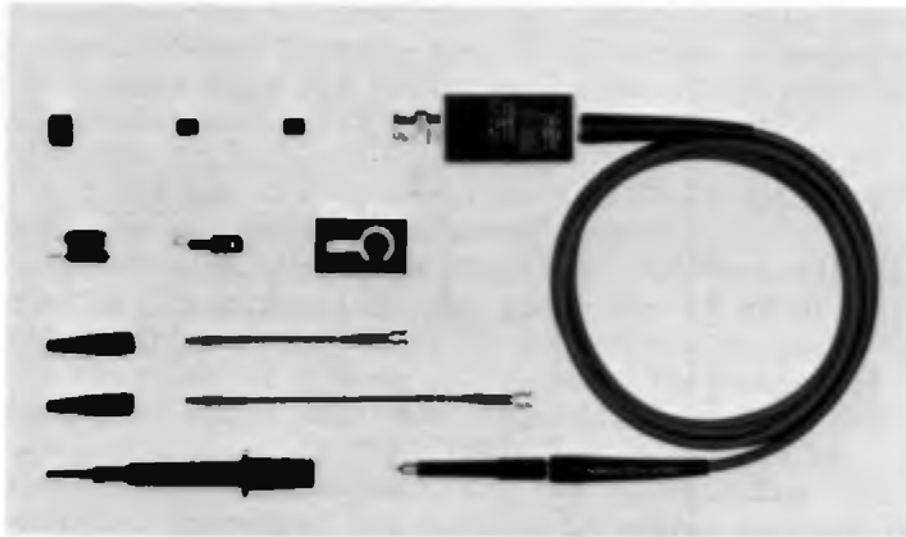
Fig. 3-9. 100X  $Z_0$  probe.

Fig. 3-10.

As noted earlier, the  $Z_0$  probe can be used with conventional oscilloscopes. The oscilloscope input must be terminated in  $50\ \Omega$ . Fig. 3-8 is a diagram of the proper arrangement.

A 100X  $Z_0$  probe is also available. Attenuation is achieved by using a  $4950\ \Omega$  resistor in the probe head. The larger resistance slows probe risetime to 200 ps and increases DC input voltage rating to 50 V. Fig. 3-9 is a schematic for the 100X probe. The 100X is similar to the 10X probe with the value of  $R_1$  being the only major change. Calibration and measurement considerations are the same for both probes.

The probes discussed previously are primarily designed for use with sampling systems but could be used with conventional oscilloscopes. A  $Z_0$  probe that is primarily intended for use with conventional oscilloscopes is also available. A photograph is shown in Fig. 3-10. Physically, this

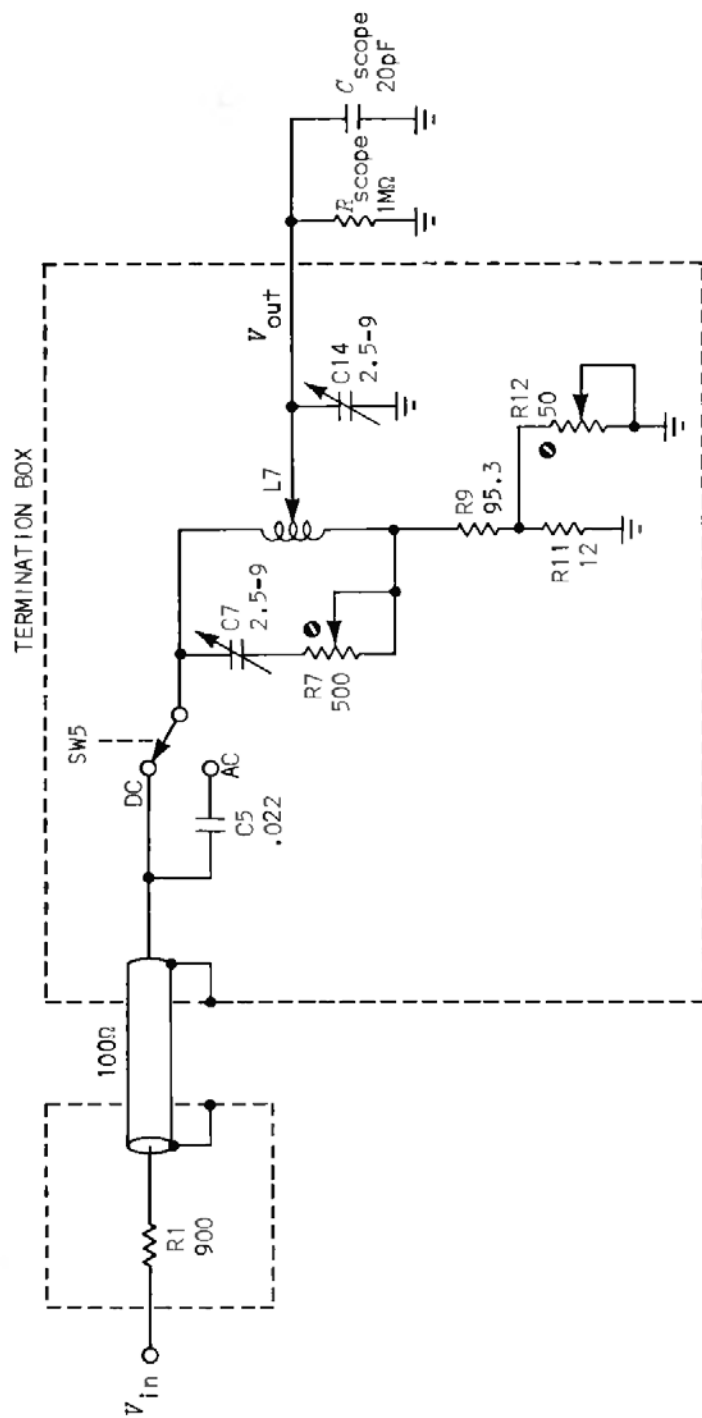


Fig. 3-11. 10X probe schematic.



probe more closely resembles a passive voltage probe than the probes discussed previously and is less delicate than the other  $Z_0$  probes. The circuitry in the termination box is designed to couple the output of the  $100\ \Omega$  cable to the input of a conventional oscilloscope. An AC-DC selector switch and the AC coupling capacitor are contained in the termination box. The probe provides 10X attenuation and has a DC input resistance of  $1\ \text{k}\Omega$ .

A T-coil type termination is used to transfer the signal from the  $100\ \Omega$  environment of the cable to the oscilloscope input. Fig. 3-11 is the schematic. R1 in series with the cable and termination provides 10:1 attenuation. R12 is made adjustable to compensate for cable tolerances,  $100\ \Omega \pm 3\%$ . R9, R11 and R12 are the termination for the T-coil. C7 and R7 are adjusted for minimum aberrations consistent with specified risetime. C14 provides a method of "compensating" the probe for 15 pF or 20 pF oscilloscopes. C14, C7 and R7 are adjusted for a risetime of 3.5 ns or faster. The output of a fast rise pulse generator terminated in  $50\ \Omega$  and an oscilloscope with risetime of at least 2.4 ns are used for calibration. C5, the AC coupling capacitor, increases the maximum DC input voltage to 200 V from 20 V, DC coupled. The high voltage derating and  $F_p$ - $X_p$  curves are similar to the curves of the  $50\ \Omega$  probes.

The  $Z_0$  probes provide a low capacitive load to high frequency signals. The low input resistance of the  $Z_0$  probes can cause excessive loading to DC signals. An AC coupling capacitor is provided to minimize DC loading. In general, the  $Z_0$  probes have faster risetimes than passive voltage probes and are primarily used to investigate high speed signals.

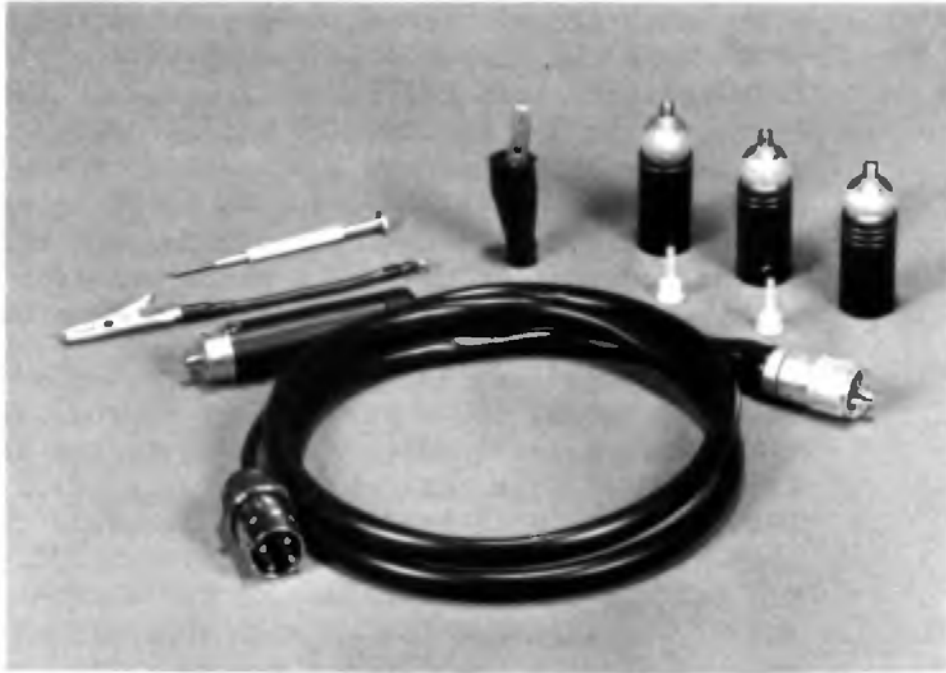


Fig. 4-1.

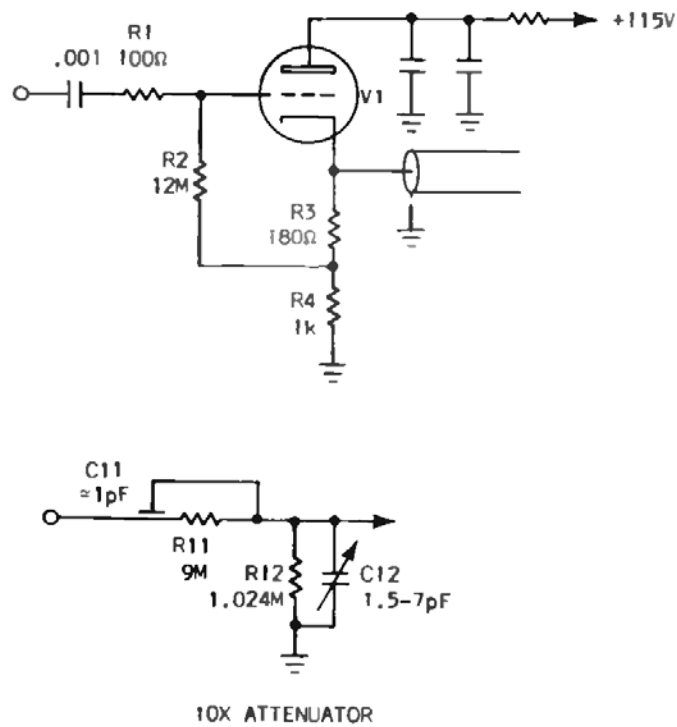


Fig. 4-2. AC coupled CF probe.

## 4

## ACTIVE VOLTAGE PROBES

The active voltage probe was designed to provide a better method of coupling high speed signals to the oscilloscope. Active voltage probes have higher input impedances with less attenuation than the passive voltage probes. Although these probes are more expensive and usually more bulky, their use extends the measurement capabilities of the probe-oscilloscope combination.

cathode  
and source  
followers

There are two types of active voltage probes: (1) the cathode follower probe and (2) the source follower probe. Of the two, the cathode follower probes are the bulkier even though miniature tubes are used. A photograph of an early CF probe is shown in Fig. 4-1. The attenuator heads are included with the probe as is the screwdriver for adjusting attenuation ratios. This probe has a risetime faster than 2 ns.

1 k  $\rightarrow$  28 M

A schematic of an AC coupled cathode follower probe is shown in Fig. 4-2. The input circuit is designed to present a very high input resistance. The input resistor, R2, is tied to the junction of cathode resistors, R3 and R4. About 70% of the input signal is developed across R4 and only 30% of the input signal is dropped across R2. Since R2 drops 30% of the input signal, there must be a 28 M $\Omega$  resistor that drops the remaining 70%. R4 does this and therefore appears as a 28 M $\Omega$  resistor to the source. Hence, the input resistance of this probe is 40 M $\Omega$ .

The input capacity is 4 pF which is comparable to the input capacity of a 100X passive probe. This active probe has a higher input impedance than the 100X passive probes with an attenuation factor of 1.25. The output impedance of the CF is designed to drive a 170  $\Omega$  cable which is terminated in 170  $\Omega$  at the oscilloscope input.

A 10X attenuator, compensated voltage divider, is available which increases maximum input voltage from 5 V to 50 V. C12 is the compensating capacitor and is adjusted using the oscilloscope calibrator signal. The input impedance is  $10\text{ M}\Omega$  shunted by 2 pF with the 10X attenuator.

circuit  
description

Cathode follower probes can be used to provide a high input impedance for  $50\ \Omega$  systems such as sampling oscilloscopes. Fig. 4-3 is a schematic of such a probe. Probe gain is about  $1/3$  with the  $50\ \Omega$  cable terminated in  $50\ \Omega$  at the oscilloscope input. R3 is adjusted for proper gain when the probe is driven by a low speed square wave of known amplitude. L5 provides high frequency peaking. The output DC level is about  $1/2\text{ V}$ , depending on the self-bias of V1. D3, a silicon diode, provides a voltage offset between V1 cathode and the output. The section of  $50\ \Omega$  coax tied to V1 cathode damps ringing in the high frequency response. The coax is terminated in an open circuit to reflect a short circuit  $1/4$  wavelength from the termination. Since voltage does not exist at a short circuit, the ringing is damped.

This probe is normally used with an attenuator because the maximum input voltage of the probe is only 500 mV. Attenuators from 10X to 1000X in a 1, 2, 5 sequence are available to extend maximum input voltage rating from 1.5 V to 150 V. The attenuation ratios are designed for the  $1/3$  gain of the probe. Probe risetime is 0.4 ns and bandpass is 850 MHz with any attenuator. This is the fastest cathode follower probe built by Tektronix.

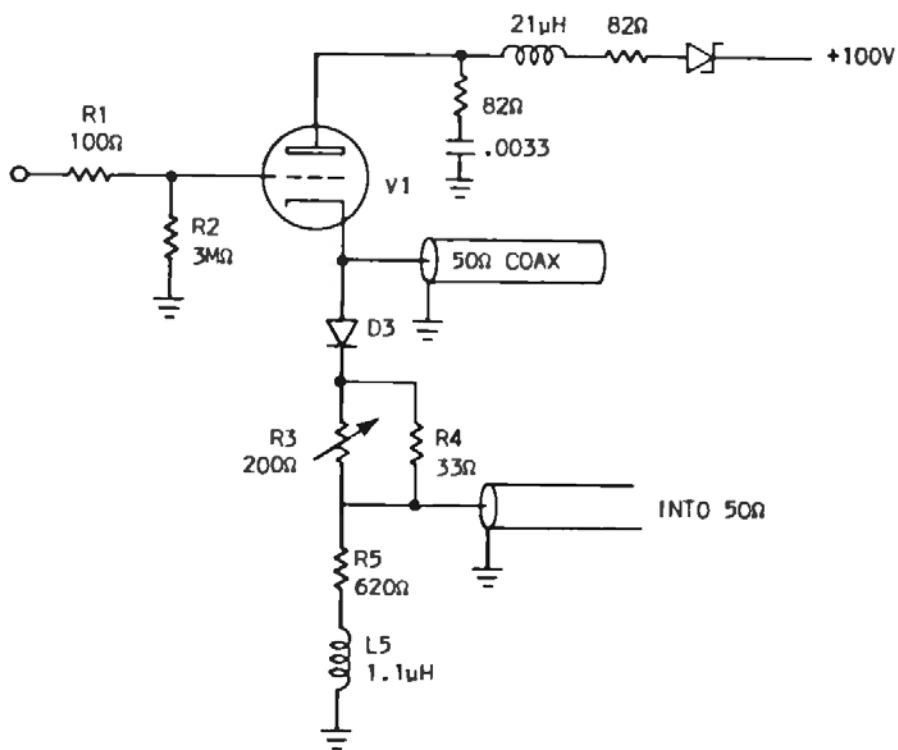


Fig. 4-3. Wide band CF probe.

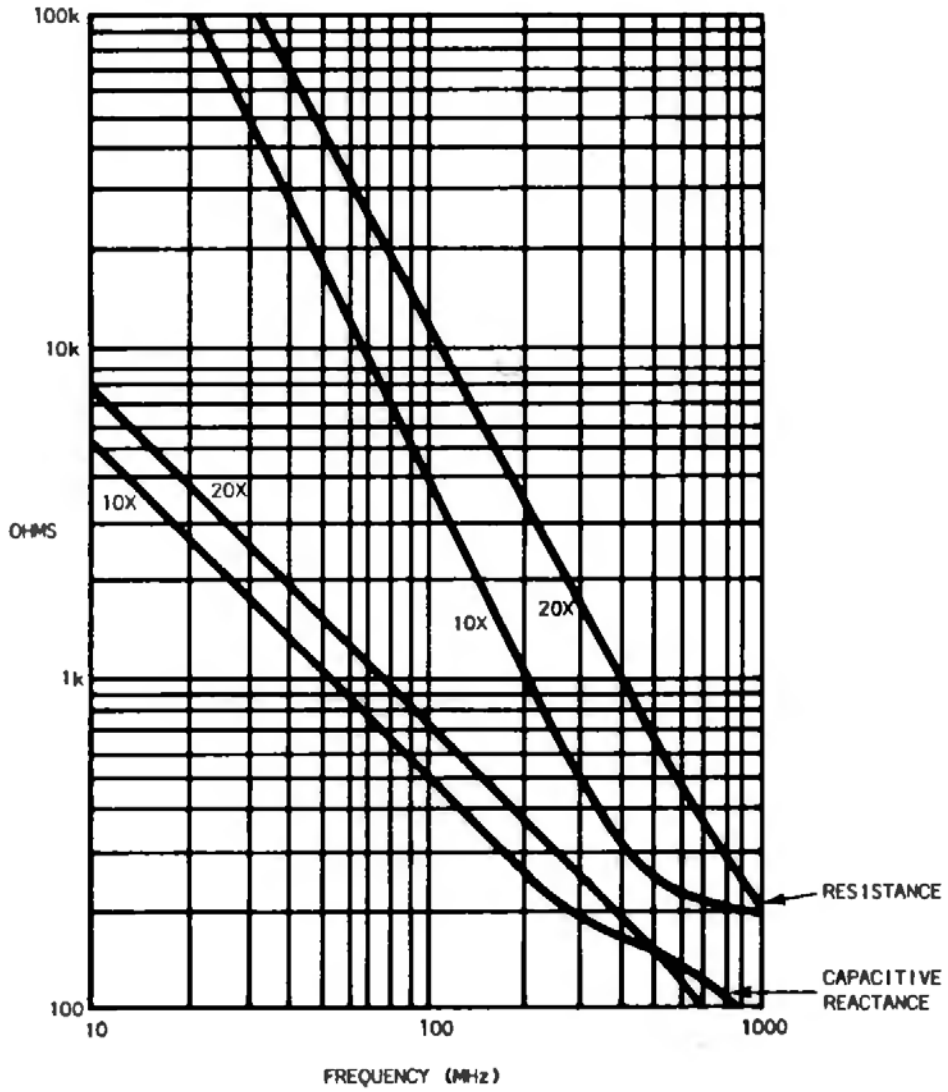


Fig. 4-4A. Frequency derating.

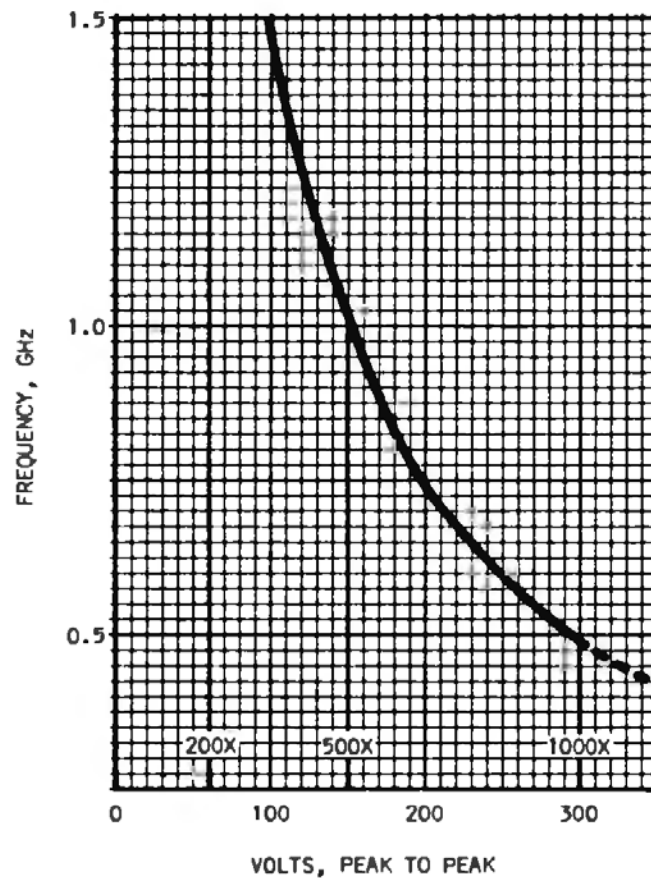


Fig. 4-4B. Voltage derating.

derating

The active voltage probes have limitations similar to the limitations of passive voltage probes. We have mentioned maximum DC input voltage which is normally smaller than passive probes. High frequency derating curves are also given for active voltage probes as are  $R_p$ - $X_p$  curves. Some examples are shown in Fig. 4-4.

As expected, these curves are similar to the corresponding curves for passive voltage probes. Both types of probes have an RC input circuit, with the cathode follower probe having the higher input impedance. The CF probes are larger which makes them awkward for probing high density circuits. We can get a reduction in size by using an FET input circuit. Since semiconductors are quite small, a number of active devices can be placed in the probe head.

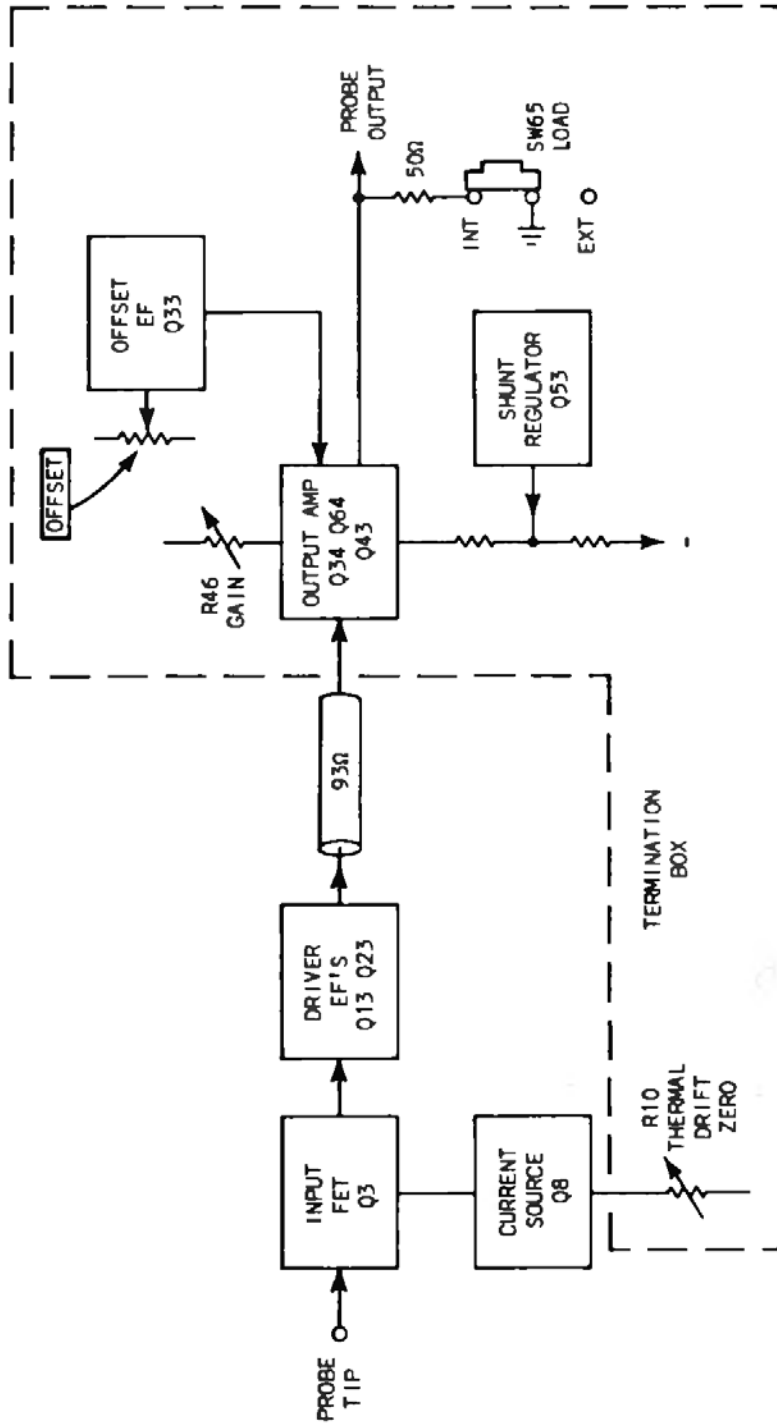


Fig. 4-5.



block  
description

A block diagram of an FET probe is shown in Fig. 4-5. The probe has three basic parts: the probe head, cable and termination box. The probe head circuitry is Source Follower, Q3, with Current Source, Q8 and Driver EF's, Q13 and Q23. The Thermal Drift Zero, R10 located in the termination box, is adjusted to set the drain current of Q3 such that the thermal characteristics of Q3-Q8 are equal and opposite to the thermal characteristics of Q13-Q23. In addition to providing temperature compensation, the use of two Driver EF's also provides a low driving impedance (near zero ohms) to the 93  $\Omega$  cable which is terminated in the termination box.

Common Base Amplifier Q34 is the input stage and its emitter voltage is controlled by the OFFSET via Offset EF Q33. The affects of the DC signal at the probe tip are canceled by making the emitter voltage of Q34 track the DC signal. The Gain control, R46, is located in the emitter circuitry of Q43, a common emitter amplifier. The common emitter output stage, Q64, is designed to drive a 50  $\Omega$  load. Gain from probe tip to probe output, terminated in 50  $\Omega$ , is unity.

50  $\Omega$  load

The LOAD switch allows an internal or external 50  $\Omega$  load to be selected. When the probe is used with a conventional high impedance oscilloscope, the internal position is selected. When the probe is used with an oscilloscope whose input impedance is 50  $\Omega$ , the external position is selected. Thus the labels Internal and External identify the location of the 50  $\Omega$  termination.

Shunt Regulator Q53 maintains a constant load to the power supply. This allows a wide variation in output current without affecting the power supply voltage. Not indicated in the block diagram are the five high frequency adjustments that are located in the emitter circuits of Q43 and Q64.

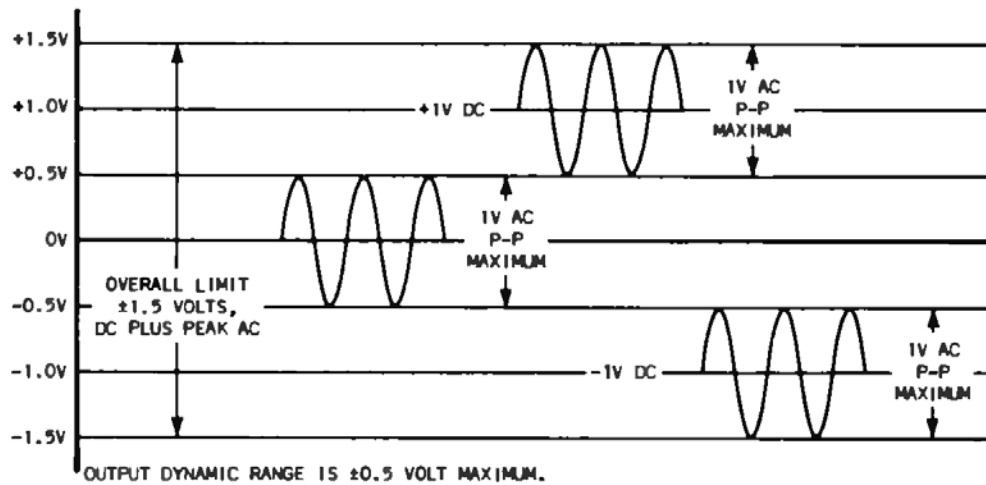


Fig. 4-6A.

	PEAK TO PEAK MAXIMUM	DC PLUS PEAK AC MAXIMUM
P6045 ALONE	1 VOLT	±1.5 VOLTS
WITH COUPLING CAPACITOR	1 VOLT	±200 VOLTS DC WITH 1 VOLT AC
WITH 10X ATTEN	10 VOLTS	±15 VOLTS
WITH 10X ATTEN AND COUPLING CAPACITOR	10 VOLTS	±185 VOLTS DC WITH 15 VOLTS AC OR ±200 VOLTS DC WITH 1 VOLT AC
WITH 100X ATTEN	100 VOLTS	±100 VOLTS (BREAK-DOWN LIMITED)
WITH 100X ATTEN AND COUPLING CAPACITOR	100 VOLTS	±100 VOLTS DC WITH 100 VOLTS AC OR ±200 VOLTS WITH 1 VOLT AC

Fig. 4-6B.

signal  
input  
limits

Offset range is  $\pm 1$  V and the dynamic range of the probe amplifier is  $\pm 500$  mV. Maximum DC plus peak AC input voltage of *probe only* is 1.5 V. A coupling capacitor can be used to extend the DC limits to 200 V with 1 V of AC signal. A 10X and 100X attenuator are available to extend the input signal limits. Fig. 4-6A graphically indicates input signal limits of the *probe only* and Fig. 4-6B indicates input signal limits with the various combinations of coupling capacitor and attenuators.

circuit  
description

A schematic of the circuitry contained within the probe head is shown in Fig. 4-7. The series resistance of R1 and R2 make up the 10 MΩ input resistance of the probe. The capacity at the probe tip is less than 5.5 pF. Since R1 is in series with Q3 gate, the FET is protected from overload signals. C1 is a high frequency bypass capacitor. R5-C5 maintain the input characteristics of the Source Follower over a wide frequency range. A portion of the fast signals are shunted to Q3 source via C2 to compensate for the slow response of the FET.

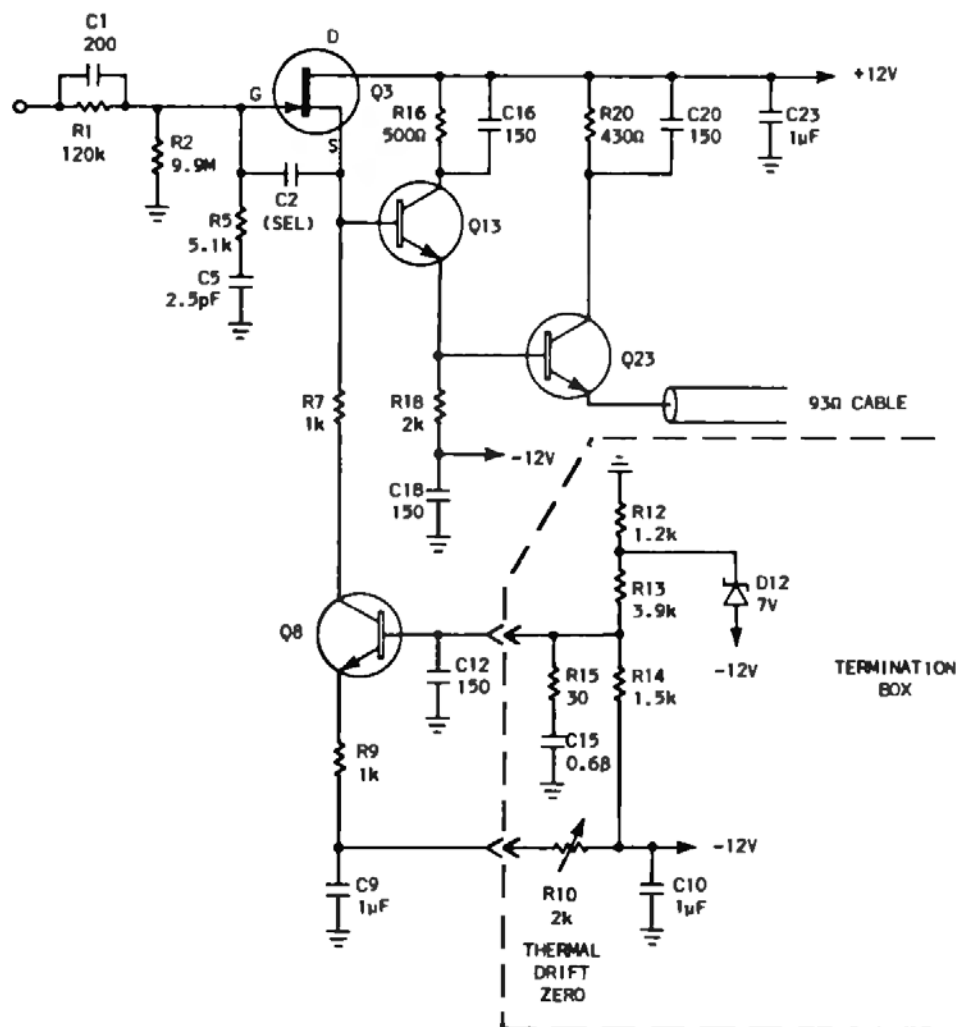


Fig. 4-7. Probe head circuitry.

thermal  
drift

The probe is calibrated for thermal drift less than 0.5 mV/°C by proper adjustment of the Thermal Drift R10 located in the termination box. The probe is placed in a 25°C environment for about 5 minutes and then placed in a 50°C environment for another 5 minutes. Total trace shift should be less than 12.5 mV. If the trace shift is greater than 12.5 mV, the temperature characteristics of Q3-Q8 do not match Q13-Q23. R10 is adjusted to move the trace in the opposite direction about 6 CRT divisions for each division of trace shift. The check is repeated until trace shift is less than 12.5 mV. Gain of the probe head circuitry, from probe tip to cable input, is about 0.95. Q23 is a near zero impedance source to the 93 Ω cable.

circuit  
description

A schematic of the termination box circuitry is shown in Fig. 4-8. The cable termination is R30-C30 in series with Q34 emitter impedance. The collector load of Q34 is the equivalent resistance of R41 and R42 or about 240 Ω. Stage gain is approximately 2.6 which is the ratio of the 240 Ω equivalent collector resistance to the total emitter circuit resistance including Q23 output resistance. LR40 is a high frequency peaking circuit. The output of Q34 drives Q43, an inverting amplifier whose gain is less than unity. Gain control R5, located in the emitter circuit of Q43, is adjusted for an overall gain of unity. Q43 drives output stage Q64, whose gain is about 0.8. R67 provides temperature compensation and C67 maintains pulse response. Current shunt Q53 maintains a constant current through R58 which allows wide variations in output current without adversely affecting the -12 V supply.

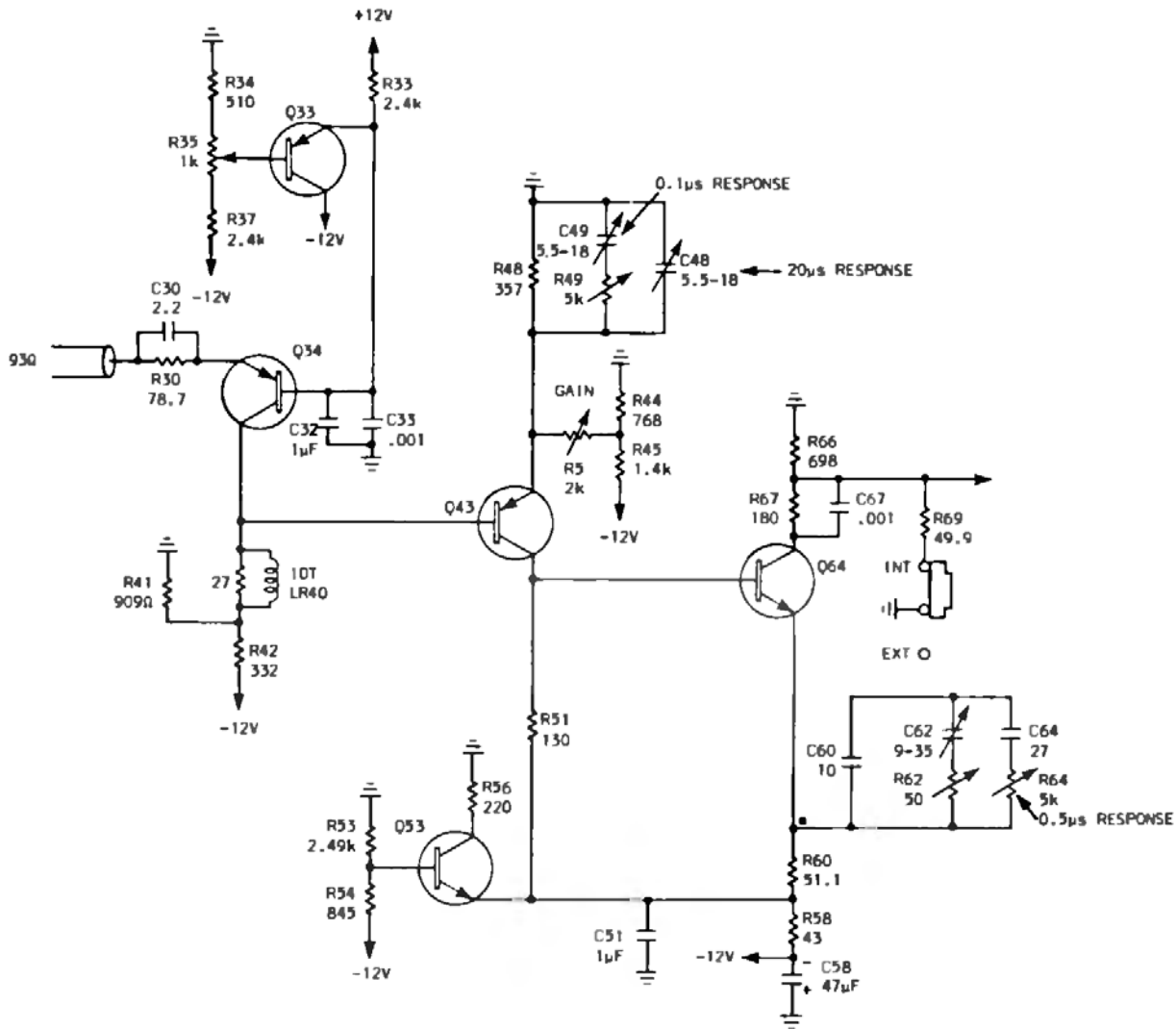


Fig. 4-8. Termination box circuitry.



Fig. 4-9. Probe adapter.

probe  
adapter

There are instances when we would like to take advantage of the high input impedance and the low output impedance of an active voltage probe without suffering the expense of the active voltage probe. A probe adapter that couples passive voltage probes to  $50\ \Omega$  environments is available. The probe adapter has an input impedance of  $1\ \text{M}\Omega$  shunted by  $16.5\ \text{pF}$ . See Fig. 4-9. The BNC type input connector mates with the output connector of the passive voltage probe and the GR type output connector mates with the input connector of a sampling oscilloscope. Power is supplied via the four pronged jack which mates with the probe power connector of a sampling vertical plug-in.

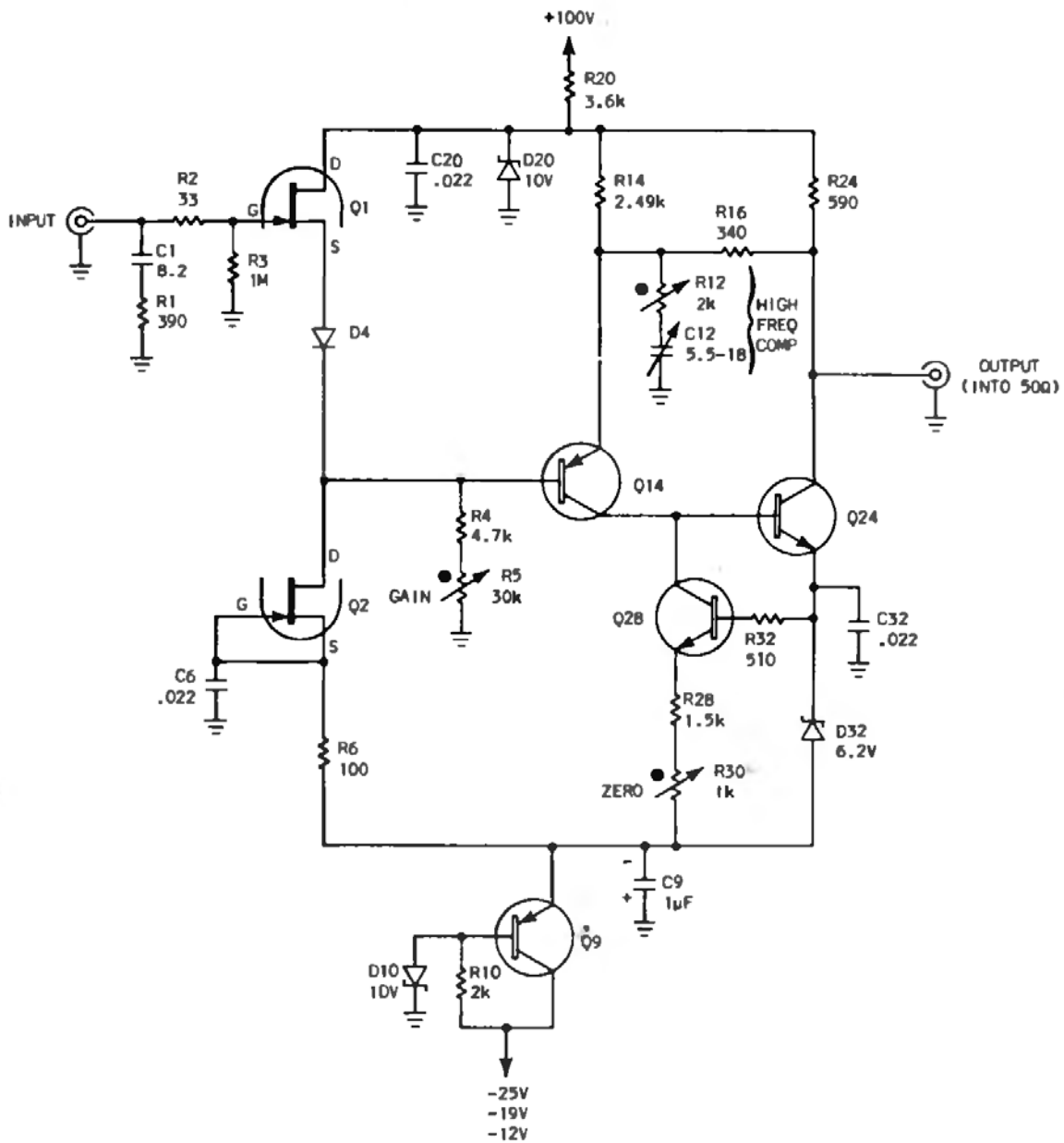


Fig. 4-10. Probe adapter circuitry.



circuit  
description

The schematic is shown in Fig. 4-10. Q1 is the input Source Follower and Q2 is a current source. Q14 and Q24 form a non-inverting feedback amplifier. Voltage gain is unity for the probe adapter. Q28 is a current source for Q14 and is a convenient location for the Output Voltage Zero adjustment. Q9 is a regulator that converts -25 V, -19 V and -12 V to the -10 V supply required by the probe adapter circuitry.

The design of the input circuit provides temperature compensation by taking advantage of the inherent temperature characteristics of the FET's. Note Q1 and Q2 have a quiescent voltage drop which almost equals the supply voltage. (The voltage drop across R6 and D4 is negligible.) The source to drain voltage of FET's tends to change when ambient temperature changes. In this circuit, the voltage across the FET's cannot change because there are no other elements that drop voltage in the circuit.

R5, the Gain Adjustment, and R4 shunt Q2. Thus an adjustment of R5 changes the resistance in series with the gate to source resistance of Q1. The resistance of R5 is adjusted to set the voltage divider ratio of Q1 gate to source resistance to the shunt combination of Q2 and R4-R5. R5 is adjusted to provide an overall gain of unity.

Q14 and Q24 are high gain inverting amplifiers. The output of Q24 is fed back to the emitter of Q14 via R16. The output signal is developed across R14 and R16. Stage gain is defined as:

$$A_V = \frac{R14 + R16}{R16} .$$

At high frequencies, R12-C12 shunt R14 to increase stage gain. R12-C12 are adjusted for best transient without overshoot.

At quiescence, Q14 current is held constant by the feedback loop. Thus, Q28 current can be changed by adjusting R30, which causes Q24 base current to change which maintains Q14 current. The quiescent output voltage can be set to zero by adjustment of R30.

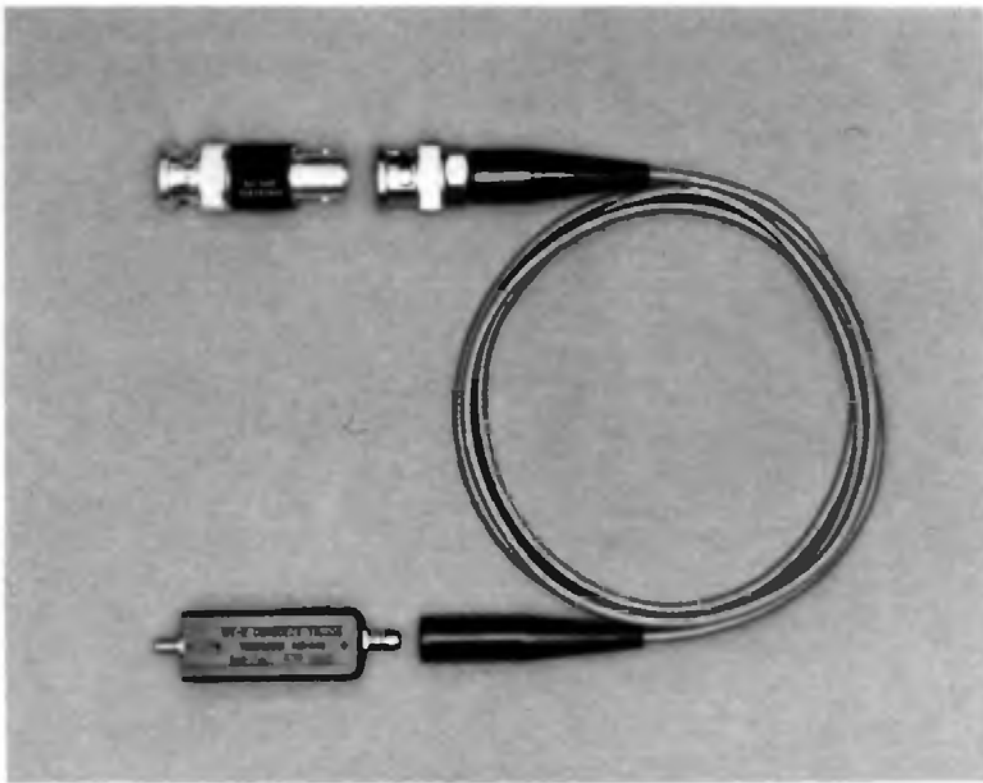


Fig. 5-1. Closed core current probe.

## 5

## CURRENT PROBES

The current probe provides an alternate method of coupling the signal to the oscilloscope. The signal is inductively coupled to the oscilloscope and a direct electrical connection is not required. The current probe, unlike the voltage probe, is a low impedance device and thus should be driven by a high impedance generator.

There are three classes of Tektronix current probes:

- (1) Closed core passive probes. These probes require that we break the circuit to place the conductor through the core. Closed core probes sense changes in current only and have passive terminations. These have the widest frequency response of Tektronix current probes.
- (2) Split core passive probes. These probes can be clipped around the conductor without breaking the circuit. Again, only changes in current are sensed by this probe. Active or passive terminations may be used.
- (3) Split core active probe. This probe contains a Hall generator which extends measurement capability to DC. This probe can be clipped around a conductor without breaking the circuit. Active termination is used with this type.

A photograph of a closed core current probe is shown in Fig. 5-1. The current sensing device is called a current transformer and the connecting cable is called a probe, although both parts are required to complete the system. The current transformer is a 25 turn coil wound on a ferrite core which is shunted by a 50  $\Omega$  resistor. The conductor under test is a single turn primary. The signal is connected to the oscilloscope via the terminated 50  $\Omega$  cable.

closed  
core  
probes

insertion  
or  
reflected  
impedance

Fig. 5-2 is a schematic of a current transformer with insertion impedance indicated. Insertion impedance is the impedance reflected into the conductor under test by the current probe. In conventional transformer applications reflected impedance is treated as an inherent circuit component because it is always there. In this case, the current probe is a foreign element which inserts an impedance into the conductor under test. Thus, probe loading is defined by the insertion impedance.

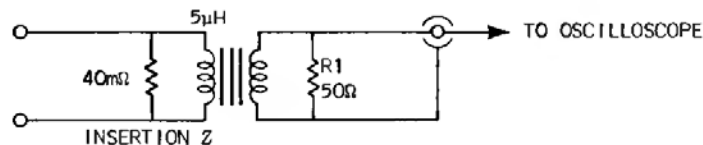


Fig. 5-2. Closed core schematic.

The insertion impedance can be determined from basic transformer theory if the secondary impedance (current probe impedance) is known. From the schematic we note R1 shunts the terminated 50 Ω cable. Thus, the secondary drives a 25 Ω load. The inductance of the secondary can be determined from the formula:

$$L_s = k \frac{N^2 A \mu_e}{l} \text{ where,}$$

$L_s$  = self-inductance in henrys,

$N$  = number of turns in the coil,

$A$  = effective cross-sectional area of the core -- depends on construction techniques,

$l$  = effective length of magnetic path in core,

$\mu_e$  = effective permeability,

$k$  = the required constants.

The indicated quantities are readily determined except for  $l$  and  $\mu_e$ . The effective length of the magnetic path is reasonably approximated by using the center-line length of the core. This approximation holds for regular shapes, such as toroidal or rectangular cores. The value of  $\mu_e$  is more difficult to determine. The methods used to find the proper value of  $\mu_e$  will depend on how we define inductance. This definition will depend on the application of the inductor. Some examples will clarify the situation.

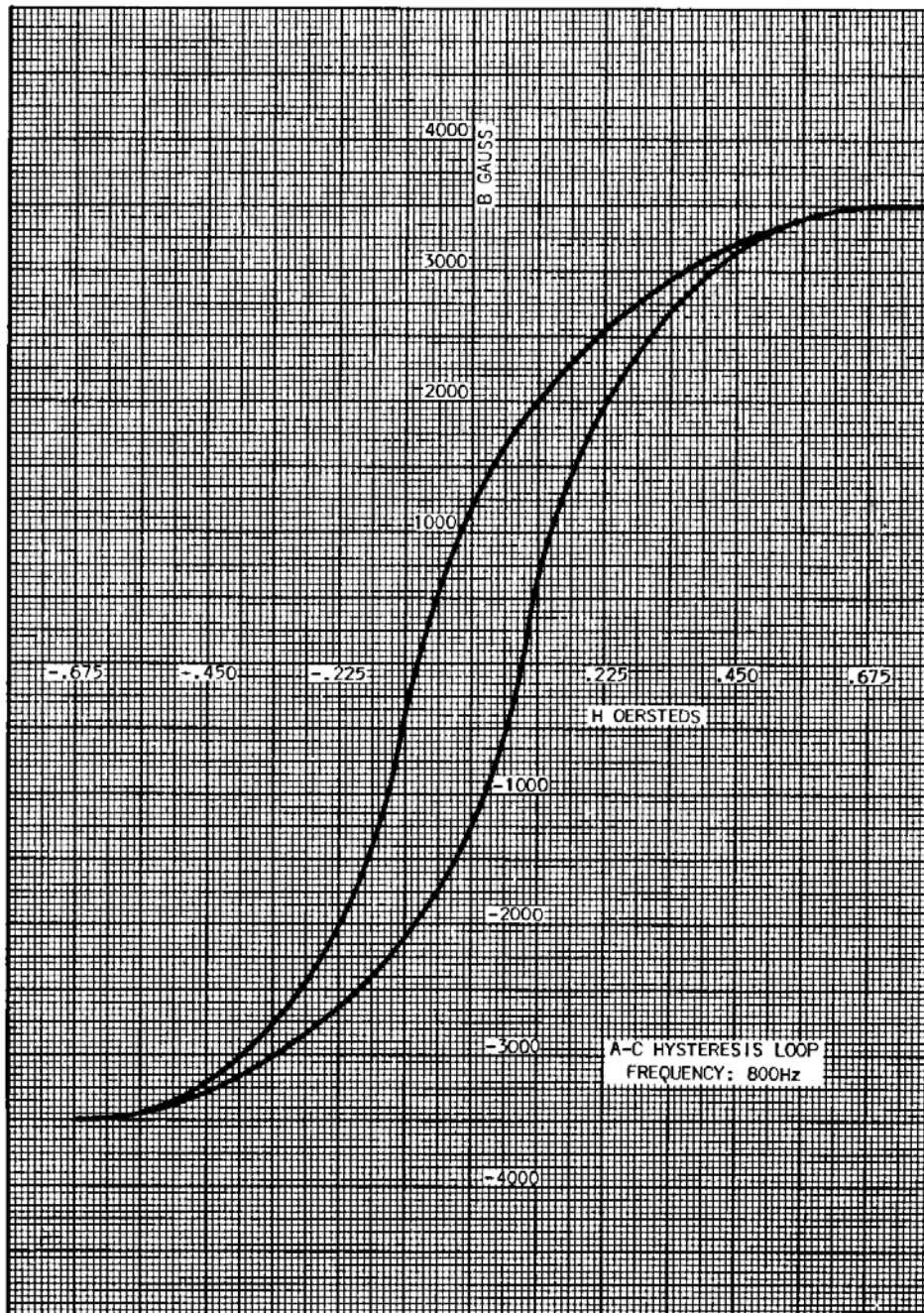


Fig. 5-3.

core  
permeability

Fig. 5-3 is the B-H or hysteresis loop of a magnetic core material used in Tektronix current probes. Permeability, the ratio of B to H at any point on the graph, is largest near the origin. We can expect best signal transfer when the probe monitors small signals that do not have a DC component. In this case the probe application is similar to a line-filter inductor and the quantity we are concerned with is the initial permeability. We

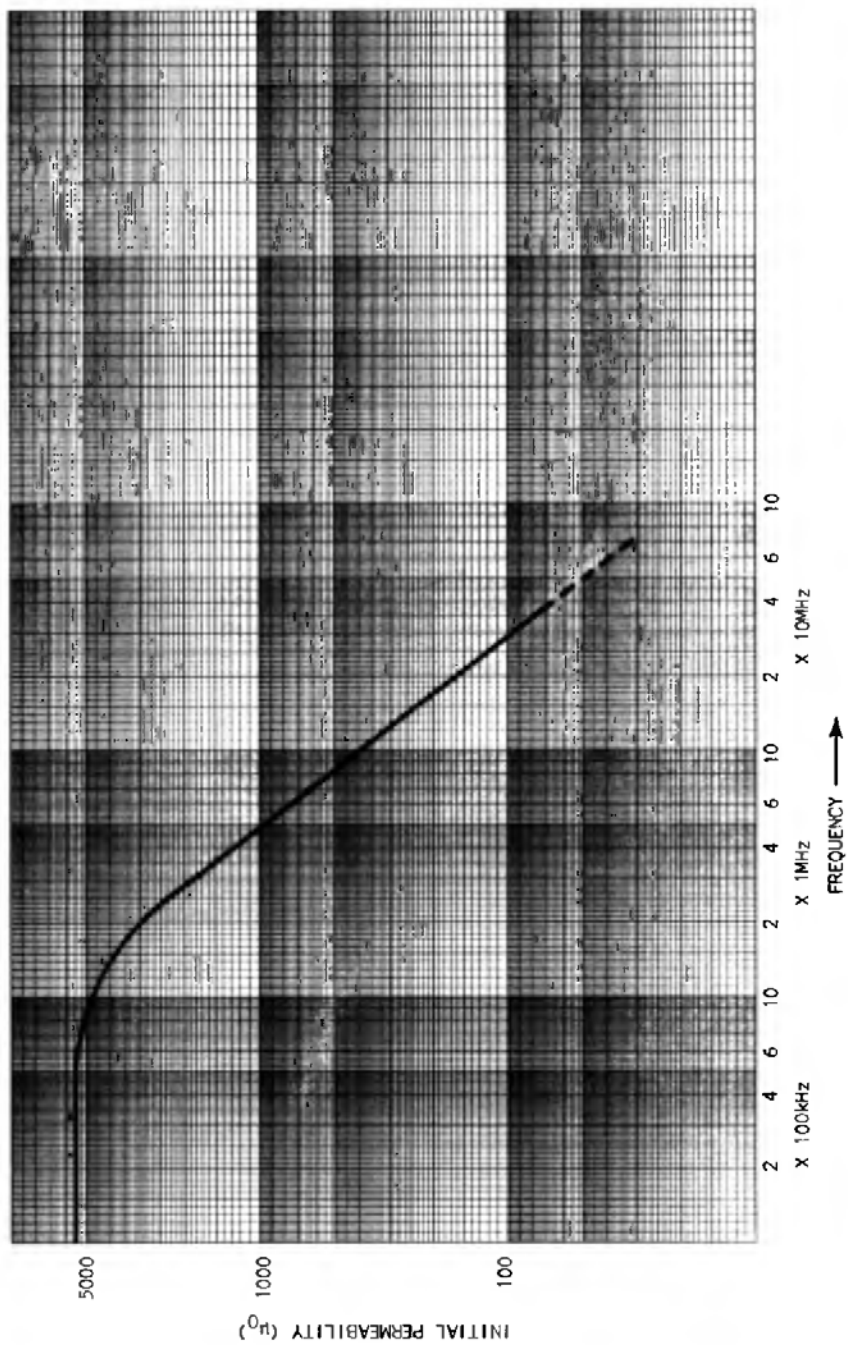


Fig. 5-4. Initial permeability vs frequency (taken at constant impedance source).

will find however, that initial permeability changes with frequency. See Fig. 5-4. This graph was obtained by using a signal generator with constant output current and output impedance as the output frequency was varied. We note that permeability decreases with an increase in applied frequency. This change in permeability must be considered in wide band applications such as the current probe.

The term "initial permeability" assumes no DC signal component. When we monitor the output of a transistorized amplifier, we must consider the effects of small DC currents, say 10 mA or less. Here the current probe operation is similar to an interstage transformer application. Again we refer to the core B-H loop, Fig. 5-3, and find that the quiescent current shifts the operating point of the transformer. The operating point is displaced from the origin by the magnetization caused by the DC current. For small AC signals, we are concerned with the incremental permeability. Fig. 5-5 is a graph of permeability versus DC milliampere turns.

shifting  
operating  
point

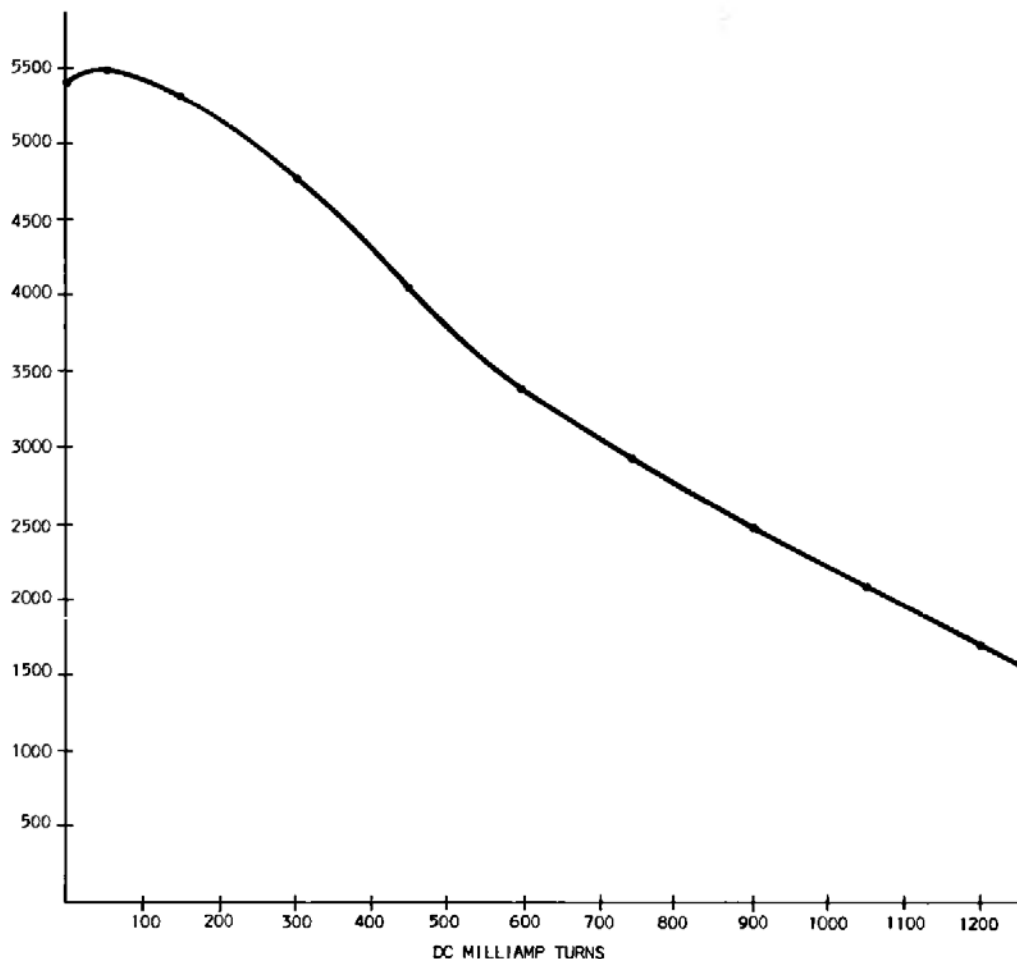


Fig. 5-5. Permeability vs DC mA turns.

worst  
case  
insertion  
impedance

We find permeability decreases as the DC signal increases, excluding the small increase at 80 milliampere turns. For large, low frequency signals we are interested in another definition of inductance. We want to know how much energy can be applied to the transformer without saturating the core. This signal condition is similar to the operating conditions of a power supply filter choke. The inductance of the transformer secondary is dependent on the measurement conditions. The largest value of  $\mu_e$  we obtain is the initial permeability. Using this for  $\mu_e$  we will obtain the largest value for inductance. Therefore, we will use this number to calculate worst case insertion impedance.

Let us return to the circuit of Fig. 5-2. The inductance of the secondary is 3.125 mH and the secondary drives a 25  $\Omega$  load,  $R_1$  in shunt with the terminated 50  $\Omega$  cable. From basic transformer theory, the reflected resistance and inductance are given by:

$$R_1 = R_s \frac{N_p^2}{N_s^2} \text{ and } L_1 = L_s \frac{N_p^2}{N_s^2} \text{ where,}$$

$R_1$  = insertion resistance,  
 $R_s$  = secondary resistance,  
 $L_1$  = insertion inductance,  
 $L_s$  = secondary inductance,  
 $N_p$  = primary turns,  
 $N_s$  = secondary turns.

Hence,

$$R_1 = 25 \Omega \frac{1}{25^2} = 40 \text{ m}\Omega,$$

$$L_1 = 3.125 \text{ mH} \frac{1}{25^2} = 5 \text{ }\mu\text{H}.$$

Thus, the current probe inserts 40 m $\Omega$  shunted by 5  $\mu$ H in series with the circuit under test.

primary  
current  
and CRT  
deflection

The oscilloscope deflection factor is given in volts/division but this probe is a current sensing device. We need a method to determine primary current from CRT deflection. Assume 1 division of vertical deflection with a 50 mV/div deflection factor. Then the current through  $R_{scope}$  is 1 mA



and a 1 mA current also flows through R<sub>1</sub>. The current in the transformer secondary is 2 mA. From basic transformer theory, we know the relationship between primary and secondary current is given by:

$$I_p = I_s \frac{N_s}{N_p} \text{ where,}$$

$I_p$  = primary current,  
 $I_s$  = secondary current,  
 $N_p$  = primary turns,  
 $N_s$  = secondary turns.

Thus,

$$I_p = 2 \text{ mA} \frac{25}{1} = 50 \text{ mA.}$$

In this example, we have 50 mA of primary current and 50 mV of CRT deflection. Thus, probe sensitivity is 50 mA/50 mV or 1 mA/mV.

We have calculated the so called mid-frequency transfer characteristic of the current probe. We know that the transformer is a frequency sensitive device, i.e., the secondary voltage is dependent on the input current rate of change. For our discussion of frequency response we will use a 5 turn secondary with inductance of 125  $\mu$ H. The 5 turn secondary drives a terminated 50  $\Omega$  cable which is shunted by a 50  $\Omega$  resistor. The low frequency 30% down point is easily determined. We solve for the frequency where  $X_L$  equals  $R$ , in this case 25  $\Omega$ .

$$f = 25/2\pi L \approx 35 \text{ kHz with } L = 125 \mu\text{H.}$$

flux  
and  
frequency

In our discussion of current probes, we have implied ideal transformer operation. That is, all the primary flux links the secondary and all the secondary flux links the primary. As the frequency of the input signal is increased without bound, we find this condition cannot be true. Assume we have a signal generator which supplies a constant current as its output frequency is varied from zero toward infinity. As the output frequency is increased from zero, we would find the secondary voltage also increases from zero. This action would continue

leakage  
inductance

through the 30% down frequency until we reach some frequency which we call  $f_1$ . As the input frequency goes above  $f_1$ , the secondary voltage remains nearly constant. The secondary volts will remain constant until we reach a point we call  $f_2$ . At  $f_2$  the secondary voltage will decrease as the input frequency continues to increase. At frequencies between  $f_1$  and  $f_2$ , an increase in current rate of change did not cause an increase in secondary voltage. And above  $f_2$ , secondary voltage decreases as the current rate of change increases. Thus, we have a signal loss which is caused by a component we call leakage inductance.

Leakage inductance describes the primary flux that does not link the secondary windings. At low frequencies, leakage inductance is negligible. Between  $f_1$  and  $f_2$ , an increase in leakage inductance is matched by the increase in current rate of change and the secondary voltage is constant. Above  $f_2$ , leakage inductance increases more rapidly than the current rate of change and the secondary voltage goes toward zero. The effects of leakage inductance are graphically illustrated in Fig. 5-6. The graph is a plot of secondary voltage (shown as a percentage of maximum voltage) versus input frequency. We have plotted probe response under three conditions of initial magnetization as indicated by the DC signal

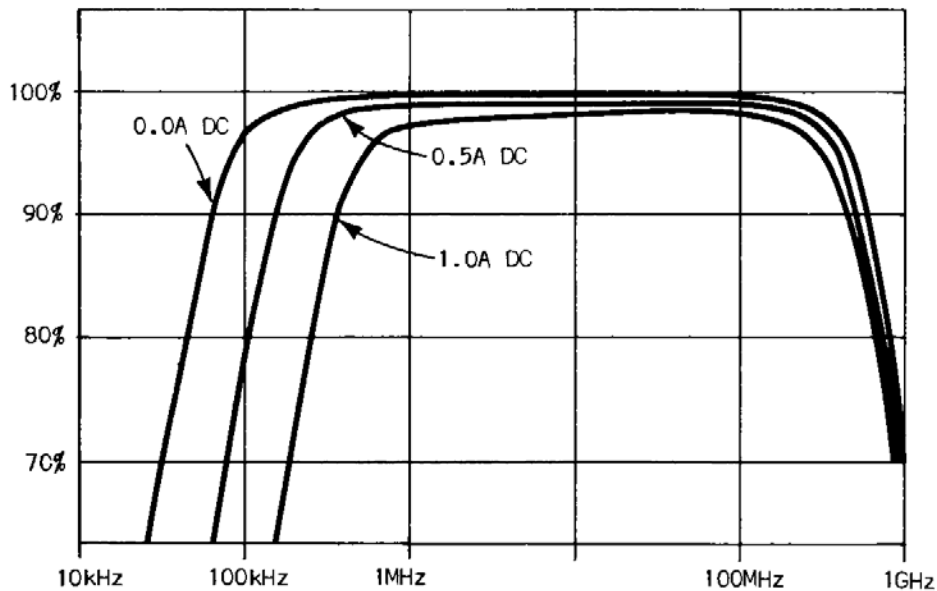


Fig. 5-6. Typical frequency response.

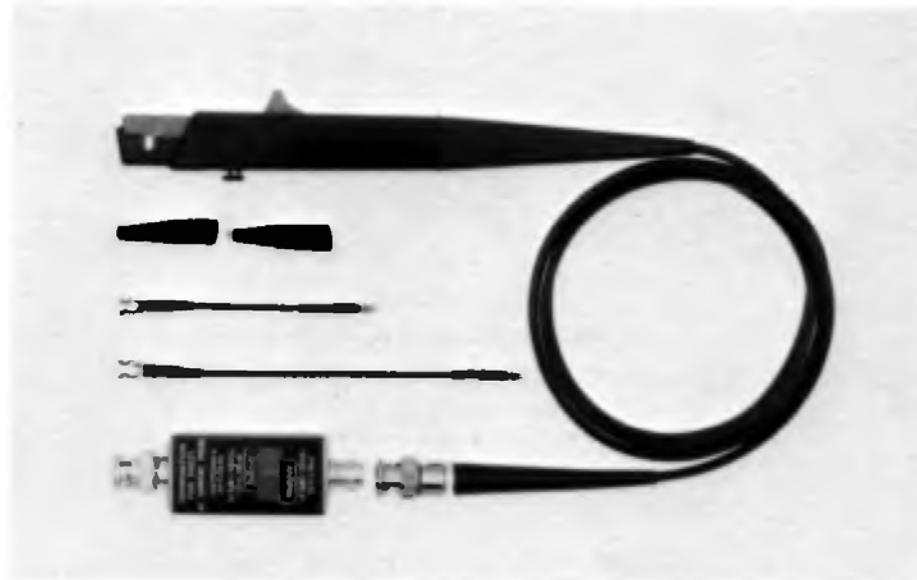


Fig. 5-7. Split core current probe.

in the conductor under test. As we would expect from our discussion of permeability, a DC signal component does not improve probe response.

split  
core  
probes

The closed core passive current probe is the fastest Tektronix current probe. Its main disadvantage is that the circuit under test must be broken to monitor the signal. The split core probes eliminate this problem. A photograph is shown in Fig. 5-7.

The split core probe is opened by operating the thumb controlled slide switch. The probe is then clipped around a conductor and the slide switch moved forward. With the exception of the first Tektronix split core probe, P6016, a spring returns the slide switch to the closed position. On occasion, these probes have functioned as crude, but effective, wire cutters.

termination

The passive termination is included in the probe package, the active termination is purchased separately. The passive termination is designed for one probe only as the labeling indicates. The signal is coupled from the probe head to the termination via a 62.5  $\Omega$  coaxial cable. The design

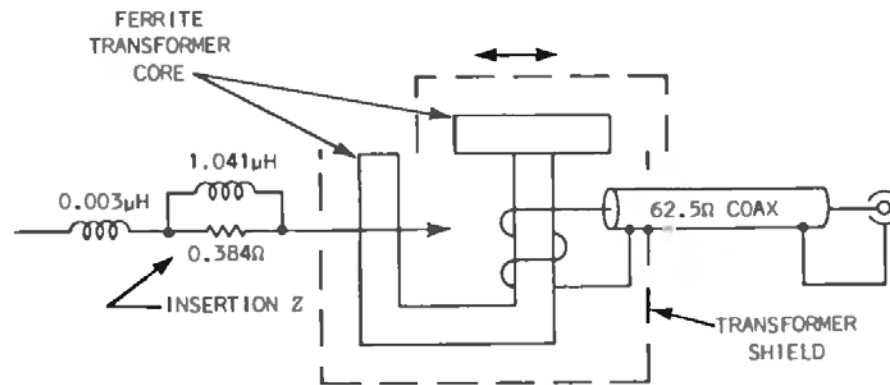


Fig. 5-8. Split core schematic.

of the probe head circuitry has undergone two stages of evolution. Fig. 5-8 is a schematic of the first probe head and insertion impedance. The core is a stationary U-piece and a movable flat piece. The sliding arrangement provides a wiping action which tends to clear foreign matter from the exposed portion of the core. The grounded transformer shield is exposed when the switch is opened. Care must be exercised when clipping the probe around a bare wire.

For our purposes, the current in the conductor under test is the input signal and the voltage across the transformer secondary is the output signal. The voltage loss across the insertion impedance is negligible because the insertion impedance is about  $1 \Omega$  at 20 MHz, the 30% down frequency of this probe. Conductor to probe capacitance is a function of the size and composition of the conductor. When the probe is clipped around a 1/2 watt metal film resistor, this capacitance is about 2.8 pF and clipped around a No. 16 insulated wire, about 0.8 pF. Maximum voltage from conductor to probe is 600 V DC plus peak AC.

voltage  
loss and  
insertion  
impedance

passive  
termination

Probe sensitivity is 2 mA/mV or 10 mA/mV when used with the passive termination. A schematic of the passive termination is shown in Fig. 5-9. Switch 1 is shown in the 2 mA/mV position. R2 and R3 terminate the cable in its  $Z_0$ , 62.5  $\Omega$ . The shunt combination of R6 and R7 in series with R8 isolate the input capacitance from the termination resistors, R2 and R3. R9 and C9 tend to minimize aberrations at the front corner of a fast pulse. L7 provides high speed peaking.

When Switch 1 is set to the 10 mA position, R2 and R3 form a voltage divider which reduces the signal to 20% of the transformer voltage. L5 shunts R2 to provide better low frequency response. As the frequency of the signal goes toward zero, the voltage across the transformer goes toward zero. The shunting effect of L5 reduces the voltage divider ratio to extend the low frequency response.

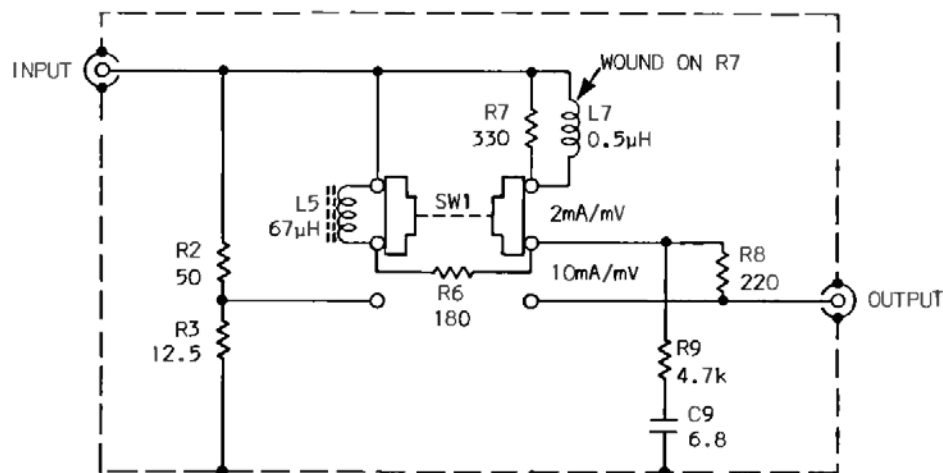


Fig. 5-9. Passive termination.

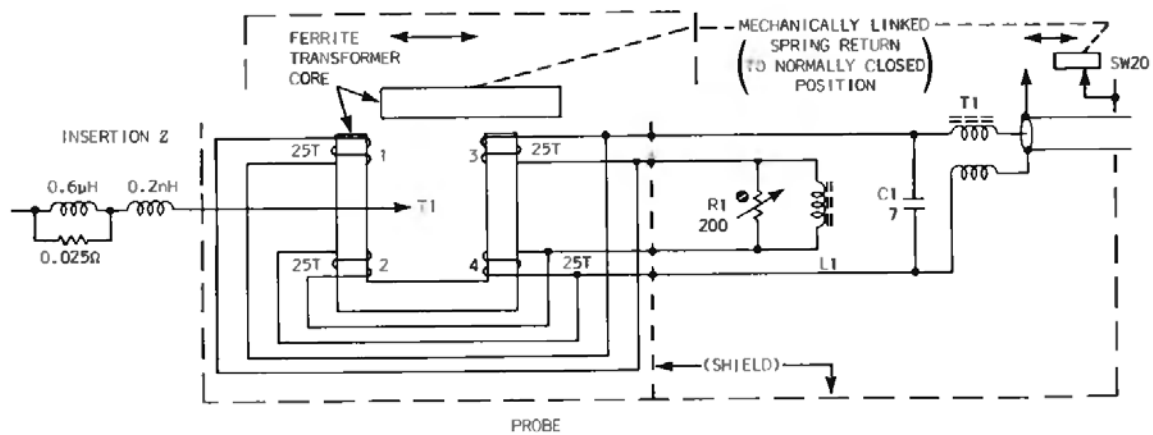


Fig. 5-10.

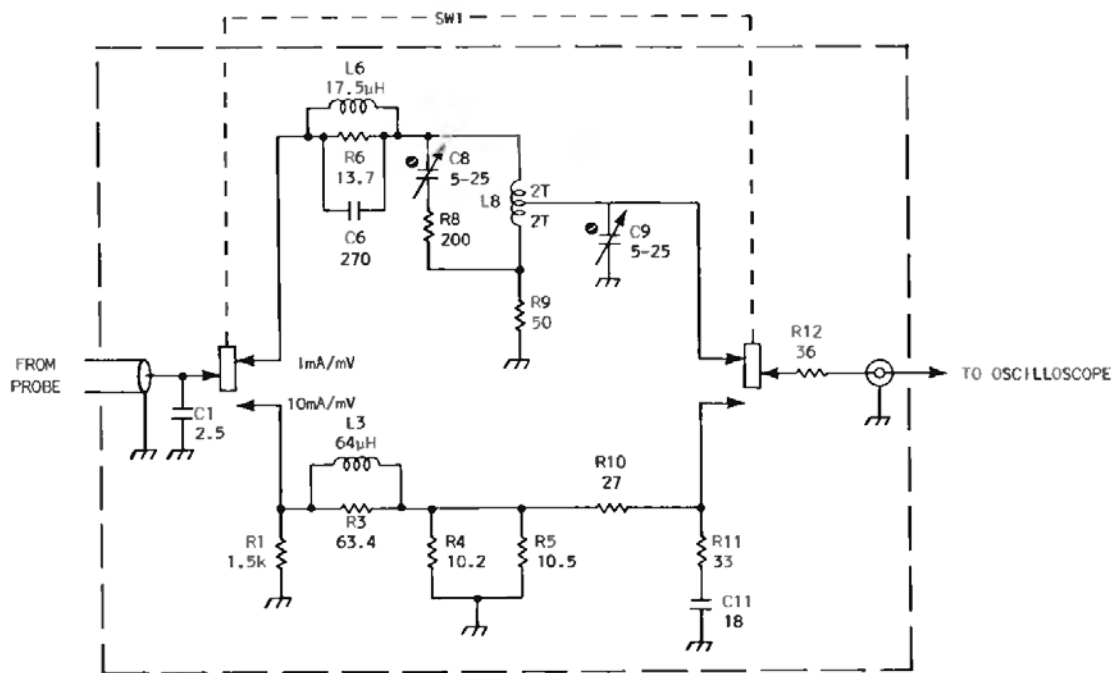


Fig. 5-11. Termination for 50-turn current probe.

- frequency response and fast rise pulse
- This particular probe-termination has a high frequency 30% down frequency of 20 MHz. When a fast rise pulse is applied to the probe, the front corner can be peaked or rolled off by moving the conductor within the probe opening. This problem is minimized by adding compensation circuitry in the probe head. A schematic of this type probe and insertion impedance is shown in Fig. 5-10.
- 50 turn circuit description
- The secondary has 50 windings which are divided into two 25 turn sections. Parallel windings No. 1 and No. 3 are in series with parallel windings No. 2 and No. 4. R1 and L1 are in series with the two groups of windings to provide damping consistent with good pulse response. C1 is a high frequency bypass capacitor which minimizes peaking. T14 rejects signal currents common to the ground path and the cable center conductor. The ground connection to the transformer shield is disconnected when Switch 20 is opened.
- 50 turn termination
- The output of the probe head is applied to the termination via the 62.5  $\Omega$  cable. The schematic of the termination is shown in Fig. 5-11. Deflection sensitivity is selectable at either 1 mA/mV or 10 mA/mV. Switch 1 is shown in the 1 mA/mV position. The cable termination is R6 in series with the 50  $\Omega$  T-coil network. The T-coil network, L8, C8 and R8, is terminated by R9. R6 and the terminated T-coil form a voltage divider which transfers 80% of the probe head signal to the oscilloscope. This arrangement reduces probe sensitivity to maintain a convenient deflection sensitivity.
- sensitivity
- Probe sensitivity is a function of the turns ratio and secondary impedance. With 50 mA of current in the conductor under test, the secondary current is 1 mA. A 1 mA current through 62.5  $\Omega$  is 62.5 mV across the transformer secondary. With these conditions the probe sensitivity is 1 mA/1.25 mV, 50 mA/62.5 mV. Ergo, a divider is used in the termination to eliminate the awkward deflection sensitivity.
- circuit description
- L6 in shunt with R6 extends the low frequency response by reducing the divider ratio as the transformer output decreases when the applied frequency goes toward zero. C6 serves a similar function as the applied frequency is increased.

The T-coil is used to isolate the cable from the input capacitance of the oscilloscope. C9 is adjusted to maintain the capacitive loading of the T-coil constant regardless of the input capacitance of the oscilloscope. C8 and C9 are adjusted for minimum aberrations on the pulse response. After the capacitors are properly adjusted, R1 in the probe head is adjusted for minimum pulse aberrations.

When Switch 1 is set to the 10 mA/mV position, 8% of the transformer signal is applied to the oscilloscope. The cable termination is R3 in series with the shunt combination of R4 and R5. L3 provides low frequency boost. The voltage divider in conjunction with R10, R11 and C11, is sufficient to isolate the cable from the input capacitance of the oscilloscope in the 10 mA/mV position.

The bandpass of this particular probe-termination is from 935 Hz to 120 MHz in the 10 mA/mV position. The high frequency response is quite respectable but the low frequency response leaves something to be desired. The low frequency response is improved by building a transformer with a higher inductance secondary. The schematic of a 125 turn secondary probe head is shown in Fig. 5-12.

125 turns  
for low  
frequency

125 turn  
circuit  
description

Three sets of shunt windings are used in the 125 turn secondary with damping provided by R1-L1 and R2-L2. Pulse fidelity is maintained by roll-off network R3-C3, peaking network R4-L4, and roll-off capacitor C4. The three networks have different time constants and control different portions of the response. T4 rejects signals common to the ground path and cable center conductor. The transformer signal is connected to the termination via a 62.5  $\Omega$  cable. The transformer shield is ungrounded when the slide switch is opened.

calibration

The pulse response of the probe is calibrated separately from the termination to avoid interaction between the controls. The oscilloscope input is terminated in 50  $\Omega$  and an impedance matching network is required to transfer the signal from the 62.5  $\Omega$  cable to 50  $\Omega$ . A schematic is shown in Fig. 5-13A. A GR\* insertion unit is a convenient housing for the impedance matching network. Adjust R1, R2 and C3 for best pulse response as indicated in Fig. 5-13B. The probe-termination system can now be calibrated.

\*General Radio Corp., West Concord, Mass.



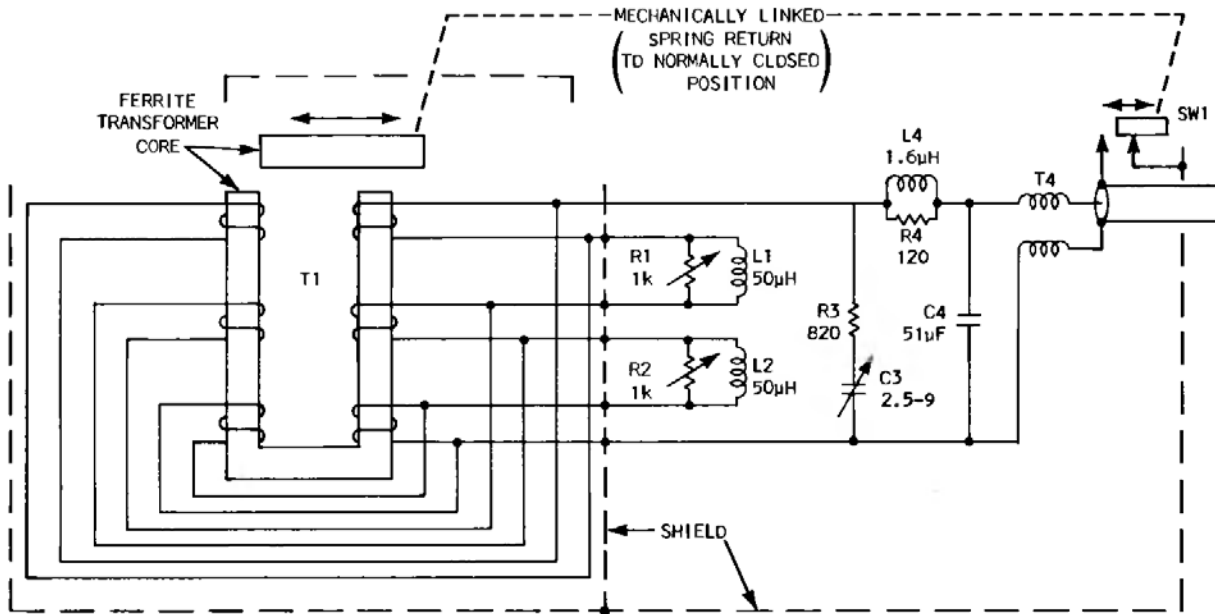
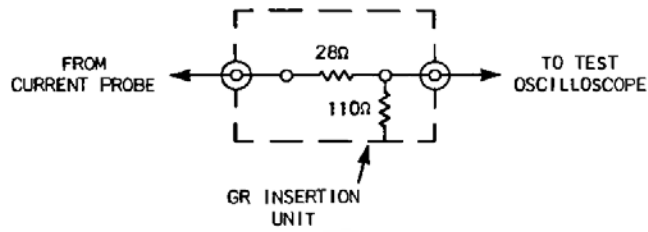
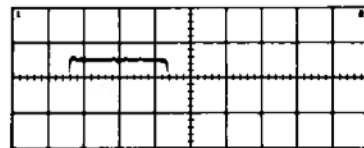


Fig. 5-12. 125-turn probe.



IMPEDANCE MATCHING NETWORK CONSTRUCTED IN GR INSERTION UNIT

(A)



PROBE RESPONSE PROPERLY ADJUSTED

(B)

Fig. 5-13.

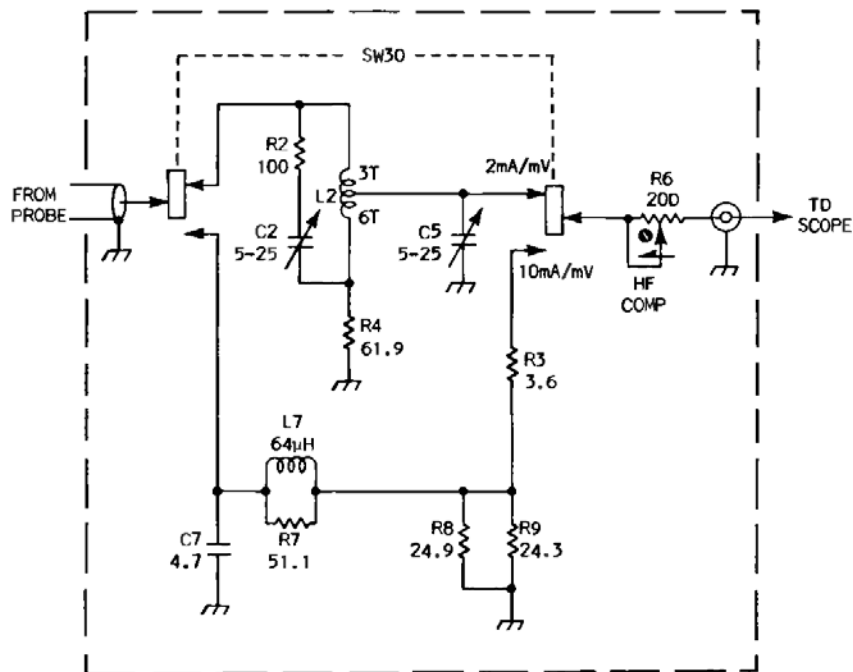


Fig. 5-14. 125-turn termination.

circuit  
description

A schematic of the termination is shown in Fig. 5-14. Deflection sensitivity is selectable at either 2 mA/mV or 10 mA/mV. Switch 30 is shown in the 2 mA/mV position. The cable is terminated by the 62.5  $\Omega$  T-coil which in turn is terminated by R4. C5 compensates for the different input capacitance of various oscilloscopes and R6 is adjusted to eliminate peaking in the pulse response. In the 10 mA/mV position, 20% of the transformer voltage is applied to the oscilloscope via voltage divider R7 and the shunt combination of R8 and R9.

This probe-termination has a bandpass of 120 Hz to 54 MHz in the 10 mA/mV position. The added inductance of the 125 turn secondary extends low frequency response but not to the point where line frequency signals are above the 30% down frequency. An active termination can be built that will have a low input impedance to low frequency signals. The low input impedance increases the secondary L/R ratio to extend low frequency response beyond the power line frequencies.

active  
termination

A schematic of an active termination is shown in Fig. 5-15. This active termination, which is designed for use with the probe head of Fig. 5-8, has an attenuator, amplifier and -15 V supply. The attenuator provides deflection sensitivities from 1 mA/div to 1 A/div with a bandpass from 30 Hz to 17 MHz. The amplifier is designed to work into an oscilloscope deflection factor of 50 mV/div.

circuit  
description

In the 1 mA/div position, R1 in series with the emitter impedance of Q54 ( $10 \Omega$ ) terminates the cable. In the 2 mA/div position, R8A in shunt with the series combination of R8B and Q54 input impedance form a 2:1 current divider. Thus, the deflection sensitivities are provided by increasing the resistance in series with Q54. At low frequencies, L1 shunts R1 and the input resistance of the amplifier goes to  $10 \Omega$ . We have decreased the terminating resistance which increases the ratio of secondary inductance to terminating resistance. This increase of L/R time constant extends probe response beyond the line frequencies.

Common base amplifier Q54 is the first stage of amplification. R52 in the base circuit provides adjustment of the input impedance to insure that the cable is properly terminated. The voltage at Q54 emitter changes when the resistance of R52 is varied. Thus, R54 current changes and the emitter resistance of Q54 changes. With the deflection sensitivity at 1 mA/div, a signal is applied to the current probe and the CRT deflection is noted. The

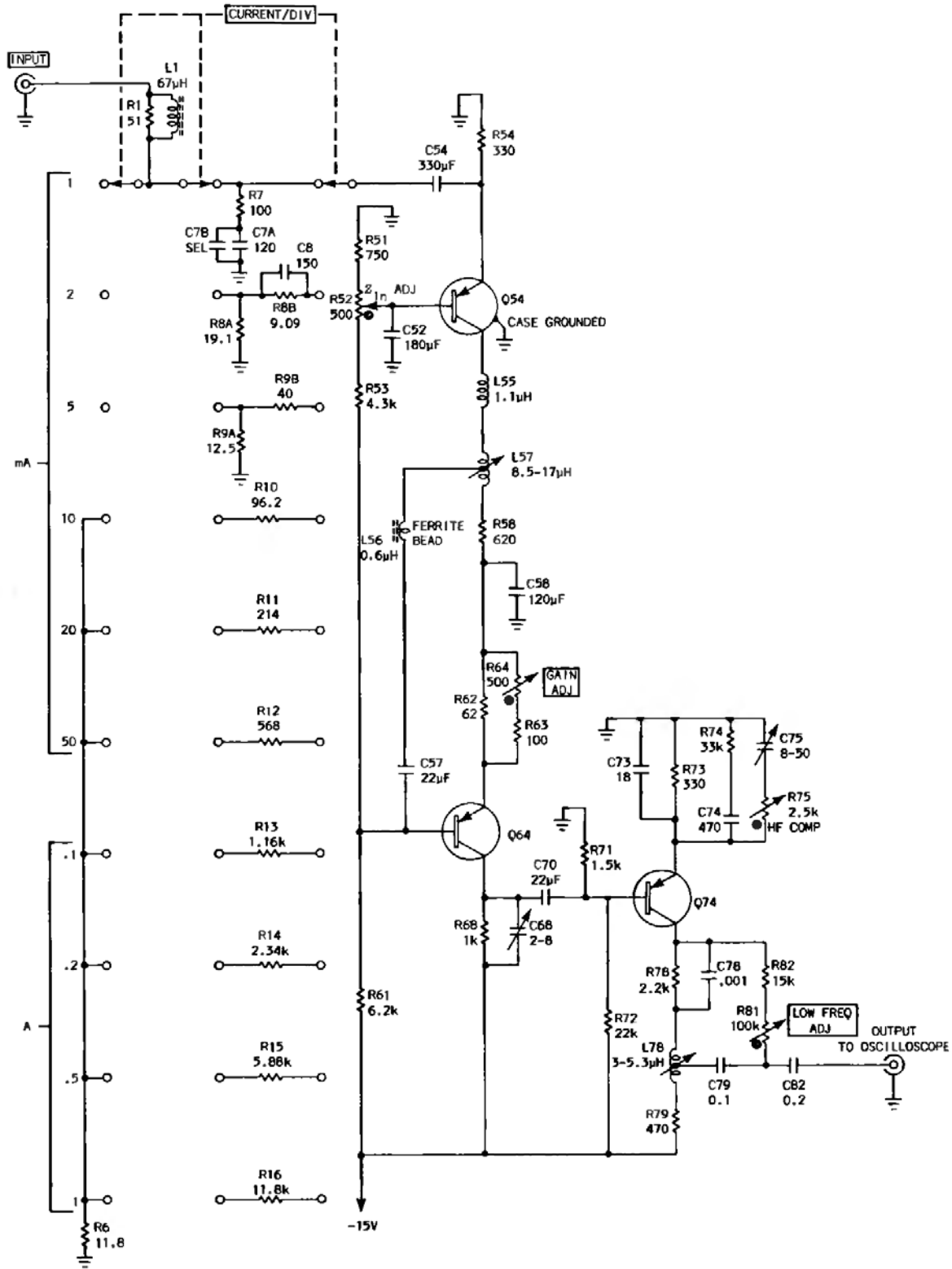
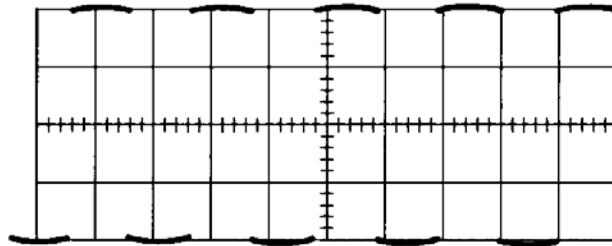


Fig. 5-15. Active termination.

sensitivity is switched to 2 mA/div, and R52 is adjusted for 1/2 the previously noted CRT deflection. A change in R52 value will change the deflection in the 1 mA/div position, ergo, repeat the above procedure to eliminate interaction. This adjustment is made before amplifier gain is calibrated.

The output of Q54 is taken at the center tap of L57 and applied to Q64 base via L56 and C57. C58 is a signal ground for the collector circuit of Q54 and the emitter circuit of Q64. (If Q54 is replaced, C7B may require selection for proper high frequency response at 1 mA/div.) R64 is adjusted to calibrate amplifier gain by setting the collector to emitter resistance ratio of Q64. C68 provides a means of minimizing high frequency ringing. The output of Q64 is applied to the base of Q74 via C70.



NORMAL CALIBRATOR CURRENT WAVEFORM OBTAINED WHEN THE LOW FREQUENCY ADJUSTMENT AND GAIN ADJUSTMENT CONTROLS ARE SET CORRECTLY. SWEEP RATE IS 0.5 ms/DIV.

Fig. 5-16.

high  
and low  
output

Q74 has two output paths: at high frequencies the signal is coupled to the oscilloscope from the center tap of L78 via C79 and C82 and at low frequencies the signal is coupled to the oscilloscope from Q74 collector via R81, R82 and C82. R81 is adjusted to make the corners of the calibrator signal the same amplitude. See Fig. 5-16.

After the low frequency response is adjusted, the amplifier gain should be checked and readjusted if necessary. High frequency compensation is provided by L78 (collector peaking) and C75-R75 (emitter peaking). Displayed noise of probe-amplifier is equivalent to 100  $\mu$ A P-P signal or 1/10th CRT division at the 1 mA/div setting.

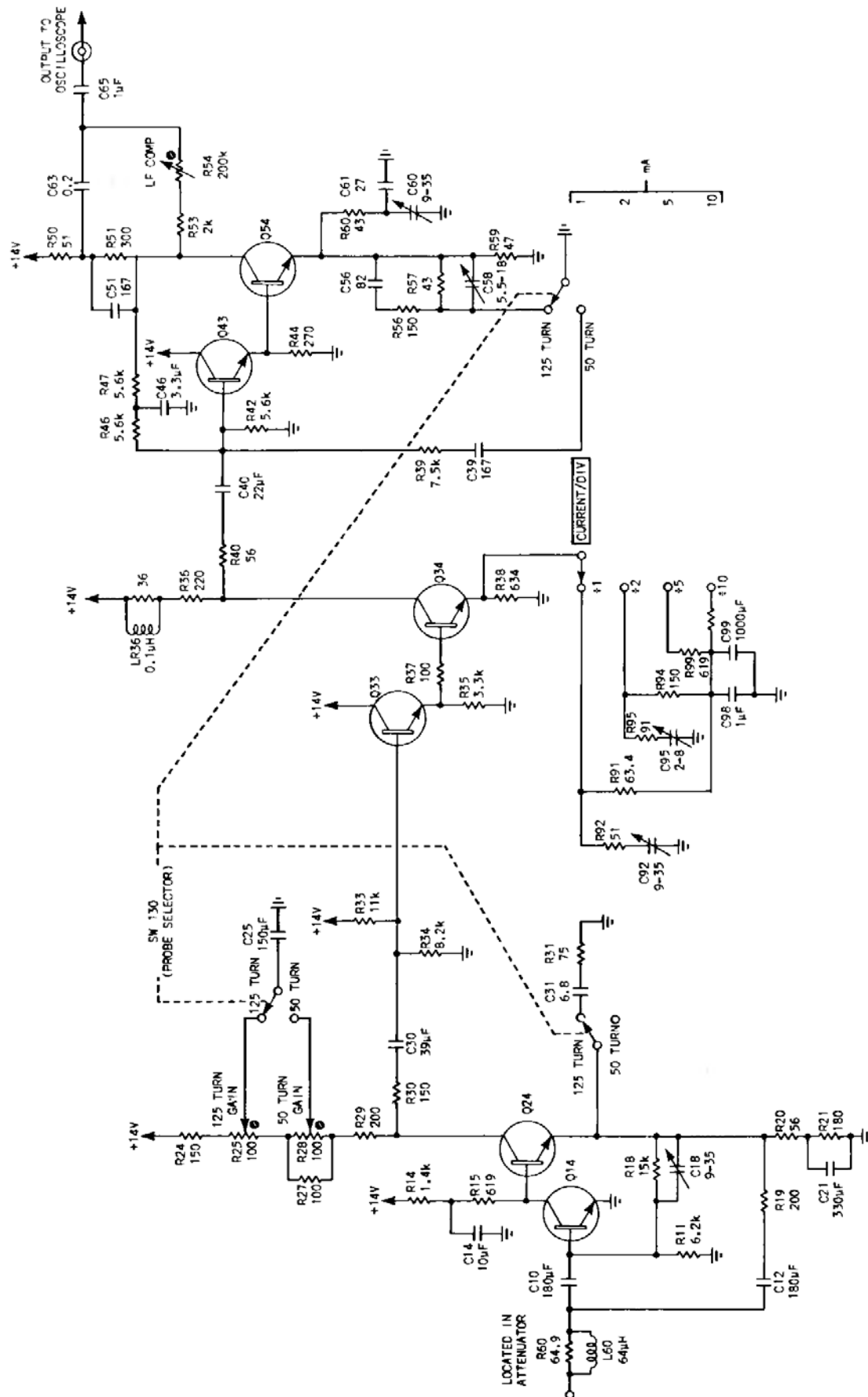


Fig. S-17. Current probe amplifier.

combination  
termination

The previous active termination is not designed to accept signals from a 50 turn secondary probe. An active termination that can be used with both types of current probes is shown in Fig. 5-17. The amplifier consists of an attenuator, a Feedback Input pair, Q14 and Q24, Isolation EF, Q33, Switched Gain Amplifier, Q34, and Output pair, Q43 and Q54. The attenuator is not shown in detail because it is similar to the attenuator circuit discussed previously. The attenuator in conjunction with the gain switching provides deflection sensitivities from 1 mA/div to 1 A/div.

circuit  
description

The base of Q14 is a virtual ground point and the amplifier input impedance is about  $2 \Omega$ . The negative feedback path from the emitter of Q24 is through R19 and C12. Signal voltage at Q24 emitter is a function of R19 divided by the attenuator resistance. At low frequencies, the attenuator resistance is bypassed and the transformer secondary drives the virtual ground input. The low frequency response is extended to 12 Hz (30% down frequency) with the 125 turn secondary. The probe selection switch changes the signal gain at Q24 collector so that amplifier gain is correct for the 50 turn or 125 turn probes. The probe selection switch connects a signal ground via C25 to the selected potentiometer, R25 or R28. The DC characteristics of the circuit are not affected by probe selection.

R11 and R18 form a divider circuit which establishes DC operating characteristics and provides temperature compensation. C18 provides high frequency compensation. In the 125 turn position, C31 and R31 are switched into Q24 emitter circuit to give high frequency peaking. Q24 output is applied to the base of Q33 via R30 and C30. Q33 isolates Q24 collector circuit from effects of gain switching in Q34 emitter circuit. Gain switching is accomplished by changing Q34 emitter resistance. High frequency peaking is provided by LR36 in Q34 collector circuit. Q34 output is applied to Q43 via R40 and C40.

output  
description

Emitter Follower, Q43, drives the unity gain output amp, Q54. Divider circuit R46, R47 and R42 establish the DC operating conditions. R51 provides thermal compensation and C51 retains the high frequency response. R60, C60 and C61 are the peaking circuit for the 50 turn probe. C39 and R39 are switched into Q43 base circuit to compensate for overpeaking

introduced by LR36 for the 50 turn probe. R56, R57, C56 and C58 provide peaking for the 125 turn probe. At low frequencies, Q54 output is applied to the oscilloscope via R53, R54 and C65. The Low Frequency compensation is adjusted as described for the previous amplifier.

DC and  
low  
frequency  
signals

The clip-on passive current probe is a convenient tool with a wide performance range. However, probe performance is a little weak at the low frequency end. In our discussion of permeability we noted that DC signal components degraded probe response, especially at the low frequency end. We can get back good low frequency response if we null the effects of the primary DC signal.

Consider the circuit of Fig. 5-2. We have returned to our old friend the 25 turn coil that drives a 25  $\Omega$  load, R1 in shunt with the terminated 50  $\Omega$  cable. In our previous discussion of this circuit, we found that the secondary current was 1/25 of the primary current. The ratio of secondary current to primary current is inversely proportional to the secondary to primary turns ratio:

$$\frac{N_p}{N_s} = \frac{I_s}{I_p}.$$

We cross-multiply the equation to get:

$$I_p N_p = I_s N_s.$$

flux  
nulling

Thus, we state that the primary ampere-turns product is equal to the secondary ampere-turns product. This means that the secondary current produces a magnetic flux with magnitude equal to the flux magnitude produced by the primary ampere-turns. The polarity of the secondary flux is opposite to the polarity of the primary flux. Under *ideal* transformer conditions, all the primary flux links the secondary and vice versa, with the net flux in the core being zero! Thus the transformer is inherently a flux nulling device.

adding  
DC to  
secondary

We can take advantage of this transformer property by introducing a DC signal into the transformer secondary. The situation is shown in Fig. 5-18. We have added a DC generator in series with the transformer secondary. The amplitude of the DC



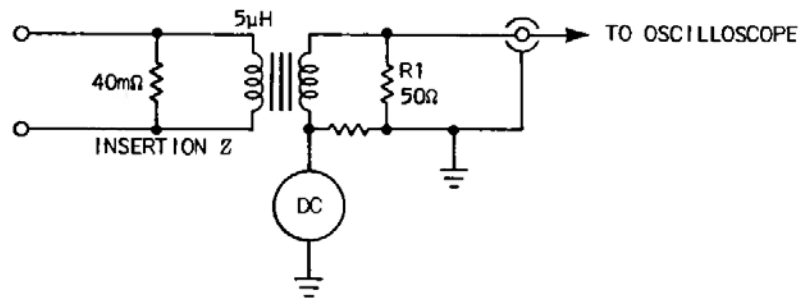


Fig. 5-18.

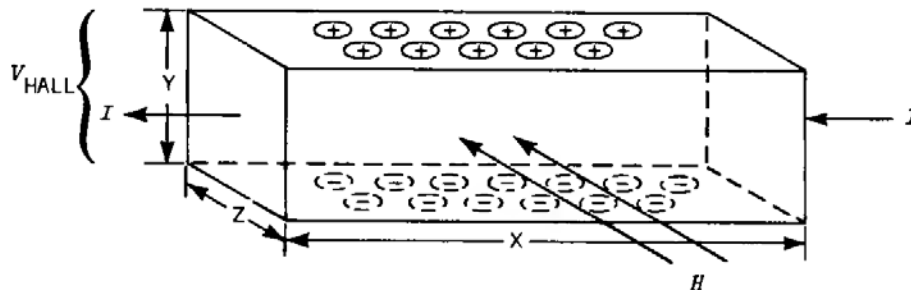


Fig. 5-19. Hall effect.

generator signal must be 1/25 of the primary DC signal. The direction of the DC signal must be such that the flux produced by the secondary DC signal has opposite polarity to flux produced by the primary DC signal. We note that the secondary DC signal also is applied to the parallel combination of  $R_1$  and the terminated  $50\ \Omega$  cable. The voltage dropped across  $R_1$  is a function of the DC and the AC signals. Not only have we canceled the DC flux, we have extended current measurement capability to DC!

Hall  
effect

All we need is a device that is capable of sensing and responding to flux produced by DC signals. The Hall Generator is such a device. The design of the Hall Generator is based on an interaction between moving charges and an unrelated magnetic field. See Fig. 5-19. The particular phenomena we are interested in is called the "Hall effect."

We have a conducting element with dimensions  $x$ ,  $y$  and  $z$ . In the absence of an externally applied

Hall  
voltage

magnetic field,  $H=0$ , we can think of the direct current,  $I$ , as a relatively uniform distribution of charged particles moving along the x axis. When an external magnetic field is applied, the charged particles are displaced. In Fig. 5-19, the holes (+) are shown displaced upward and the electrons (-) displaced downward. The action is similar to electromagnetic deflection in a CRT. Thus a difference in potential exists along the y axis. This difference in potential is called the Hall voltage.

character-  
istics

The amplitude of the Hall voltage is a function of the charge concentration in the conducting element, the magnitude of the magnetic field, the type of material and the thickness of the element. We note as the z dimension approaches zero, the displaced charged particles are more tightly concentrated at the boundaries of the element. We will also find that a semiconductor has a better transfer ratio than a conductor. We can reverse the polarity of the Hall voltage by reversing the direction of the direct current or the direction of the magnetic field. Thus the Hall Generator is amplitude and polarity sensitive.

application

We will mount the Hall Generator in the core of a current probe. The Hall Generator will sense the flux produced by current in the conductor under test. A block diagram of the new probe is shown in Fig. 5-20. The output of the Hall Generator is coupled to  $R_L$  via Low Frequency Amplifier A1 and the transformer secondary. The voltage across  $R_L$ ,

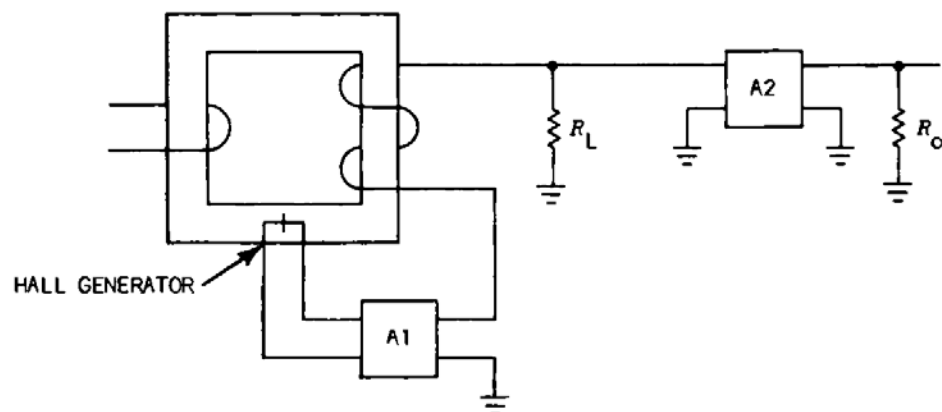


Fig. 5-20. DC current probe.

which is the sum of the transformer voltage and A1 output voltage, is amplified by Wide Band Amplifier A2. The output of A2 is applied to the oscilloscope via a terminated 50  $\Omega$  cable.

We have restricted the bandpass of the Hall Generator by placing Low Frequency Amplifier A1 in the system. In theory, the Hall Generator can have a bandpass from DC to gigahertz frequencies. The Hall Generator is an active device and like other active devices, it generates random noise. We can minimize noise effects by restricting bandpass. Thus, A1 improves signal to noise ratio which gives us a more sensitive system. This DC probe is designed to provide deflection sensitivities from 1 mA/div to 1 A/div with an oscilloscope deflection factor of 50 mV/div. Bandpass is from DC to 50 MHz.

noise

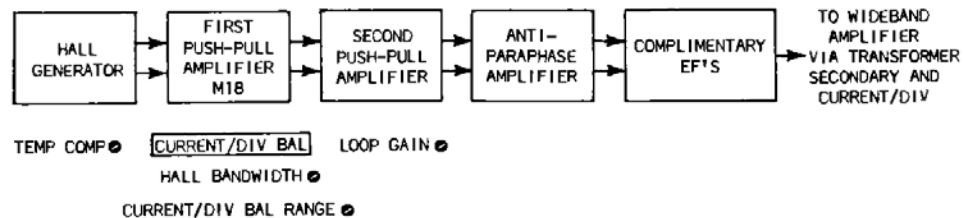


Fig. 5-21. Low-Frequency Amplifier block diagram.

The output of the Hall Generator is a push-pull signal that drives the Low Frequency Amplifier. See Figure 5-21. The 1st Push-pull Amplifier is an integrated circuit, M18, that drives the 2nd Push-pull Amplifier. The CURRENT/DIV BALANCE (located in the external circuitry of the integrated circuit) provides a method of canceling trace shift introduced by changing the CURRENT/DIV switch setting. The Current/Div Balance Range and Hall Bandwidth adjustments are part of the 1st Push-pull Amplifier circuitry.

first  
push-pull

The 2nd Push-pull Amp is a convenience stage. It provides a means of changing the average DC level of the signal such that the Low Frequency Amplifier output is at ground potential under "no signal" conditions. The 2nd Push-pull Amp is also a convenient location for the Loop Gain adjustment.

second  
push-pull

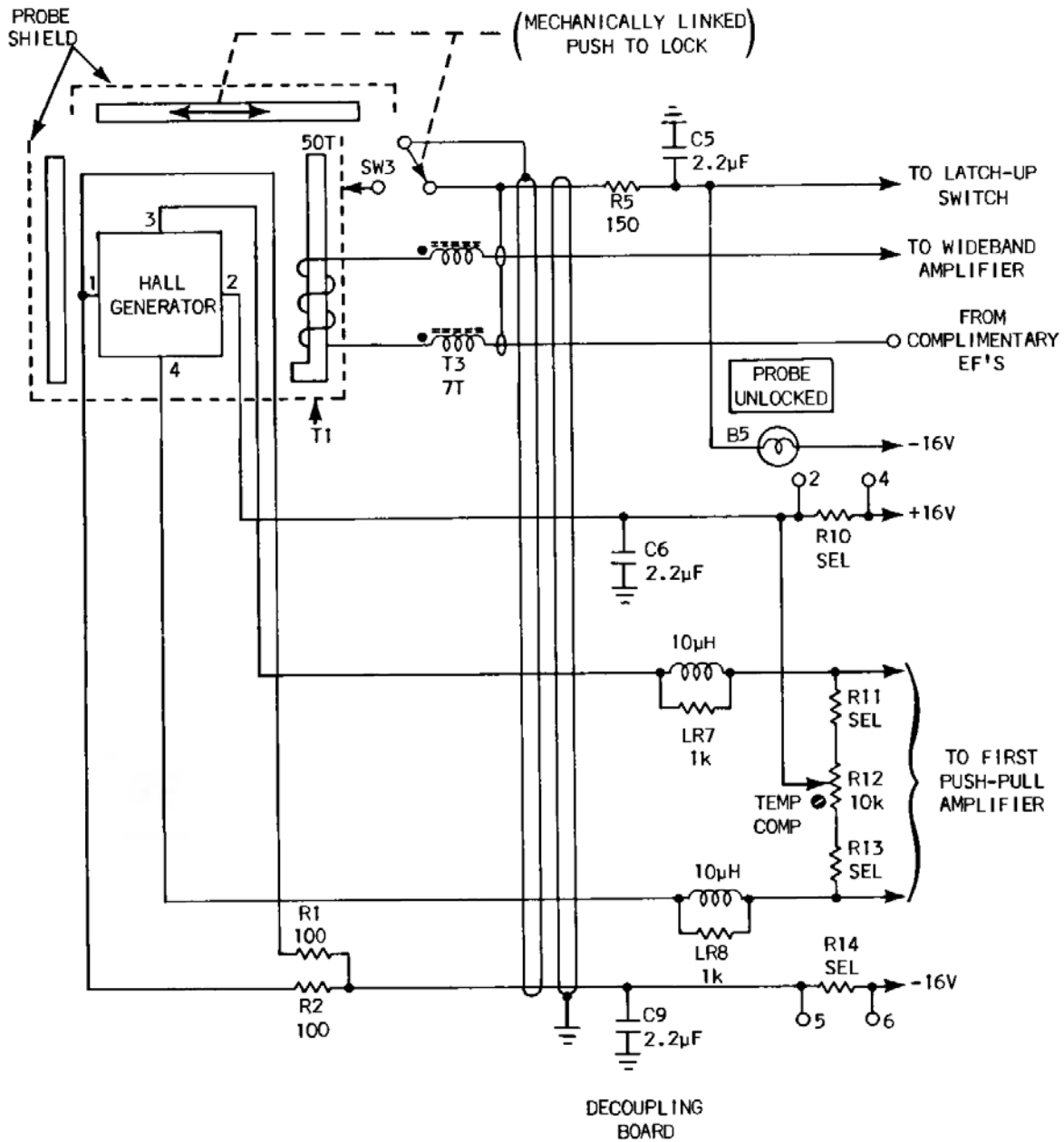


Fig. 5-22. Hall generator.

anti-  
paraphase

The output of the 2nd Push-pull Amp is applied to the Anti-paraphase Amplifier that converts the push-pull signal to a single-ended signal. Under "no signal" conditions, the output of the Anti-paraphase Amp is at ground potential. The Anti-paraphase Amp drives the Complimentary EF's which provide a  $50 \Omega$  source impedance to the transformer secondary, the  $50 \Omega$  cable and the Wide Band Amp which is terminated in  $50 \Omega$ . The use of Complimentary EF's provides a means of processing positive going or negative going signals equally well.

feedback

The output current of the Low Frequency Amplifier flows through the transformer secondary which generates a flux that opposes the primary flux sensed by the Hall Generator. When the secondary flux equals the primary flux, an equilibrium or null condition exists. Thus, the conductor under test, the Hall Generator, the Low Frequency Amplifier and the transformer secondary form a feedback system. The output of the Hall Generator will increase until the primary and secondary ampere-turns products are equal. Thus, the DC signal transfer characteristic is defined by the transformer turns ratio.

Hall  
Generator

Before discussing the circuits of the individual amplifiers, we will discuss the Hall Generator and its associated circuitry. See Fig. 5-22. The control current of the Hall Generator flows through R14, R1, the Hall Generator and R10. The resistance of R14 and R10 is selected for an excitation current of 20 mA with the output terminals, No. 3 and No. 4, near 0 V under "no signal" conditions. The Hall Generator transfer characteristic is approximately  $70 \mu\text{V}/1 \text{ mA}$  in the conductor under test with a 20 mA control current.

temperature  
compensation

Like other semiconductors, the Hall Generator is sensitive to thermal environment. A change in temperature will cause the output terminal voltages to move in the same direction with equal amplitudes. A change in temperature generates a common mode signal which is felt across voltage divider R12 and R11 or R13. (Either R11 or R13 is selected for proper temperature compensation.) Since the common mode signal is across the divider, the voltage at the center tap of R12 equals the common mode signal. The center tap of R12 is tied to the junction of R10 and No. 2 of the Hall Generator. The resultant

change in voltage at No. 2 tends to nullify the thermally generated signal. R11 and R13, along with R10 and R14, are selected to match the characteristics of the Hall Generator. (If the Hall Generator should require replacement, these resistors will be included in the new parts package.)

The output leads of the Hall Generator link about 1/2 the core flux. As the frequency of the applied signal is increased, the output of the Hall Generator will increase due to the 1/2 turn secondary formed by its output leads. The Hall Generator signal is coupled to the 1st Push-pull Amplifier via LR7 and LR8. The LR networks form a voltage divider with the input circuitry of the 1st Push-pull Amplifier to attenuate the higher frequency signals.

unlocked  
bulb

When the thumb controlled slide switch is opened to clamp the probe around a conductor, the ground is removed from the transformer shield. The ground is connected to the left side of R5 which completes the circuit for B5. The front panel lamp will glow to indicate that the probe head is unlocked.

circuit  
description

The 1st Push-pull Amplifier, an integrated circuit, and its associated circuitry are shown in Fig. 5-23. The components that are not contained in the IC package are: R16, R17, R18, C18, R19, R20 and R21. Q1 is the current source for the input pair, Q2 and Q3, and R2 and R3 longtail the output pair, Q4 and Q5. The negative voltage supply (-16 V) is connected to the integrated circuit via R21 and the positive voltage supply (+16 V), via R20.

current  
balance

R4 and R5 emitter coupling resistors are shunted by the CURRENT/DIV BALANCE control circuit. The CURRENT/DIV BALANCE control must be able to move the trace 3 divisions above and 3 divisions below the center graticule line. Either R16 or R17 is adjusted to provide the required range for the CURRENT/DIV BALANCE control. Any signal unbalance introduced by the CURRENT/DIV BALANCE circuit is canceled in the output pair.

bandwidth

The input signal is applied to pins 2 and 10 of the integrated circuit and the output signal is taken at pins 4 and 8. Pins 3 and 9, the collectors of the input pair, are connected via R18 and C18, the Hall Bandwidth adjustment circuit. As the frequency of the applied signal is increased, the impedance of the Hall Bandwidth circuit decreases which shunts the signal from the output pair.

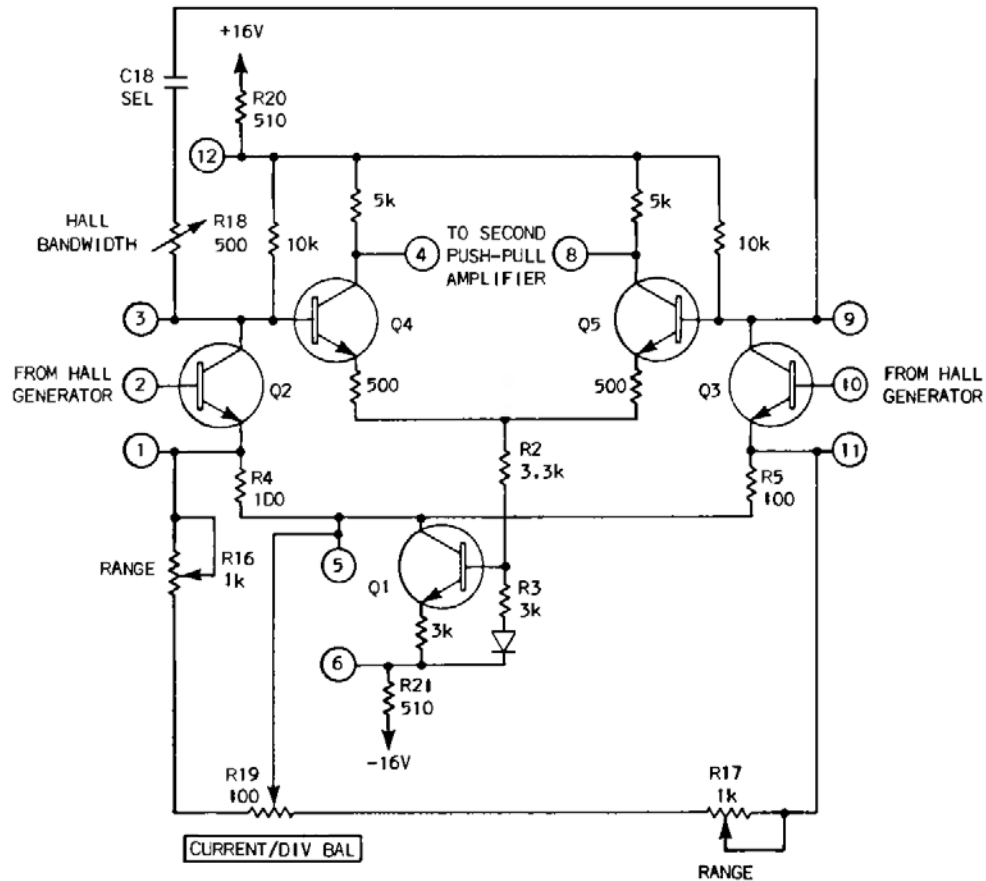


Fig. 5-23. First push-pull amplifier.

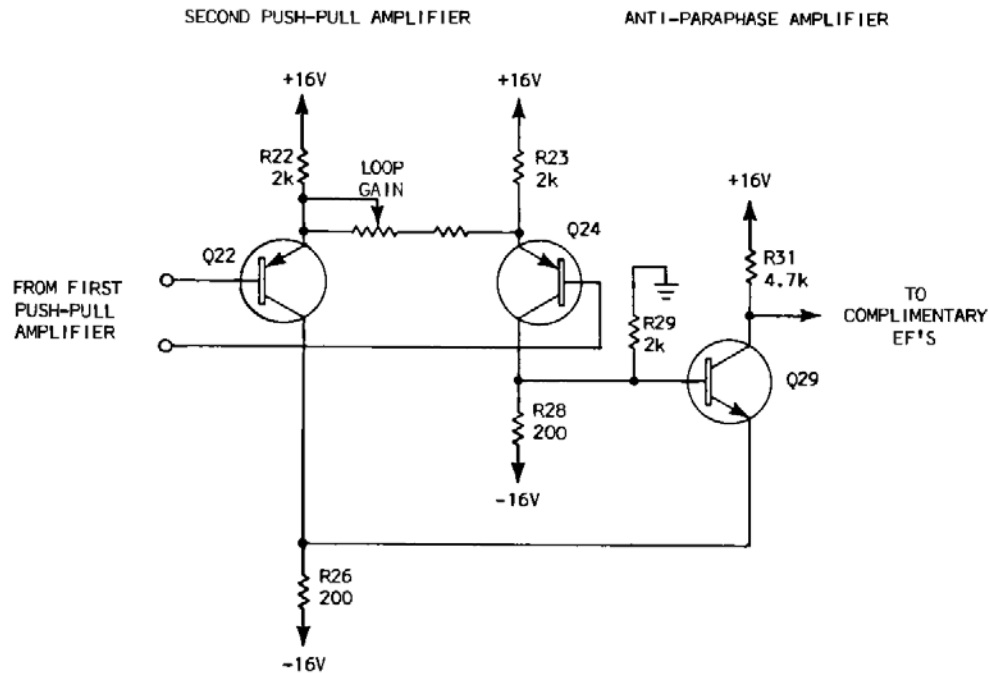


Fig. 5-24.

second  
push-pull  
circuit  
description

The 2nd Push-pull Amp and the Anti-paraphase Amp are shown in Fig. 5-24. The output of the IC drives Q22 and Q24, an emitter coupled amplifier, which in turn drives Q29, the Anti-paraphase Amp. The Loop Gain adjustment, R24, sets the emitter coupling resistance of Q22 and Q24 to control loop gain. The signals applied to the base and emitter of Q29 are out of phase. Ergo, Q29 is a summing circuit that converts the push-pull signal to a single-ended signal. The output of Q29 is applied to the Complimentary EF's.

complimentary  
EF circuit  
description

Fig. 5-25 is a schematic of the Complimentary EF's. Positive going signals are processed by Q44-Q45 and negative going signals by Q53-Q54. Q42 is the current source for Q44 and Q55 is the current source for Q53. A negative going signal from Q29 causes Q44 emitter to go negative which cuts off Q45. This signal causes Q53 emitter to go negative which increases the current through Q54. The signal is applied to the transformer secondary from Q54 emitter via LR58. The Complimentary EF pairs are in the same thermal environment to provide temperature compensation. The output of the Complimentary EF's is applied to the Wide Band Amplifier via the transformer secondary and the 50  $\Omega$  cable.



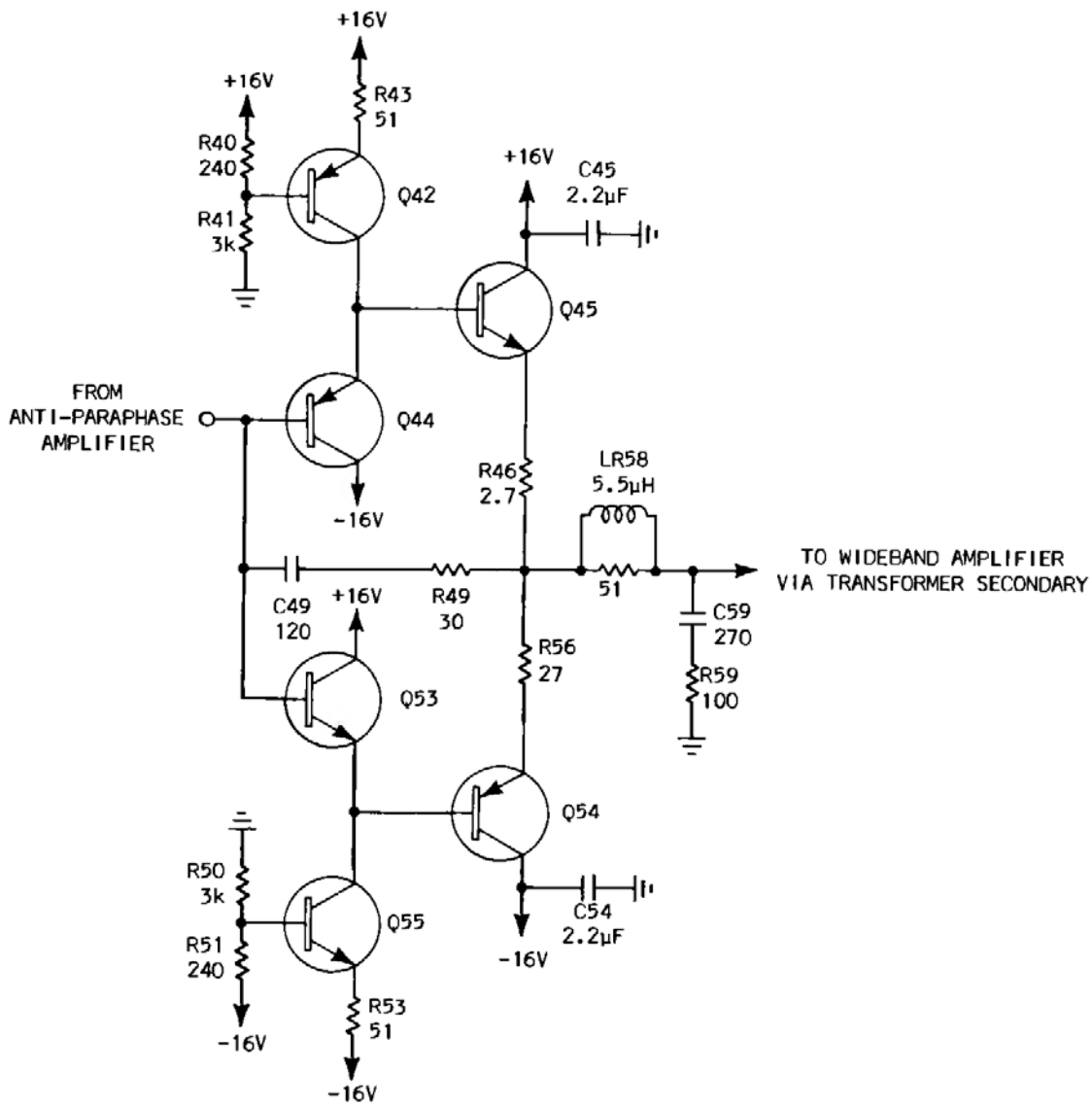


Fig. 5-25. Complimentary EF's.

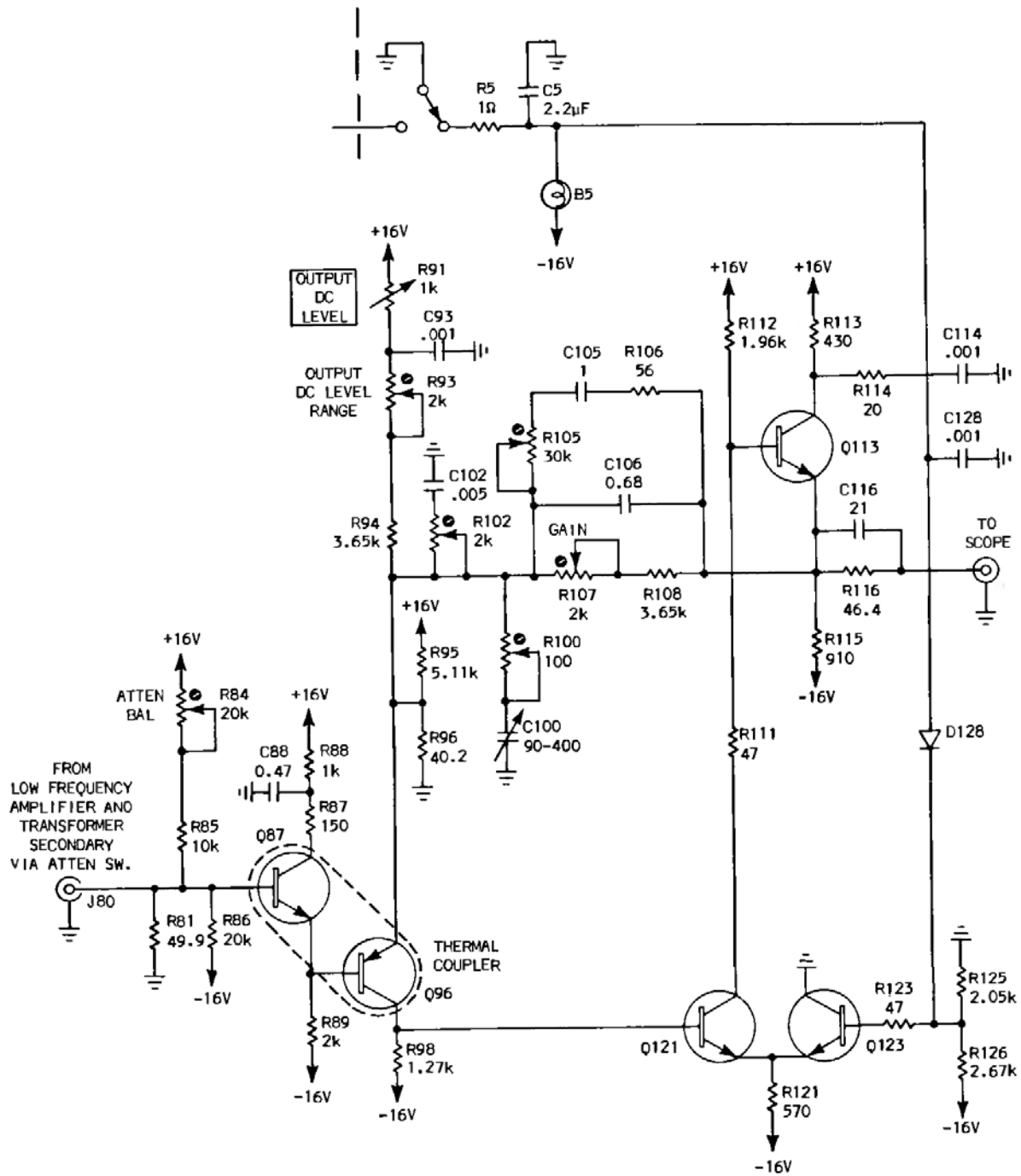


Fig. 5-26. Wideband amplifier.

A schematic of the Wide Band Amplifier is shown in Fig. 5-26. The Wide Band Amplifier is a single stage feedback amplifier. Emitter Follower Q87 isolates Q96 from the CURRENT/DIV switch. Q96 and Q121 are high gain inverters and Q113 is the Driver EF. The signal at Q113 emitter is in phase with the signal at Q96 base. Feedback is from Q113 emitter to Q96 emitter via R108 and R107. The feedback causes 100% follower action at Q96 emitter and there is essentially no change in the emitter-base voltage of Q96. R96 in conjunction with R107 and R108 determines stage gain. At high frequencies, R96 is shunted by R100-C100 and R102-C102 and R107 and R108 are shunted by C106 and series network R105, C105 and R106. The high frequency adjustments are set so the RC time constant associated with R107 and R108 is equal to the RC time constant associated with R96.

feedback

high  
frequency  
compensationDC  
level

The Atten Bal control is adjusted for 0 V at Q87 base with J80 disconnected. The OUTPUT DC LEVEL, a front panel control, provides positioning capability. The Output DC Level Range adjustment is set so that the OUTPUT DC LEVEL control can move the trace 3 divisions above and 3 divisions below the center graticule line. (Oscilloscope deflection factor is 50 mV/div.)

open  
probe --  
no  
signal

Latch Up switch Q123 is cut off when the probe head is closed. D128 anode is tied to -16 V through the probe unlocked light and is reverse biased. When the probe head is opened, the left side of R5 is grounded. D128 is forward biased and Q123 base becomes more positive than Q121 base. Thus Q123 robs all the current through R121. Q121 cuts off to prevent signal processing while the probe head is open.

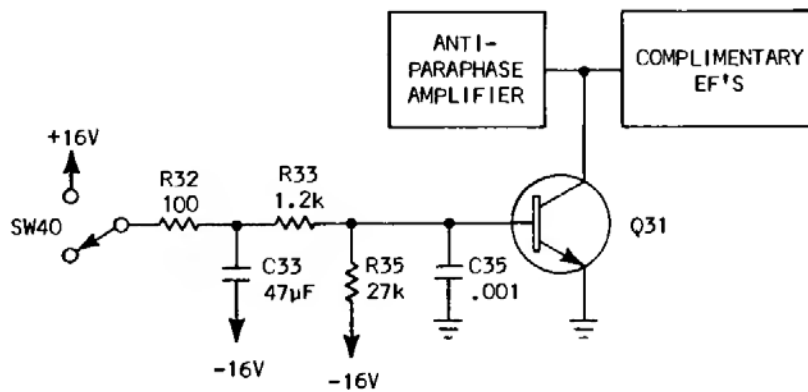
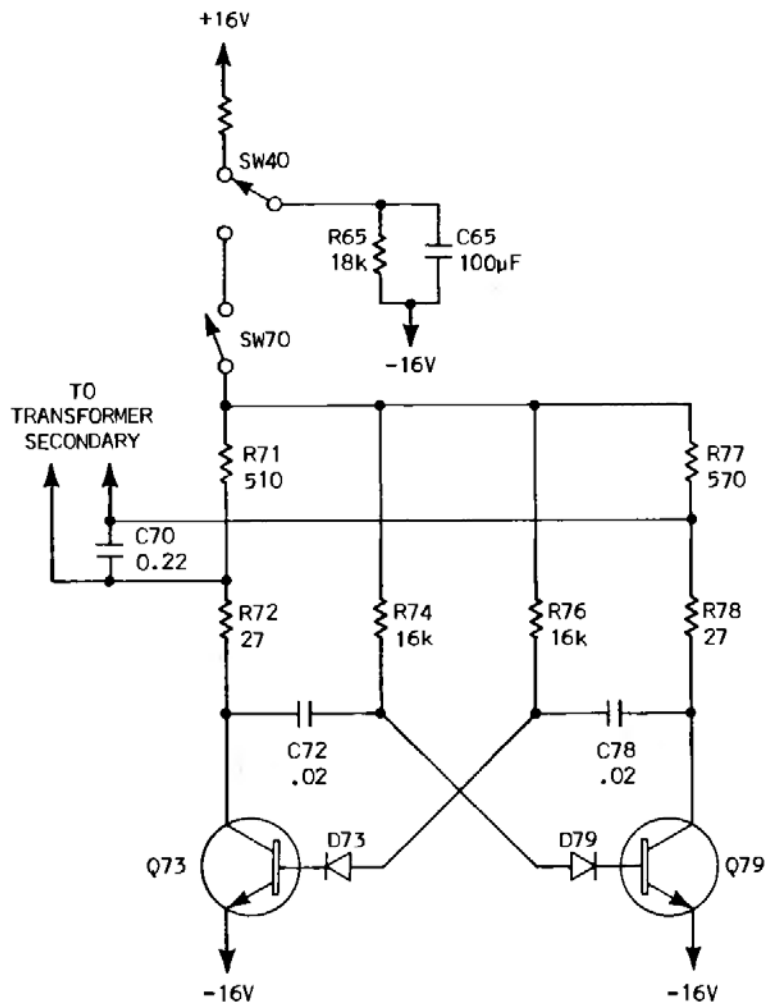


Fig. 5-27. Degaussing circuits.

degaussing

Since the Hall Generator is a flux sensing device, any residual flux in the core will be processed as an input signal. Degaussing is required to prevent erroneous display information. A front panel receptacle acts as a probe holder and a degaussing point. See Fig. 5-27. The Degauss Multivibrator, Q73 and Q79, will oscillate only when SW70 and SW40 are closed. SW70 is located within the front panel receptacle and is closed by the probe head. When SW40 is opened, C65 charges to the +16 V supply. When the DEGAUSS control, SW40, is depressed, the multivibrator will oscillate until the voltage across C65 becomes too small to sustain oscillations. C70 shunts the transformer secondary and the core is driven by a series of damped sine waves. The DEGAUSS switch forward biases Q31 which clamps Q29 collector at 0 V. (Q29 is located in the Low Frequency Amplifier.) This action breaks the feedback loop in the Low Frequency Amplifier to insure successful degaussing. When the DEGAUSS switch is released, C33 charges to -16 V through R33 and R35 which causes Q31 to cut off slowly. Q31 cannot generate a fast transient which would magnetize the core.

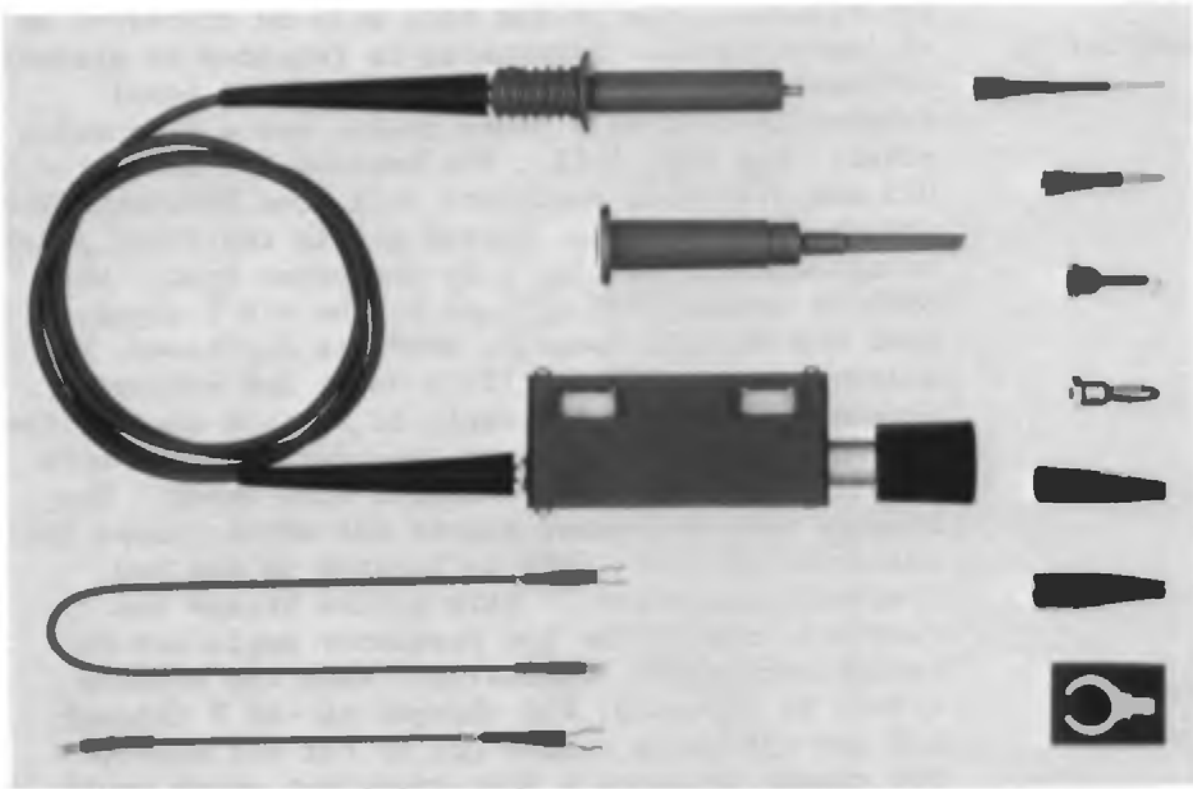


Fig. 6-1. Passive differential probe.

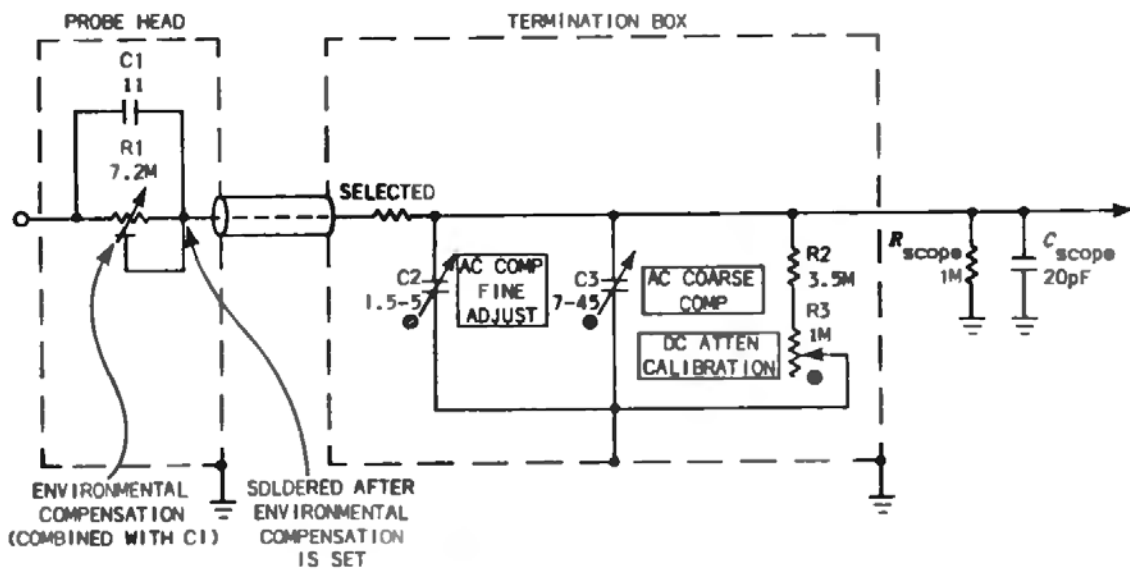


Fig. 6-2. Passive differential probe schematic.

## 6

## DIFFERENTIAL PROBES

Differential probes are designed to provide a matched coupling network between the signal source and the oscilloscope. At Tektronix, common-mode rejection is specified at the input of the differential amplifier. The use of a differential probe extends common-mode measurement capability to the probe tip with little or no loss in performance.

termination  
box

Fig. 6-1 is a photograph of a passive differential probe. The probe head assembly is typical for passive voltage probes. A resistive center conductor cable is used to couple the signal to the termination box. The termination box has three adjustments: the two thumb-wheel adjustments shown in the photograph and a screwdriver adjustment that is located on the back side of the termination box. The screwdriver adjustment is an AC Coarse Compensation and is equivalent to the compensation capacitor of a passive voltage probe. The thumb-wheel adjustments are DC Attenuation Calibration and AC Fine Compensation Adjustment which provide a means of matching one probe-oscilloscope input to the other probe-oscilloscope input. Note two probes are required for a differential measurement.

circuit  
description

The schematic is shown in Fig. 6-2. The environmental capacitance in conjunction with C1 sets the input capacitance at 12 pF. Probe input resistance is 8 M $\Omega$  with a 1 M $\Omega$  input resistance oscilloscope. The signal is coupled to the termination via a resistive cable. The selected resistor terminates the cable at high frequencies. R2 and R3 shunt the oscilloscope input resistance which lowers the nominal resistance at the oscilloscope input to 800 k $\Omega$ . The resistive shunt allows compensation of the 1% tolerance in the 1 M $\Omega$  input resistors. Attenuation ratio can vary from 9.75:1 to 10.25:1 with adjustment of R3 but the input resistance and attenuation ratio can be made equal at the probe tips.

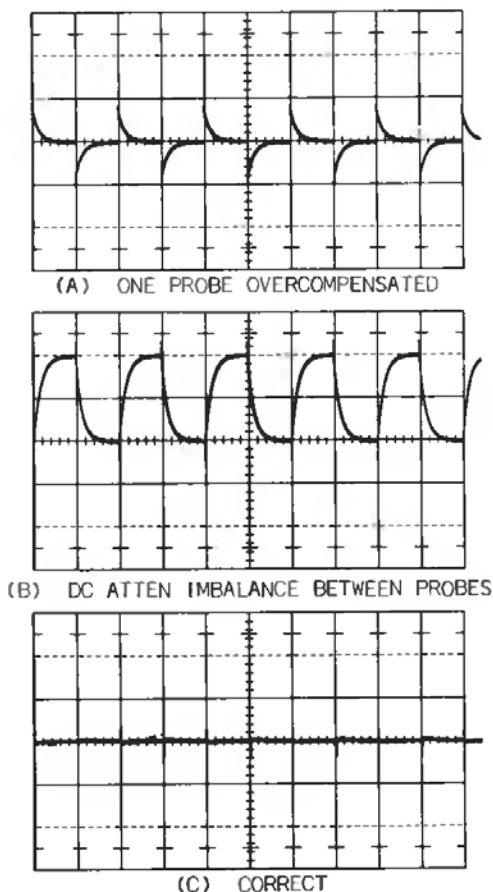


Fig. 6-3. Probe compensation.

compensation

The AC Coarse Comp and the AC Comp Fine Adjust for one probe are calibrated for proper response to the 1 kHz calibrator square wave similar to the compensation of a passive voltage probe. The DC Atten Calibration and the compensating capacitors of the other probe are adjusted differentially for best common-mode rejection as shown in Fig. 6-3.

use  
below 5 MHz

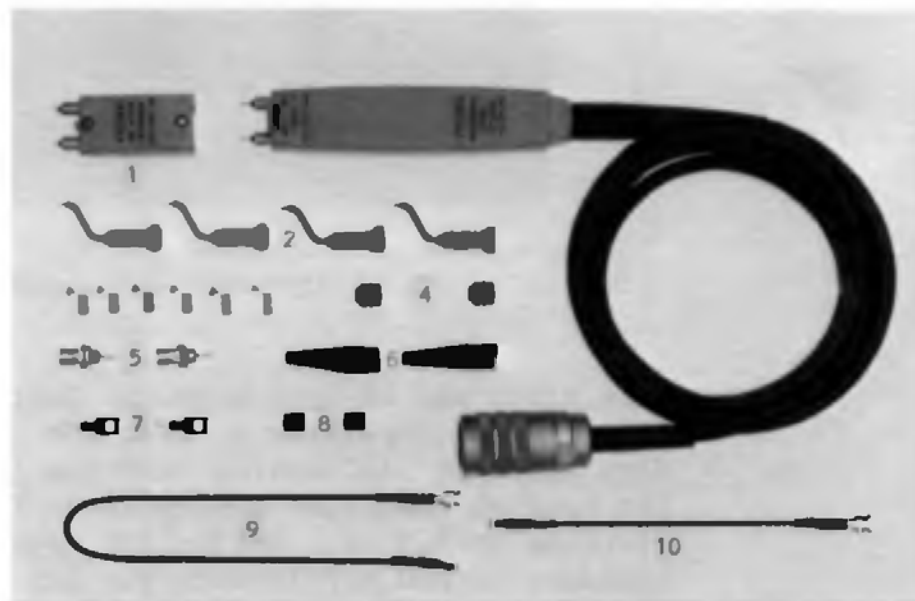
The passive differential probe is very useful for low-frequency measurements. Its common-mode performance decreases at high frequencies because the input time constants are not matched. Neither  $R_1$  nor its shunt capacity is adjustable after the probe leaves the factory. Above 5 MHz, a small difference between probes can noticeably reduce common-mode rejection.

We can build a differential amplifier with well matched input circuitry in the probe head. Thus we will have an active differential probe with minimal



differential  
head

distance between the test point and the amplifier. A photograph of an active differential probe is shown in Fig. 6-4. The probe is shown with its standard accessories which are identified in the list accompanying the photograph. Note that both input connectors are located on the probe head to minimize propagation time between the test point and the input circuit. This probe has a 5 V common-mode performance range and a 25 V maximum input rating. A common ground between the probe and the test chassis is imperative to reduce the possibility of destroying input circuit components.



- |                         |                        |
|-------------------------|------------------------|
| 1. DUAL ATTENUATOR HEAD | 6. ALLIGATOR CLIPS     |
| 2. SWIVEL TIPS          | 7. HOOK TIPS           |
| 3. SPRING GROUND CLIPS  | 8. INSULATION TUBES    |
| 4. SPECIAL GROUND TIPS  | 9. 12-INCH GROUND LEAD |
| 5. TEST JACKS           | 10. 5-INCH GROUND LEAD |

Fig. 6-4. Active differential probe.



Fig. 6-5. Differential probe amplifier.

This probe may be used with the Tektronix Type 1A5 Differential Amplifier or with its own amplifier. With the Type 1A5, system risetime is 7 ns and with the special amplifier system shown in Fig. 6-5, risetime is 3.5 ns. Deflection factors are from 1 mV/div to 200 mV/div with the 1A5 or with the amplifier driving an oscilloscope with 10 mV/div deflection factor.

Fig. 6-6 is a block diagram of the probe head circuits. The terms "+ input" and "- input" indicate that a positive going signal at the + input will be deflected upward on the CRT and a positive going input at the - input will be deflected downward on the CRT. The Input Amplifiers are source coupled and the active device in the longtail, Q224, enhances common-mode rejection. The Bootstrap Driver provides a signal to minimize changes in the gate to drain voltage of the Input FET's. The Output Amplifiers, Q154-Q164 and Q254-Q264, are cascode amplifiers. Gain of the probe head is unity for single-ended signals.

block  
diagram

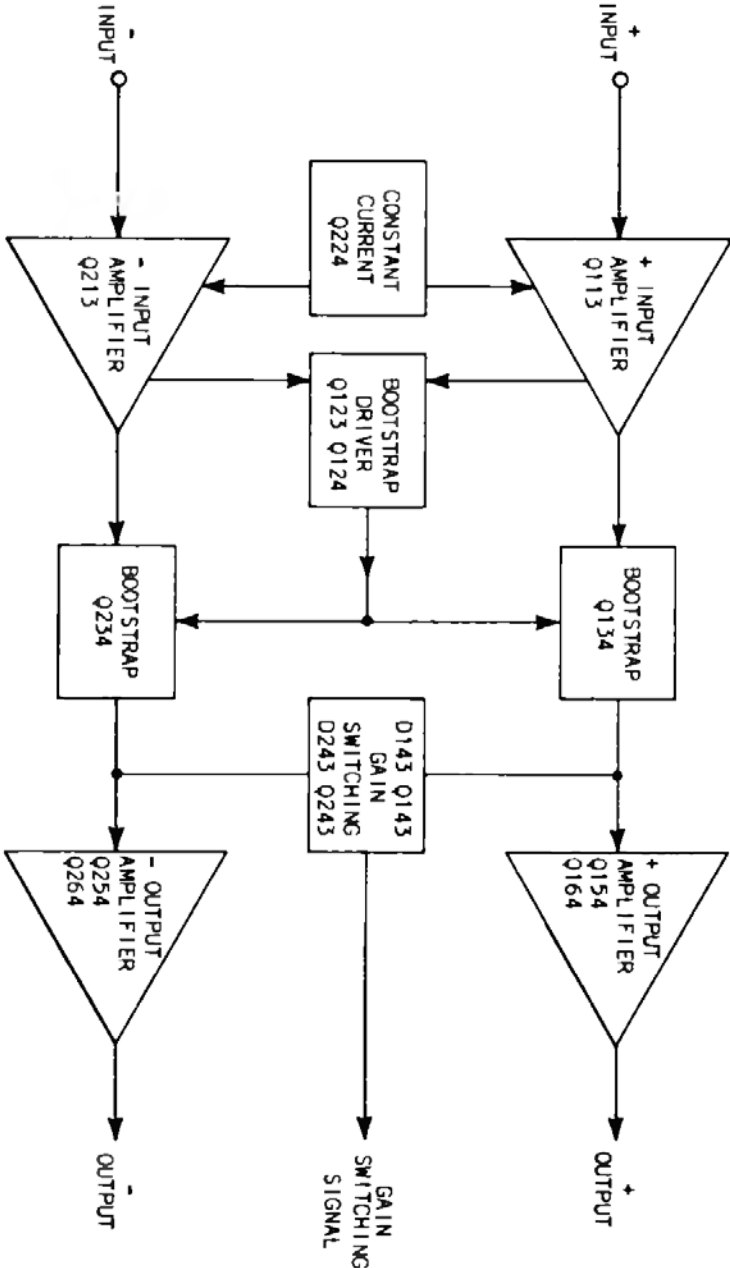


Fig. 6-6. Active differential probe.

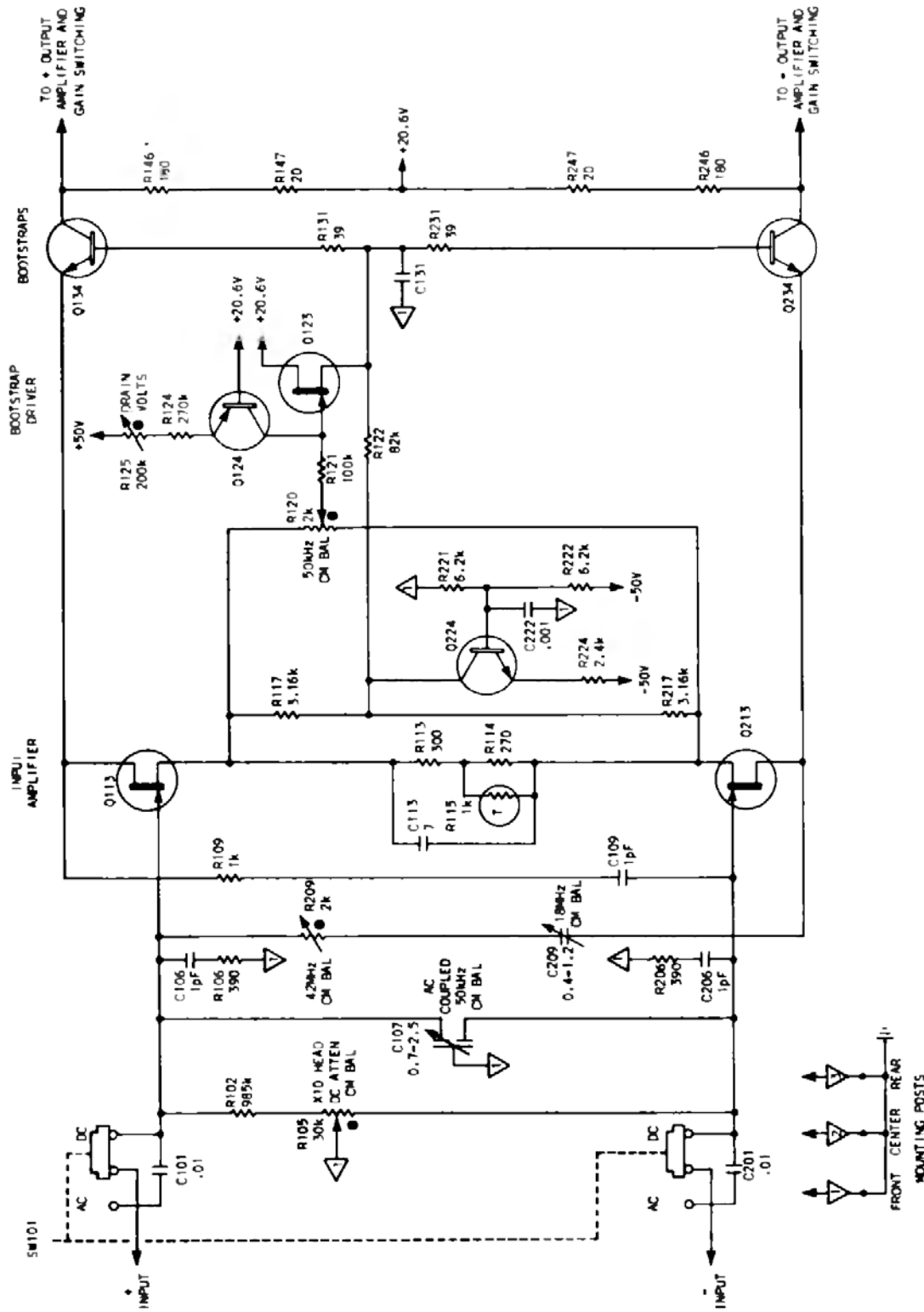
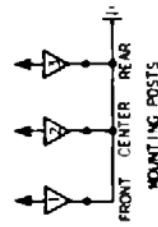


Fig. 6-7. Active differential probe schematic (partial).



A schematic of the Input Amplifiers, Bootstraps, Current Source and Bootstrap Drivers is shown in Fig. 6-7. Current Source, Q224, supplies 10 mA which is divided between Q113 and Q213 at quiescence. Source coupling resistors R113, R114 and R115 are gain determining components. Thermistor R115 also provides temperature compensation for the Input Amplifiers. C113 is a source-peaking capacitor. Q113 and Q213 current drive Q134 and Q234.

single-ended gain For single-ended signals the Input Amplifiers and the Bootstraps form a cascode amplifier with the output voltages developed across R146-R147 and R246-R247. Single-ended signal gain from the gate of Q113 to the collector of Q134 is about 0.4 (R146 plus R147 divided by the source coupling resistance, R113, R114, R115 and the FET source resistance).

common-mode gain Common-mode gain is a different matter. Assume the voltage at Q113 and Q123 sources becomes 1 V more positive. The voltage across Q224 will increase 1 V. The resulting change in FET current is 1 V divided by the resistance of the current source. Q224 is a high resistance source and the common-mode signal current is small. This small current flows through the 200- $\Omega$  load at the collector of Q134 and Q234 to produce a nearly negligible common signal. The change in voltage at Q113 and Q213 sources is seen at the center tap of R120, the 50 kHz CM Bal. A portion of this signal is coupled to the gate of Q123 via R121. Q123 source follows the change in voltage at Q113 and Q213 sources. Q123 output is applied to the bases of Q134 and Q234. The emitters of Q134 and Q234 follow the change in base voltage to move the drain voltage of Q113 and Q213 in the same direction as the input signal. This action keeps the gate to drain voltage constant.

balancing input Drain Volts adjust, R125, in Q124 circuit is adjusted so that the drains of Q113 and Q213 are at 10.5 V regardless of the self bias of the Input Amplifier FET's. R109 and C109 with C209, 18 MHz CM Bal, and R209, 42 MHz CM Bal, provide a method of balancing the gate to drain capacitance of the Input FET's. C107, AC Coupled 50 kHz CM Bal, cancels any difference in the AC Coupling capacitors, C101 and C201. R105, 10X head DC Atten CM Bal, cancels any difference in the input resistance when the 10X Attenuator is used.

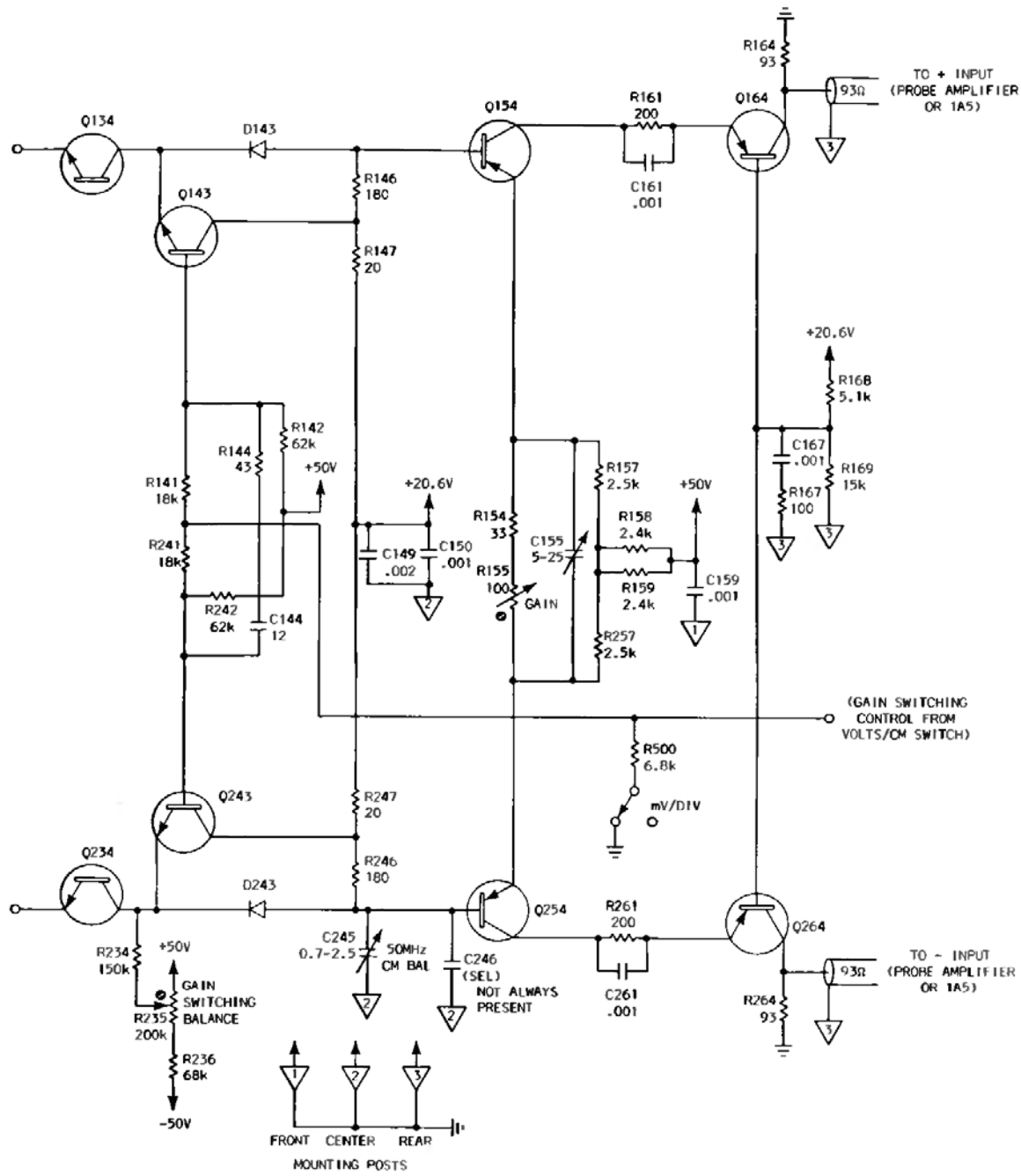


Fig. 6-8. Active differential probe schematic (partial).

caution There is no protection circuitry for the inputs of Q113 and Q213. The protection circuits were omitted because additional circuitry tends to reduce common-mode performance. It is impossible to perfectly match components. Thus, protection circuitry would require more adjustments and we would have less common-mode rejection. The probe input is sensitive to large signals and care should be taken to avoid damaging the input FET's.

chassis grounds We note all grounds are connected to mounting posts for the probe head covers. The inside of the covers are gold plated to provide chassis ground and shielding. When the covers are removed, chassis ground for the probe head is also removed! Special covers are available which allow access to the screwdriver adjustments for probe calibration. Consult the instrument instruction manual for proper maintenance techniques.

gain switching The Output Amplifier and Gain Switching circuitry is shown in Fig. 6-8. Gain switches, Q143 and Q243, are normally cutoff. Thus the signal from the Bootstrap is developed across R146-R147 and R246-R247. When the mVOLTS/DIV control of the 1A5 or probe amplifier is set at 50, 100, or 200 mV/div, the ground is removed from R500. The bases of Q143 and Q243 go from 17 V to 20.5 V. The transistors are driven into saturation and the voltage across the silicon diodes D143-D243 falls below 0.5 V. The diodes are reverse biased and Bootstrap output current is routed through R147 and R247. The collector load is decreased by a factor of ten which reduces single-ended and common-mode signal gain by a factor of ten. Gain Switching Balance, R235, is adjusted for no-trace shift as the mVOLTS/DIV control is rotated throughout its range. C245, 50 MHz CM Bal, compensates for differences in circuit capacitance when the gain switching diodes are reverse biased. Selected capacitor C246 insures that C245 has sufficient adjustment range.

output  
stage

Q154-Q164 and Q254-Q264 are the cascode output stage. The collector load is R164 and the 93- $\Omega$  cable for Q164 and R264 and the 93- $\Omega$  cable for Q264. Gain adjust, R155, sets the collector-emitter resistance ratio to unity gain from input to output. A 1 V single-ended input signal will produce opposite phase 1/2 V signals at the collectors of Q164 and Q264. Any common-mode signal applied to the bases of Q154 and Q254 is canceled by the low common-mode transfer characteristic of the Output Amplifier. Common-mode gain is the collector load divided by R157 plus R158 for Q154-Q164 and R257 plus R258 for Q254-Q264. Thus common-mode gain is less than 1/50.

C155 provides emitter peaking. R161 and R261 temperature compensates Q154 and Q254 respectively. C161 and C261 retain transient response. The output of the probe is applied to the 1A5 plug-in or the probe amplifier.

probe  
amplifier

A block diagram of the probe amplifier is shown in Fig. 6-9. The input stage Emitter Followers, Q413 and Q513, isolate the probe circuitry from the 1st Amplifier. The gain of the 1st Amplifier is switched by the mVOLTS/DIV control from 5 mV/div to 200 mV/div. The output of the 1st Amplifier is applied to the 2nd Amplifier via Emitter Followers, Q443 and Q543. The gain of the 2nd Amplifier is switched in the 1, 2 and 5 mV/div positions. The output signal is taken from the + side of the 2nd Amplifier and applied to the oscilloscope via Output Emitter Followers, Q463 and Q473. Not indicated in the block diagram is the gain-switching signal from the mVOLT/DIV control to the probe. Probe gain is reduced to 1/10 in the 50, 100 and 200 mV/div positions.



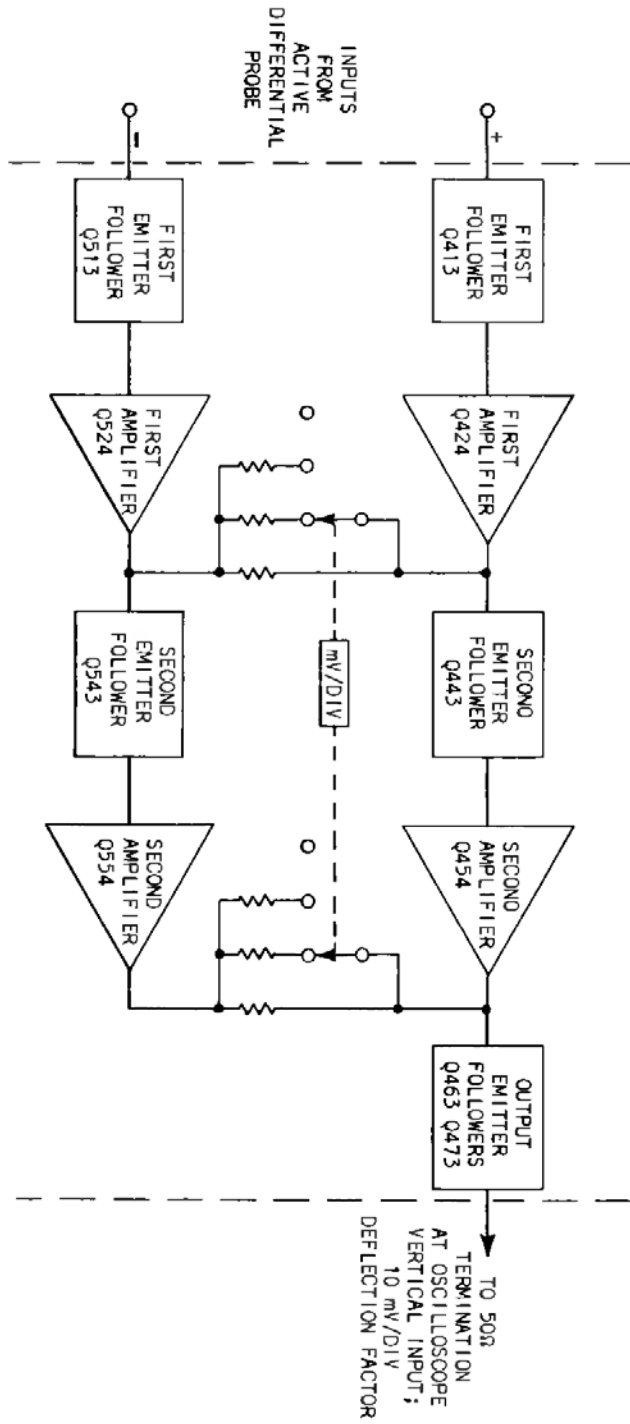


Fig. 6-9. Probe amplifier.

+ side

A schematic of the Input Emitter Followers and the 1st Amplifier is shown in Fig. 6-10. We will discuss only the + side circuitry because the circuitry on both sides of the amplifier is the same. The probe signal is applied to the base of Q413. D5 clamps Q413 base when the probe is disconnected. The clamping action prevents the base-to-collector voltage from exceeding breakdown potential. The Atten Bal, R2, is adjusted for no-trace shift as the mVOLT/DIV control is rotated. The adjustment is made in conjunction with Gain Switching Balance in the probe. Q413 output is applied to the base of Q424.

gain

The gain determining resistance in the emitter circuit is R22 in series with Q424 emitter resistance, a total of about 30  $\Omega$ . The collector resistance is R24 in parallel with 1/2 of R29. R29 is driven push-pull and its center point is at signal ground. C20-R20 provide emitter peaking and in the 1 and 2 mV/div positions, adjustable capacitor C21 provides additional peaking. Q424 output drives Q443, the 2nd EF.

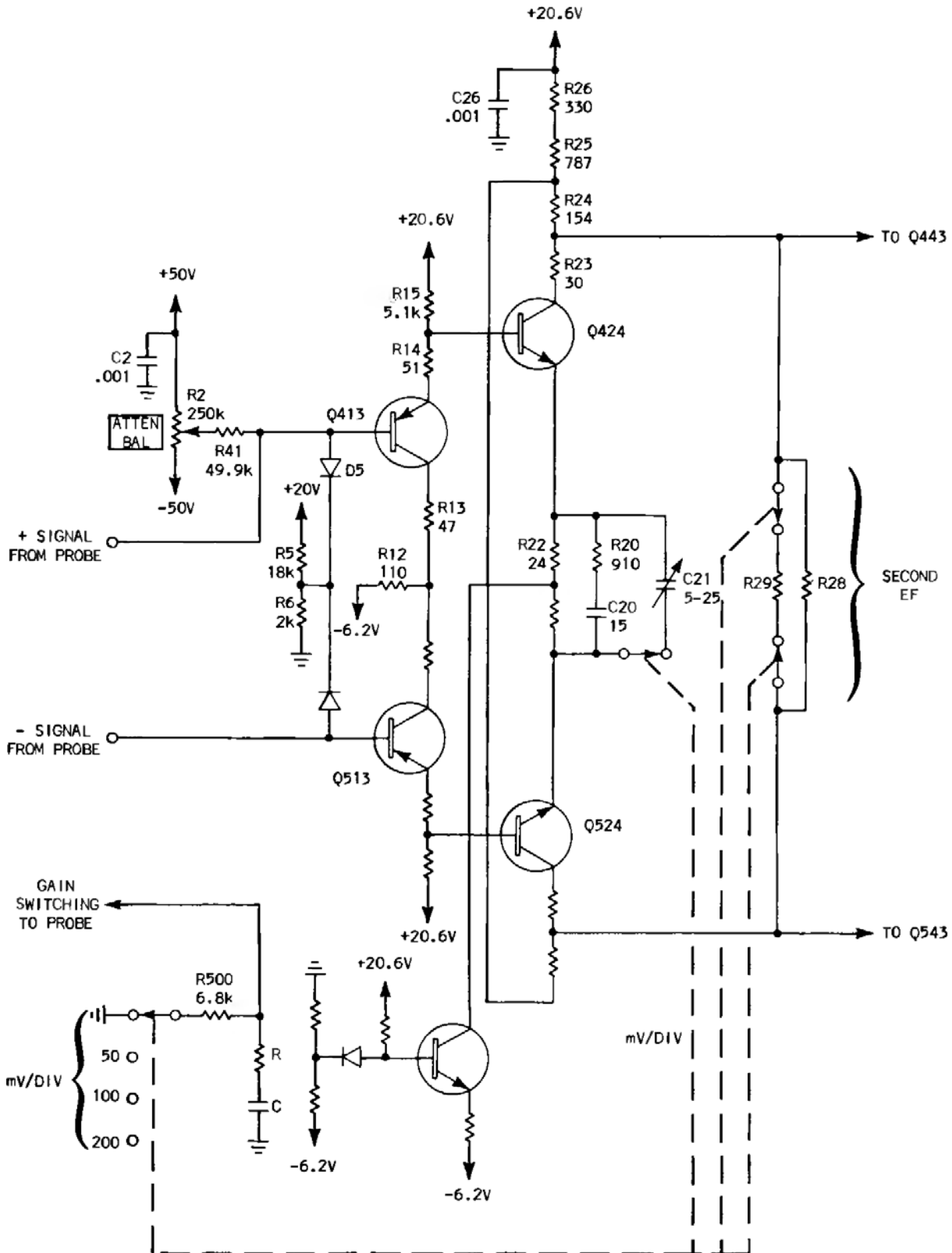


Fig. 6-10. Probe amplifier schematic (partial).

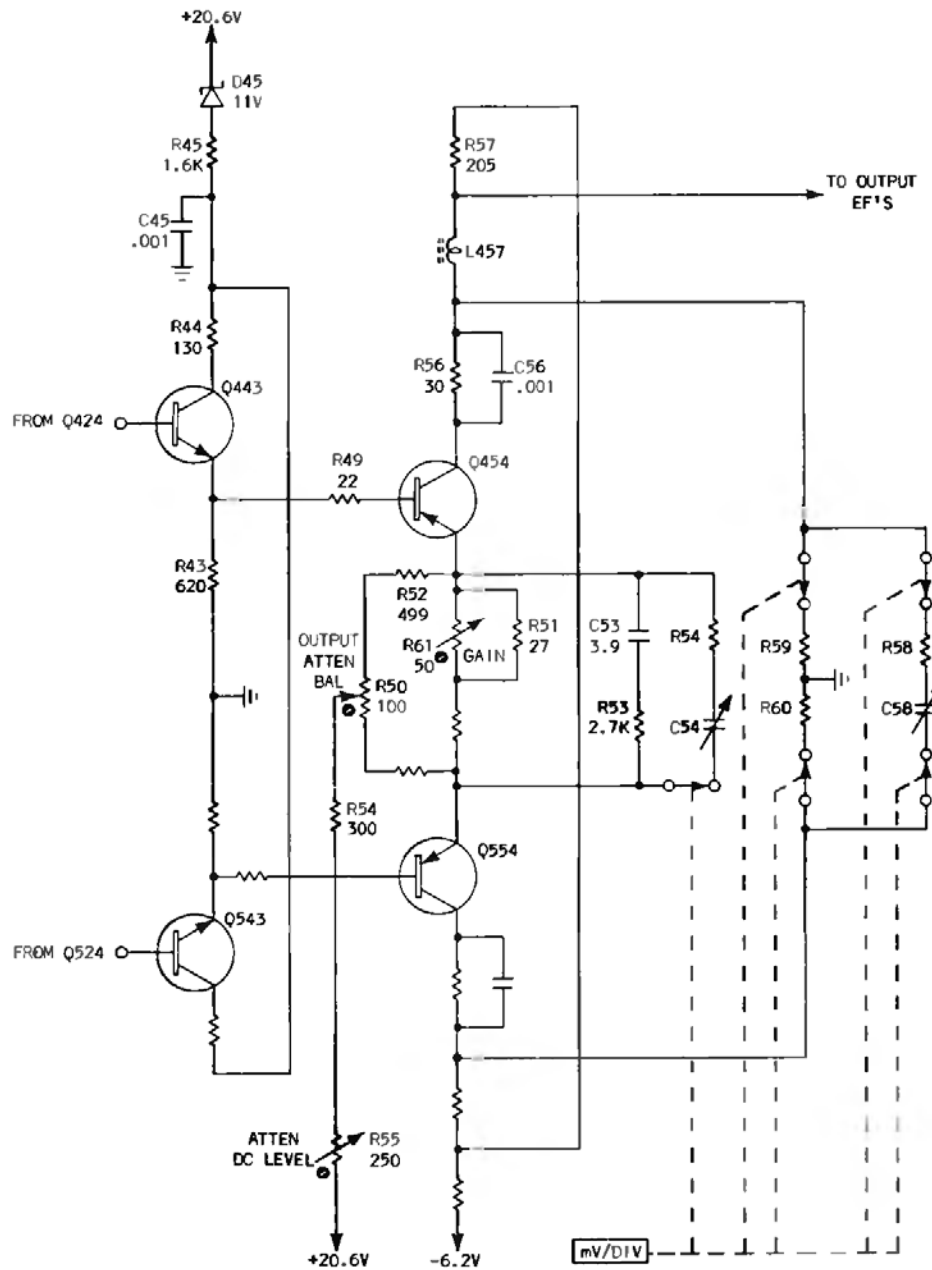


Fig. 6-11. Probe amplifier schematic (partial).

2nd EF's  
and  
amplifier

A schematic of the 2nd Emitter Followers and the 2nd Amplifier is shown in Fig. 6-11. Q443 couples Q424 output to the base of Q454. Gain adjust, R61, is adjusted to calibrate overall amplifier gain. The collector load is R57 in parallel with R59. R58-C58 provide collector high-frequency compensation in the 10 mV/div position. R53-C53, along with switched network R54-C54, provide emitter

peaking at all deflection factors. R56 is for temperature compensation and C56 retains transient response.

calibration  
 After the Output DC Level in the Output EF circuitry is adjusted for 0-V output at quiescence, the Atten DC Level, R55, and the Output Atten Bal, R50, are adjusted. The output terminal of Q454, the junction of R56 and R57, is connected to the output terminal of Q554 via a 10- $\Omega$  test resistor. The Atten DC Level, R55, is then adjusted to set the trace at center screen. The 10- $\Omega$  test resistor is removed from the output terminals and is used to connect the bases of Q454 and Q554. The Output Atten Bal, R50, is adjusted to set the trace at center screen.

output  
 EF circuitry  
 The Output EF circuitry is shown in Fig. 6-12. Q463 isolates Q473 from the gain switching. Q473 emitter resistance in series with R79 provides a 50- $\Omega$  source impedance to drive a terminated 50- $\Omega$  cable. The Output DC Level is adjusted for 0 V with Q463 base tied to ground via a 10- $\Omega$  test resistor. R73 in Q473 collector provides temperature compensation to the output stage.

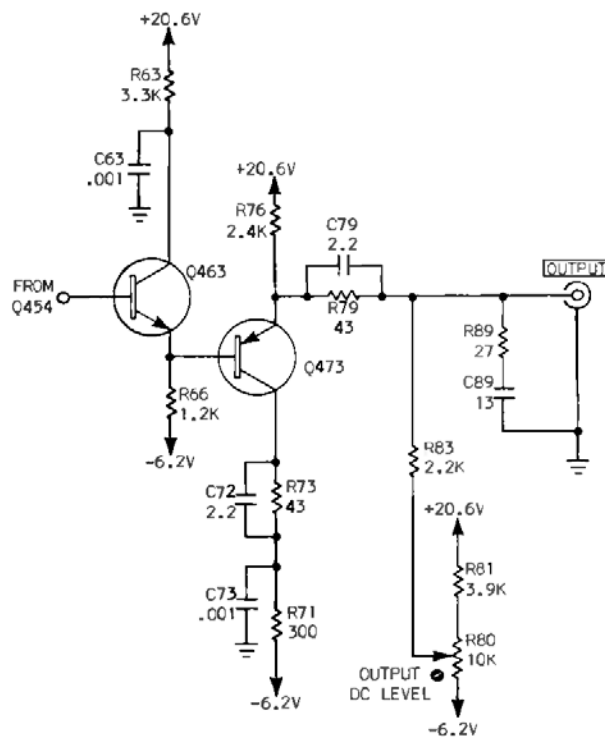
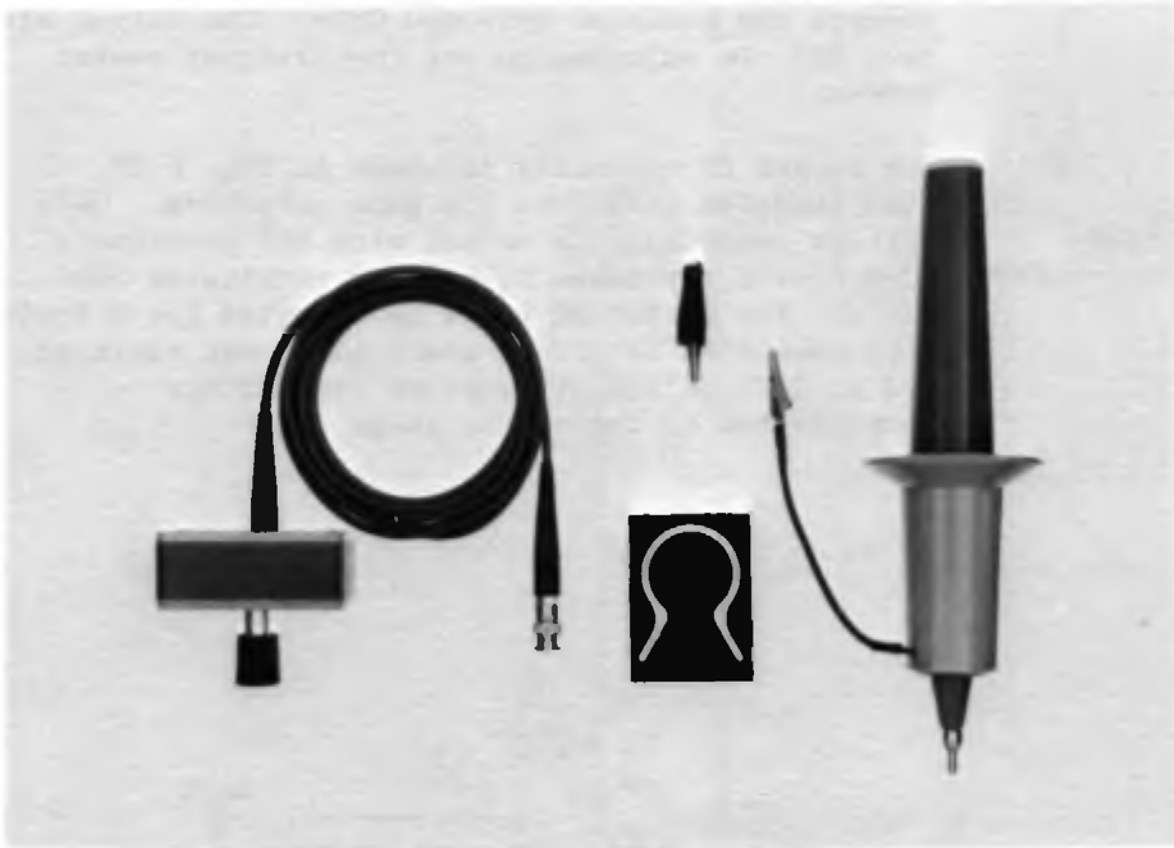


Fig. 6-12. Probe amplifier output.



**Fig. 7-1. High voltage probe.**

## 7

## HIGH VOLTAGE PROBES

general  
description

High voltage probes are designed to couple kilovolt signals to conventional oscilloscopes. The probe head is made of high impact strength thermoplastic material that provides mechanical protection for the internal components and electrical protection for the user. A photograph of a high voltage probe is shown in Fig. 7-1. The flaring on the probe head assembly is designed to isolate sweaty hands from the high voltage. Please connect the ground lead before monitoring high voltage test points. The resistive center conductor coaxial cable is detachable to facilitate storage. The output connector on the termination box is a locking BNC type.

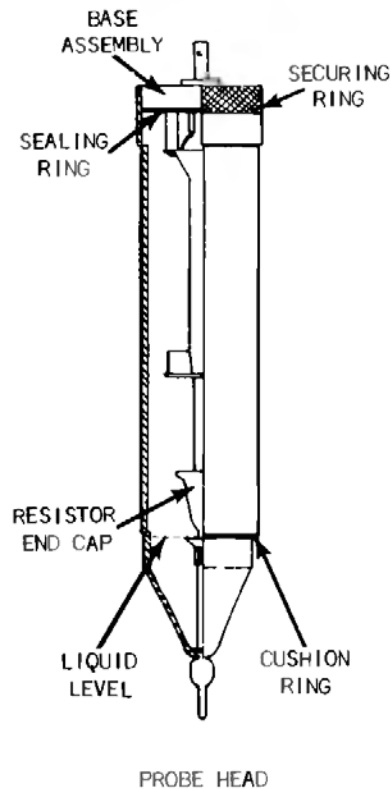


Fig. 7-2. Cutaway drawing of 40 kV probe.

fluorocarbon  
114

This probe has a maximum voltage rating of 12 kV. A second probe with an input voltage rating of 40 kV is also available. The probes have similar appearance except the 40-kV probe has a clear plastic tip. A cutaway drawing is shown in Fig. 7-2. The high voltage properties of this probe are enhanced through the use of a dielectric fluid, fluorocarbon 114. This fluid has a saturation pressure of approximately 2 atmospheres at 25°C. At room temperature a small amount of liquid should be visible at the clear probe tip. If the liquid is not seen at the level indicated in the drawing, the probe must be recharged. Refer to the instrument instruction manual for the proper procedure for dismantling and recharging the probe.

input  
impedance

A schematic is shown in Fig. 7-3. The input impedance is 100 MΩ shunted by 3 pF with a 10-ft cable. The input capacitance rises to 3.5 pF with a 25-ft cable. The 100-MΩ resistor is four inches long which makes the input RC time constant complex. The distributed nature of the input capacitance is indicated in the schematic.



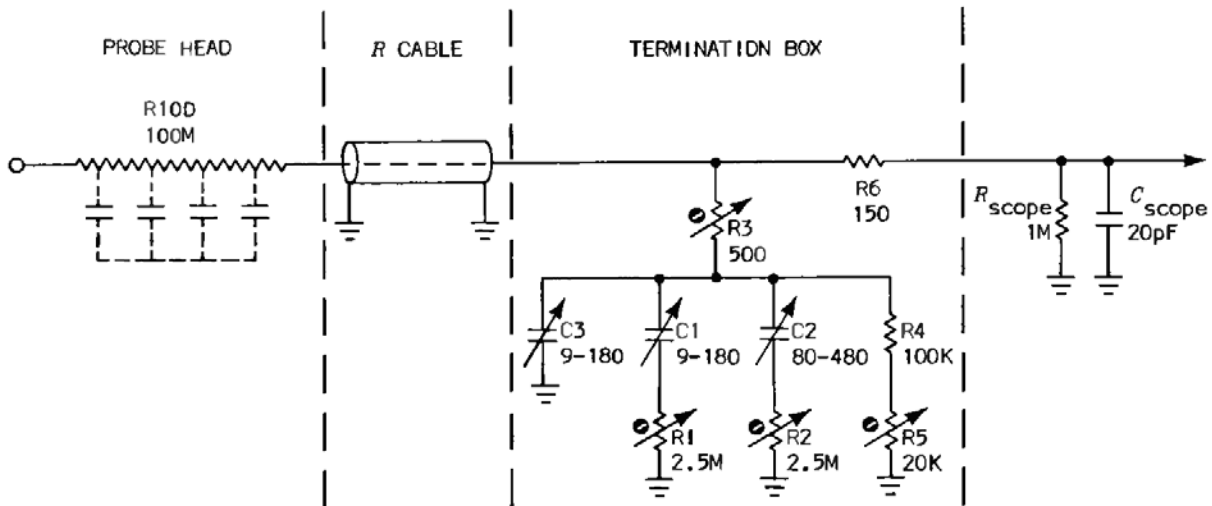


Fig. 7-3. High voltage probe.

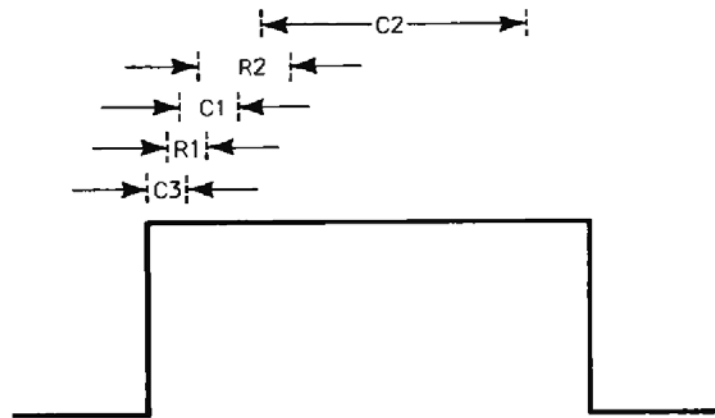


Fig. 7-4. Pulse area affected by each compensation adjustment.

compensation

The signal is coupled to the termination box via a resistive center conductor cable. R5 is adjusted for 1000X attenuation with a 1-M $\Omega$  input-resistance oscilloscope. R1, C1, R2, C2 and C3 are adjusted to compensate the probe to the input time constant of the oscilloscope. Probe compensation range exceeds 50 pF at the oscilloscope input. Fig. 7-4 shows the area of the calibrator signal affected by each compensation adjustment. R5 should be adjusted for proper attenuation before the compensation is adjusted. R6 and R3 terminate the probe cable at high frequencies. R3 is adjusted for no overshoot on the front corner with the adjustment set as far counterclockwise as possible. Risettime of the 40-kV probe is 4 ns and risetime of the 12-kV probe is 7 ns.

temperature  
and  
frequency  
derating

The high voltage capability of the probe is affected by environmental temperature and the frequency of the input signal. A derating curve for the 12-kV probe is shown in Fig. 7-5. Note at 50°C, the input voltage rating falls to 8 kV. From DC to 100 kHz the 100-M $\Omega$  input resistor defines the input limits. Above 100 kHz, the shunting effect of the input capacitance causes a noticeable increase in cable current.

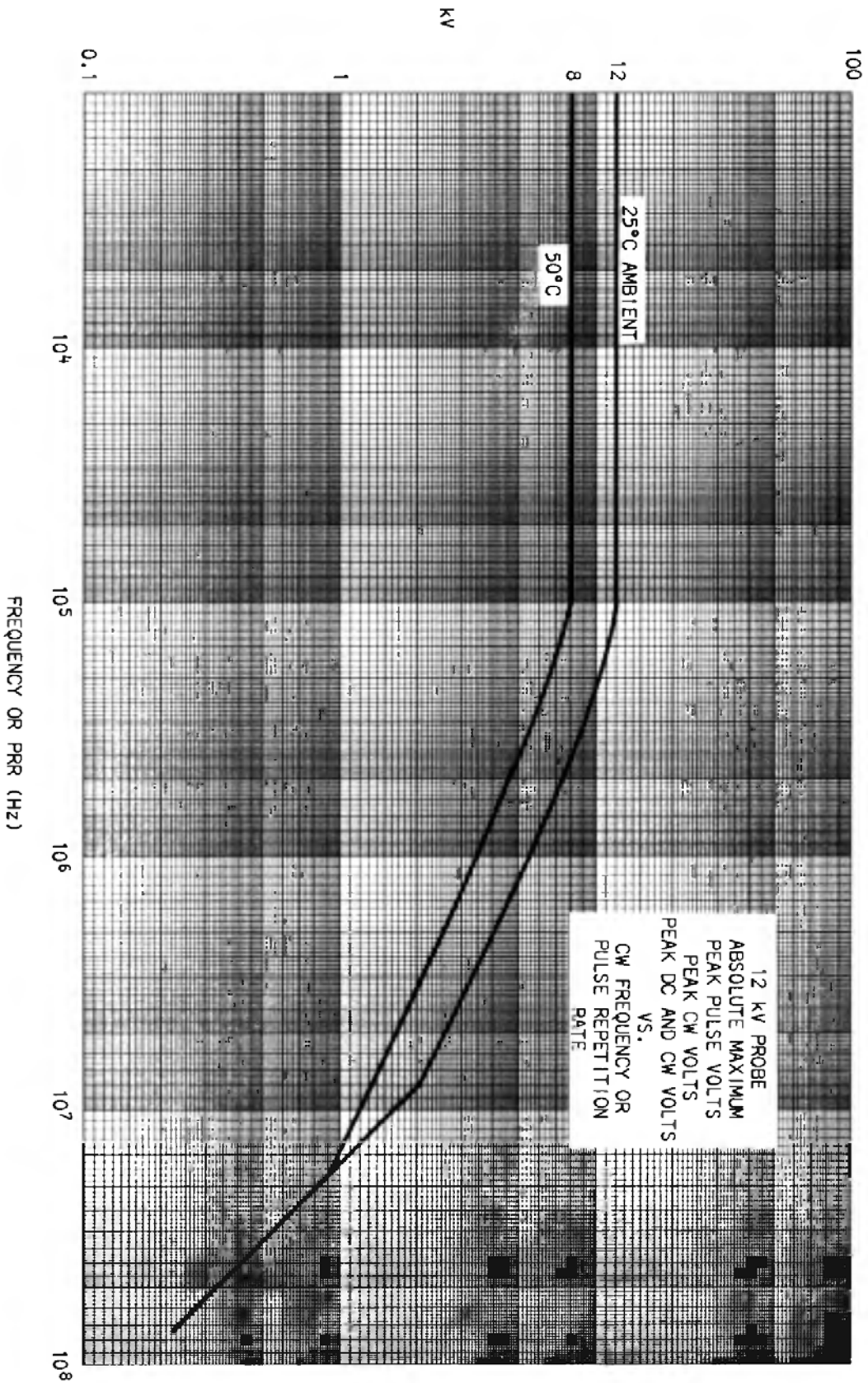


Fig. 7-5. 12 kV probe derating curves.

## NOTES

## INDEX

- Adapter, probe, 50-53
- Amplifier, probe, 74-76, 79-87, 94, 100-105
- Cables,
  - lossless ( $Z_0$ ), 11-14
  - $R$ , 15-18
- Capacitor, DC blocking, 33
- Compensation,
  - input, 11-14, 18-25, 29-31, 92, 110
  - thermal, 47-48, 81-82, 97, 105
- Core,
  - closed, 55-63
  - split, 55, 63-89
- Degaussing, 88-89
- Derating,
  - frequency, 26-27, 32-33, 42-43, 62, 110-111
  - temperature, 110-111
- FET probe, 43-49
- Flux, 61-63, 76-78
- Hall effect, 77-78
- Hall Generator, 77-89
- Hysteresis loop, 57
- Impedance, insertion, 56, 60, 64, 66-67
- Inductance, leakage, 62-63
- Offset voltage, 46
- Permeability, 55-60
- Probes,
  - current, 55-89
  - determining response of, 7
  - differential, 91-105
  - general classes, 4-5
  - high voltage, 106-111
- Probes (continued),
  - voltage,
    - active, 38-53
    - passive, 9-27
    - $Z_0$ , 28-37
- Termination,
  - active, 48-49, 71-76
  - functions, 6
  - passive, 65-70, 91
  - $Z_0$  probes, 36-37
  - 50-turn, 67-68
  - 125-turn, 68-70

## NOTES

## BOOKS IN THIS SERIES:

### CIRCUIT CONCEPTS

<i>title</i>	<i>part number</i>
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
Spectrum 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

# **K4XL's** **BAMA**

This manual is provided **FREE OF CHARGE** from the "BoatAnchor Manual Archive" as a service to the Boatanchor community.

It was uploaded by someone who wanted to help you repair and maintain your equipment.

If you paid anyone other than BAMA for this manual, you paid someone who is making a profit from the free labor of others without asking their permission.

You may pass on copies of this manual to anyone who needs it. But do it without charge.

Thousands of files are available without charge from BAMA. Visit us at <http://bama.sbc.edu>