Service Scope

USEFUL INFORMATION FOR USERS OF TEKTRONIX INSTRUMENTS

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A PRACTICAL APPROACH TO TRANSISTOR AND VACUUM TUBE AMPLIFIERS

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Tektronix, Inc.

Electronic Instrumentation Group

Display Devices Development

Two articles published in past issues of Service Scope contained information that, in our experience, is of particular benefit in analyzing circuits. The first article was "Simplifying Transistor Linear-Amplifier Analysis" (issue #29, December, 1964). It describes a method for doing an adequate circuit analysis for trouble-shooting or evaluation purposes on transistor circuits. It employs the "Transresistance" concept rather than the complicated characteristic-family parameters. The second article was "Understanding and Using Thevenin's Theorem" (issue #40, October, 1966). It offers a step-by-step explanation on how to apply the principles of Thevenin's Theorem to analyze and understand how a circuit operates.

Now, in this issue of Service Scope, we present the first of three articles that will offer a practical approach to transistor and vacuum-tube amplifiers based on a simple DC analysis. These articles will, by virtue of additional information and the tying together of some loose ends, combine and bring into better focus the concepts of "transresistance" and the principles of Thevenin's Theorem. We suggest that a "refresher" reading of the two previous articles will enable our readers to more readily follow the information in this and the two following issues of Service Scope.

The Editor

Part I

THE TRANSISTOR AMPLIFIER

INTRODUCTION

Tubes and transistors are often used together to achieve a particular result. Vacuum tubes still serve an important role in electronics and will do so for many years to come despite a determined move towards solid state circuits.

Whether a circuit is designed around vacuum tubes or transistors or both, it is important to recognize the fact that the two are in many ways complementary. It is wrong to divorce vacuum tubes and transistors as separate identities each peculiar to their own mode of operation. Indeed, as this series of articles will show there is an analogy between the two. It is true of course, that the two are entirely different in concept; but, so often we come across a situation where one can be explained in terms of the other that it is very desirable to recognize this fact.

Transistor and vacuum tube data give us very little help in the practical sense. Parameter Curves and electrical data show the behavior of these devices under very defined conditions. In short, they are more useful to the designer than the technician. We are often reduced to explaining most circuits in terms of an ohms law approach; so, it seems pointless not to pursue this approach to its logical conclusion.

In this first article we will look at a transistor amplifier as a simple DC model; and then, in the second article, look at a vacuum-tube amplifier in a similar light. We will assume that both devices are operated as linear amplifiers and then use the results in a practical way.

One must bear in mind that this approach cannot be assumed in all cases. It is, as it is meant to be, a simple analysis but the results will prove to be a valuable tool in trouble-shooting and understanding circuits.

Let us consider the general equation for current through a P.N. diode junction.

$$I = I_o \left[\exp \frac{V}{\rho V_e} - 1\right] \qquad (1)$$

where V = applied volts

I_o = reverse bias current

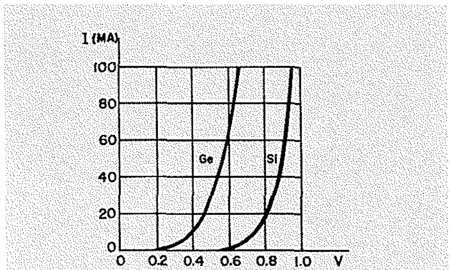
ρ = constant between 1 & 2

and $V_e = \dfrac{kT}{q}$ where k = Boltzmans Const., 1.38×10^{-23} Joule/°Kelvin

T = absolute temperature in degree Kelvin at room temperature, i.e., $T = 300°K$

q = electronic charge 1.602×10^{-19} Coulomb.

$$V_e = \frac{300}{11600} = 0.026 \text{ volts}$$



THE FORWARD VOLT-AMPERE CHARACTERISTICS OF A GERMANIUM AND SILICON DIODE

Figure 1.

Figure 1 shows a typical forward volt/amp characteristic for germanium and silicon diodes. Figure 2 is a plot of the collector current or the base current versus the base-to-emitter voltage of a transistor; point A on this curve is a typical operating point.

OBJECTIVE

The objective of this paper is to present a practical approach to Transistor and Vacuum-tube amplifiers based on a simple DC analysis.

The articles will be published in the following sequence.

1. The Transistor Amplifier.
2. The Vacuum-tube Amplifier.
3. An analysis of a typical Tektronix hybrid circuit (Type 545B vertical) based on conclusions reached in (1) and (2).

As a corollary they will bring forward some important relationships between vacuum tubes and transistors.

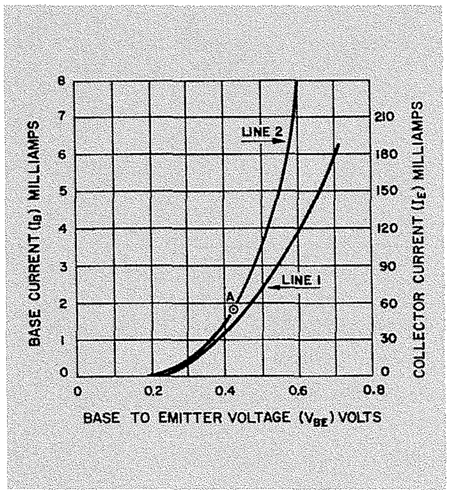


Figure 2. Line (1) is a plot of the base current versus the base-to-emitter voltage (V_{BE}). Line (2) is a plot of the collector current versus the base-to-emitter voltage (V_{BE}). Point "A" is a typical operating point.

One is quite justified in looking at a transistor in terms of the two-diode concept, refer to Figure 3. Therefore, assuming diode A to be forward biased and diode B to be reverse biased, as would be the case if we were to operate the transistor as a linear amplifier, the current through diode A will conform to equation (1). Let us take a closer look at Figure 2.

We define conductance in the general case as

$$g = \frac{I}{V}$$

and therefore at our operating point "A" the dynamic conductance

$$g' = \frac{\Delta I}{\Delta V} \qquad (2)$$

hence

$$g' = \frac{I_o \exp \dfrac{V}{\rho V_e}}{\rho V_e}$$

$$= \frac{I + I_o}{\rho V_e} \qquad (3)$$

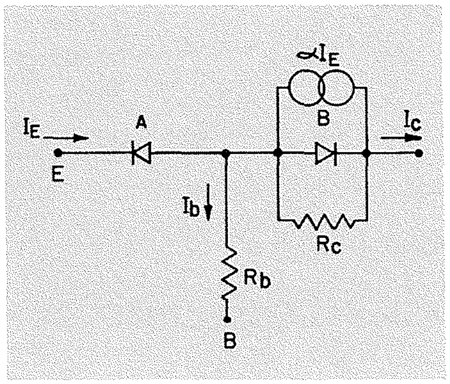


Figure 3. Illustration of the two-diode concept of a transistor.

but $I >> I_0$ then $g' = \dfrac{I}{\rho V_e}$

$$\text{or } g' = \frac{I}{0.026\rho} \text{ mhos} \tag{4}$$

The term "ρ" takes into account the recombination of carriers in the junction region. It is approximately unity for germanium and approximately 2 for silicon. At a typical operating point this term can usually be neglected. Therefore, we may say that

$$g' = \frac{I}{26} \text{ if I is in milliamps.} \tag{5}$$

Now resistance is the reciprocal of conductance and therefore the value of conductance at point "A" can be given in terms of resistance

$$r_e = \frac{26}{I}\ \Omega\text{'s} \tag{6}$$

This resistance (r_e) is commonly known as the dynamic emitter resistance.

At this point we will depart from our simple model and look at the transistor in another form; but, bear in mind our first thoughts. Transistor parameters are derived from various equivalent circuits depending upon the configuration i.e., common emitter, common base, or common collector. We will not consider any detailed analysis in this approach; but, to understand the approach it is necessary to know how these parameters are derived. It will be simple enough to derive another set of parameters once we have our basic model constructed.

The simplest and easiest equivalent circuit of a transistor is the "Tee" equivalent. It is a very good approximation about the behavior of a transistor, especially at DC and low frequencies. We can also represent either the common emitter or the common base simply by interchanging R_b and R_e. Figure 4 is a "Tee" equivalent circuit of the common emitter and the common base configurations.

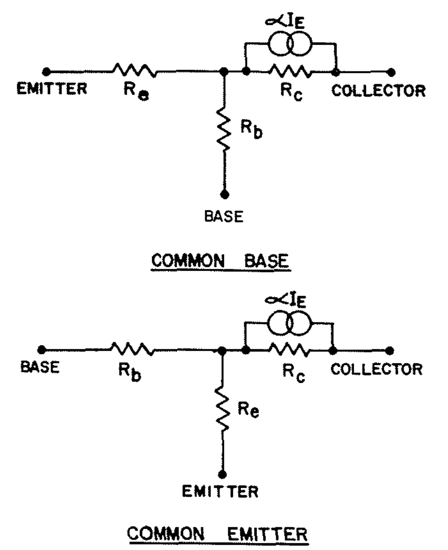


Figure 4. "Tee" equivalent circuits for the common-base and common-emitter configurations of transistors.

Firstly, let us define the term β (the small-signal current gain) as

$$\beta = \frac{\Delta I_c}{\Delta I_b} \tag{7}$$

and since $I_E = I_c + I_b$

$$\text{then } I_E = I_c\left(1 + \frac{1}{\beta}\right) \tag{8}$$

usually $\beta >> 1$ then $I_E \approx I_c$

Equation (8) shows us that only $\dfrac{1}{\beta}$ of the emitter current flows into the base. Hence, it is reasonable to suppose that any impedance in the emitter, when viewed from the base, will be β times as great; and, any impedance in the base, when viewed from the emitter, will be β times as small. That is to say, the dynamic resistance multiplied by β must equal R_e in our equivalent "Tee" circuit.

Hence $R_e = \beta re$

Our equivalent circuit shows a resistance R_b. This resistance is known as the base-spreading resistance. It is a physical quantity and can be expressed in terms of resistivity associated with the base-emitter junction. It can vary between a few ohms to hundreds of ohms, depending upon the type of transistor; and therefore, must be taken into consideration. Looking into the emitter we see it as an impedance whose value is divided by β and appears in series with the dynamic emitter resistance (r_e). Hence the emitter current encounters an impedance in the base/emitter junction which is equal to the sum of the dynamic resistance plus $\dfrac{R_b}{\beta}$, the latter term we will designate R_r and the sum of these two resistances we will designate R_t.

$$\text{Hence } R_t = r_e + R_r \tag{9}$$

The value of R_r can vary anywhere between $2\,\Omega$ to $24\,\Omega$ depending on the value of R_b. R_b is difficult to measure and rarely given in electrical data on transistors. A figure of 250 Ω's is a typical value at low frequencies. Therefore, if β were 50 then R_r would be 5 Ω's.

Now if we look into the base in the common emitter or the common collector configuration it is reasonable to suppose we will see the resistance (R_t)—plus any other impedance which may be wired to the emitter terminal—multiplied by β, then

$$R_{in} = \beta(R_t + R_E) \tag{10}$$

where R_E = the external emitter resistance.

If $R_E >> R_t$ then $R_{in} = \beta R_E$

So far we have had very little to say about R_e shunted by the current generator $\propto I_E$. If our equivalent "Tee" circuit consisted of resistances alone, it would be passive; i.e., it could supply no energy of its own. But a transistor can amplify energy to the signal. To represent this we have shown a current generator shunting R_c. The value of R_c will depend on the circuit configuration; i.e., tens of kilohms for a common emitter configuration, to many megohms for a common base configuration. In our approach it is not necessary to pursue this matter any further since we will not be considering a transistor in any extreme condition.

Now in a more practical sense, let us look at Figure 5, a typical common-emitter configuration.

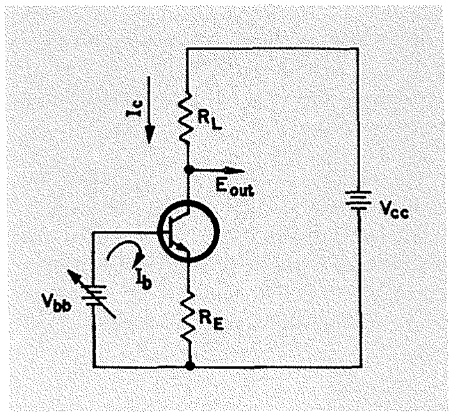


Figure 5. A typical common-emitter circuit.

Now we will assume $R_c >> R_L$.

Now by inspection

$$E_{out} = V_{cc} - \Delta I_c R_L \tag{11}$$

$$\text{hence } \Delta E_{out} = -\Delta I_c R_L \tag{12}$$

The input impedance we see looking into the base of a transistor in the common emitter configuration is

$$R_{in} = \beta(R_E + R_t) \tag{10}$$

$$\text{also } \Delta I_b = \frac{\Delta V_{bb}}{R_{in}}$$

$$= \frac{\Delta V_{bb}}{\beta(R_E + R_t)} \tag{13}$$

we also recall that

$$\beta = \frac{\Delta I_c}{\Delta I_b} \tag{7}$$

$$\text{hence } \Delta I_c = \beta \Delta I_b \tag{14}$$

Therefore substituting equation (13) in equation (14)

$$\Delta I_c = \beta \frac{\Delta V_{bb}}{\beta(R_E + R_t)} \tag{15}$$

and from equation (15)

$$\Delta V_{bb} = \Delta I_c\,(R_E + R_t) \tag{16}$$

we define the voltage gain as

$$A_{(v)} = \frac{\Delta E_{out}}{\Delta V_{bb}}$$

Then from equation (12) and equation (16)

$$A_{(v)} = -\frac{\Delta I_c \; R_L}{\Delta I_c \; (R_E + R_t)} \qquad (17)$$

$$= -\frac{R_L}{R_E + R_t}$$

and if $R_E >> R_t$ then

$$A_{(v)} = -\frac{R_L}{R_E} \qquad (18)$$

If we analyze the common-base configuration in a similar manner we arrive at the same result with the one exception that the sign is positive.

The conclusion we can draw from this analysis is that the gain of a transistor stage is set by external conditions provided that the emitter resistance is sufficiently great enough to "swamp" our internal resistance (R_t). In the absence of an emitter resistance

$$A_{(v)} = \frac{R_L}{R_t}$$

There is one very important fact we must remember about R_E. R_E will be that impedance in which the signal current will flow to the AC ground. We define an AC ground point as that point in a circuit at which the power level of the signal has been reduced to zero.

We normally encounter three types of an AC ground:

1. An Actual AC Ground.

This is the chassis point or the DC ground point. It is as well to remember the

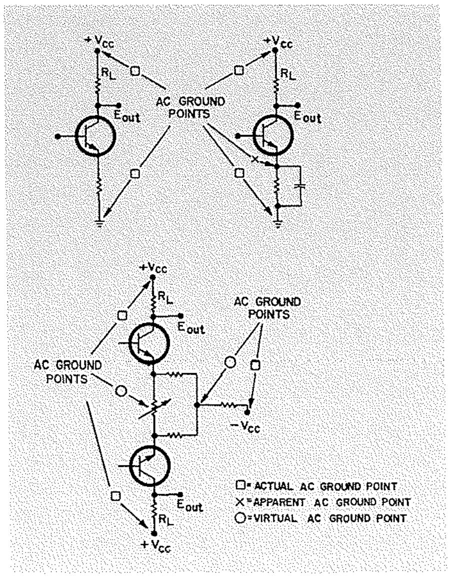


Figure 6. Illustrating the three types of AC ground normally encountered in electronic circuits.

power supply can be placed in this category so far as the signal is concerned.

2. An Apparent AC Ground.

The apparent AC ground may be represented by any point in a circuit which acts as to represent a low impedance between that point and the actual AC ground thereby bypassing the signal to an actual AC fround. A large value capacitor is a typical example should one side be returned to an actual AC ground.

3. The Virtual AC Ground.

The virtual A.C. ground point is perhaps the most difficult to recognize. It may best be explained as that point in a circuit where we have two signals of equal amplitude and frequency but exactly opposite in phase. Figure 6 will help clarify these points.

Figure 8 summarizes the results of our DC analysis of the common emitter, common base and common collector.

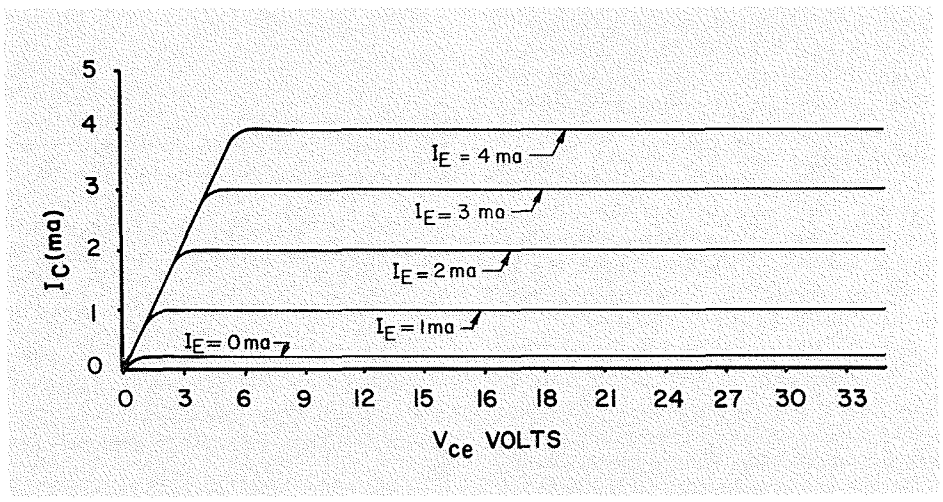


Figure 7. We define the parameter R_c in the common-base "Tee" configuration as;

$$R_c = \left. \frac{\Delta V_{ce}}{\Delta I_c} \right|_{I_E} \text{ohms}$$

Where ΔV_{ce} is the change in the collector voltage because of the change in collector current ΔI_c, when we hold the emitter current I_E constant.

Once the collector becomes saturated, the change in I_c is very small for a large change in V_{ce}. Hence, R_c is a very large resistance and does not modify the DC equivalent circuit to any extent. For this reason it was omitted from Figure 7. Therefore; $R_{out} \cong R_L$ (Common Base).

LIST OF SYMBOLS

$A_{(v)}$	Voltage gain defined as $\frac{\Delta E_{out}}{\Delta E_{in}}$
I_b	Base current
I_c	Collector current
I_E	Emitter current
R_c	Collector resistance (Tee Equivalent)
R_b	Base spreading resistance (Tee Equivalent)
R_e	Emitter resistance (Tee Equivalent)
R_E	External Emitter resistance (refer to text)
$R_{E(s)}$	The equivalent resistance between the signal source and the emitter terminal of the transistor in the common base configuration.
R_L	Load resistance
R_r	$\frac{R_b}{\beta}$
R_t	The "Transresistance" resistance (re + R_r)
re	dynamic emitter resistance
V_{bb}	Base voltage
V_{cc}	Supply voltage
V_{ce}	Collector to emitter voltage
Δ	(Delta) the change in the variable with which it is associated.

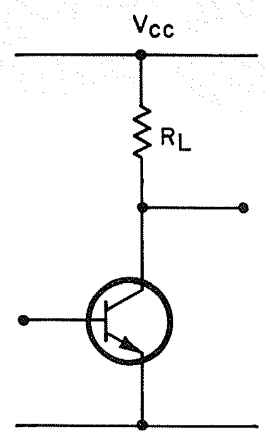
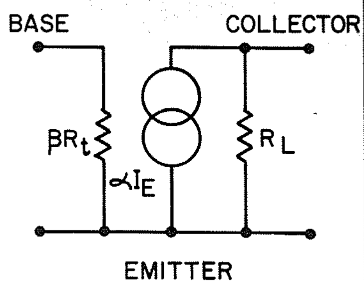
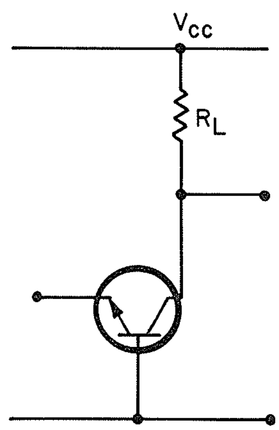
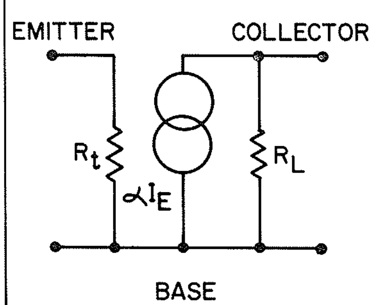
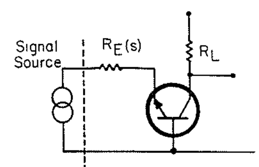
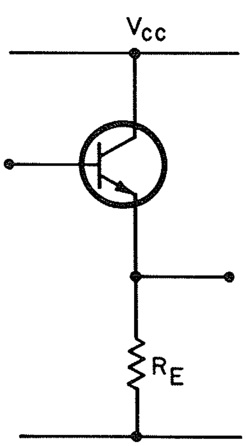
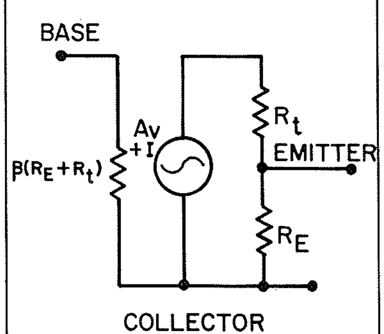
BASIC CIRCUIT	EQUIVALENT CIRCUIT	EQUATIONS	REMARKS
		GAIN $= A(v) = -\frac{R_L}{R_t}$ RIN $= \beta R_t$	A resistance (R_E) between the emitter terminal of the transistor to the AC ground will modify the gain equation and the input impedance; then, $A_{(v)} = -\frac{R_L}{R_t + R_E}$ and $R_{in} = \beta\,(R_E + R_t)$.
	COMMON EMITTER		
		GAIN $= A(v) = +\frac{R_L}{R_t}$ RIN $= R_t$ ROUT $= R_L$	The equivalent resistance $R_{E(s)}$ between the input signal source and the emitter terminal of the transistor will modify the gain equation and the input impedance as seen from the signal source; then, $A_{(v)} = +\frac{R_L}{R_t + R_{E(s)}}$ and $R_{in} = R_t + R_{E(s)}$. 
	COMMON BASE		
		GAIN $= A(v) = \frac{R_E}{R_E + R_t}$ RIN $= \beta\,(R_E + R_t)$ ROUT $= (R_t + \frac{I}{\beta})$ in parallel with R_E	The actual value of R_{out} will depend on what resistance is connected to the base. Let us assume the base is directly coupled to the preceding stage. The equivalent output impedance of the preceding stage becomes the numerator over beta in the second term in the parenthesis and the output impedance of the stage under consideration R_{out} is modified accordingly; eg., if the output impedance of the previous stage is 100 Ω, then $R_{out} = (R_t + \frac{100}{\beta})$ in parallel with R_E.
	COMMON COLLECTOR		

Figure 8.

SERVICE NOTES

SILVER-BEARING SOLDER AND SILVER SOLDER: TWO DIFFERENT THINGS

Many components in Tektronix instruments are mounted on ceramic strips. The notches in these strips are lined with a silver alloy and repeated use of ordinary tin-lead solder will breakdown the silver-to-ceramic bond. For this reason, we recommend the use of a silver-*bearing* solder containing 3% silver when performing service or maintenance work that requires soldering on these ceramic strips. This type of solder is used frequently in printed circuits and should be readily available from radio-supply houses.*

Silver-*bearing* solder should not be confused with silver solder. They are two different things!

The use of silver-*bearing* solder in the construction and maintenance and repair of electronic circuits is a safe and accepted practice. The silver-*bearing* solder used and recommended by Tektronix for ceramic strip soldering, melts at about 365 degrees Fahrenheit, and is applied with an ordinary soldering iron. It is composed of 60% tin, 37% lead, and 3% silver. It contains absolutely no cadmium! It produces no toxic or lethal fumes!

Silver solder, on the other hand, is a brazing alloy and is most commonly used by welders. It is composed essentially of silver, copper, zinc, and sometimes cadmium. When the alloy is composed of 45% silver, 30% copper and 25% zinc it requires approximately 1340 degrees Fahrenheit to melt it and it is usually applied with an acetylene torch. Should either the silver solder or the metals to which it is being applied contain cadmium, this high temperature will cause the cadmium to vaporize and release fumes. These fumes will be toxic and they can be lethal.

In summary, let us repeat; Silver-*bearing* solder and silver solder are two different things:

Silver-*bearing* solder is used primarily in the soldering of electronic circuits. Silver solder is an alloy used in the brazing and welding of metals.

Silver-*bearing* solder is applied with a soldering iron and requires only relatively low temperature to melt it. Silver solder is applied with an acetylene torch and requires a high temperature to melt it.

Silver-*bearing* solder absolutely does not produce toxic fumes. Silver solder, if it contains cadmium or is used on metal containing cadmium, does produce fumes that are toxic and can be lethal.

Positively no silver solder is used in any instrument produced by Tektronix, Inc.

*If you prefer you can order this solder directly through your local Tektronix Field Office, Field Engineer, Field Representative, or Distributor. Order Tektronix part number 251-0515-00.

OOPS! WRONG PART NUMBER

In the December 1966 issue of Service Scope, we transposed two figures in the Tektronix part numbers for the probe identification tags. The part number for the identification tags for use on the smaller (0.125" diameter) cables is 334-0798-00, and the number for the larger (0.178 to 0.185" diameter) cables is 334-0798-01.

COMPONENT LUBRICATION KIT FOR TEKTRONIX INSTRUMENTS

We have available a component lubrication kit for Tektronix instruments. The kit contains: a detent lubricant in a container-applicator; a switch-contact lubricant in a container-applicator; a pot lubricant in a container-applicator; 12 each detent-ball replacements (for lost or worn detent balls) in the following sizes—5/32", 3/16", and 7/32"; a #3 brush, and an instruction book.

The instruction book contains information on the cleaning and washing of Tektronix instruments and when an instrument needs lubrication. It also contains illustrations showing the different types of switches used in Tektronix instruments and tells how to lubricate them and replace worn or lost detent balls. The lubrication of potentiometers and fan motors and the care of air filters are also covered. Suggestions for the lubrication of rackmount tracks are given.

You may order the kit through your local Tektronix Field Office, Field Engineer, Field Representative, or Distributor. Specify Tektronix part number 003-0342-00.

TYPE 1L5, TYPE 1L10, TYPE 1L20, AND TYPE 1L30 PLUG-IN SPECTRUM ANALYZERS WITH A TYPE 132 POWER SUPPLY

These spectrum analyzers can be used in conjunction with a Type 132 Plug-In Unit Power Supply and the output displayed on any Tektronix oscilloscope that has a Sawtooth-Out sweep voltage available on the front panel.

Positive output-polarity voltage from the Type 132 can be applied to the DC-coupled input of the oscilloscope. Centering of the oscilloscope sweep is performed with the oscilloscope vertical-position control prior to RF signal application to the analyzer. The analyzer vertical-position control can then be used for trace positioning.

The Sawtooth-Out sweep voltage from the oscilloscope is applied to the Sweep-Input connector.

TYPE 1L10, TYPE 1L20, AND TYPE 1L30 PLUG-IN SPECTRUM ANALYZERS—VERTICAL TRACE SHIFT

If a vertical trace shift is encountered when a Type 1L10, Type 1L20, or Type 1L30 Plug-In Spectrum Analyzer is switched between linear and log mode, suspect a gassy input tube in the indicator (oscilloscope) vertical amplifier. The output impedance of the analyzer unit is much higher in the log mode than it is in the linear mode. If grid current is present in the input tube, this current will give a different voltage drop across the input resistance (analyzer output impedance); consequently, a DC shift of the trace will result.

simplify waveform measurements

Tektronix Type 502A

100 μV/cm dual-beam oscilloscope

- ☐ **Measure stimulus and reaction on the same time base.**
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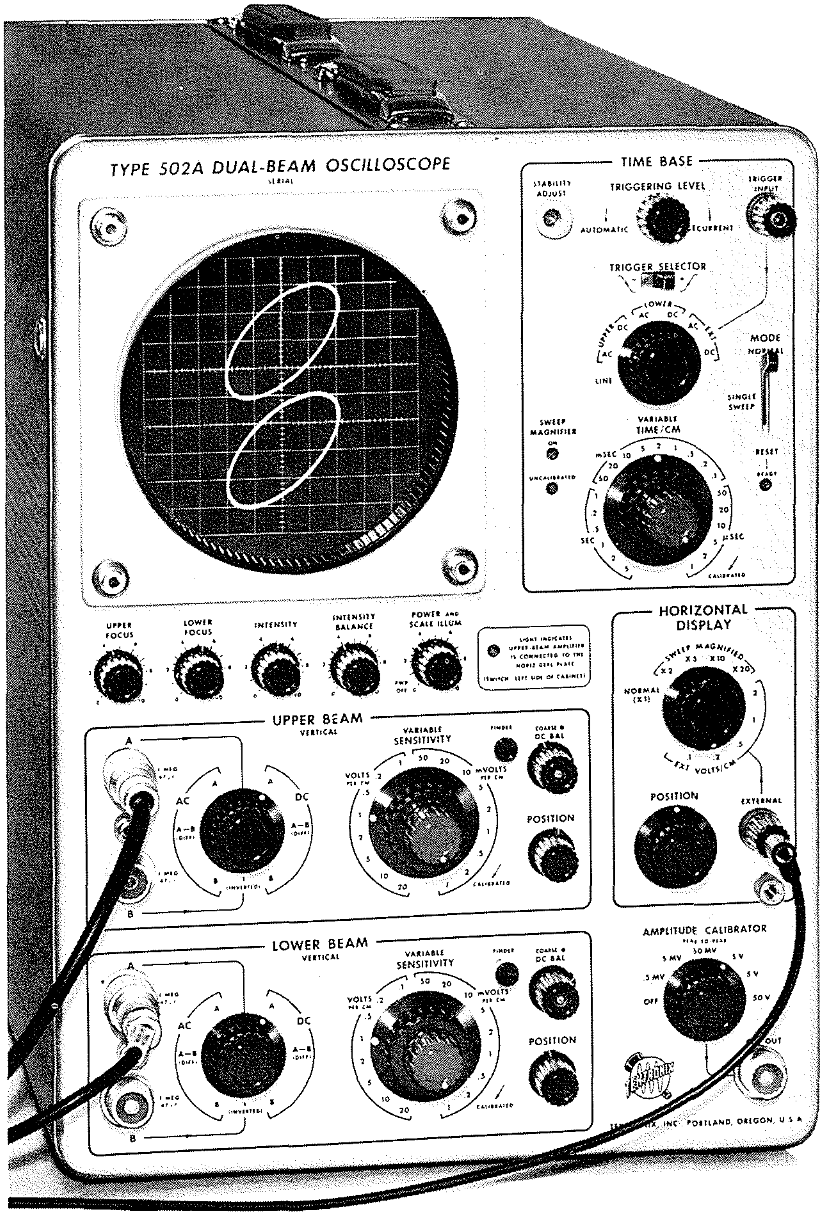

The Type 502A combines the performance capabilities unique to dual-beam oscilloscopes with operational features designed to simplify and speed up your measurements.

You can examine two waveforms simultaneously by applying input signals to both of the identical vertical amplifiers. You can use each vertical amplifier in a differential display mode to examine the difference between two signals. You can also use the Type 502A as a single-beam X-Y oscilloscope or as a dual-beam X-Y oscilloscope with both traces plotted on the same X scale.

This performance is combined with operating conveniences which include pushbutton beam finders for quick location of off-screen signals, vertical signal outputs, intensity balance for identification of upper and lower beams, single-sweep operation, Z-axis input, variable control of vertical and horizontal deflection factors, electronically-regulated power supplies for stable operation, and other refinements.

performance characteristics include:

Bandwidth from DC to 100 kHz at 100 μV/cm, increasing to DC to 1 MHz from 5 mV/cm to 20 V/cm • Calibrated deflection factors from 100 μV/cm to 20 V/cm in 17 steps; continuously variable between steps, uncalibrated, and to 50 V/cm • Common-mode rejection of at least 50,000:1 from DC to 50 kHz ••Phase difference between amplifiers less than 1 degree from DC to 100 kHz • Calibrated sweep rates from 1 μs/cm to 5 s/cm in 21 steps • 2X, 5X, 10X, 20X sweep magnification • Flexible trigger facilities • Amplitude Calibrator • 10 cm by 10 cm display area.

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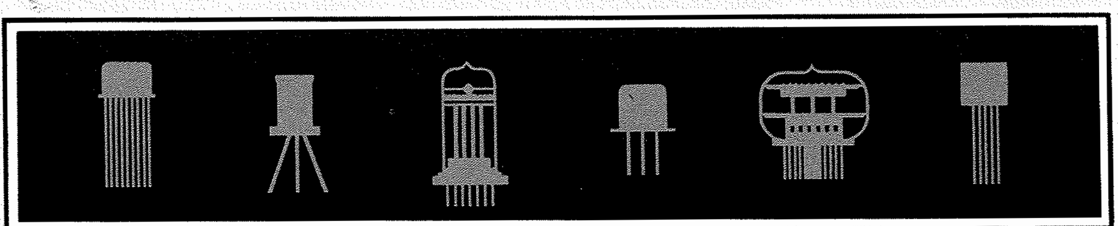
PRACTICAL APPROACH TO TRANSISTOR AND VACUUM TUBE AMPLIFIERS

BY F. J. BECKETT
TEKTRONIX, INC.
ELECTRONIC INSTRUMENTATION GROUP
DISPLAY DEVICES DEVELOPMENT

PART 2
THE VACUUM TUBE AMPLIFIER



This is the second in a series of three articles offering a new approach to transistor and vacuum-tube amplifiers. This new approach is based on a simple DC analysis that incorporates the concepts of "transresistance" and the principles of Thévenin's Theorem.

Part 1, "The Transistor Amplifier", which appeared in the February, 1967 issue of SERVICE SCOPE considered the transistor amplifier as a simple DC model. This second article looks at the vacuum-tube amplifier in a similar light and sees some striking similarities in the two devices.

In the previous article (Part I, "The Transistor Amplifier) of this series, it was shown that the gain of a linear transistor amplifier is set by external conditions. The same reasoning can also be applied to vacuum tubes. The equivalent circuit of a vacuum-tube amplifier is shown in Figure 9. The current that is produced in the plate circuit by the signal (E_g) acting on the grid is taken into account by postulating that the plate circuit can be replaced by a generator, $-\mu E_g$ having an internal resistance (rp). We may also consider a vacuum-tube amplifier in terms of the constant-current form by replacing the voltage generator in the constant-voltage form with a current generator (gm E_g) shunting the internal resistance (rp).

These two approaches are valid in every respect but they do not convey much to us in the practical sense. Let us now consider a vacuum-tube amplifier from another approach.

In an amplifier which has its grid referenced to ground all plate-circuit impedances, R_L and rp, when viewed from the cathode are multiplied by the term $\frac{1}{\mu + 1}$. Also, by the same reasoning, the cathode impedances when viewed from the plate circuit are multiplied by the term $(\mu + 1)$. Therefore, the impedance we see looking into the cathode must be

$\frac{rp + R_L}{\mu + 1}$, where μ equals the amplification factor of the tube.

Hence it is reasonable to suppose that the voltage E_c, reference Figure 10, appears across this impedance we see looking into the cathode.

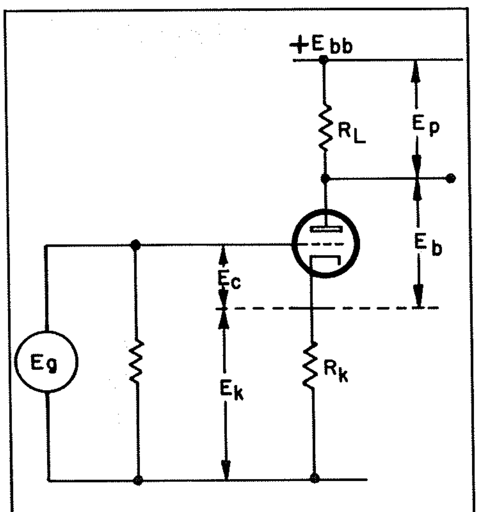


Figure 10. A vacuum tube amplifier in the grounded cathode configuration showing the various voltage measurements around the circuit.

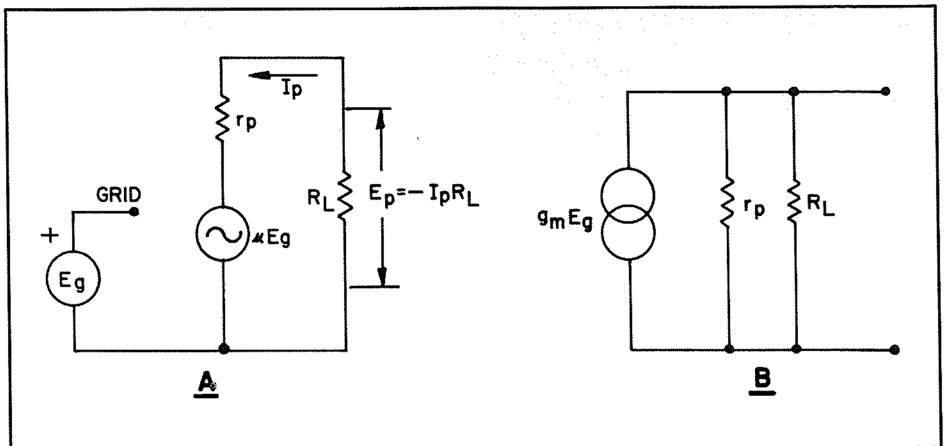


Figure 9. Illustrating the more familiar equivalent circuit of a vacuum tube amplifier.
(a) The constant voltage generator form or the Thévenin equivalent.
(b) The constant current generator form or the Norton equivalent.

The Triode Amplifier (Ground Cathode)

We will now look at a triode amplifier in terms related to our equivalent circuit. The common component is of course, the plate current. The change in this current due to the action of a control grid will determine the output voltage across the load impedance (R_L).

Now $E_g = E_c + E_k$ (19)

That is to say

$$E_g = I_p\left[\frac{rp + R_L}{\mu + 1}\right] + I_pR_k$$

$$\text{Or, } E_g = I_p\left[\left(\frac{rp + R_L}{\mu + 1}\right) + R_k\right] \quad (20)$$

Also, $E_{bb} = E_b + E_p + E_k$ (21)

or $E_b = E_{bb} - E_p - E_k$ (22)

and $E_p = -I_pR_L$ (23)

We define the voltage gain $A_{(v)}$ as

$$A_{(v)} = \frac{E_p}{E_g} \quad (24)$$

$$\text{Then } A_{(v)} = -\frac{I_pR_L}{I_p\left[\left(\frac{rp + R_L}{\mu + 1}\right) + R_k\right]}$$

$$= -\frac{R_L}{\left(\frac{rp + R_L}{\mu + 1}\right) + R_k} \quad (25)$$

We now have arrived at an equation for gain which is a ratio of impedances. The same approach may be applied to the grounded-grid configuration and we arrive at a similar result, except the sign is positive.

The Pentode Amplifier

In the triode amplifier all the cathode current will flow through the output load impedance (R_L). However, in the case of the pentode and other multigrid tubes, some of this current is diverted into the screen. Equation (23) defines the output voltage in terms of the plate current. Therefore, to derive the actual gain figure we must determine the actual amount of cathode current which will finally reach the plate and become signal current. This figure can be arrived at from a graphical analysis of the mutual-conductance curves. In most cases, about 72% of the cathode current reaches the plate to become signal current. A typical example is a type 12BY7 pentode. However, this figure can be as high as 90% for some types—for example a 7788 pentode. The ratio of the plate current (I_p) to the cathode current (I_k) is the plate efficiency factor, i.e., $\eta = \frac{I_p}{I_k}$.

Now let is reexamine what effect this fact must have on the gain of a pentode amplifier as compared to a triode amplifier. The impedance we see looking into the cathode of a pentode is the same as for a triode.

That is $\frac{rp + R_L}{\mu + 1}$

however rp >> R_L and therefore R_L can usually be neglected in this equation.

That is to say $\frac{rp}{\mu + 1} \approx \frac{1}{gm}$

and since conductance is the reciprocal of resistance we will call this impedance r_k.

$$\text{i.e. } r_k = \frac{1}{gm} \quad (26)$$

We have seen that the gain equation of the triode amplifier is defined in terms of the parameters μ and rp. We should not lose sight of the fact that μ and rp are related to the plate current and therefore when these parameters are transferred to cathode dimensions these terms must be multiplied by the plate efficiency factor (η). That is to say the impedance we see looking into the cathode r_k must be multiplied by (η). With these facts in mind let us now derive the gain equation for a pentode amplifier.

We recall that:

$$E_b = E_{bb} - E_p - E_k \tag{22}$$

and $$E_p = -I_pR_L \tag{23}$$

also $$E_g = E_c + E_k \tag{19}$$

$$= \eta r_k I_k + I_k R_k \tag{27}$$

but $$I_k = \frac{I_p}{\eta} \tag{28}$$

Therefore substituting equation (28) in equation (27)

$$E_g = \frac{\eta r_k I_p}{\eta} + \frac{I_p R_k}{\eta}$$

$$= I_p \left(r_k + \frac{R_k}{\eta}\right) \tag{29}$$

and since the voltage gain

$$A_{(v)} = \frac{E_p}{E_g}$$

$$= -\frac{I_pR_L}{I_p\left(r_k + \dfrac{R_k}{\eta}\right)}$$

$$= -\frac{R_L}{r_k + \dfrac{R_k}{\eta}} \tag{30}$$

The same remarks we made about the external emitter resistor R_E (refer to Part No. 1, The Transistor Amplifier) apply equally as well to the cathode resistor, R_k; *namely, R_k will be that impedance in which the signal current will flow to the AC ground.*

In the case of the grounded plate (the cathode follower) we do not need to consider the plate efficiency factor if the amplifier is triode connected, therefore, the "gain" can be considered in terms of a simple divider network which can never be greater than unity.

$$A_{(v)} = \frac{R_k}{R_k + r_k} \tag{31}$$

The Push-Pull Amplifier

We can view a push-pull amplifier in a similar light by recognizing the existence of a virtual AC ground point between the cathodes of $V_{(1)}$ and $V_{(2)}$ as shown in Figure 11. Therefore, the gain of a push-pull triode amplifier will be:

$$A_{(v)} = \frac{R_{L(1)} + R_{L(2)}}{r_{k(1)} + r_{k(2)} + R_{k(1)} + R_{k(2)}} \tag{32}$$

where subscripts (1) and (2) are associated with $V_{(1)}$ and $V_{(2)}$.

And if:

$$R_{k(1)} = R_{k(2)}$$

$$\text{and } r_{k(1)} = r_{k(2)}$$

which is usually the case; then,

$$A_{(v)} = \frac{R_{L(1)} + R_{L(2)}}{2r_k + 2R_k} \tag{33}$$

Where $r_k = \dfrac{r_p + R_L}{\mu + 1}$ (either $V_{(1)}$ or $V_{(2)}$)

and $R_k = R_{k(1)}$ or $R_{k(2)}$

With a push-pull pentode amplifier we must consider the plate-efficiency factor (η). Therefore,

$$A_{(v)}\ \text{pentode} = \frac{R_{L(1)} + R_{L(2)}}{2r_k + \dfrac{2R_k}{\eta}} \tag{34}$$

where $r_k = \dfrac{1}{gm}$ either $V_{(1)}$ or $V_{(2)}$

$R_k = R_{k(1)}$ or $R_{k(2)}$

η = plate-efficiency factor of either $V_{(1)}$ or $V_{(2)}$.

The Cascode Amplifier

The cascode amplifier fundamentally consists of two tubes connected in series, see Figure 12. Normally we usually fix the grid of $V_{(1)}$ at some positive voltage.

The key to understanding this type of circuit is to consider $V_{(2)}$ as a voltage-activated current generator. All the current delivered by $V_{(2)}$ passes through the output load impedance R_L. Any change in voltage appearing at the grid of $V_{(2)}$ appears as a change in current across R_L. We can derive the gain equation in the same way as we did for a pentode amplifier. There is no need to consider (η) if both tubes are triodes.

$$A_{(v)}\ (\text{stage}) = \frac{R_{L(1)}}{R_{k(2)} + r_{k(2)}} \tag{35}$$

$$\text{where } r_{k(2)} = \frac{r_{p(2)}}{\mu_{(2)} + 1}$$

$$= \frac{1}{gm_{(2)}}$$

where the subscripts (1) and (2) are associated with $V_{(1)}$ and $V_{(2)}$.

One of the advantages of this type of circuit is that the internal impedance which shunts R_L is extremely high.

In this respect the triode cascode amplifier closely approximates a pentode amplifier. If we compare the plate-current versus plate-voltage curves of both devices we see a close resemblance.

The Hybrid Cascode Amplifier

Figure 13 is a typical configuration consisting of a vacuum tube V_1 and a transistor, Q_1, connected in series. We can apply much the same approach as we did for the cascode vacuum-tube amplifier. Let us assume the base to emitter junction of Q_1 to be forward biased. The collector current of Q_1 becomes the plate current of V_1. Therefore, any change occurring at the base of Q_1 is reflected as a change in plate current in V_1.

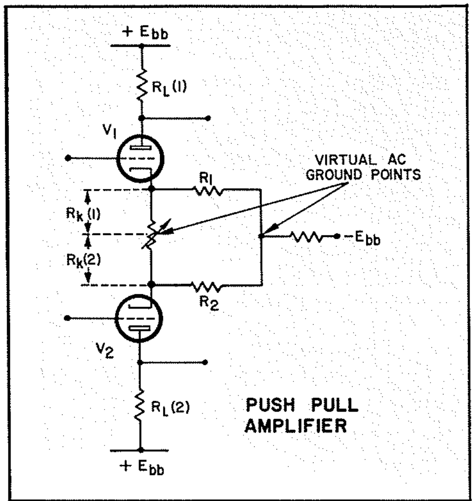


Figure 11. A typical push-pull triode amplifier. We normally encounter two virtual AC ground points between the cathodes V_1 and V_2. It may be necessary to consider the effect of the virtual AC ground point at the junction of R_1 and R_2. If R_1 or R_2 is large in value compared respectively to $R_{k(1)}$ or $R_{k(2)}$ then we can neglect this virtual AC ground and consider R_k in terms of $R_{k(1)}$ or $R_{k(2)}$. However, if this is not so, R_k will be the parallel combination of $R_{k(1)}$ and R_1 or $R_{k(2)}$ and R_2.

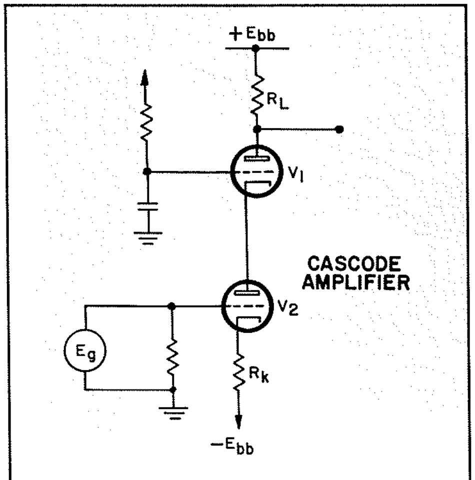


Figure 12. Illustrating a cascode amplifier using two triodes.

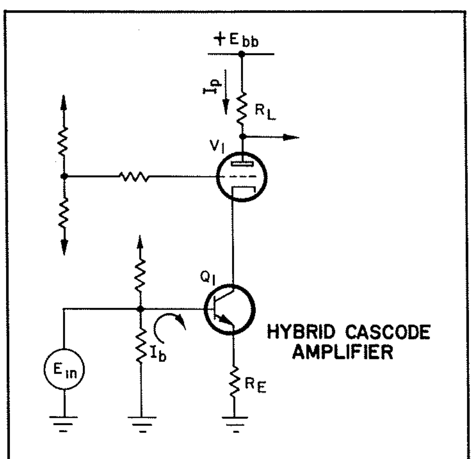


Figure 13. A typical hybrid cascode amplifier using a transistor and a vacuum tube.

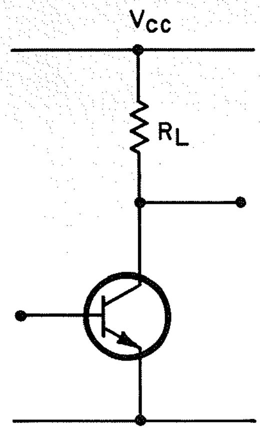
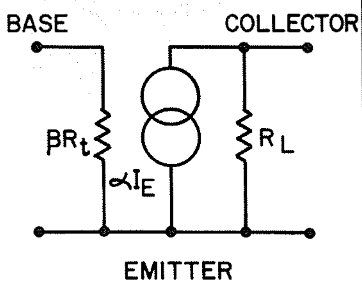
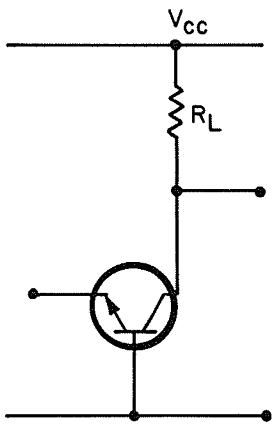
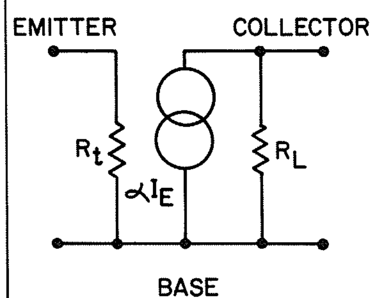
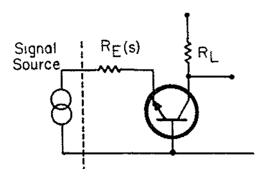
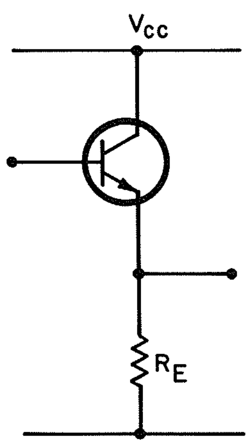
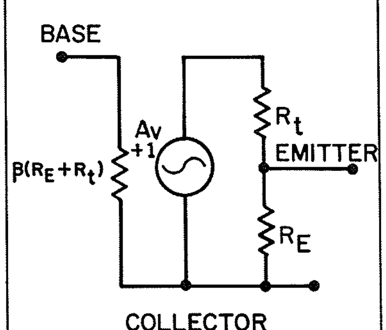
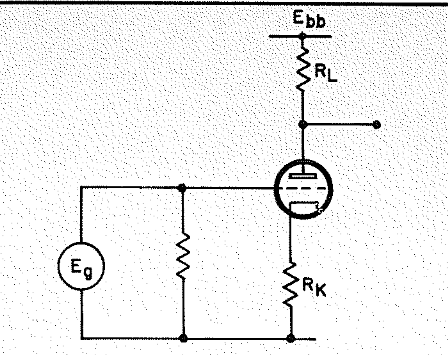
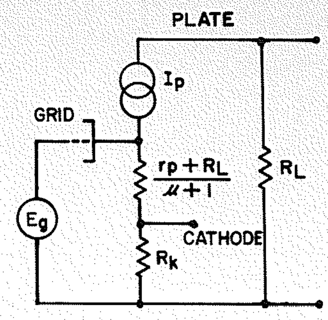
BASIC CIRCUIT	EQUIVALENT CIRCUIT	EQUATIONS	REMARKS
		GAIN $= A(v) = -\frac{R_L}{R_t}$ $R_{IN} = \beta R_t$	A resistance (R_E) between the emitter terminal of the transistor to the AC ground will modify the gain equation and the input impedance; then, $A(v) = -\frac{R_L}{R_t + R_E}$ and $R_{in} = \beta\ (R_E + R_t)$.
	COMMON EMITTER		
		GAIN $= A(v) = +\frac{R_L}{R_t}$ $R_{IN} = R_t$ $R_{OUT} = R_L$	The equivalent resistance $R_{E(s)}$ between the input signal source and the emitter terminal of the transistor will modify the gain equation and the input impedance as seen from the signal source; then, $A(v) = +\frac{R_L}{R_t + R_{E(s)}}$ and $R_{in} = R_t + R_{E(s)}$. 
	COMMON BASE		
		GAIN $= A(v) = \frac{R_E}{R_E + R_t}$ $R_{IN} = \beta\ (R_E + R_t)$ $R_{OUT} = (R_t + \frac{1}{\beta})$ in parallel with R_E	The actual value of R_{out} will depend on what resistance is connected to the base. Let us assume the base is directly coupled to the preceding stage. The equivalent output impedance of the preceding stage becomes the numerator over beta in the second term in the parenthesis and the output impedance of the stage under consideration R_{out} is modified accordingly; eg., if the output impedance of the previous stage is 100 Ω, then $R_{out} = (R_t + \frac{100}{\beta})$ in parallel with R_E.
	COMMON COLLECTOR		

Figure 8.

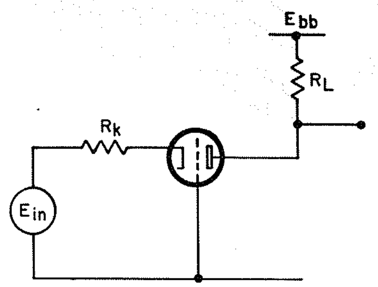
BASIC CIRCUIT (TRIODE)	EQUIVALENT CIRCUIT	EQUATIONS

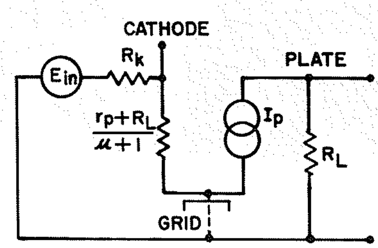




GROUNDED CATHODE

$$\text{GAIN} = A_{(v)} = -\frac{R_L}{\dfrac{r_p + R_L}{\mu + 1} + R_k}$$

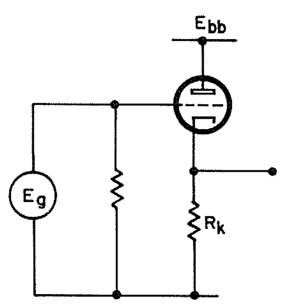


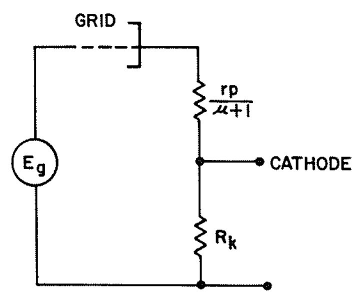


GROUNDED GRID

$$\text{GAIN} = A_{(v)} = +\frac{R_L}{\dfrac{r_p + R_L}{\mu + 1} + R_k}$$

$$R_{IN} = R_k + \frac{r_p + R_L}{\mu + 1}$$





GROUNDED PLATE

$$\text{GAIN} = A_{(v)} = +\frac{R_k}{R_k + r_k}$$

$R_{OUT} = r_k$ in parallel with R_k

WHERE: $r_k = \dfrac{r_p}{\mu + 1}$

PENTODE AMPLIFIER

$$\text{GAIN} = A_{(v)} = \frac{R_L}{r_k + \dfrac{R_k}{\eta}}$$

WHERE:

R_L = LOAD RESISTANCE

$r_k = \dfrac{1}{gm}$

R_k = CATHODE RESISTOR (Refer Text)

η = PLATE EFFICIENCY FACTOR

PUSH PULL AMPLIFIER

TRIODE PAIR

$$\text{GAIN} = A_{(v)} = \frac{R_{L(1)} + R_{L(2)}}{2r_k + 2R_k}$$

WHERE:

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

PENTODE PAIR

$$\text{GAIN} = A_{(v)} = \frac{R_{L(1)} + R_{L(2)}}{2r_k + \dfrac{2R_k}{\eta}}$$

WHERE: $r_k = \dfrac{1}{gm}$

SUBSCRIPTS (1) AND (2) ARE ASSOCIATED WITH V_1 AND V_2

CASCODE AMPLIFIER

$$\text{GAIN} = A_{(v)} = \frac{R_{L(1)}}{r_{k(2)} + R_{k(2)}}$$

WHERE:

$$r_k = \frac{1}{gm\,(2)}$$

HYBRID CASCODE AMPLIFIER

$$\text{GAIN} = A_{(v)} = \frac{R_L}{R_E + R_t}$$

WHERE:

R_L = LOAD RESISTANCE

* $R_t = r_e + R_r$

* R_E = EXTERNAL EMITTER RESISTANCE

* REFER PART I "THE TRANSISTOR AMPLIFIER"

Figure 16

We recall (Part 1, The Transistor Amplifier, Eq. 10) that the input impedance we see looking into the base of a transistor in the common-emitter configuration is:

$$R_{in} = \beta (R_E + R_t) \qquad (10)$$

Now $E_{in} = I_b R_{in}$

$$= I_b \, \beta (R_E + R_t) \qquad (36)$$

also $\beta = \dfrac{I_c}{I_b}$

$$\text{or } I_c = \beta \, I_b \qquad (37)$$

therefore substituting equation *(37)* in equation (36)

$$E_{in} = I_c (R_E + R_t) \qquad (38)$$

now the collector current Q_1 becomes the plate current of V_1. Then,

$$E_{in} = I_p (R_E + R_t) \text{ since } I_p = I_c \quad (39)$$

$$\text{also } E_p = -I_p R_L \qquad (23)$$

and since

$$A_{(v)} \text{ (stage)} = \frac{E_p}{E_{in}}$$

then from equations (23) and (39)

$$A_{(v)} \text{ (stage)} = -\frac{I_p R_L}{I_p (R_E + R_t)}$$

$$= -\frac{R_L}{R_E + R_t} \qquad (40)$$

If the vacuum tube is not a triode but some other multigrid tube such as a pentode, the gain equation will have to be multiplied by the plate efficiency factor (η).

The same remarks concerning the output impedance of the vacuum-tube cascode amplifier can be applied to the hybrid counterpart.

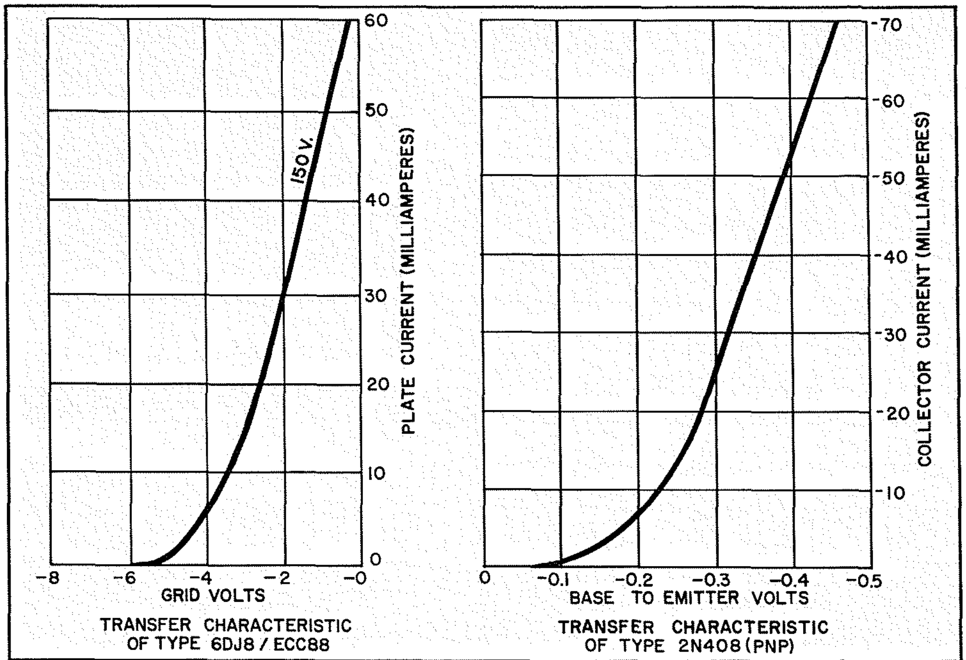


Figure 14. The transfer characteristic curves of a vacuum tube (6DJ8) and a PNP transistor (2N408), illustrating the basic similarity between vacuum tubes and transistors.

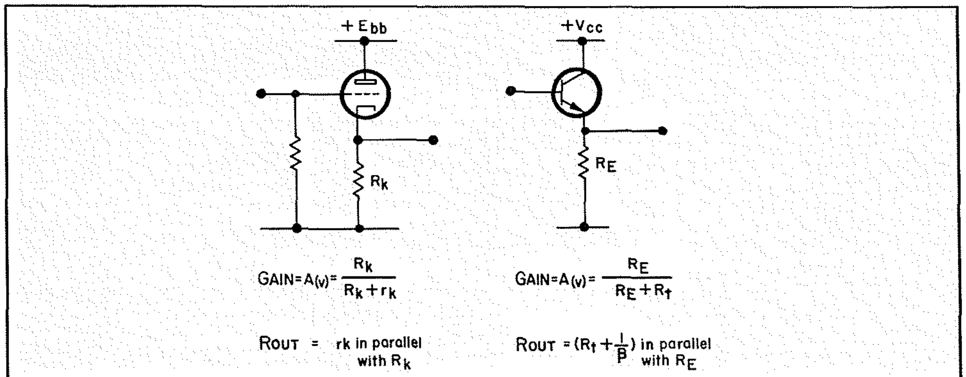


Figure 15. The analogy between the cathode follower (grounded plate) and the emitter follower (the common collector) in terms of "gain" and output impedances of both devices.

Summary

We have shown that the gain of a linear amplifier, transistor or vacuum tube, is a ratio of impedances. We can, of course, derive the gain equations for both devices in terms of mutual conductance. In fact, if we compare the transfer curves of both devices, Figure 14, we see a striking similarity. V_{BE} and E_g can be thought of in the same terms and in like manner I_p and I_c perform identical functions. Our analysis of both devices has shown that this fact is not coincidence.

It is not unreasonable to say that when we compare the cathode-follower (grounded-plate) against the common-collector configuration, Figure 15, we can think of both devices as being identical in operation—differing only in concept. The same argument can be put forward about the common-base amplifier and the grounded-grid amplifier. So too, the common-emitter amplifier and the grounded-cathode amplifier if we chose to ignore the input impedances of both devices.

Figure 16 (see page 5) summarizes the results of our analysis of the grounded cathode, grounded grid, and grounded plate amplifiers. Opposite this Figure we have reprinted Figure No. 8 from the previous article (Part I, The Transistor Amplifier) which summarized the results of the analysis on the three types of transistor amplifiers. These two charts will assist you to follow more closely our analysis of the 545B vertical amplifier (appearing in the next issue of SERVICE SCOPE) and to make a comparison between transistor and vacuum tube amplifiers.

It is not surprising we sometime find ourselves explaining one device in terms of another. Nature has a charming way of making most things interdependent upon one another. Recognize this fact and most tasks become a little easier.

The third and concluding article in this series will appear in the June, 1967 issue of SERVICE SCOPE. That article will present an analysis of a typical Tektronix hybrid circuit—a Type 545B Oscilloscope's vertical amplifier.

The analysis will be based on conclusions reached in Part 1 (February, 1967 issue) and Part 2 (this issue) of the series of articles.

ERRATA

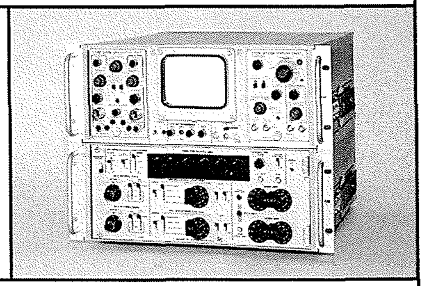
We call your attention to a typographical error in the caption under Figure 7 in the February issue of SERVICE SCOPE. The Figure referred to in the last line of this caption should be Figure 8—not Figure 7.

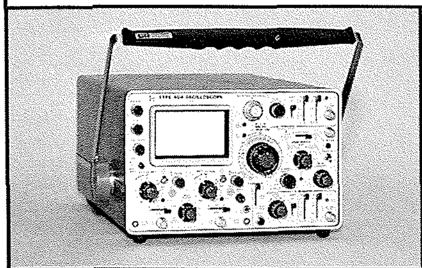
New From Tektronix, Inc. In 1967

For complete information, contact your Field Engineer, Field Representative, or Distributor.

DIGITAL READOUT OSCILLOSCOPE

The Type 568 Readout Oscilloscope accepts sampling and real-time amplifiers and time-base units. Used with the Type 230 Digital Unit, digital readout of measurements (in addition to the analog CRT display) is provided, allowing faster and more accurate answers than using the CRT alone. A wide variety of repetitive pulse measurements can be made digitally, without operator error: pulse voltage, risetime, delay time, storage time, and pulse width, among others. Measurement limits may be selected to provide Go/No-Go indicators.



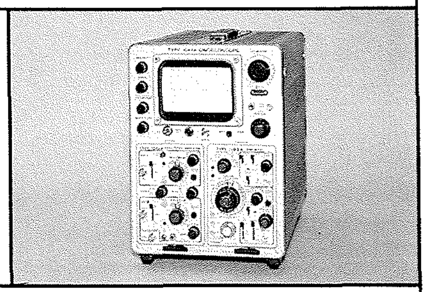


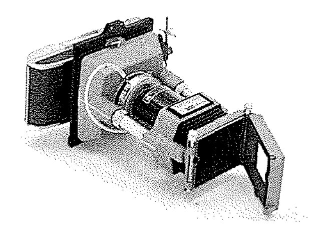
DC-to-150 MHz PORTABLE OSCILLOSCOPE

The Type 454 provides accurate dual-trace displays of fast-rise pulses and high-frequency signals previously beyond the capability of most real-time oscilloscopes. Risetime with included 10X probes is 2.4 ns. Other features include X-Y displays to 5 mV/div, chopped or alternate switching between inputs, calibrated sweep delay, and rugged design for environmental extremes.

DC-to-100 MHz OSCILLOSCOPE

The Type 647A with choice of amplifier and time-base units provides accurate displays over a wide range of temperature and other environmental extremes. Bandwidth and triggering extend to 100 MHz with the new Type 10A2A Dual-Trace Plug-In and Type 11B2A Sweep-Delay Time Base Plug-In. Differential comparator and single time-base plug-ins are also available. The Type 647A has 14-kV accelerating potential for bright 6 x 10-cm displays.





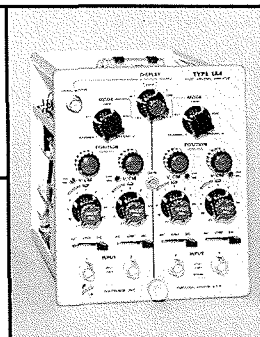
HIGH-SPEED CAMERA

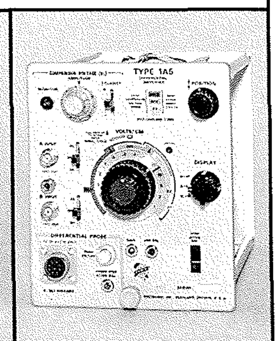
The C-40 is a high-performance camera for Tektronix portable oscilloscopes. The f/1.3—1:0.5 lens and Polaroid* Roll-Film back for 10,000-speed film provide the writing speed necesary to record single-shot events on the Type 454 Oscilloscope.

*Registered Trade-Mark Polaroid Corporation

DC-to-50 MHz FOUR-TRACE AMPLIFIER

The Type 1A4 Plug-In Unit for Type 530, 540, 550, and (with adapter) 580-Series Oscilloscopes provides the equivalent of two wide-band, dual-trace units connected to a third wide-band, dual-trace unit. Unique display logic provides unprecedented display flexibility including four-channel adding (±1 ±2) + (±3 ±4). Deflection factor is 10 mV/cm to 20 V/cm.





DC-to-50 MHz DIFFERENTIAL AMPLIFIER

The Type 1A5 Plug-In Unit for Type 530, 540, 550, and (with adapter) 580-Series Oscilloscopes achieves a new high in common-mode rejection. Gain-bandwidth products exceed those previously available in a differential amplifier. Bandwidths from DC to 50 MHz can be achieved at 5 mV/cm, DC to 45 MHz at 2 mV/cm, and DC to 40 MHz at 1 mV/cm. A ±5-V comparison voltage is built in.

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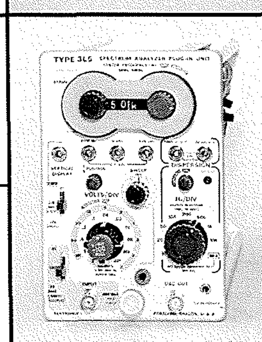
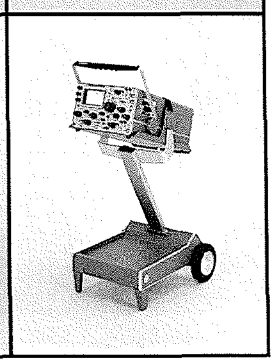
The Type 3L5 Plug-In Unit for Type 561A and 564 Oscilloscopes operates over a center-frequency range of 50 Hz to 1 MHz, and provides accurate spectral and time-based displays from 10 Hz to 1 MHz. Deflection factors extend to 10 μV/div RMS for spectral displays, and to 1 mV/div P to P for time-base displays. Dispersion is calibrated from 10 Hz/div to 100 kHz/div. Resolution bandwidth is $\leq$10 Hz to $\geq$500 Hz.





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Model 200-1 holds Type 454 or other portable instruments. Friction locks provide tilting from 0 to 60 degrees. Cart occupies less than 18 inches of aisle space, goes up and down stairs easily. Model 200-2 is similar, holds Type 422. Model 205-2 and 205-3 hold Type 568 or other instruments of similar size. Plug-in compartments are provided for three Letter Series or 1-Series plug-ins, or four 2- or 3-Series plug-ins.

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A PRACTICAL APPROACH TO TRANSISTOR AND VACUUM TUBE AMPLIFIERS

BY F. J. BECKETT

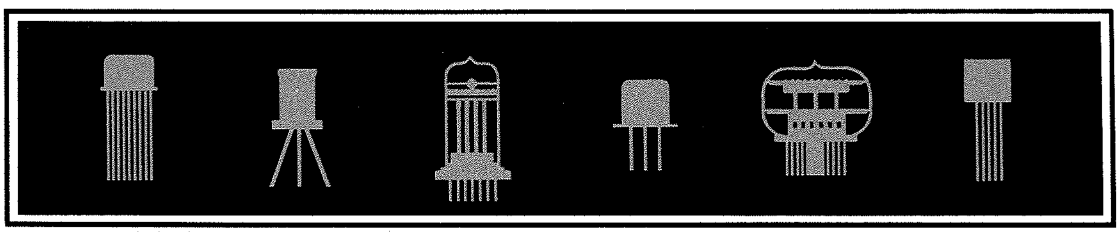
TEKTRONIX, INC.

ELECTRONIC INSTRUMENTATION GROUP

DISPLAY DEVICES DEVELOPMENT

PART 3

A DC ANALYSIS OF A TYPICAL TEKTRONIX HYBRID CIRCUIT



This is the third in a series of three articles offering a new approach to transistor and vacuum-tube amplifiers. This new approach is based on a simple DC analysis that incorporates the concepts of "trans-resistance" and the principles of Thévenin's Theorem.

In this article, conclusions reached in Part 1, "The Transistor Amplifier" (Service Scope #42, February 1967) and Part 2, "The Vacuum Tube Amplifier" (Service Scope #43, April 1967) form the basis for a DC analysis of a typical Tektronix, Inc. hybrid-amplifier circuit.

As a typical example of a Tektronix, Inc. hybrid circuit on which to demonstrate our DC analysis, we have chosen the vertical amplifier of a Type 545B Oscilloscope. This circuit is representative of the hybrid circuit one encounters so often in electronic instrumentation today.

The Type 545B vertical amplifier is a hybrid push-pull amplifier operating in a class A mode. It incorporates a few extra circuits such as trigger pick-off amplifiers necessary to accomplish its function, but, basically it is a hybrid push-pull amplifier.

To begin our analysis of the amplifier, the first thing we must do is select a portion of the amplifier circuit which will give us the information necessary for us to make our first calculation. We are going to analyze the whole circuit so we can choose our point of entry. The input circuit is as good a point as any. Bear in mind that, for our purpose, this is not the only point of entry. Any point on the circuit which will give us useful information would do.

A quiescent DC voltage of +67 volts is the nominal voltage at the output of the plug-in amplifiers used in the Type 545B oscilloscope. This voltage appears at terminals 1 and 3 of J11 in Figure 17, and thus, at the grids of V494A and V494B, a 6DJ8 dual triode. The input cathode follower (V494 A & B) has a bias of about 4 volts; therefore, both cathodes will be at +71 volts. The base voltage of Q514 and Q524 is then fixed at 71 volts. This sets the emitter voltages of Q514 and Q524 at one junction drop more negative (they are both NPN transistors) than the base. Therefore, the voltage at the emitter of Q514 and Q524 is 70.5 volts. T500 is a small toroidal transformer used for high-frequency common-mode rejection. The DC BALANCE Control, R495, sets the quiescent condition. We mean by this that the trace is centered.

We have made certain assumptions about the bias of a vacuum tube and the base-to-emitter voltage drop of a transistor. This is quite justifiable since we know what function the device performs. One helpful hint about transistors is that you can expect a base-to-emitter voltage drop of about 0.5 to 0.6 volts for a silicon transistor and about 0.2 volts for a germanium transistor.

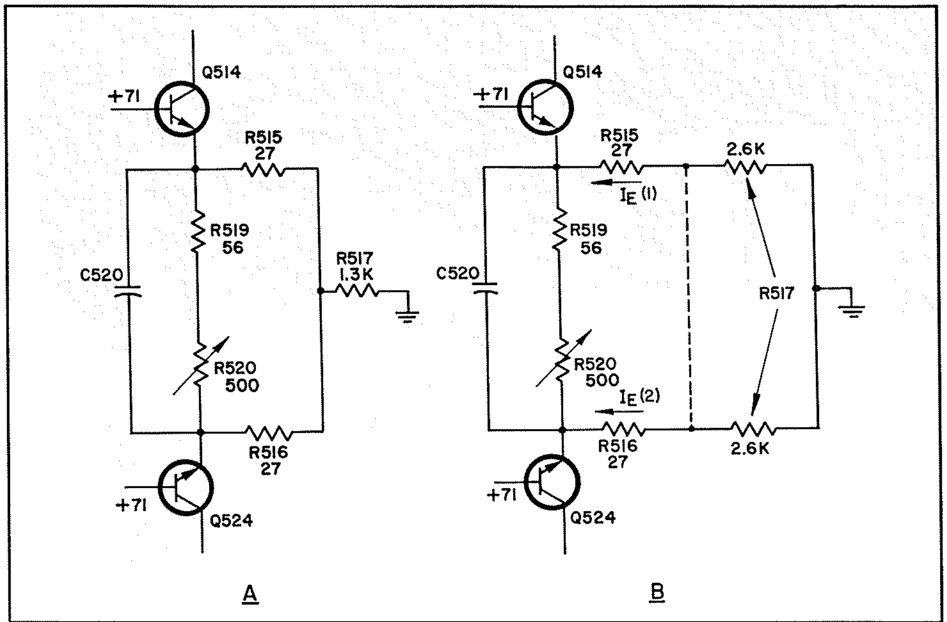


Figure 18. The circuit which will determine the DC emitter currents for either Q514 or Q524. (A)—The actual circuit as shown in Figure 17. (B)—The equivalent DC circuit considering R517 as two resistors through which the individual emitter currents will flow.

We are now able to calculate the emitter current of either Q514 or Q524. The DC-emitter current will flow through R515 or R516 and into R517 to ground. Since the emitter currents of Q514 and Q524 both pass through R517, we may think of R517 being made up of two resistors, each of 2.6 kΩ in value, in which the individual emitter currents will flow, refer to Figure 18: Therefore,

$$I_E\ (1) \text{ or } (2) = \frac{70.5 \times 10^3}{2.627 \times 10^3} \text{ mA}$$

$$= 27 \text{ mA}$$

We can now calculate the value of r_e, the dynamic-emitter resistance,

$$r_e = \frac{26}{I_E} = \frac{26}{27}$$

$$= 0.96\ \Omega$$

to this we can add our constant, R_r, of say, 4 Ω. We recall that:

$$R_t = r_e + R_r \qquad (9)$$

therefore:

$$R_t = 0.96 + 4 = 4.96\ \Omega$$

or approximately 5 Ω. We have now established the value of the emitter current and the value of R_t for Q514 and Q524.

Our next step is to find the value of R_E. We must know this value in order to calculate gain. You will recall that R_E will be that impedance through which the signal current will flow to the AC ground. Let us take another look at the resistive network between the emitters of Q514 and Q524. The signal currents flowing in this circuit will be equal and opposite at two points, refer to Figure 19. These points are virtual AC-ground points; therefore, the impedance seen by the signal current from the emitters of Q514 or Q524 will be the parallel combination of 153 Ω and 27 Ω or approximately 23 Ω to the AC ground points. Hence, R_E for Q514 or Q524 will be 23 Ω.

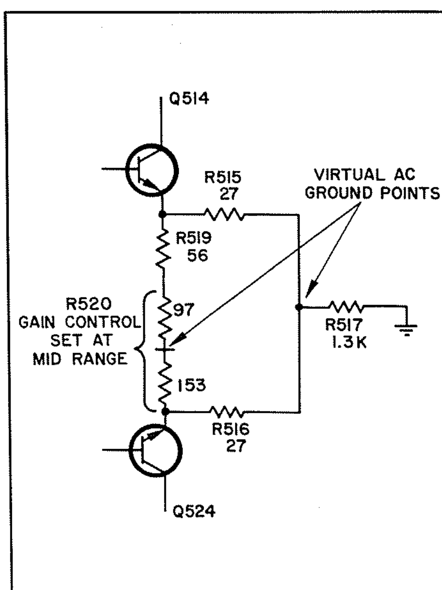


Figure 19. The location of the virtual AC ground points between the emitters of Q514 and Q524.

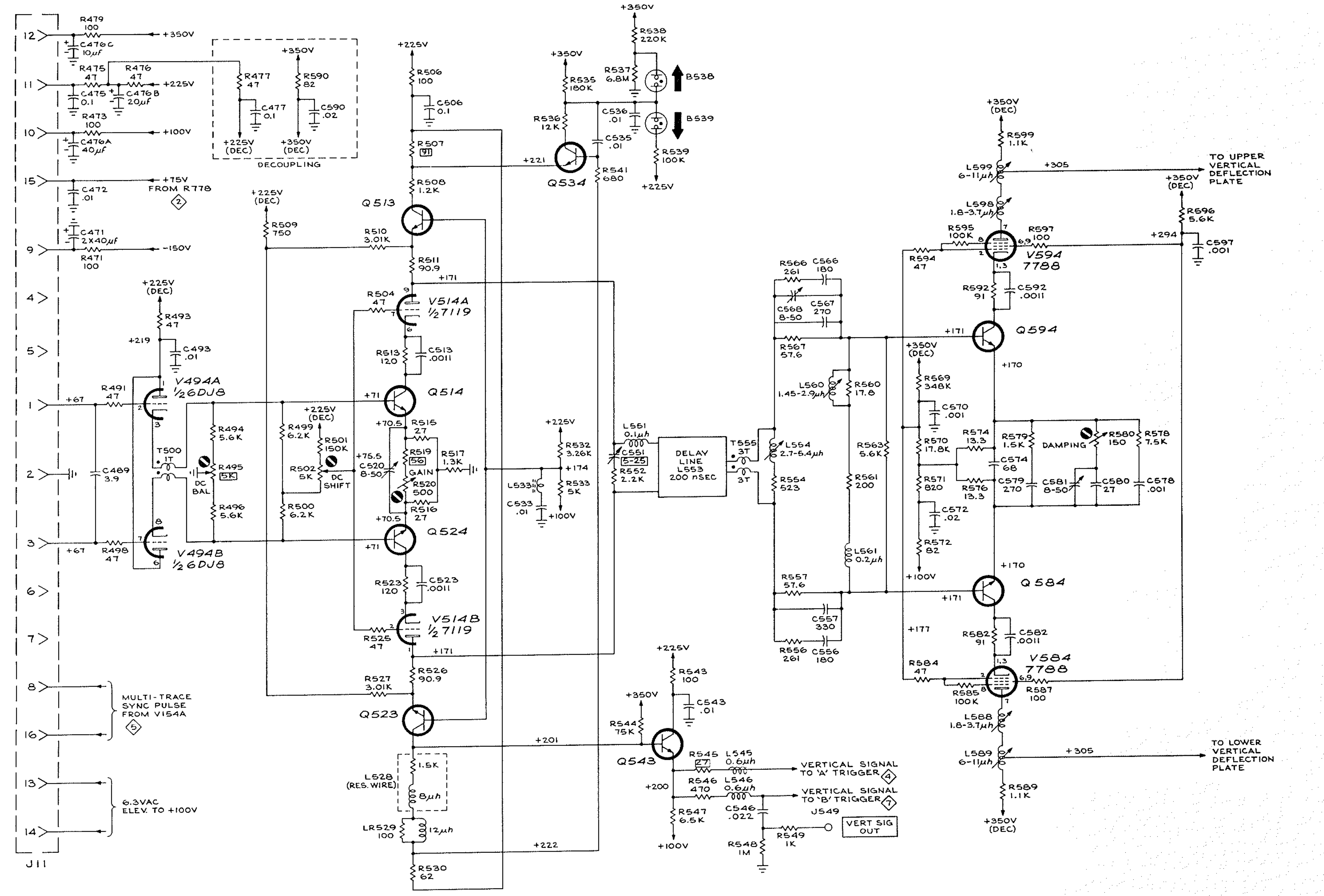


Figure 17.   TYPE 545B OSCILLOSCOPE   VERTICAL AMPLIFIER

We have now calculated from this part of the circuit all of the information we need to progress further into the circuit. Let us turn our attention to the circuit around Q513 and Q523. The first thing we notice is that the base of Q513 and Q523 are tied together at an AC-ground point. You will recall that the impedance we see looking into the emitter of the common-base configuration is R_t. In order to calculate R_t we must, of course, calculate r_e and add our constant for R_r of 4 Ω; r_e will be a function of the actual value of current flowing into the emitter. 27 milliamps has been set in the emitter circuit of Q514 and Q524; but not all of this current will flow into the

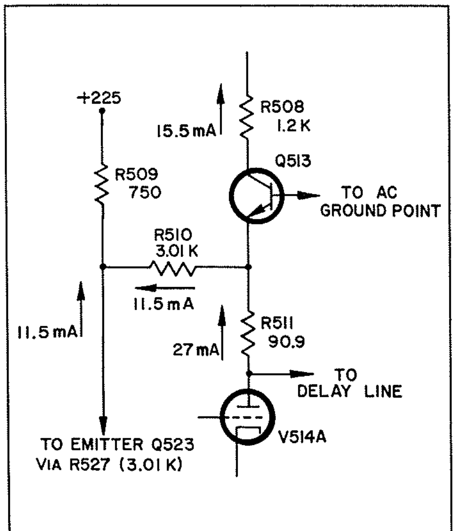


Figure 20. The DC current paths in the emitter circuit of Q513 or Q523.

emitter of Q513 and Q523. 11.5 milliamps will flow through R510 and R527, refer to Figure 20. The actual value of current into Q513 or Q523 will be 15.5 milliamps. Therefore, the impedance (R_t) we see looking into the emitter of Q513 and Q523 will be

$$R_t = r_e + R_r \qquad (9)$$

$$= \frac{26}{15.5} + 4\,\Omega$$

$$= 5.68\,\Omega$$

This impedance of 5.68 Ω plus R511 or R526 (90.9 Ω) constitutes part of the load impedance of the hybrid cascode amplifier Q514, V514A or Q524, V514B and the necessary matching impedance for the delay line.

There is one point we should make clear here. We have assumed a value of 4 Ω for R_r which you will recall is equal to $\frac{R_b}{\beta}$. R_r can vary from between 2 Ω to 24 Ω depending upon the type of transistor (refer to Part 1, "The Transistor Amplifier" SERVICE SCOPE #42, February 1967). This is one of those few times we should be really a bit more specific about assuming a value of R_r. The sum of the impedances 5.68 Ω and 90.9 Ω should be equal to 93 Ω since our delay line is a 186 Ω balanced line. Therefore, we have a difference of 3.58 Ω between the theoretical value and the calculated value, or an error of approximately 3.7%. This error has been due in part to our presupposed value of R_r to be 4 Ω. Such an error could not be tolerated in design work but it is acceptable here for our purpose of DC analysis. Bear this limitation in mind when you apply this analysis.

There is another point we must clear up. What is the load impedance of the hybrid cascode amplifier Q514, V514A or Q524, V514B? Clearly it will be that impedance or impedances connected from the plate of V514A or V514B to the AC ground. We are using a balanced delay line of 186 Ω, (93 Ω to a side), referenced to the AC ground. Therefore, the delay line impedance (93 Ω) must shunt R511 in series with R_t (or R526 in series with R_t) making an effective load impedance of approximately 47 Ω in the plate circuit of V514A or V514B. We now have all the necessary information to calculate the gain to this point.

$$A_{(v)} = \frac{R_{L(1)} + R_{L(2)}}{R_{E(1)} + R_{E(2)} + R_{t(1)} + R_{t(2)}}$$

$$= \frac{47 + 47}{23 + 23 + 5 + 5}$$

$$= \frac{94}{56}$$

$$A_{(v)} = 1.68$$

Q523 is the trigger pick-off amplifier and Q543 is an emitter follower providing isolation between the vertical amplifier and the trigger circuits.

The trigger pick-off amplifier Q523 is one part of a transistor cascode amplifier. The input stage is Q514 and Q524. Normally, the gain of a transistor cascode amplifier is the ratio of R_L to $R_E + R_t$. The gain in this case must be multiplied by 0.5 for the following reason. The signal current is equally divided at the plate of V514B, half of the signal current will flow through the delay line impedance (93 Ω) and the other half through R526 and finally through the load impedance of Q523. The load impedance will be that impedance which is connected to the AC ground. The collector of Q523 is connected to the base of Q543. The impedance we see looking into the base of Q543 is

$$R_{in} = \beta\,(R_E + R_t) \qquad (10)$$

If we choose to neglect the input circuit of the trigger amplifier we see that R_E in this case is R547 6.5 kΩ. A beta of 50 is a close figure to use for Q543, and since $R_E >> R_t$ then,

$$R_{in} = \beta\,R_E$$

$$= 50 \times 6500\,\Omega$$

$$= 325\,\text{k}\Omega$$

This impedance shunts R544 (75 kΩ) and L528 a 1.5 kΩ wire-wound resistor. We may then, for all practical purposes, consider L528 the collector load resistance (R_L); therefore,

$$A_{(v)} = 0.5\left[\frac{R_L}{R_{E(1)} + R_{E(2)} + R_{t(1)} + R_{t(2)}}\right]$$

$$= 0.5\left[\frac{1500}{23 + 23 + 5 + 5}\right]$$

$$= 13.3$$

Q534 is the beam-indicator amplifier. Its function is to drive two neon lamps situated above the CRT on the front panel of the oscilloscope. These neons indicate the position of the trace in a vertical direction. In the quiescent condition the voltage at the junction of R535 and R536 is 287 volts. Both indicator neons, B538 and B539, have 62 volts across them, not enough voltage to strike either neon. (This type of neon has a striking voltage in excess of 68 volts.)

When we apply a negative signal to the vertical input of the oscilloscope, the base of Q524 is driven negative and the base of Q514 moves in a positive direction by a similar amount. Therefore, the current through R530 decreases and the current through R507 increases. The voltage at the emitter of Q534 increases and the voltage at the base of Q534 decreases. As a result, the base-to-emitter junction of Q534 becomes reverse biased and Q534 ceases to conduct. Therefore, the voltage at the junction of R535 and R536 rises towards 350 volts striking neon B539 which indicates trace has shifted down.

R513 and R523 and the DC SHIFT control R502 are thermal-compensation networks. The thermal time constants are long and the visible result appears on the CRT display as a DC shift in trace position after a step function. The DC SHIFT control is adjusted for the best dynamic thermal compensation, typically about 1% tilt.

We will now analyze the output circuits to the right of the delay line, refer to Figure 17. The first thing we must do is to calculate the voltage at the base of Q594 or Q584. The voltage at the junction of R532 and R533 (174 volts) will set the base voltage of Q513 and Q523. Assuming a junction drop of 0.5 volt the voltage at the emitter of Q513 and Q523 will be 173.5 volts. The current through R511 and R526 is 27 milliamps, hence the voltage drop across these resistors will be

$$\frac{90.9 \times 27}{1000}$$

$$\approx 2.5 \text{ volts}$$

therefore, the voltage at the plate of V514A and V514B is

$$173.5 - 2.5 = 171 \text{ volts.}$$

This 171 volts is directly coupled to the base of Q594 and Q584 via the delay line. The voltage at the emitter of both Q594 and Q584 is then 170.5 volts. We will now calculate the current flowing into the emitter of Q594 or Q584. Figure 21 shows a step-by-step approach in solving this problem. The simplest approach is to use Thévenin's Theorem to simplify the resistive network R569, R570, R571 and R572. The result is we have a V_{oc} of +100 volts and a Z_{th} of 900 Ω to the junction of R574 and R576. Therefore, looking from the emitter of either Q594 or Q584 we see an impedance of 13.3 Ω in series with 1800 Ω to +100 volts.

$$I_E = \frac{(170.5 - 100)\,10^3}{1.8 \times 10^3} \text{ mA}$$

$$= \frac{70.5}{1.8}$$

$$= 39 \text{ mA}$$

we now calculate r_e

$$r_e = \frac{26}{I_E} = \frac{26}{39}$$

$$\approx 0.7\ \Omega$$

and to this we add our constant R_r of 4 Ω; therefore,

$$R_t = r_e + R_r \qquad (9)$$

$$= 0.7 + 4.0$$

$$= 4.7\ \Omega\text{'s}$$

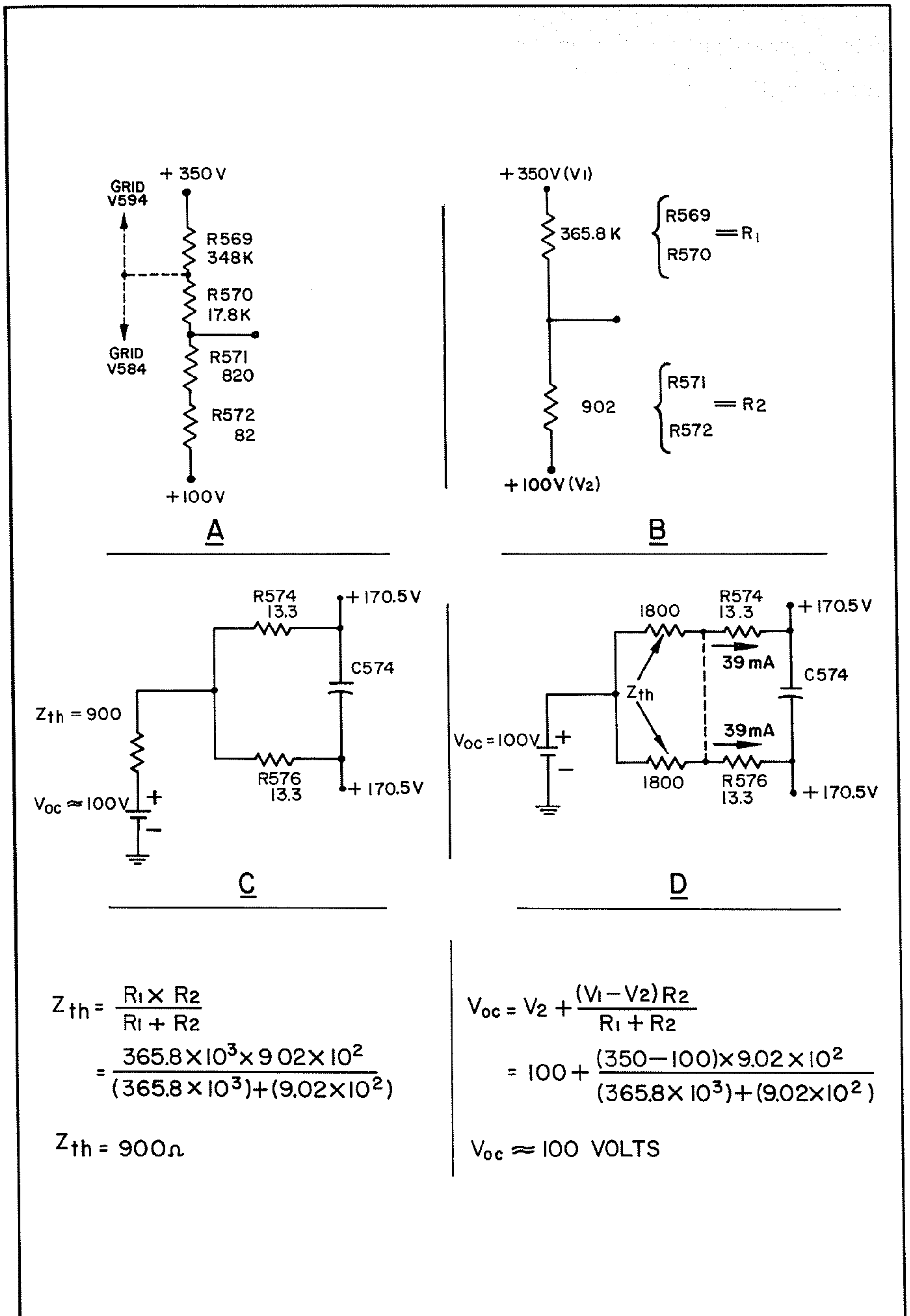


Figure 21. Illustrating the use of Thévenin's Theorem to simplify a network consisting of a voltage source and a resistive network. (A)—The network whose Thévenin equivalent is to be determined. (B)—Determining the equivalent source impedance (Z_{th}) and the equivalent voltage source (V_{oc}). (C)—The Thévenin equivalent network of (A) connected to the junction of R574 and R576. (D)—The equivalent circuit considering Z_{th} as two resistors through which the individual emitter currents will flow.

We have only one point in this circuit (a virtual AC ground point) at which the signal currents will be equal and opposite. That point is the junction of R574 and R576 (13.3 Ω resistors). This fact sets R_E at 13.3 Ω. The purpose of the RC network to the right of R574, R576 is to compensate the high frequencies.

The input impedance we see looking into the base of Q594 or Q584 is

$$R_{in} = \beta\ (R_E + R_t) \qquad (10)$$

A beta of 75 for this type of transistor is a close figure to use for practical purposes. Therefore,

$$R_{in} = 75\ (13.3 + 4.7)\ \Omega\text{'s}$$

$$= 1350\ \Omega$$

The value of R_{in} is part of a resistive network which will terminate the delay line in its correct impedance. Therefore, before we leave this section we must check to see if our value of R_{in} is within practical limits. Figure 22 shows a progressive breakdown of this network.

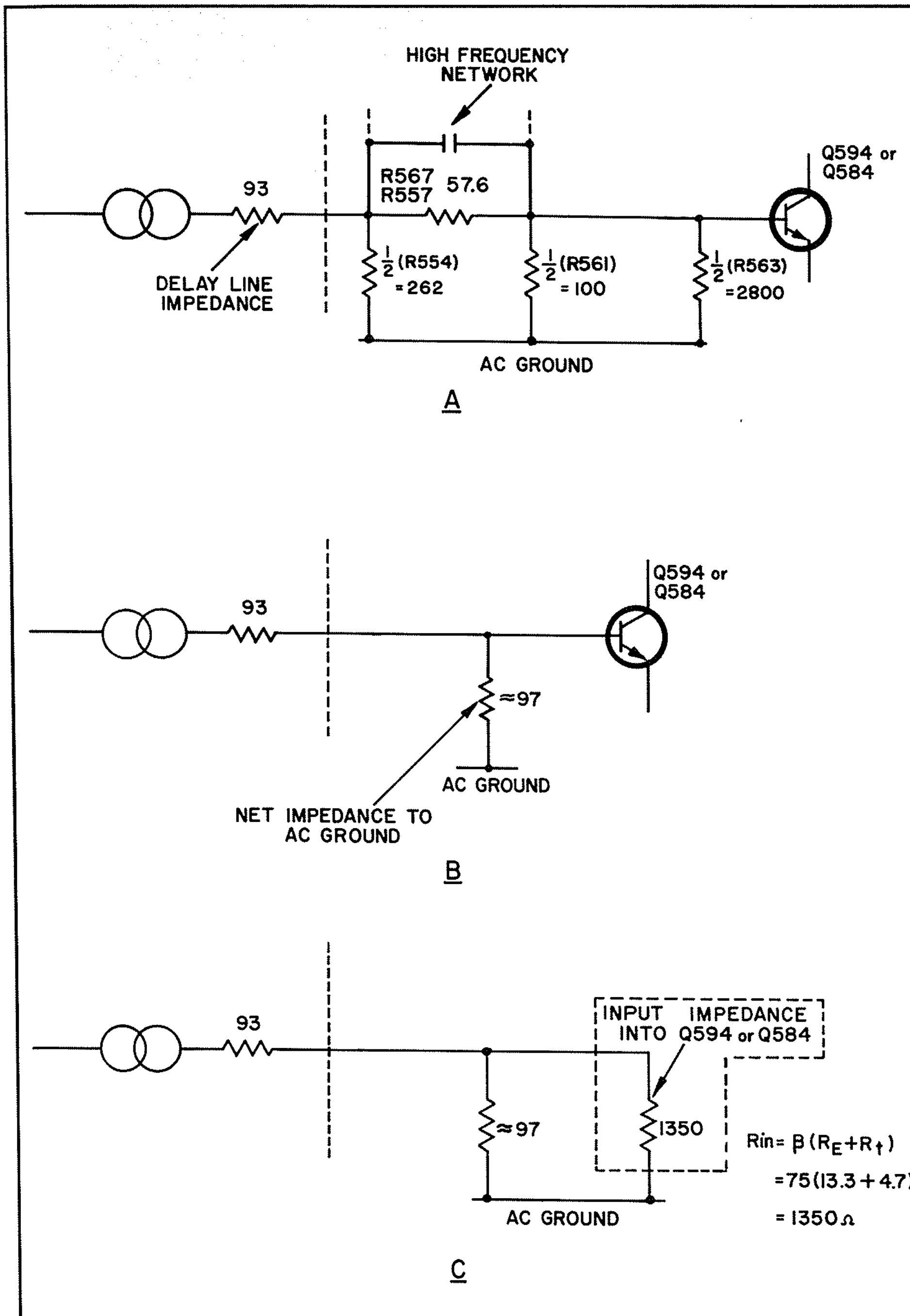


Figure 22. Illustrating a step-by-step approach for analyzing the circuit between the output of the delay line and the base of Q594 or Q584. (A)—The circuit between the output of the delay line and the base of Q594 or Q584. (B)—The net impedance of the circuit in (A) to the AC ground between the output of the delay line and the base of Q594 or Q584. (C)—The input impedance into Q594 or Q584 and the net impedance, as shown in (B), providing the terminating impedance for the delay line.

This network will induce a loss between the two stages. The signal is reduced in amplitude by a factor of 0.64 because of the voltage divider network consisting of 57.6 Ω and the parallel combination of 100 Ω, 2800 Ω, and the input impedance into Q594 or Q584.

The gain of the output stage is

$$A_{(v)} = \left[\frac{R_{L(1)} + R_{L(2)}}{R_{E(1)} + R_{E(2)} + R_{t(1)} + R_{t(2)}}\right]\eta$$

$$= \left[\frac{1100 + 1100}{13.3 + 13.3 + 4.7 + 4.7}\right]\eta$$

$$= \left[\frac{2200}{36}\right]\eta$$

$$= 61\ \eta$$

You recall that the gain equation of a hydrid cascode amplifier (refer part 2, "The Vacuum Tube Amplifier," Service Scope #43, April 1967) must be multiplied by the plate efficiency factor (η) if the vacuum tube is not a triode. The plate efficiency factor (η) normally varies from between 0.7 to 0.9. In this case (η) is approximately 0.9 - 0.88 to be exact. So finally,

$$A_{(v)} = 61 \times \frac{9}{10}$$

$$= 54.9$$

The gain of the complete Type 545B vertical amplifier is

$$A_{(v)}\ (\text{total}) = 54.9 \times 1.68 \times 0.64$$

$$= 59$$

Summary

This brings to a close this series of three articles dealing with a practical approach to transistor and vacuum-tube amplifiers. This approach has been offered as a direct method of trouble shooting and understanding circuits. There are limitations as to its application as we have seen. However, these limitations do not impair the practical approach we must apply to our everyday maintenance and trouble shooting problems.

The Tektronix Type 454 is an advanced new portable oscilloscope with DC-to-150 MHz bandwidth and 2.4-ns risetime performance where you use it—at the probe tip. It is designed to let you make convenient measurements of fast-rise pulses and high-frequency signals previously outside the range of conventional oscilloscopes.

The Type 454 is a complete instrument package with dual-trace vertical, high-performance triggering, 5-ns/div delayed sweep and solid-state design, all in a rugged 31-lb. instrument. You also can make 1 mV/div single-trace measurements and 5 mV/div X-Y measurements with the Type 454.

The 2.4-ns risetime and DC-to-150 MHz bandwidth are specified at the tip of the new miniature P6047 10X Attenuator Probe. The dual-trace amplifiers provide the following capabilities with or without probes:

*Deflection Factor**	*Risetime*	*Bandwidth*
20 mV to 10 V/div	2.4 ns	DC to 150 MHz
10 mV/div	3.5 ns	DC to 100 MHz
5 mV/div	5.9 ns	DC to 60 MHz

*Front panel reading. Deflection factor with P6047 is 10X panel reading.

The Type 454 features a new CRT with distributed vertical deflection plates and a 14-kV accelerating potential. It has a 6 by 10 div (0.8 cm/div) viewing area, a bright P-31 phosphor and an illuminated, no-parallax, internal graticule. The Type C-30 and the New Type C-40 (high writing speed) cameras mount directly on the oscilloscope.

The instrument can trigger to above 150 MHz internally, and provides 5-ns/div sweep speeds in either normal or delayed sweep operation. The calibrated sweep range is from 50 ns/div to 5 s/div, extending to 5 ns/div with the X10 magnifier. Calibrated delay range is from 1 μs to 50 seconds.

The Type 454 is designed to be carried and has the rugged environmental characteristics required of a portable instrument. A rackmount, the 7-inch-high Type R454 oscilloscope, is available with the same high performance features. Also available is the new Type 200-1 Scope-Mobile® Cart.

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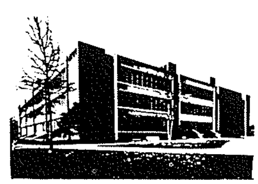

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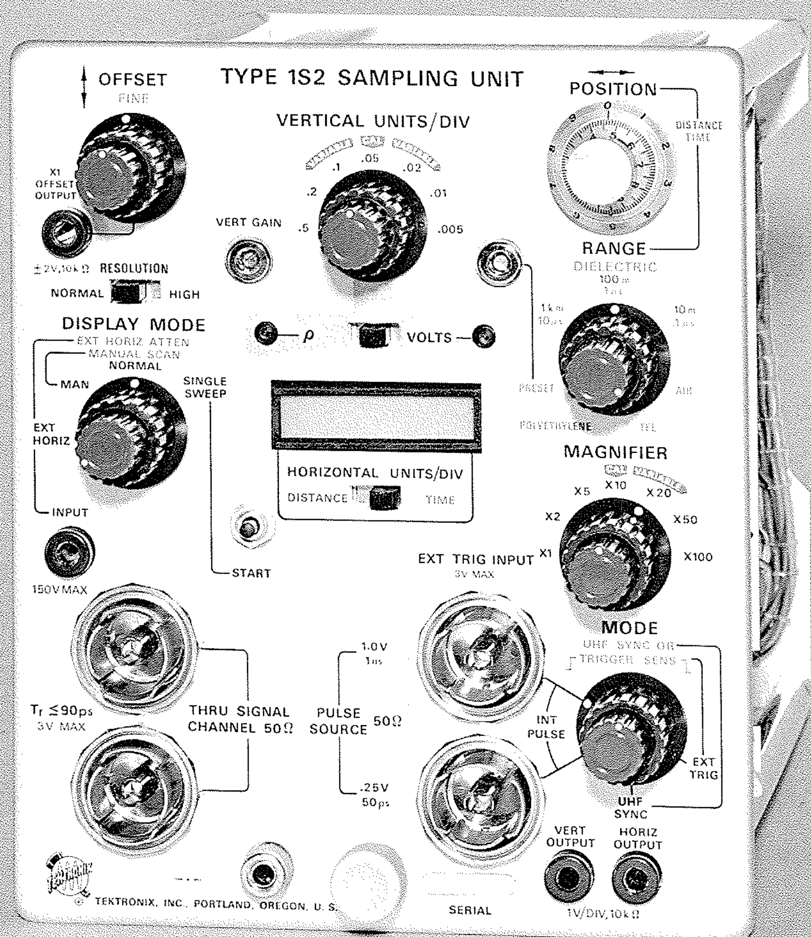


A Discussion of Time Domain Reflectometry Theory and Coaxial Cable Testing.

TIME-DOMAIN REFLECTOMETRY THEORY AND THE TESTING OF COAXIAL TRANSMISSION LINES

INTRODUCTION:

Maintaining the fidelity of electronic signals that of necessity have to be transmitted from point to point is of primary concern to those that design, build and maintain electronic equipment. The simple, inexpensive coaxial transmission line is perhaps the most common method used to accomplish this task. The techniques for determining transmission line performance vary from simple visual inspection to elaborate instrumentation set-ups that require a great deal of skill and time. The availability of instruments such as the Tektronix Type 1S2 TDR Plug-In Unit have simplified the testing of transmission line performance.

This article begins with a comparison between two methods of testing transmission lines - Sinewave testing and Voltage step-function testing. The Sinewave testing method is known as Frequency-Domain Reflectometry (FDR) and the Voltage step-function method is known as Time-Domain Reflectometry (TDR). The FDR-TDR comparison is followed by a basic description of TDR testing principles; reflections from capacitors and inductors; reflections from resistive discontinuities; coaxial-cable response to a step signal; and finally, special applications.

The waveforms illustrated throughout this article were taken with a C-12 Camera using a Type 547 Oscilloscope and the Type 1S2 TDR Plug-In. The Type 1S2 Plug-In converts any Tektronix 530, 540, 550-Series Oscilloscope to a TDR measurement system.

FDR-TDR COMPARISON

Frequency domain reflectometers, the slotted line and bridges, drive and observe the input terminals of a transmission line as a function of frequency. They do not locate discontinuities on a distance basis. As a result, measurement techniques and the unique advantages of such devices differ from those of TDR.

A pure resistance measured by either time domain or frequency domain devices will appear as an infinitely long lossless transmission line. Thus, a perfectly terminated short length of lossless line will yield the same information to both kinds of testing, and neither test system can locate the termination. However, if the termination includes a small inductive or capacitive reactance, both systems will indicate its presence, but the TDR system will show where in the line the reactance is located.

The following comparisons of TDR and frequency domain (FDR) devices are supported by four specific examples and illustrations.

1. FDR measures Standing Wave Ratio (SWR) directly, but a TDR display can speed FDR testing by locating resonant frequencies of resonant networks prior to FDR testing.

2. TDR locates discrete discontinuities and permits analysis of their value. But FDR will indicate two different resonant discontinuities which may be located very close together when TDR may not.

3. FDR measures an antenna standing wave ratio directly while TDR will not. But TDR will locate faults more quickly and identify the type of fault more rapidly than will FDR, should a change in SWR indicate problems. The time domain display will validate a transmission line to an antenna, while frequency domain reflectometry cannot, unless the antenna is disconnected and the transmission line terminated.

4. TDR can locate small changes in transmission line surge impedance (such as a too-tight clamp holding a flexible line) while FDR will show whether or not the SWR is acceptable.

5. Both test systems will quantitatively evaluate single discrete reactances, with higher degree of accuracy possible with FD.

6. Both TDR and FDR have advantages, each being very valuable in its own way. Thus, the two systems complement each other and both aid where observations and measurements are required.

TDR vs FDR Measurements

A one pF discrete capacitor inserted in parallel with a transmission line will produce almost no TDR indication if the step pulse has a risetime of 1 nanosecond. The same capacitor will produce a significant reflection if the step pulse has a risetime of 150 picoseconds. A FDR test will produce a large SWR at the series resonant frequency determined by the capitance and its lead inductance. Such a discontinuity would require considerable time for proper FDR testing due to the numerous frequency test points, but with a fast rise TDR system the capacitance and resonant frequency can be quickly determined.

Fig 1 shows waveforms and SWR curves of first a single capacitor and then two capacitors inserted in parallel with a transmission line. Note that the FDR measurement on the right side of the figure plainly shows the two resonant circuits of the two closely spaced small capacitors, while the TDR display at the left shows two resonant frequencies, but not in a manner to permit separation of the two capacitors.

The single capacitor of this example was made of 1/4 inch wide strip copper, 5/8 inch long, with one end soldered to the side of a component insertion unit (Tektronix Part No. 017-0030-00) and the other end near the center conductor. The insertion unit was modified to have a continuous center conductor using three inner transition pieces (Tektronix Part No. 358-0175-00). One of the inner transition pieces was shortened to fit between the two mounted end pieces, and then soldered in place. The second capacitor (resonant at 2.1 GHz) was a 0.5 to 1.5 pF piston trimmer with a total lead length of about 5/16 inch, and it was adjusted to about 1.2 pF. The piston capacitor was soldered in place in parallel with the strip copper capacitor about 1/8 inch away. It is obvious from both testing methods that neither capacitor was critically damped by the characteristic impedance of the transmission line. The physical and equivalent circuit of the single shunt capacitor is shown in Fig 2. The single capacitor test was made with a shield in place completely covering both openings.

Fig 3 shows the ability of TDR to locate an off-impedance point in a transmission line, and quickly resolve its value. The same through-connected insertion unit used in example number 1 was tested without any component inserted in it. The shield was in place for both TDR and FDR testing.

The TDR display of Fig 3 shows the increased surge impedance due to the increased diameter of the outer conductor at the two cutout access slots. Such a TDR display will permit rather rapid correction to be made to the center conductor diameter if one desires to make a truly constant impedance through the length of the insertion unit.

The SWR curve shows some changes from a constant impedance transmission line, but does not help to locate an aberration if it is inside a continuous piece of cable. Either FDR or TDR would help one to make the unit have a constant impedance if such a unit were being designed.

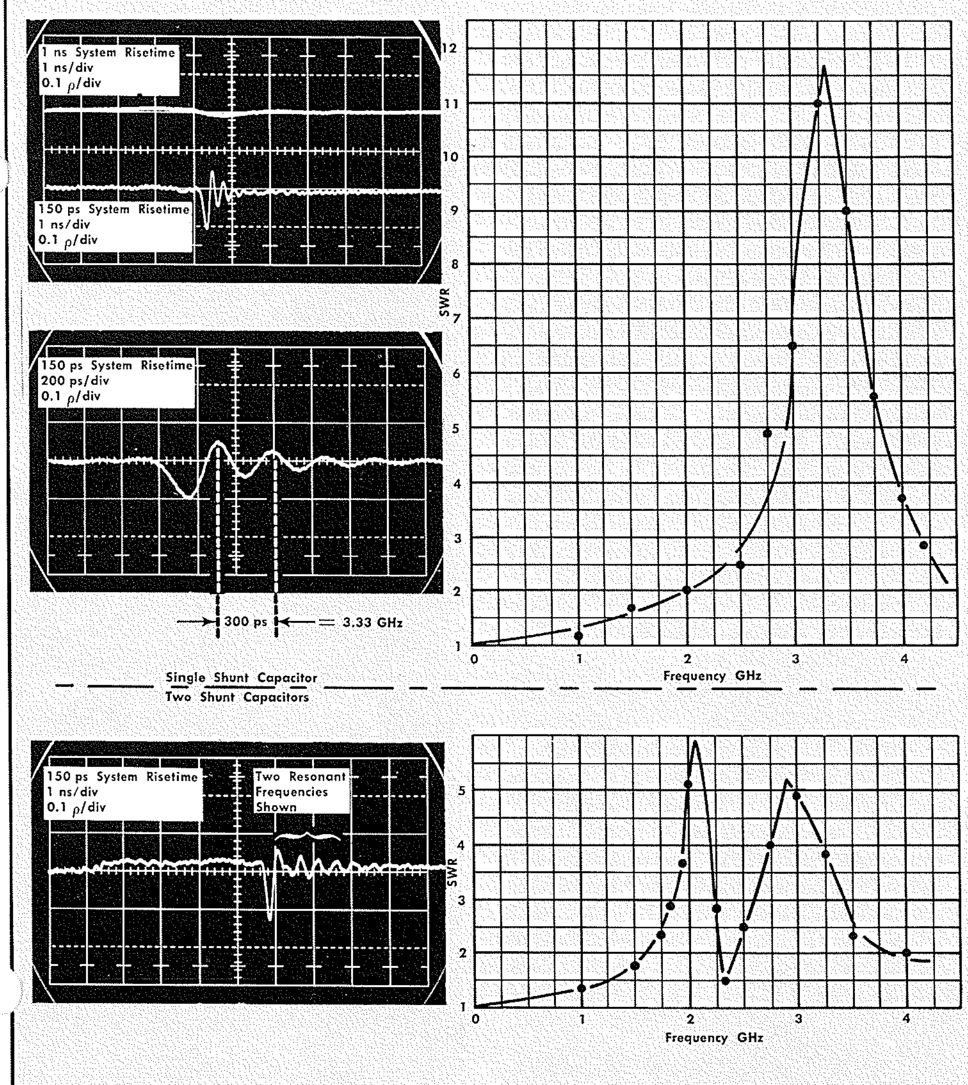


Fig 1. Two examples of discrete shunt capacitors.

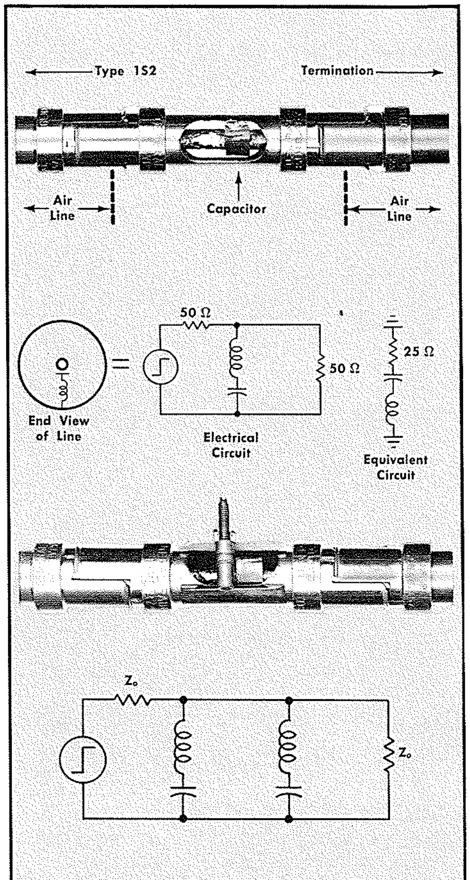


Fig 2. Capacitors measured in Fig 1.

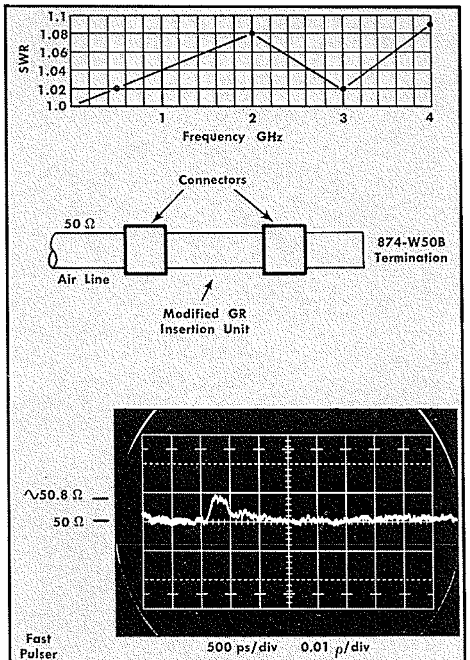


Fig 3. Modified (through-connected) Tektronix Insertion unit for testing small components in parallel with 50 Ω line.

Fig 4 shows two TDR and two SWR plots of a simple dipole antenna. The TDR waveforms at the left were photographed first, quickly locating the two radiating resonant frequencies and permitting a saving in time for the FDR testing. The SWR curves permit a direct evaluation of the antenna radiation resistance ($\frac{R_L}{Z_0} = \frac{V_{max}}{V_{min}}$ if R_L is purely resistive), while the TDR display tells only the transmission line quality and the radiating resonant frequencies of the non-shorting type antenna. An antenna design engineer could use the SWR data and FDR test equipment to test a compensating network to be located at the antenna to minimize standing waves in the transmission line. The TDR system cannot be used for such design assistance.

Fig 5 shows both TDR and FDR tests of a General Radio Type 874-K series blocking capacitor. The upper TDR display permits direct calculation of the series capacitance, in this case approximately 6.2 nanofarads ($0.0062\mu F$).

The SWR curve shows that the series capacitor does not upset the transmission line significantly except for low frequencies. The middle TDR waveform shows the change in surge impedance due to the physical shape of the series capacitor. Note that the disc capacitor reduces the transmission line surge impedance to approximately 49 ohms for onl a very short period of time. The same display also permits the precise location of adjacent discontinuities that affect the high frequency performance. The combined TDR and FDR data tells more about the series capacitor unit than either testing method does alone.

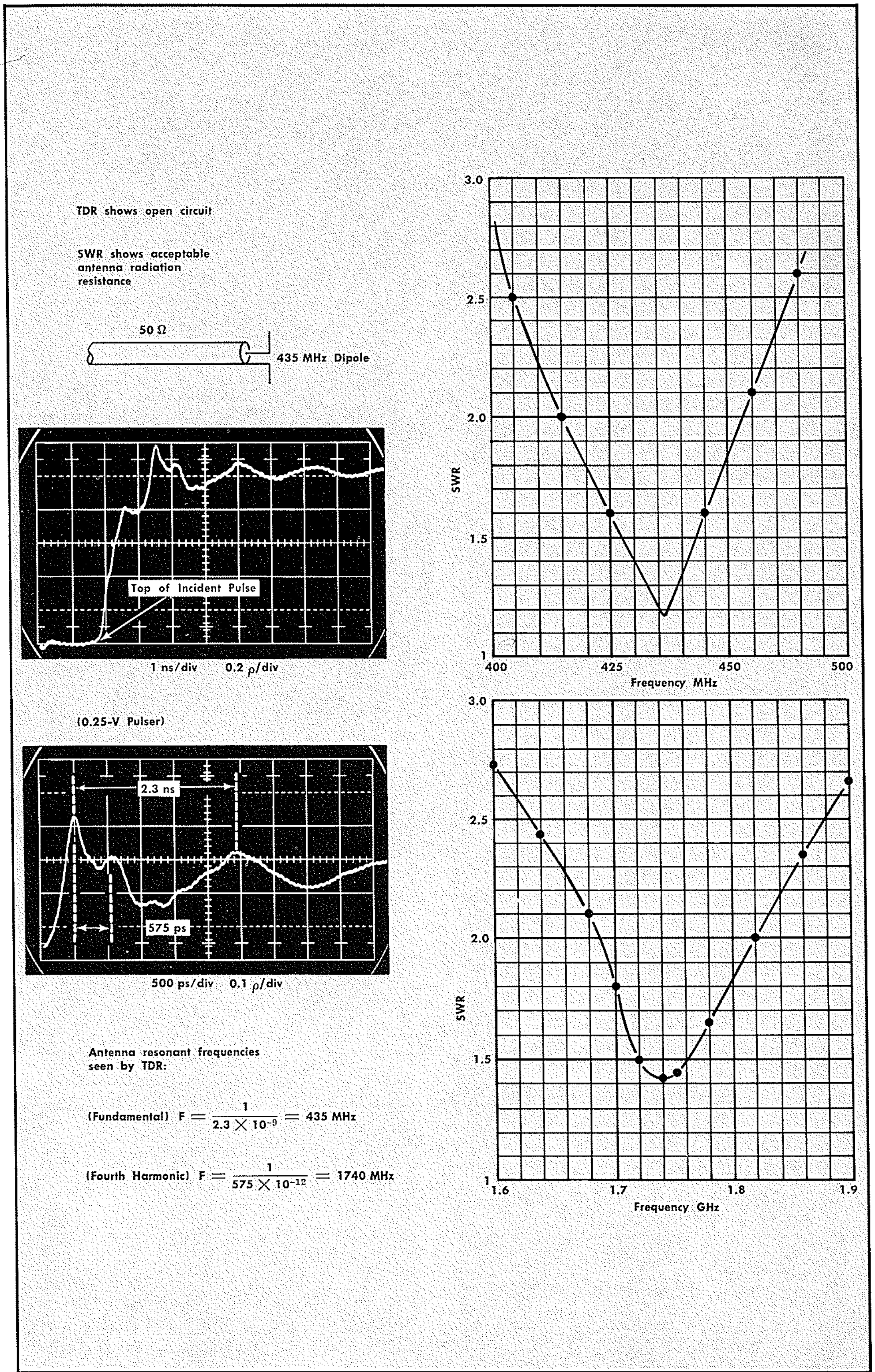


Fig 4. Two plots of 435 MHz dipole antenna.

BASIC APPROACH TO TDR

Time Domain Reflectiometry can be understood most easily if its operation is first compared with a DC circuit.

DC Analogy

Fig 6 shows three simple circuits that can be related to transmission lines and TDR. Fig 6A is the diagram of an ordinary resistance voltage divider, where the voltage across

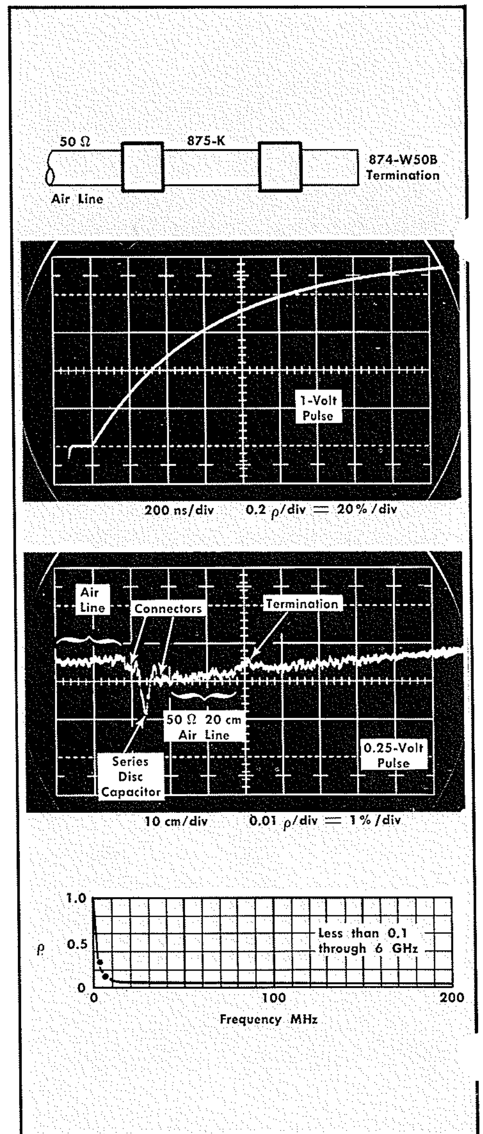


Fig 5. Series blocking capacitor: General Radio Type 874-K.

R_2 is $E_{R2} = \frac{R_2}{R_1 + R_2} \times E$ of the battery. (1)

Fig 6B substitutes R_{line} (or Z_o) for R_2, and substitutes R_g (generator resistance) for R_1. It is assumed the battery has zero internal resistance and that R_g is an inserted series generator resistance. If the battery is 1 volt and if $R_g = R_{line}$, then a voltmeter across R_{line} will indicate 0.5 volt when the switch is closed.

Fig 6C indicates a pair of zero resistance wires of same length physically connecting R_{line} to the battery and switch. A voltmeter across R_{line} will still indicate 0.5 volt when the switch is closed.

Adding the Time Dimension

Fig 7 substitutes a step generator for the battery and switch of Fig 6. The generator has zero source resistance so R_g is again added in series with the generator. The generator and R_g drive a finite length transmission line that has a characteristic impedance of Z_o. The transmission line has output terminals that permit connecting a load R_L. An oscilloscope voltmeter measures the voltage signal(s) at the input end of the transmission line.

Assume that no load resistance is connected to the transmission line output terminals ($R_L = \infty$) and that $R_g = Z_o$ (Z_o acts exactly as if it were the DC resistor R_{line} of Fig 6). As the zero impedance step generator applies its 1-volt step signal to R_g, the oscilloscope voltmeter indicates 0.5 volt. The oscilloscope voltmeter will continue to indicate a 0.5 volt signal until the wave has traveled down the line to the open end, doubled in amplitude due to no current into $R_L = \infty$, and reflected back to the generator end of the line. The oscilloscope finally indicates a signal of 1 volt after the measurable period of time required for the step signal to travel down and back the finite length of open ended transmission line.

Reflection Signal Amplitudes

Fig 8 shows TDR oscilloscope (voltmeter) displays related to the value of R_L vs the value of the transmission line Z_o. Apply resistance values of 50 Ω to R_g and Z_o, and 75 Ω to R_L of Fig 7. By formula (1), the oscilloscope display of the reflection amplitude will be 0.6 volt. The actual reflection, however, is only 0.1 volt added to the 0.5-volt incident step.

Reflection Coefficient

A somewhat more convenient method of handling signal reflections than has just been suggested, is to consider the reflection as having been added to or subtracted from the incident pulse. Thus the reflection amplitude is not measured from zero volts, but is referenced to the incident signal amplitude. This permits establishing a ratio between the incident and reflected signals which is called the reflection coefficient, rho (ρ). The value of ρ is simply the reflected pulse amplitude (the display total amplitude minus the incident pulse amplitude) divided by the incident pulse amplitude. Fig. 9 shows the two parts of the display appropriately labeled to identify the incident and reflected signals.

When $\rho = 0$, the transmission line is terminated in a resistance equal to its characteristic impedance Z_o. If the line is terminated in $R_L > Z_o$, then ρ is positive. If the line is terminated in $R_L < Z_o$, then ρ is negative. The dependence of ρ on the transmission line load is

$$\rho = \frac{R_L - Z_o}{R_L + Z_o} \qquad (2)$$

If ρ is known, R_L can be found by rearranging formula (2);

$$R_L = Z_o \left(\frac{1 - \rho}{1 + \rho}\right) \qquad (3)$$

Formula (3) applies to any display that results from a purely resistive load. The load shown in Fig 9 is assumed to be at the end of a lossless coaxial transmission line.

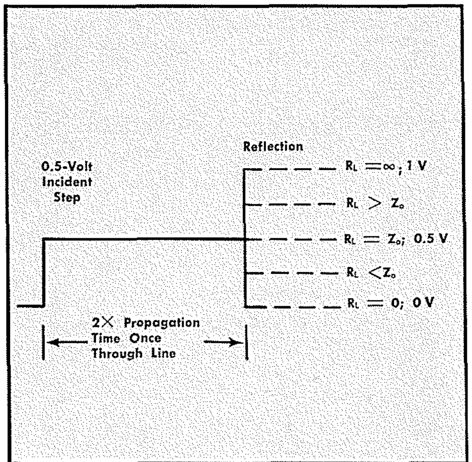


Fig 8. Oscilloscope voltmeter displays for circuit of Fig 7, dependent upon value of R_L vs. Z_o.

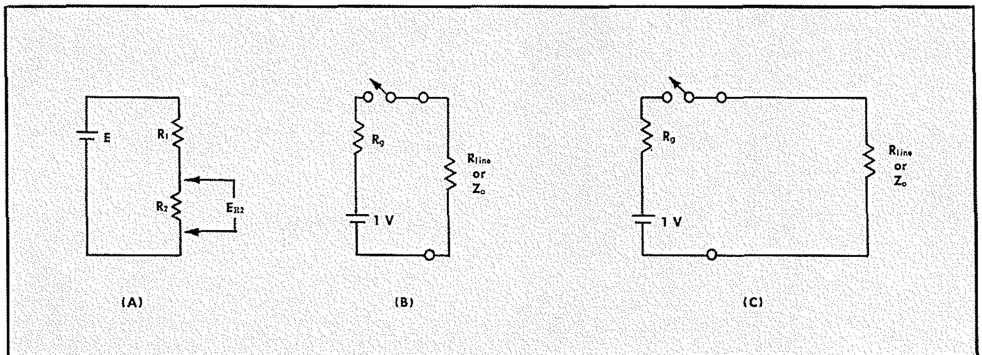


Fig. 6. Circuits showing DC analogy of TDR.

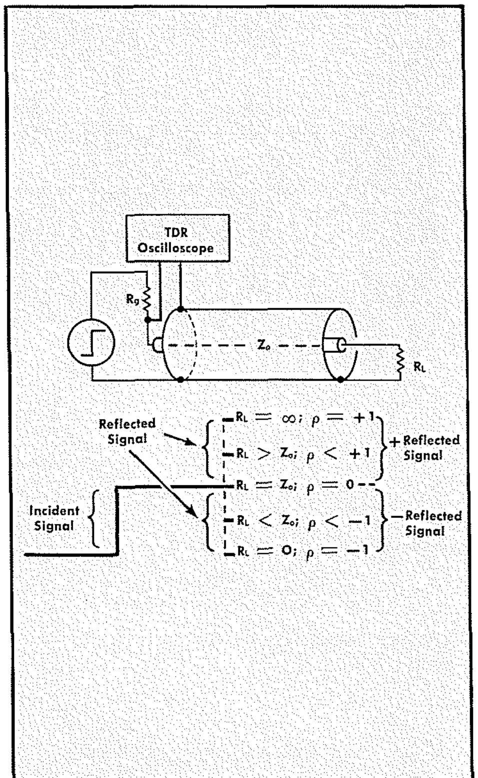


Fig 9. TDR oscilloscope displays for various values of R_L vs Z_o.

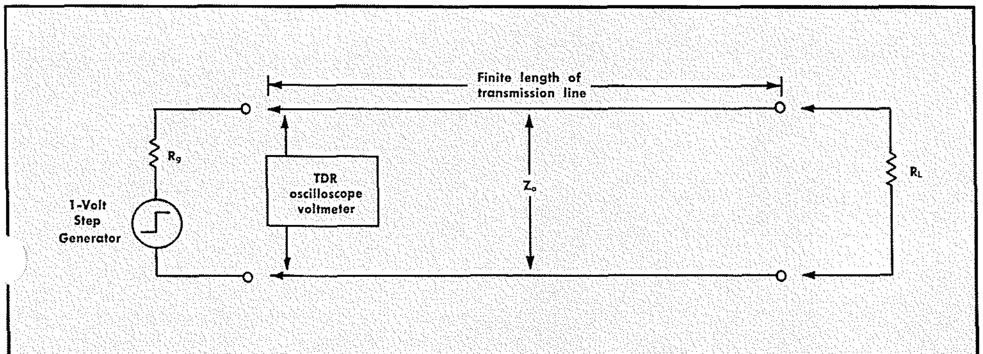


Fig 7. Adding the time dimension to the circuit of Fig 6.

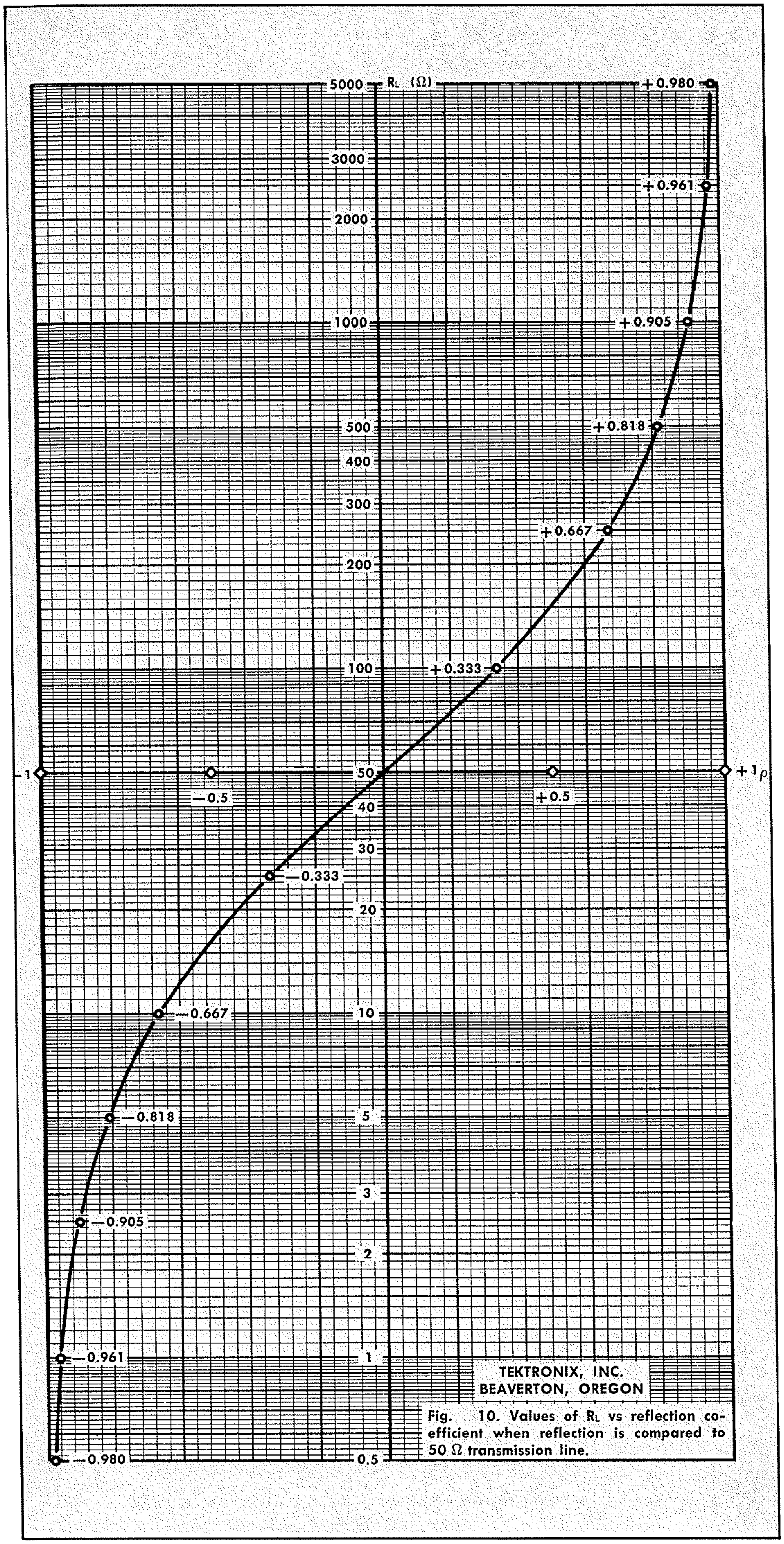


Fig. 10. Values of R_L vs reflection coefficient when reflection is compared to 50 Ω transmission line.

Substituting 50 Ω for Z_o in formula (3), calculations for small values of ρ show that each division of reflected signal is approximately equal to a certain number of ohms. Table 1 lists the ohms per division for vertical deflection factors of 0.005 ρ, 0.01 ρ and 0.02 ρ. Or, for R_L values near 50 Ω, you may use the approximation formula

$$R_L \approx 50 + 100\,\rho.$$

This approximation formula has an error of $\leq$ 2.2% for absolute values of $\rho \leq 0.1$ and an error of $\leq$ 8% for absolute values of $\rho \leq 0.2$.

R_L for reflections with ρ up to essentially +1 or −1 can be quickly determined using the graph of Fig 10. Fig 10 is based upon a transmission line surge impedance of 50 Ω just prior to the discontinuity that causes the reflection signal. The graph of Fig 10 may be photographically reproduced without special permission of Tektronix.

TABLE 1

R_L Approximations For Reflection Coefficients of 0.005, 0.01 and 0.02 Related to a 50 Ω Transmission Line

ρ/div	Ω/div	Error/div
0.005	½	~0.016 Ω
0.01	1	~0.066 Ω
0.02	2	~0.2 Ω

REFLECTIONS FROM CAPACITORS AND INDUCTORS

Contrary to frequency domain measurements, TDR response to a reactance is only momentary. Thus either an inductor or a capacitor located in a transmission line will give only a short duration response to the TDR incident pulse. Analysis of large reactances is relatively simple and makes use of time-constant information contained in the reflection display. Small reactances are not so simple to evaluate quantitatively, so will be treated separately.

Large Reactances

The difference between a "large" and a "small" reactance is not a fixed value of capacitance or inductance, but is instead related to the TDR display. If the displayed reflection includes a definite exponential curve that lasts long enough for one time constant to be determined, the reactance is considered "large".

Discrete (single) capacitors connected in series or parallel with a transmission line start to charge at the instant the incident pulse arrives. Inductors start to conduct current at the arrival of the incident pulse. Both forms of reactance cause an exponentially changing reflection to be sent back to the TDR unit. When a capacitor is fully charged, the TDR unit indicates an open circuit. When an inductor is fully "charged" (current through it has reached its stable state), the TDR unit indicates a short circuit. The TDR unit will indicate an inductor's series DC resistance if its value is significant in relation to Z_o. The general form of reflection and long term effect upon the TDR display by both inductors and capacitors is listed in Table 2 and Table 3.

Finding One Time Constant

In practice, TDR reactance displays usually contain aberrations of the desired pure exponential reflection. Such aberrations prevent finding the normal 63% one time-constant point of the curve accurately. (The aberrations are due to either the environment around the reactance, i.e. stray inductance in series with a capacitor, or stray capacitance in parallel with an inductor, or secondary system reflections.) However, accurate time-constant information can be obtained from less than a complete exponential curve. The principle used requires that a "clean" portion of the display must exist. The "clean" portion used must include the right-hand "end" of the displayed curve (a capacitor is then fully charged, or an inductor current has stopped changing). The "end" of the curve will appear on the display to be parallel to a horizontally scribed graticule line. Thus, aberrations that exist at the beginning of the curve can be ignored.

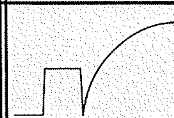
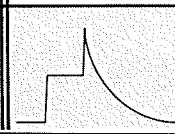
Fig 11 shows the first example of obtaining valid time-constant information from less than a full 100% exponential curve. The technique is to choose any "clean" portion of the display that includes the "end" of the exponential curve and find the half-amplitude point. The time duration from the beginning of any new 100% curve section to its 50% amplitude point is always equal to 69.3% of one time constant. Thus, the time duration for a 50% change divided by 0.693 is equal to one time constant.

Fig 11 shows the TDR displays of a capacitor placed in series with a transmission line center conductor (2 Z_o environment). This picture and the other waveform pictures shown in this article were taken with a Tektronix C-27 Camera mounted on a Tektronix Type 547 Oscilloscope with a Tektronix Type 1S2 Reflectometer and Wideband Sampling Plug-In Unit. Fig 11A waveforms comprise a double exposure with the left curve taken while the Type 1S2 RESOLUTION switch was at NORMAL and the right curve taken when the switch was at HIGH. Both curves give sufficient information to measure one time constant. Note that the top of the incident pulse is indefinite (in the displays) due to the sweep rate and short length of cable used between the Type 1S2 and the capacitor. Such a display does not have a definite beginning of the normal 100% exponential curve. This prevents 63% of the total curve from being read directly from the display. (It is also quite possible for lead inductance to cause a capacitor to ring. When a TDR display shows capacitor ringing, the ringing can sometimes be reduced by: 1. using the slower 1-Volt pulser, and/or 2. changing the transmission line environment to place a lower value Z_o in parallel with the capacitor.)

The double exposure of Fig 11B shows a full exponential curve beginning in the vicinity of 1 division from the graticule bottom. Then the same curve has been time-expanded for easier reading. The indefinite beginning of the 500 ns/DIV exponential curve prevents finding one time constant by measuring the time of 63% of the total curve amplitude. The new arbitrarily chosen 100% amplitude portion of the curve begins at the graticule center horizontal line and extends (off the right of the graticule) to the top graticule line. Three divisions were chosen for the new 100% exponential curve, with the 100% and 50% points marked. Then, dividing the time for the 50% amplitude change by 0.693 gives a total one time-constant time value of 650 ns. Since the equivalent circuit shows 2 Z_o in series with the capacitor, its value is found by formula (4) (Table 4) to be 6.5 nanofarads.

TABLE 3

Single Capacitor or Inductor TDR Displays when Connected Across End of Transmission Line

Reactance	Display	Line Impedance at Reactance
CAPACITOR		Z_o
INDUCTOR		Z_o

Large Capacitors

The difference between a "large" and a "small" capacitor is not a fixed value of capacitance, but is instead related to the TDR display. If the display includes a definite exponential curve that lasts long enough to permit one RC time constant to be determined, the capacitor value can be found by using a normal RC time-contsant formula). The actual formula varies according to the equivalent circuit in which the capacitor is located. Table 4 lists the possible configurations and their related formulae.

The first example of "large" capacitance measurement was given under the previous heading Finding One Time Constant. The large value of capacitor used is easy to measure and usually causes only one aberration to the exponential curve. That aberration is the indefinite curve beginning.

Moving A Reflection Aberration

When testing small capacitors that still produce a usable exponential curve, it may be difficult to get accurate time-constant data when there are reflections within the system.

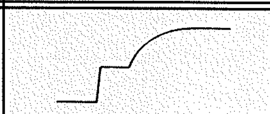
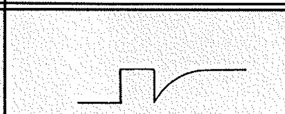
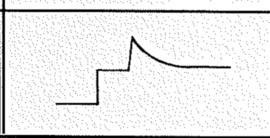
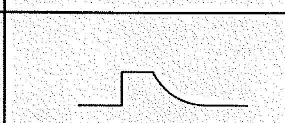
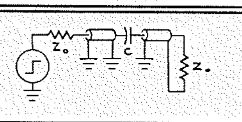
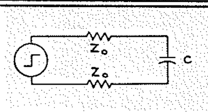
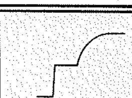
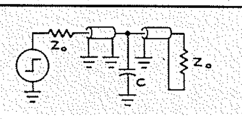
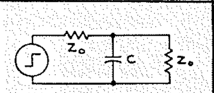
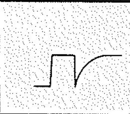
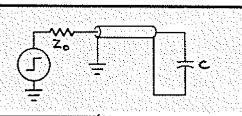
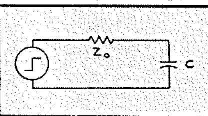
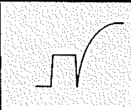
Reactance	In Series with Line	In Parallel with Line	Line Impedance at Reactance
CAPACITOR			SERIES: 2 Z_o PARALLEL: $\frac{Z_o}{2}$
INDUCTOR			SERIES: 2 Z_o PARALLEL: $\frac{Z_o}{2}$

TABLE 2 Single Capacitor or Inductor TDR Displays Related to Terminated Transmission Lines.

Circuit		Equivalent Circuit	Formula	Display
Series with terminated line			$C = \frac{1}{2}\frac{TC}{Z_o}$ (4)	
Parallel with terminated line			$C = \frac{1\ TC}{Z_o/2}$ (5)	
Across line end			$C = \frac{1\ TC}{Z_o}$ (6)	

Where C = Farads; TC = Time Constant; Z_o = Line Surge Impedance.

TABLE 4 "Large" Capacitor Circuits and Formulae.

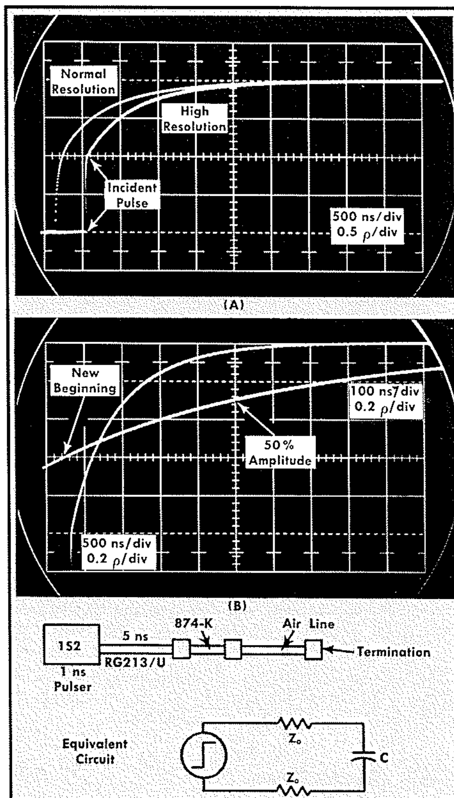


Fig 11. Exponential curves and circuit of 6.5 nF capacitor in series with terminated transmission line.

For example, a 100 pF discap was soldered into a General Ratio Radiating Line section (Fig 12). The 1-Volt pulser was used; re-reflections from the pulser distort the exponential curve at the arrow of Fig 12A. The re-reflection is moved to the right just outside the time window by placing a 20 nsec signal delay RG213/U cable between the pulser and the sampler. The acceptable waveform is shown in Fig 12B. Fig. 12C is a double exposure that shows first how the "end" of the exponential curve is set to a graticule line. Then the display is time expanded to 500 ps/DIV (leaving the vertical position as adjusted) and the new arbitrary 100% exponential curve is chosen and marked. The capacitor's value taken from the time expanded curve of Fig 12C and using formula (5) is 104 pF ($1.8 \times 10^{-9}/0.693 \div 25 = 1.04 \times 10^{-10} = 104$ pF). Note that the vertical ρ factor was changed for Fig 12C in order to make the time constant measurement from a clean section of the curve near its end.

Large Inductors

The difference between "large" and "small" inductors follows the same general display limits as large or small capacitors. A "small" inductor in series with a transmission line center conductor will give a display that does not permit normal time-constant analysis. The same inductor in parallel with a terminated transmission line may give a display that does allow normal time-constant analysis.

Ringing in the exponential TDR display is often observed when measuring inductors. It is usually caused by distributed capacitance across the coil that has not been adequately damped by transmission line surge impedance. Since an inductor with stray capacitance will ring unless adequately damped, an inductor in parallel with a transmission line ($Z_o/2$ environment) will be less likely to ring than the same inductor in series with a line ($2\ Z_o$ environment).

Fig 13 shows waveforms taken of the reflections from a seven turn ⅜ inch diameter coil. The coil was connected across the end of a 50 Ω transmission line (Z_o environment). Fig 13A was made using the Type 1S2 0.25-Volt fast pulser at High Resolution. The ringing makes it impossible to obtain an accurate time constant measurement from the display. Fig 13B was made using the Type 1S2 1-Volt pulser at Normal Resolution. Here the slower risetime incident pulse does not excite the ringing, and in addition the time averaging of fast changes by Normal-Resolution operation permits a time constant to be measured. Ringing could also have been reduced by a $Z_o/2$ environment by placing a termination across the inductor, or placing the inductor at a convenient mid point of a long line.

The triple exposure of Fig 13B includes three curves: #1, the total reflected signal at 10ns/div and 0.5 ρ/div; #2, increased vertical deflection and the exponential-curve end positioned to be one division below the graticule center horizontal line; and #3, the #2 curve time expanded to 1 ns/div for measurement of the L/R time constant. The new 100% to 50% amplitude time duration of curve #3 is shown as 3¾ ns. $3.75/0.693 = 5.41$ns for 1 time constant. Since the coil is at the end of a 50 Ω transmission line, the inductance is calculated by formula (9) of Table 5 to be 270.5 nH ($L = 50 \times (5.41 \times 10^{-9}) = 2.705 \times 10^{-7} = 270.5$ nH).

Small Reactances

"Small" reactances are here defined as series-connected inductors and shunt-connected capacitors that cause TDR reflections without apparent time-constant reaction to the incident pulse. Some small reactances are capable of being "charged" (capacitor voltage is stable; inductor current is stable) at a rate faster than the 0.25-Volt pulser incident pulse rate of rise. If the TDR display has no exponential section, normal RC and L/R calculations cannot be made. All small reactances generate TDR reflections with less than $+1\ \rho$ or $-1\ \rho$.

Small discrete capacitors with leads always include stray series inductance of a significant amount. Fig. 1 and associated discussion is an example of such a capacitor with inductive leads. Small shunt capacitors without leads may be produced by either an increase in a coaxial cable center conductor diameter or a reduction of its outer conductor diameter. Leadless capacitors are sometimes treated as a small reduction in Z_o rather than as a capacitor. Usually, such small capacitors are considered capacitance when the section of reduced Z_o line is so short physically that no level portion can be seen in the TDR display.

Small series inductors rarely have sufficient parallel (stray) capacitance to be significant in the TDR display. However, the coaxial environment around such a small inductor does affect the TDR display. Small series inductors without capacitive strays are sometimes caused by changes in diameter of a coaxial cable: decreased center conductor diameter, or increased outer conductor diameter. This form of inductor is usually treated as a small increase in Z_o rather than as an inductor. Usually, such inductors are considered to be inductance when the section of increased Z_o line is so short physically that no level portion can be seen in the TDR display.

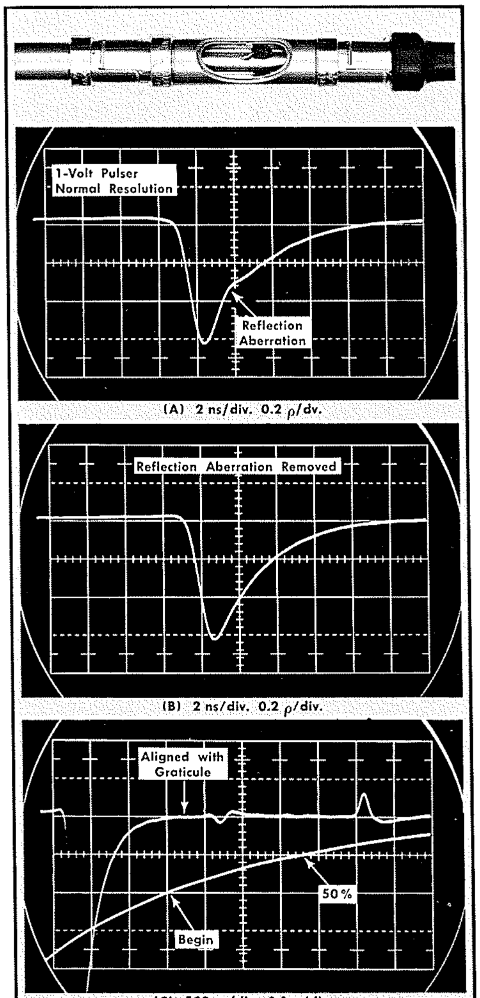


Fig 12. Example of moving display reflection aberrations to obtain a "clean" exponential curve.

Assumptions that Permit Analysis of Small Reactances

The usual TDR system does not have the required characteristics for accurately measuring small reactances. Yet small reactances can be measured provided the following assumptions are made regarding the TDR system.

1. That the actual TDR system may be adequately described by a model having a simple ramp as the pulse source and a lossless transmission line with an ideal sampler;

2. That the rounded "corners" of the actual pulse source may be ignored;

3. That the transmission line high frequency losses classed as "skin effect" or "dribble up" are not significant. ("Dribble up" is explained under Measuring Technique in connection with Fig 17).

4. That the sampler is non-loading, non-distorting and of infinitesimal risetime;

5. That parasitic (stray) reactances are insignificant.

The formula for small series inductance and small shunt capacitance in a tranmission line contain factors for (1) the system risetime at the spatial location of the reactance, (2) the observed reflection coefficient, and (3) the transmission line surge impedance.

The system risetime may be measured from the display by placing either an open circuit or a short circuit at the spatial location of the reactance.

The value for a small series inductor can be calculated using the formula

$$L = 2.5\,\alpha\,Z_o\,t_r \qquad (10)$$

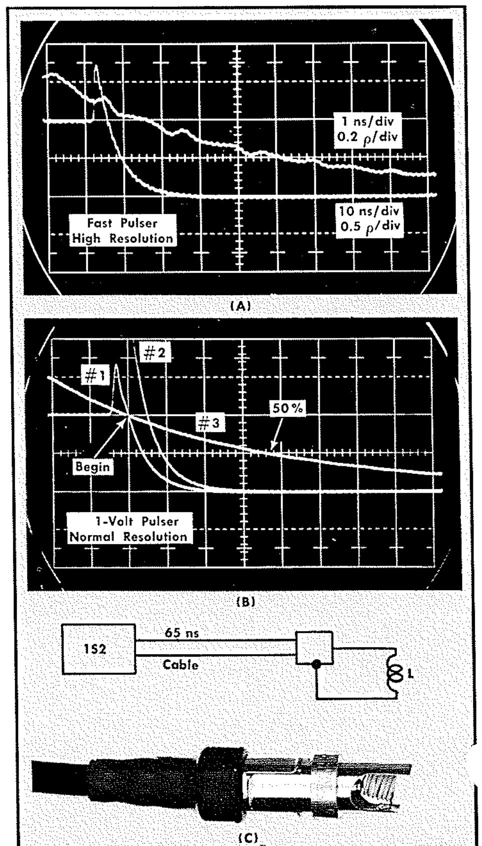


Fig 13. Seven turn coil across end of 50 Ω line.

where L is in henries, Z_o is in ohms, t_r is the system 10% to 90% risetime in seconds, and α is a dimensionless coefficient related to the observed reflection coefficient ρ by either the graph or Fig 14, or formula (11).

$$|\rho| = \alpha\left(1 - \varepsilon^{\frac{-1}{\alpha}}\right) \qquad (11)$$

A small shunt capacitor's value can be calculated using the formula

$$C = \frac{2.5\ \alpha\ t_r}{Z_o} \qquad (12)$$

where C is in farads, and the other units are as in formula (10).

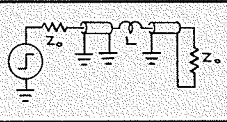
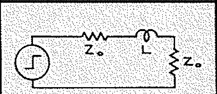
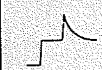
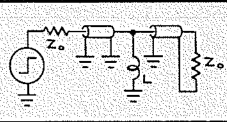
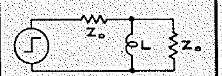
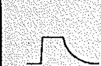
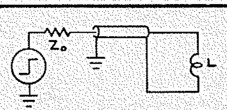
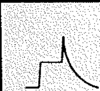
Circuit		Equivalent Circuit	Formula	Display
Series with terminated line			(7) $L = 2\ Z_o \times 1\ TC$	
Parallel with terminated line			(8) $L = \frac{Z_o}{2} \times 1\ TC$	
Across line end			(9) $L = Z_o \times 1\ TC$	

TABLE 5 "Large" Inductor Circuits and Formulae

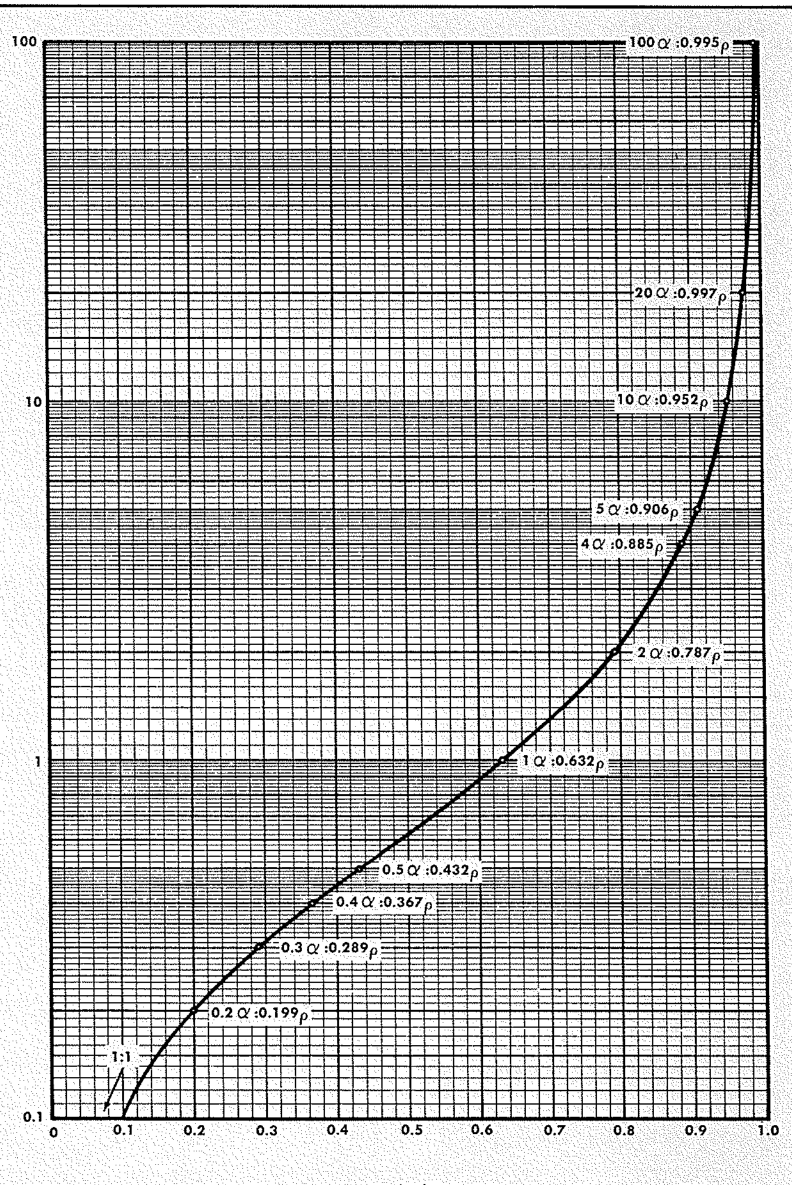


Fig 14. Graph for conversion of small reactance observed ρ to α for use in formulae (10) and (12).

Small Series Inductor

Fig 15 is an example of TDR displays from a small inductor (1¾ turn) placed in parallel with a 50 Ω line at (A), and in series with the 50 Ω line at (B). Calculations were made on Fig 15A first because the display is a clean exponential that permits L/R time constant analysis. Waveforms #1 and #2 of Fig 15A show first the full exponential decay through five CRT divisions, then at #2 the waveform was positioned vertically so the exponential end is at —1 division. Waveform #3 used the same vertical calibration, but was time expanded to obtain the new 100% to 50% time duration.

The time duration of the 50% amplitude change section of the exponential curve is 450 ps. This time divided by 0.693 produces a one time-constant time duration of 650 × 10^{-12} seconds. Then from formula (8), the value of the inductor is 16.22 nH (1.622 × 10^{-8} H).

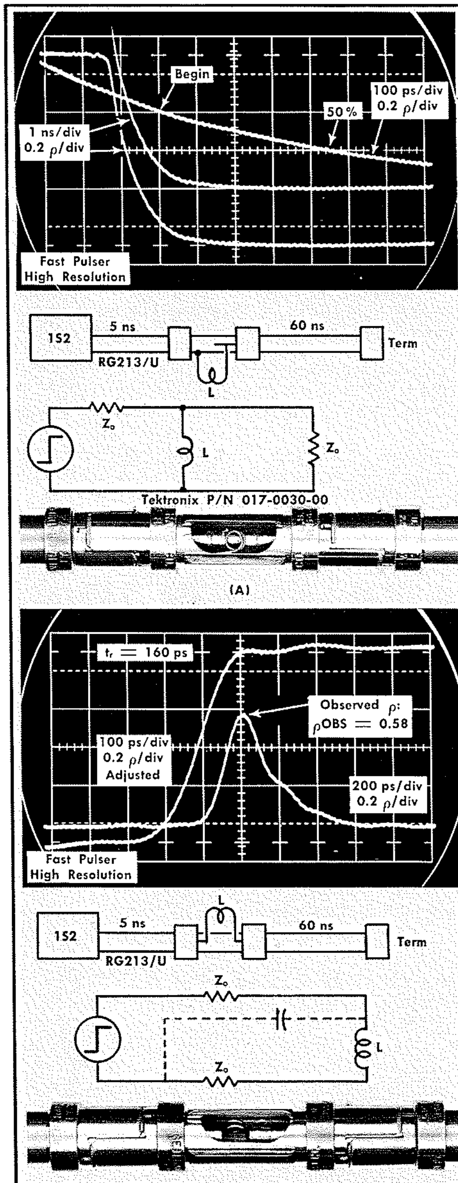


Fig 15. (A) 1¾ turn coil in parallel, (B) same coil in series, with 50 Ω coaxial line.

The waveform of Fig 15B has an observed deflection coefficient of +0.58. From the graph of Fig 14, 0.58 ρ is equal to 0.82 α. The risetime of the system was found to be 160 ps by disconnecting the insertion unit in which the inductor was located and measuring the reflection signal risetime. These figures placed into formula (11) give a value for the series inductor of 16.4 nH (1.64×10^{-8} H). This correlates very well with the previous parallel measurement.

Small Shunt Capacitor

Fig 16 is an example of a small shunt capacitor placed across a 50 Ω coaxial cable by compressing the cable outer diameter. Since the cable (RG8A/U) has normal impedance variations along its length, the peak reflection from the capacitor can only be approximated. Assuming a ρ of —1 division in Fig 16, then by formula (12), the capacitance is approximately 0.085 picofarads.

The Type 1S2 is useful for observing similar small discontinuities along transmission lines. In particular, high quality cable connectors can be evaluated for their ability to maintain a constant impedance where two cables are mated. Or, the quality of production installation of high quality connectors to flexible cable can be easily evaluated.

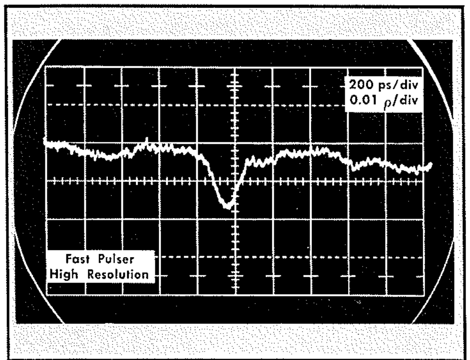


Fig 16. Shunt capacitor, ≈0.085 pF, caused by compressing RG8A/U coaxial cable with pliers.

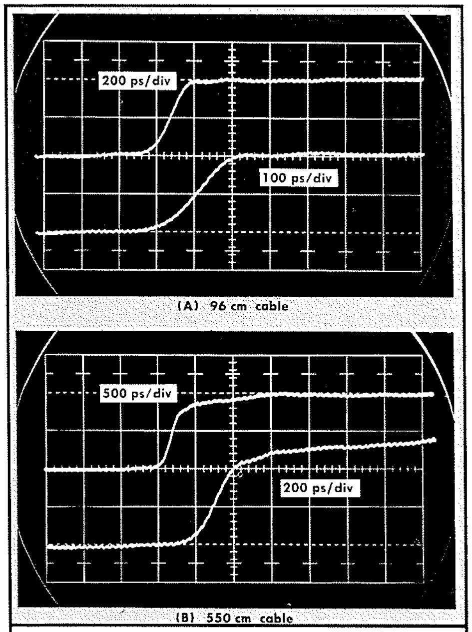


Fig 17. "Dribble up" characteristics of two lengths of RG8A/U.

Measuring Technique

The measurement of the small series inductor of Fig 15B is explained here to point out necessary techniques for measuring small reactances.

In evaluating small reactances with the TDR system, we have assumed the driving pulse to be a linear ramp; therefore, the ramp risetime must be determined for each change in the test system. The words "dribble up" refer to the characteristic of a coaxial cable to transport a step signal with distortion. The time required for the cable output signal to reach 100% of the step signal input amplitude is many times longer than the interval needed for the output signal to change from 0% to 50%. If we consider that the small reactance receives a pure ramp signal, then the rounded corners of the output pulse must be ignored.

Fig 17 shows the degradation of the Type 1S2 incident signal pulse by two different lengths of RG8A/U coaxial cable. Fig 17A is the reflection from an open cable 96 cm long, (192 cm signal path) and Fig 17B is the reflection from an open cable 550 cm long (1100 cm signal path). The upper waveform in each case was made with the Type 1S2 VERTICAL UNITS/DIV control set to 0.5 ρ/DIV, calibrated. The lower waveform in each case was made with the Type 1S2 vertical VARIABLE control advanced slightly clockwise to approximate a deflection factor of 0.5 ρ/DIV for just the ramp portion of the waveform. In each case the signal continues to rise after the inital step, but Fig 17B shows the "dribble up" characteristic very plainly. The lower waveform of Fig 17A and B does not permit an accurate measurement of the system risetime because the waveforms as shown are not large enough. However, the upper waveforms of Fig 18A and B are large enough to permit a reasonable measurement of the 10% to 90% risetime of the ramp that drives the small inductor. It is also obvious from Fig 18A and B that the series inductor peak reflection is truly caused by just the ramp portion of the driving signal and not by the "dribble up" portion.

Calculations made from Fig. 18A and B using formula (11) and the curve of Fig 14, indicate the series coil has an inductance of 16.40 nH at Fig 18A and inductance of 16.51 nH at Fig 18B. (Fig 18A: $L = (2.5)(0.82)(50)(1.60 \times 10^{-8}) = 1.64 \times 10^{-8}$ H.) (Fig 18B: $L = (2.5)(0.66)(50)(2.0 \times 10^{-8}) = 1.651 \times 10^{-8}$ H.) This indicates that an inductor in series with a coaxial transmission line can be accurately measured so long as the risetime of the ramp portion of the incident signal can be measured. Fig 18B indicates that a cable of RG8A/U a bit longer than 550 cm might make it difficult to measure the ramp risetime from the display. If a cable has sufficient length to prevent a reasonable display to measure the ramp 10% to 90% risetime, the small series inductor cannot be measured.

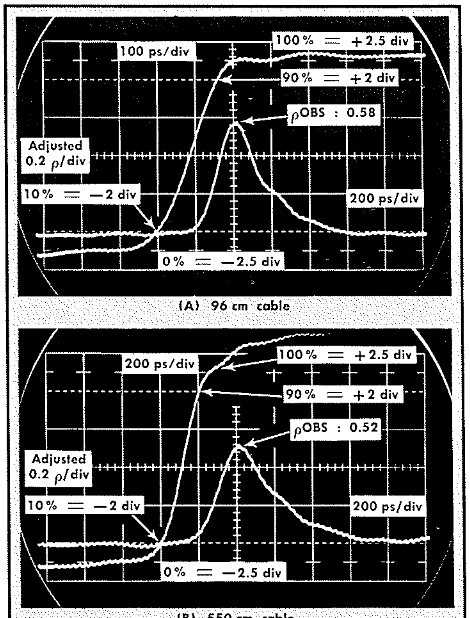


Fig 18. Small series inductor measured 96 cm and 550 cm away from Type 1S2 in RG8A/U coaxial cable.

Calculations of cable risetime will not permit small inductor measurements because the Type 1S2 vertical ρ/DIV calibration must be adjusted in each case. Once the vertical gain has been increased to measure the ramp risetime, the same new adjusted vertical ρ/DIV setting is used for measuring the observed ρ from the series inductor. If the cable is long enough to make it impossible to "see" the top of the ramp, the inductor cannot be measured. The same limitations apply when measuring small shunt capacitors.

Locating Small Reactances

The discussion of small reactances has thus far assumed that the TDR operator has access to all the cable between the TDR unit and the reactance being measured. This is, of course, not always the case. When a long length of cable indicates a fault, the reflected signal has not only been reduced in amplitude, it has also been smeared in time. The discontinuity is then located in time, closely related to the approximate 10% amplitude point or the beginning edge of the display rather than, as might be expected, at the peak of the reflection.

REFLECTIONS FROM RESISTIVE DISCONTINUITIES

Two types of reflections occur from two types or resistive discontinuity. They are a

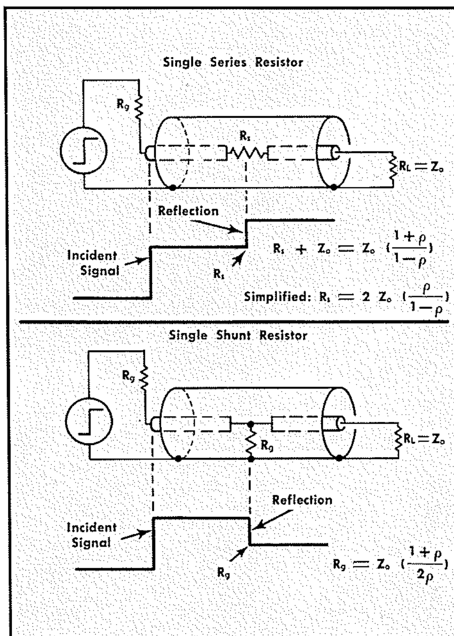


Fig 19. Single resistor discontinuities.

step reflection, or a continuously changing reflection. A resistance in series with a transmission line causes a positive reflection. A resistance in parallel with a transmission line causes a negative reflection. Discrete single resistors cause a step reflection, while distributed resistance causes a continuously changing reflection. The discrete resistor reflections are shown in ideal form in Fig 19, and the distributed resistance reflections are shown in ideal form in Fig 20.

Fig 20 has been exaggerated by showing the distributed resistance beginning at a particular point in the line. Normally, such series or shunt distributed resistance will be found in the total length of line tested by TDR.

All four forms of resistance are an indication of signal losses between the input and output ends of the transmission line. The single resistor discontinuities can occur due to discrete components or may indicate a loose connector with added series resistance. Such discontinuities can be physically located by special use of the POSITION RANGE control of the Type 1S2. Distributed losses are usually part of the particular line being tested and the TDR display can be of value for quantitative analysis of resistance per unit of line length.

No reflection should occur from a properly fabricated matched attenuator. Therefore, a TDR unit will not indicate losses when matched attenuators are used.

Distributed Resistance Examples

The examples of distributed resistance reflections that follow deal with the normal characteristics of transmission lines. Both small diameter lossy cables and moderate diameter quality cables are discussed.

Small, Lossy Cables

A small diameter 50 Ω transmission line (such as ⅛ inch diameter cable) will have sufficient DC resistance to mask "skin effect" losses. The DC resistance in its center conductor will cause a nearly exponential changing reflection. See Fig 21A. As the incident signal propagates down the line away from the TDR unit, the small series resistance causes small reflections to return to the TDR unit. If you mentally integrate the line into small sections of series resistance, you can then understand the continuous return of energy to the input end of the line. Each reflected energy "bit" is additionally attenuated on its way back to the TDR unit. This return attenuation is the factor that prevents the display from being a linear ramp, converting it into a nearly exponetial reflection. (Note the curve of the reflection between the incident signal plus step and the termination of Fig 21A.

Another way of expressing the effect of the nearly expoential reflection is to say that the transmission line input surge impedance changes with time. Fig. 21A shows the line surge impedance to be essentially 50 Ω at the beginning of the exponential reflection and to be approximately 64 Ω after 130ns ($+0.12\,\rho = 64\,\Omega$).

The long nearly exponential decay after the termination of Fig 21A is related to high frequency losses and the previously described "dribble up". The negative reflection occurs at the termination because the 50 Ω termination was driven by approximately 64 Ω. If the long exponential decay after the termination were expanded vertically, it would follow the rules for distortion to pulses by coaxial cables described with Fig 25.

If the small diameter cable is shorted at its end instead of terminated, the TDR display will appear similar to Fig 21B. A lossless line would have a full $-1\,\rho$ after the short, but the small lossy cable not only has attenuation of the signal to the short, but attenuation of the reflected signal back to the TDR unit. Again, the long nearly exponential curve after the short is caused by the cable distorting the reflected step signal.

Fig 21B also allows measuring the total cable DC resistance between the TDR unit and the short circuit of Fig 21B. The vertical distance between the incident pulse peak level and the right end flat portion of the reflected signal is due strictly to the cable DC resistance. In this case, —3.8 divisions = -0.76ρ which is equal to 6.5 Ω (from curve of Fig 10). (A bench multimeter type ohmmeter indicated 6.8 ohms for the same cable.)

Quality Cables

A quality cable, such as RG8A/U (52 Ω), RG213/U (50 Ω) or RG11/U (75 Ω) will exhibit similar characteristics to the small lossy cable just described, but the cable must be much longer to obtain a similar display of series resistance. Fig 22A and B show the same rising type of waveform caused by center conductor series resistance in RG213/U. Fig 22C shows the residual DC resistance of the line when shorted. Fig 22D is a time and voltage expansion of the (A) and (B) waveforms to show a possible use for the Type 1S2 in troubleshooting cable fabricating equipment.

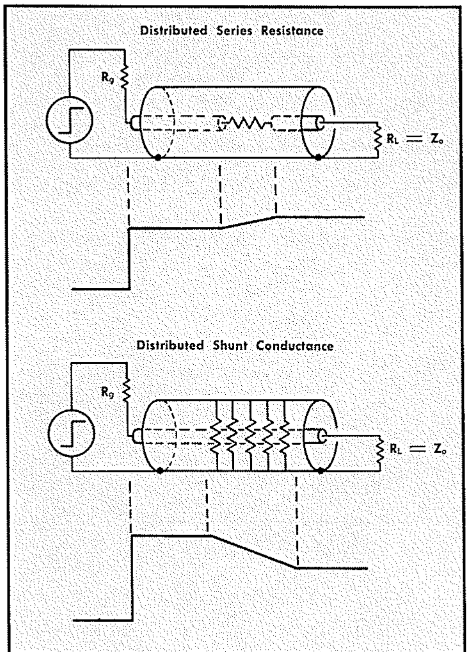


Fig 20. Distributed resistance reflections.

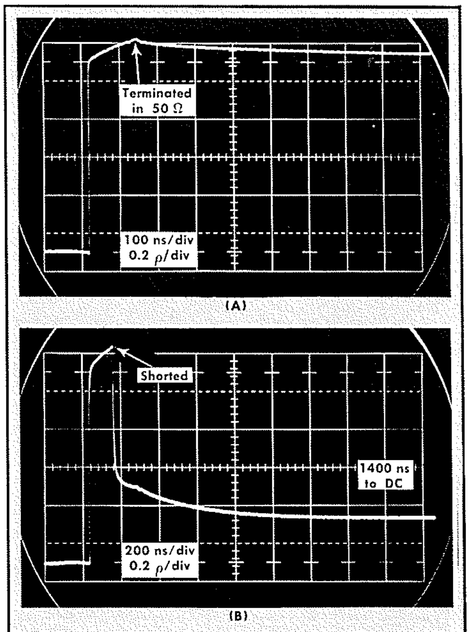


Fig 21. Waveforms of ⅛ inch diameter lossy 50 Ω cable.

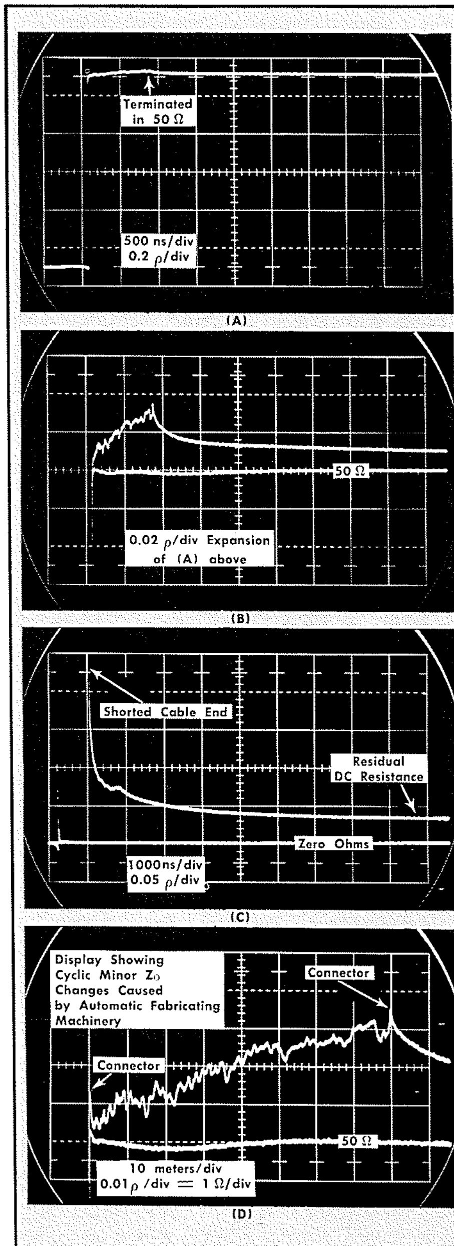


Fig 22. Quality RG213/U cable resistance and ΔZ_o characteristics. (Cable tested 260 feet long).

Fig 23 shows the same series resistance characteristics for RG11/U cable. However, instead of terminating the cable end, the series resistance was measured first with the end open, and then with the end shorted. Note the difference in slope of the waveform (apparent change in resistance) after the signal has traveled to the indicated line end. The change in slope is due to distortion of the originally flat incident pulse by traveling through the cable once. As the non-flat signal reaches the cable end, its reflection back through the cable is altered a second time. The net result is an obvious distortion to the true resistive slope of the reflected "bits" of the distributed series resistance during the 2nd half of the reflection. This example is given to show the desirability of properly terminating any line section in which you wish to measure its total distributed series resistance. (Conditions leading to this changing slope phenomenon are described by H. H. Skilling on page 397 of his text "Electronic Transmission Lines", McGraw-Hill, 1951.) Each of the three waveform pictures of Fig 23 is a double exposure with the lower waveform showing the normal Type 1S2 response to a termination resistance at (A) and (B) and a short circuit at (C).

COAXIAL CABLE RESPONSE TO A STEP SIGNAL

Coaxial cable have a step-function response that distorts the original signal. The distortion is caused by cable losses of several types which are frequency dependent. The longer the cable length, the greater the distortion. Response to a step signal can be evaluated by placing the cable in a TDR system, or by placing it between a fast rise pulser and a fast risetime sampler. (When a cable is tested by a TDR device, the signal traverses the line twice; when a cable is placed between a pulser and a sampler, the signal traverses the line once.)

Studies in the past that considered skin effect losses only[1] have indicated that some types of coaxial cables have a step-function response with decibel attenuation that varies as the square root of the frequency. Based upon this assumption (of skin effect losses only), the step response time from 0% to 50% will increase by a factor of 4 through a cable whose length is twice that of a previous test. Such is not the case in practice as seen by use of the Type 1S2. Other forms of losses due to the dielectric material between inner and outer conductors, radiation from lines whose outer conductor is braided, and reflection losses from surface variations of the conductors, are discussed in detail in an article by N. S. Nahman[2]. Nahman considers several techniques which are useful in analyzing the transient behavior of coaxial cables that have these forms of high frequency losses.

Long Cables

Distortion to pulse signals in coaxial cables is most easily evaluated (visually displayed on a CRT) when the cable is long. A long cable is here defined as one that exhibits significant losses in the system in which it is used. The tests shown in Fig 24 were made on a 100 foot section of RG11/U and a 260 foot section of RG213/U. In each case the signal traversed the line twice in a normal TDR manner. The cable far end was left an open circuit so that a return signal of $+1\rho$ could be observed. This gives the same effect as having sent the Type 1S2 signal through a line twice as long.

The term T_o, shown in Fig 24, is the length of time between the 0% amplitude and 50% amplitude points along the step rise of the cable output signal. 0% to 50% is chosen because it contains the fastest part of the transition and because it is easy to read. The usual practice of measuring risetime from 10% to 90% is perfectly valid if the display has an adequate rate of rise at the 90% point. The cables tested for Fig 24 have a 10% to 90% risetime that lasts about 18 times longer than T_o. Fig 24 shows plainly that the step response of a coaxial cable does not have the familiar Gaussian shape. For this reason the risetime of systems containing long coaxial cables cannot be calculated using the square root of the sum of the squares of the individual unit risetimes.

The length of time required for the output signal to rise to 100% of the input signal is many times longer than T_o. This distortion is called "dribble up" as first discussed earlier under Measuring Technique when measuring a small series inductor in a transmission line. Fig 25A is a double exposure using the 260 foot length of RG213/U connected between the Type 1S2 1-Volt pulser and the terminated Thru Signal Sampler. Both traces were made at 100 ns/div. The upper trace at 0.2 ρ/DIV and the lower trace at 0.05 ρ/DIV. The lower trace leads us to believe that the output pulse reaches 100% amplitude sometime between 4000 and 5000 ns after the initial step rise. More exact measurements can be made by comparing the cable output with the Type 1S2 no-cable response as shown in Fig 25B. Here both traces were made at 1000 ns/DIV and 0.02 ρ/DIV with an intentional small vertical repositioning. When the two traces become a constant distance apart, you can be relatively certain the cable output signal has reached 100% amplitude. Fig 25B indicates a possibility that the output signal had not completely reached 100% amplitude even after 8000 ns (8 μs).

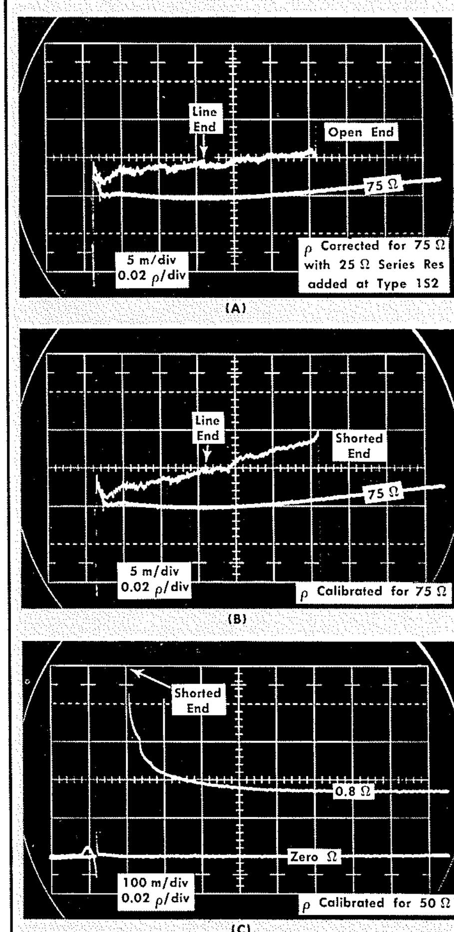


Fig 23. Quality RG11/U cable resistance and ΔZ_o characteristics. (Cable tested was 100 feet long).

Short Cables

Even though information just given on Long Cables is true for any length cable, a physically short cable can be treated as if it

[1] R. L. Wigington and N. S. Nahman, "Transient analysis of coaxial cables considering skin effect," Proc. IRE, vol. 45, pp. 166-174; February 1957. Q. Kerns, F. Kirsten and C. Winningstad, "Pulse Response of Coaxial Cables," Counting Notes, File No. CC2-1, Rad. Lab., University of California, Berkely, Calif.; March, 1956. Revised by F. Kirsten; Jan. 15, 1959.

[2] N. S. Nahman, "A Discussion on the Transient Analysis of Coaxial Cables Considering High-Frequency Losses," IRE Transactions On Circuit Theory, vol. CT-9, No. 2, pp. 144-152; June, 1962.

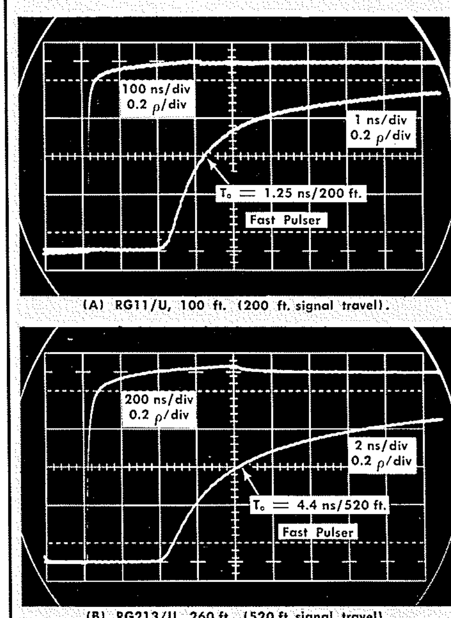


Fig 24. RG11/U and RG213/U distortion to a step signal. Waveforms are reflections from cable open end.

were Gaussian. A short cable will have a T_o sufficiently faster than the Type 1S2 fast pulser 10% to 90% risetime, that the long slow rise ("dribble up") of Fig 25 will not be evident. Under these short cable conditions, it is reasonable to assume the bandpass upper limit of a cable and its system can be approximated from the 10% to 90% risetime display. A display of 10% to 90% risetime in 100 picoseconds then approximates a sine wave upper frequency 70% amplitude of: $0.35/(1 \times 10^{-10}) = 3500$ MHz.

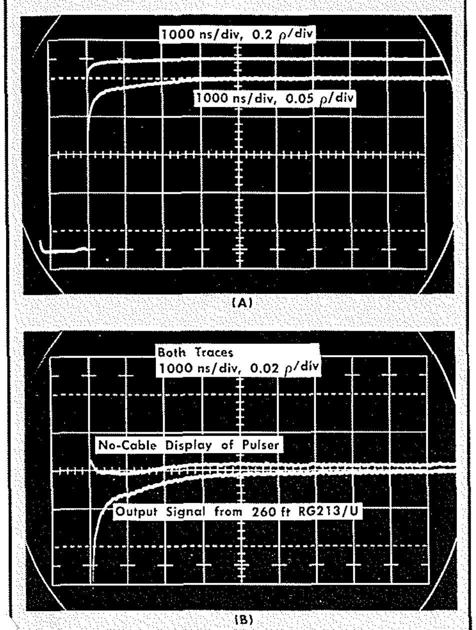


Fig 25. "Dribble up" output signals from 260 ft. RG213/U.

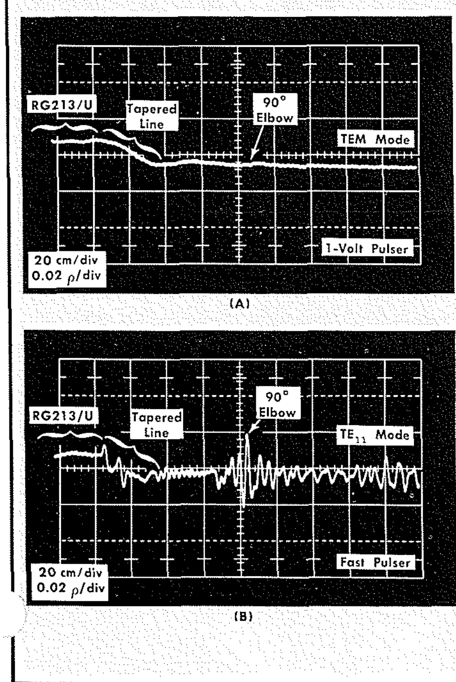


Fig 26. Propagation mode change in large diameter transmission line when driven by the Type 1S2 fast pulser.

Large Diameter Transmission Lines

Use of the Type 1S2 0.25-Volt fast-risetime pulser should be limited to use on lines whose outer conductor inner diameter is less than about one-quarter wavelength at 3500 MHz. Normal signal propagation mode in transmission lines is TEM, but will change to a waveguide mode, TE_{11}, if too high a frequency is used. Fig 26 shows both modes of propagation in a transmission line 3⅛ inch in diameter. Fig 26A picture was taken using the Type 1S2 1-Volt pulser. Fig 26B picture was taken using the Type 1S2 0.25-Volt fast pulser. The line elements were the same in each case; 1) a short section of RG213/U cable between the Type 1S2 and a tapered line section; 2) the tapered line section; and 3) a section of 3⅛ inch diameter rigid air line with a 90° elbow in the display time window. The numerous aberrations of Fig 26B are due to a change in propagation mode when the signal arrived at the 90° elbow. The resulting multiple reflections are of no value to the operator testing the line.

SPECIAL APPLICATIONS

General

Much of the previous portion of this article deals with using the Type 1S2 as a Time Domain Reflectometer. Many more uses can be made of the unit in a TDR mode, limited only by the measurement needs of the user. Listed below are suggestions of other TDR applications not yet described.

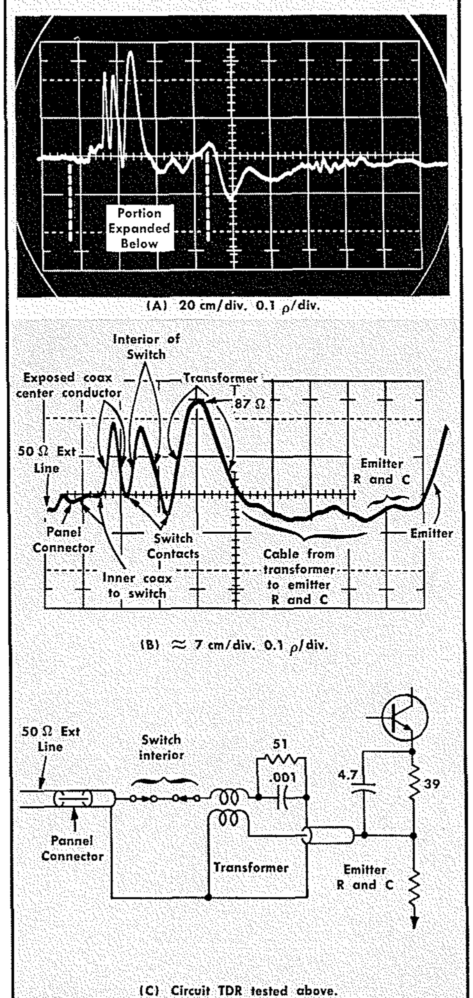


Fig 27. TDR view of broadband amplifier input circuit.

Signal Generator Output Impedance

The Type 1S2 can be connected to the output terminal of a signal generator to measure its output impedance. If the generator output signal can be turned off while keeping the output circuit active, a clean TDR can be obtained.

Broadband Amplifier Input Impedance

Fig 27 shows two pictures of a broadband amplifier input circuit. Fig 27A includes the active emitter circuit of the input common-base transistor amplifier. Fig 27B includes the parts between the input connector and the transistor emitter. The power was off when Fig 27B photo was taken to show the transistor emitter spatial location accurately.

Circuit Board Lead Impedance

Fig 28A shows changes in surge impedance of leads along an etched circuit board. (The board reverse side was fully plated.) The major dip is due to a right angle corner while the minor dip is due to a rounded corner.

Changes in surge impedance due to a change in lead width is also plainly seen by TDR. Fig. 28B shows an inductive section of line when the physical width of the line was reduced one half for a length of about 1.25 inches.

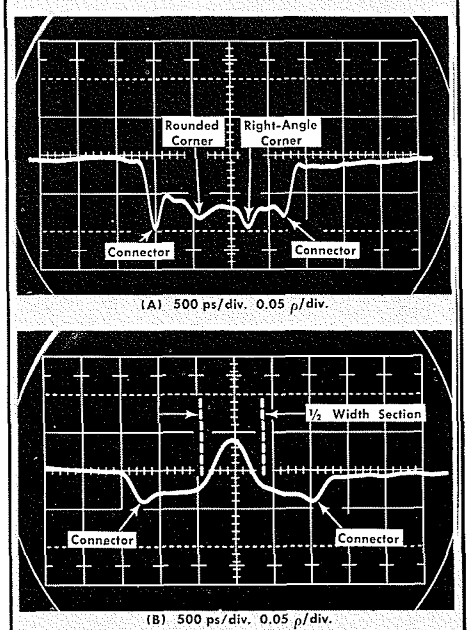


Fig 28. Etched circuit board Z_o checked by TDR.

Frequency Compensation of Lossy Cables

A lossy coaxial cable connected between one of the Type 1S2 pulsers and the sampler (terminated) permits a view of the cable output signal. Fig 29 shows the same lossy cable described earlier with Fig 21. A double exposure shows at the top how the cable distorts the 1-Volt pulser while the lower waveform is flatter due to a simple RC compensation network placed between the pulser and the cable. The TDR unit will permit testing such compensation networks.

Evaluation of Ferrite Beads and Cores

Ferrite beads and cores can be evaluated using the Type 1S2. Simple inductors wound on toroid ferrite cores are represented by an equivalent circuit which is essentially an inductance in parallel with a resistance. The resistance results from core losses and may be typically as low as 10 to 30 ohms/(turn)^2. Both the resistance and inductance characteristics of ferrites can be seen in a TDR display.

Fig 30 shows two displays and the special adapter jig used to test a ferrite bead. The adapter jig is made from one half of a Tektronix Insertion Unit (Part No. 358-0175-00). The end of the center piece was flattened and a formed piece of #10 copper wire soldered in place with a ferrite bead included. Thus, there is only a small diameter change of the 50 center conductor (pip in both displays) and one turn through the ferrite center. (Use smaller wire for smaller beads.)

Fig. 30A shows the basic display. L/R time-constant analysis is similar to that of Fig. 15 and formula (8) of Table 5, except the core R is in parallel with the driving line Z_o.

Fig 30B shows the ferrite bead resistance as $-0.16\ \rho$, or 36 Ω. (The 36 Ω is read directly from the curve of Fig 10.) The resistance value of a core is read by finding the curve knee (as marked in Fig 30B) where the inductance affect becomes obvious. The positive pip is ignored.

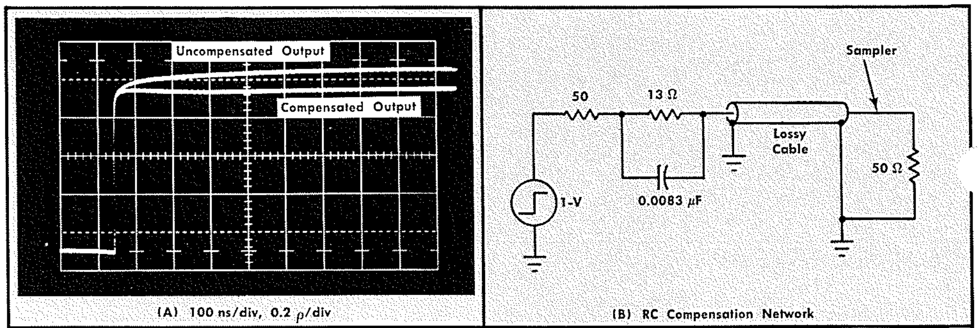


Fig 29. Simple frequency compensation for lossy coaxial cable.

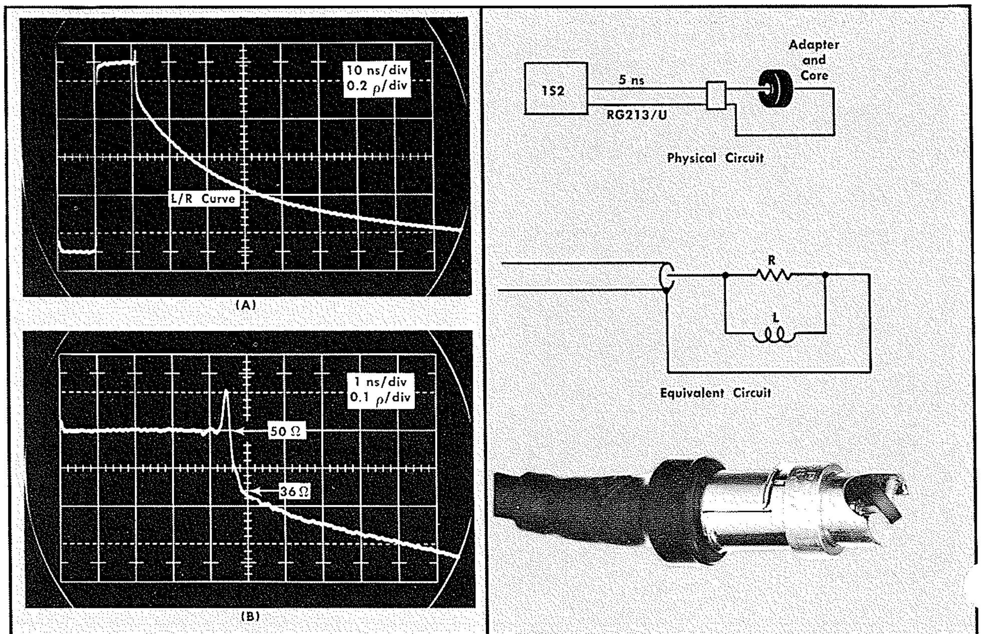
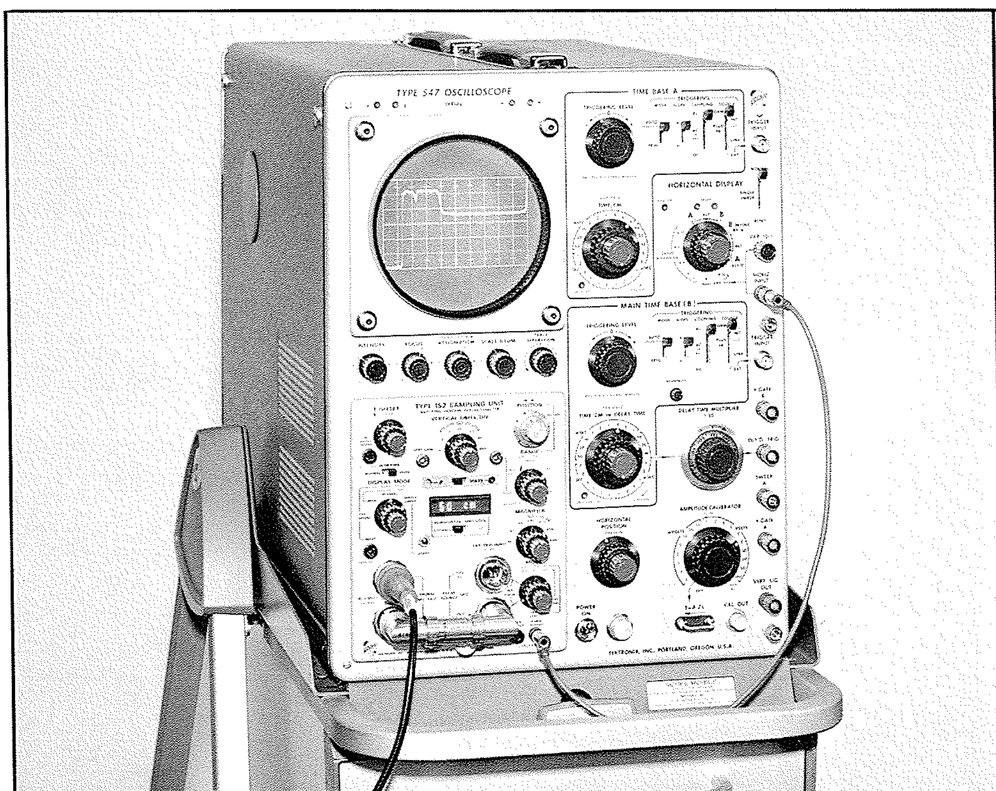


Fig 30. Evalution of a ferrite core.



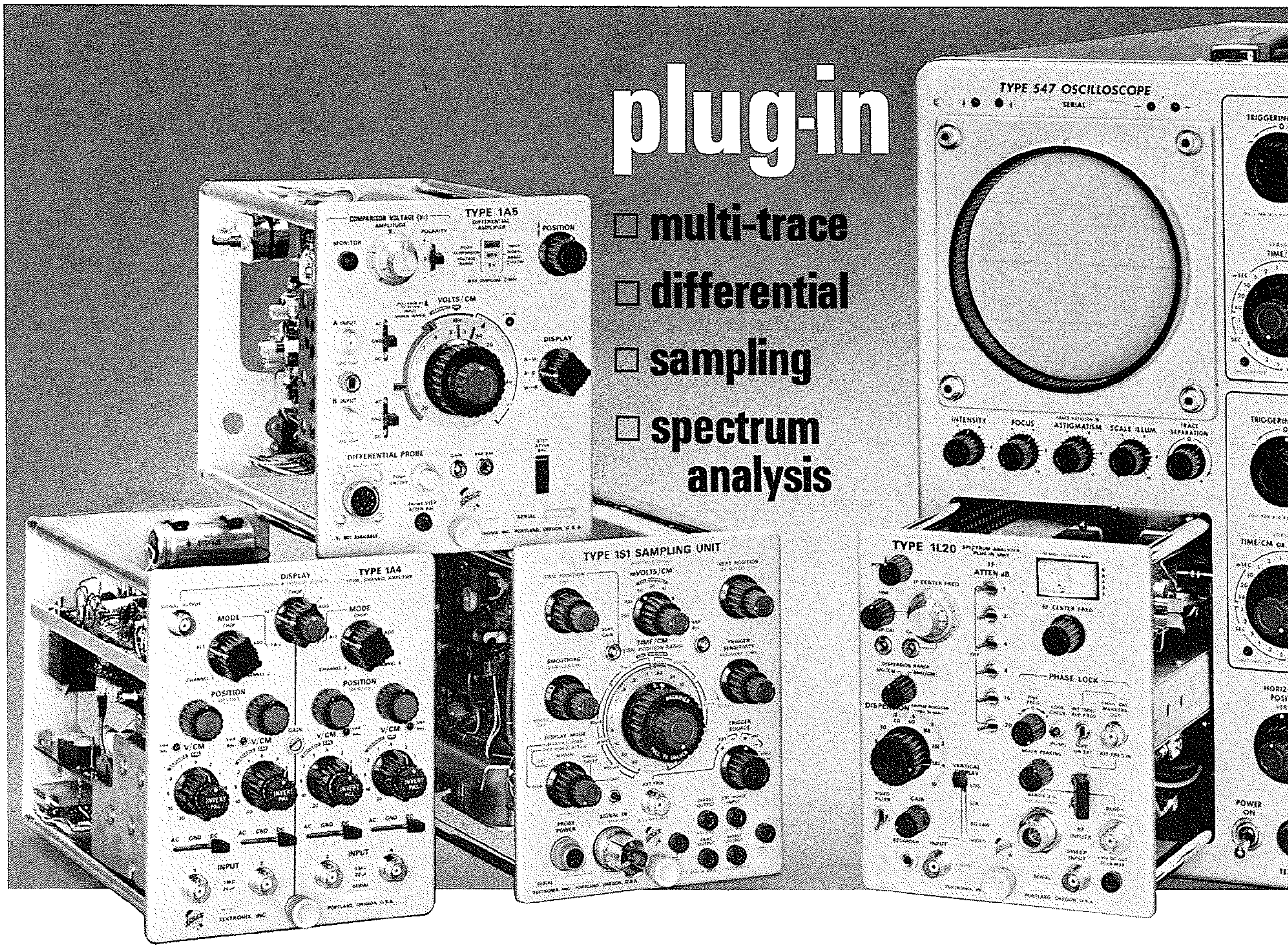
Tektronix Type 547 Oscilloscope with a Type 1S2 Sampling Plug-In Unit.

The measurements described in this article can be easily made with the Type 1S2 Plug-In Unit.

The Type 1S2 Sampling Plug-In converts any Tektronix 530, 540, 550-series oscilloscope to a time-domain reflectometry measurement system. It also has the ability to make many general sampling measurements.

As a TDR, the Type 1S2 has a system risetime of 140 ps and is calibrated in ρ (rho) from 0.005 ρ/div to 0.5 ρ/div. The horizontal is calibrated from 1 cm/div to 100 m/div for dielectrics of air, TFE and polyethylene. A 10-turn dial reads directly the one-way distance to the test-line discontinuity. Two pulse outputs provide either 50 ps T_r at 250 mV into 50 Ω or 1 ns T_r at 1 V into 50 Ω.

The 90-ps risetime, 5 mV/div deflection factor, 100ps/div sweep and built-in triggering capability make the Type 1S2 useful in many other sampling measurements.



Tektronix 530, 540 and 550-series plug-in oscilloscopes offer a wide range of performance, designed to meet your changing measurement needs. Select the performance and measurement functions you need from multi-trace, differential, sampling and spectrum analyzer plug-ins.

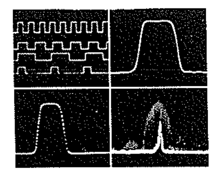
For multi-trace applications, the new Type 1A4 Four-Channel amplifier offers constant DC-to-50 MHz bandwidth and 7-ns risetime capabilities over its 10 mV/cm to 20 V/cm deflection factor range. Operating modes include alternate or chopped four channel, dual channel differential, and 2, 3, or 4 channels added or subtracted. Two dual-trace plug-ins are also available, the Type 1A1 with 28 MHz at 5 mV/cm (50 MHz at 50 mV/cm) and the Type 1A2 with 50 MHz at 50 mV/cm.

For differential applications, the new Type 1A5 Differential amplifier features 1 mV/cm deflection factor, 1,000:1 common-mode rejection ratio at 10 MHz, ± 5 V comparison voltage and 50 MHz bandwidth with 7-ns risetime at 5 mV/cm. The low-cost Type 1A6 Differential plug-in with 1 mV/cm deflection factor, 10,000:1 CMRR and 2-MHz bandwidth and the high-gain Type 1A7 Differential plug-in with 10 μV/cm deflection factor, 50,000:1 CMRR and 500 kHz bandwidth are also available.

For sampling applications, choose from two high performance plug-ins, the Type 1S1 general purpose sampling plug-in and the Type 1S2 TDR sampling plug-in. The Type 1S1 features internal triggering, 0.35-ns risetime, DC-to-1 GHz bandwidth and calibrated sweep speeds from 100 ps/cm to 50 μs/cm. The Type 1S2 is a time-domain reflectometer with a system risetime of 140 ps, 0.005 *ρ*/div deflection factor and sweep rates from 100 ps/div to 1 μs/div. With its 90-ps risetime, 5 mV/div deflection factor and built-in triggering, the Type 1S2 can be used in many other sampling applications.

Four spectrum analyzer plug-ins covering the spectrum from 50 Hz to 10.5 GHz convert your oscilloscope to a high-performance spectrum analyzer. The plug-ins cover the following frequency bands: Type 1L5 from 50 Hz to 1 MHz with 10 μV/cm deflection factor; Type 1L10 from 1 MHz to 36 MHz with —110 dBm sensitivity; Type 1L20 from 10 MHz to 4.2 GHz with—110 to—90 dBm sensitivity; and Type 1L30 from 925 MHz to 10.5 GHz with —105 to —75 dBm sensitivity.

Multi-trace *differential*



sampling *spectrum analysis*

in all Tektronix 530-540-550-series plug-in oscilloscopes

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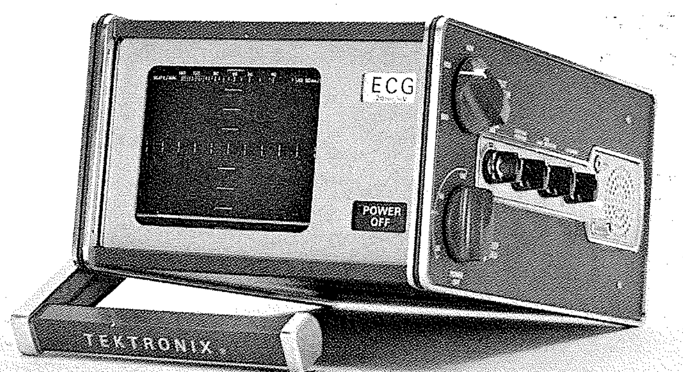
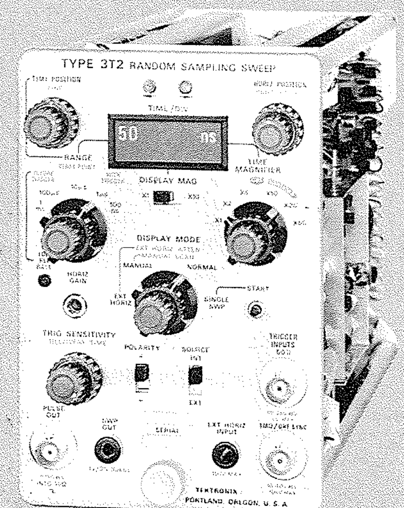


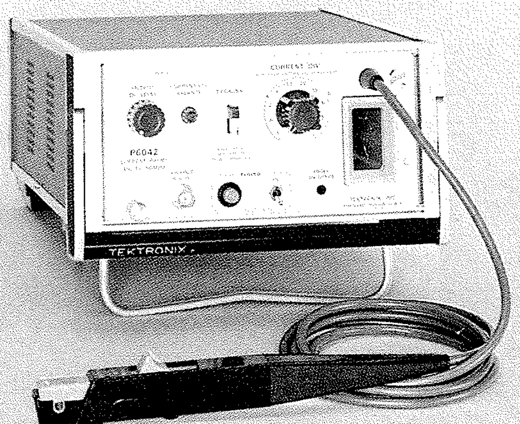
A NEW
PHYSIOLOGICAL
MONITOR

Page 2



A NEW
MEASUREMENT CONCEPT—
RANDOM SAMPLING

Page 7



A NEW
DC-TO-50 MHz
CURRENT PROBE

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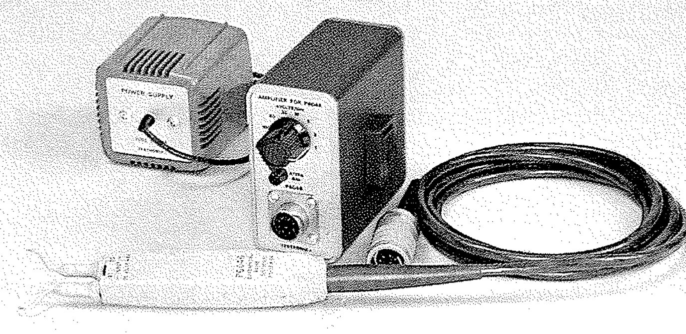
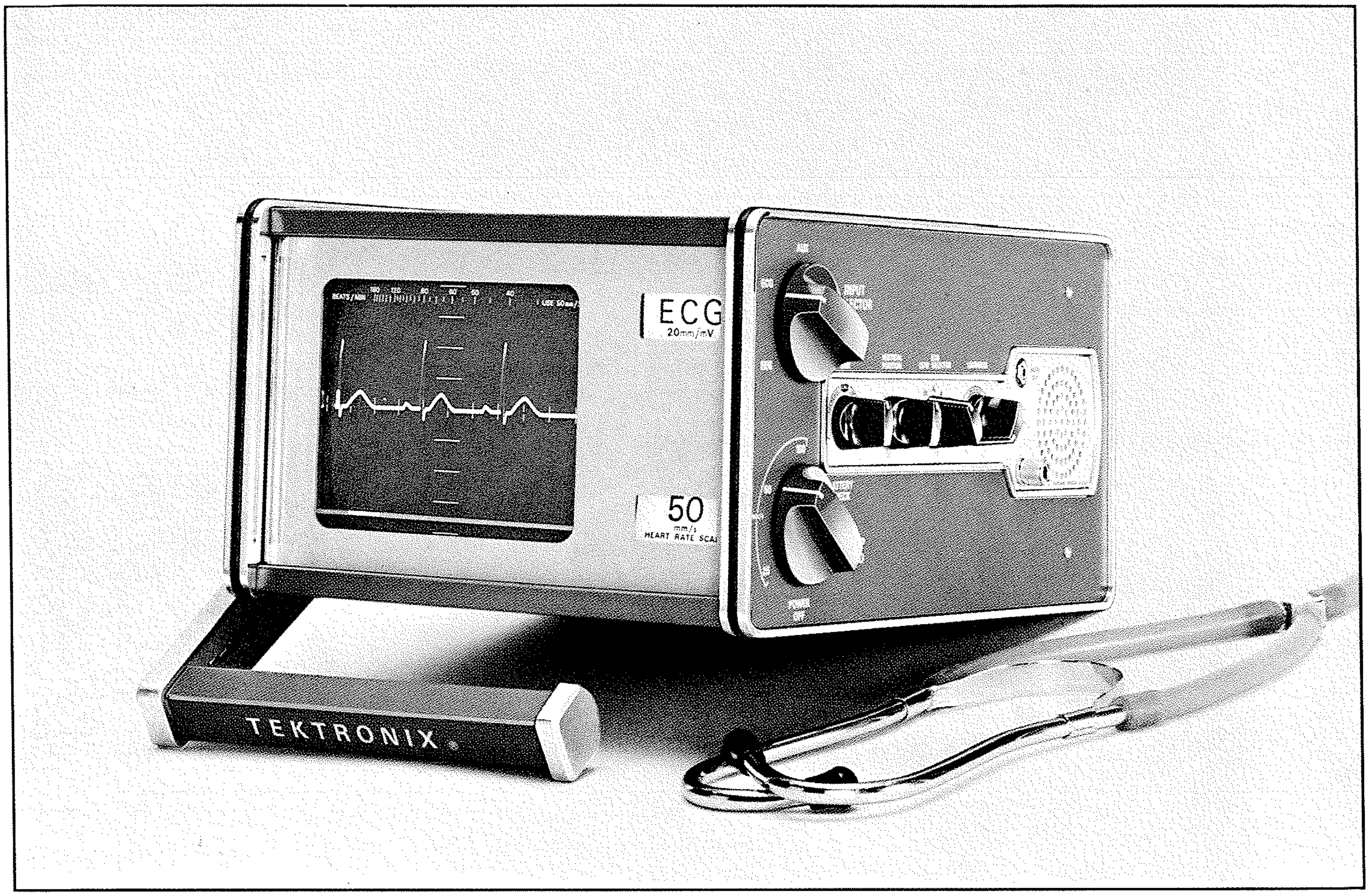


A NEW
DC-TO-100 MHz
DIFFERENTIAL PROBE
AND AMPLIFIER

Page 13

A SIMPLIFIED OSCILLOSCOPE FOR THE OPERATING ROOM

by Don L. Clark



INTRODUCTION

Tektronix recently introduced the Type 410 Physiological Monitor, a special purpose oscilloscope for use in clinical medicine. The instrument is *small* and powered by a rechargeable battery pack. Despite its compactness, it features a large 8 x 10 centimeter display area made possible by a wide-angle, magnetically-deflected cathode-ray tube (CRT). More importantly, the monitor is tailored to the unique requirements of the medical clinician.

For example, the controls are greatly simplified from those found on many oscilloscopes and are labeled in terms meaningful to medical personnel (Fig 1). The size and optional mounting fixtures permit the instrument to be used in the crowded perimeter of the surgical operating table.

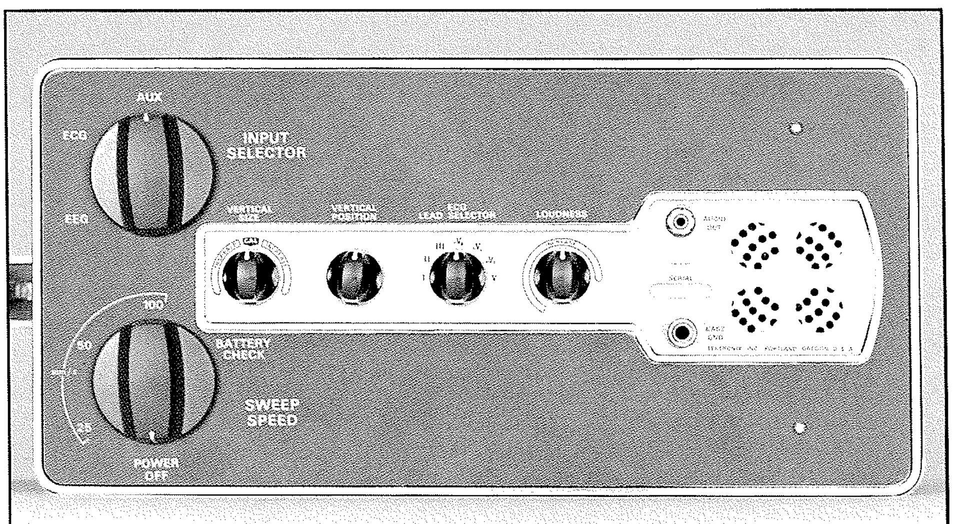


Fig 1. Type 410 Operating Controls

You can monitor any of three important physiological signals with the Type 410:

ECG—(or EKG, the Electrocardiogram) An electrical signal produced by the heart which can be detected on the surface of the body.

Pulse—Pulsations of blood sensed in the finger or elsewhere with the appropriate transducer.

EEG—(Electroencephalogram) An electrical signal produced by the brain.

The 410 was designed to be used wherever surveillance of patient condition is vital. In the operating room, a physiological monitor provides information regarding reactions to anesthesia and surgical procedures. In the recovery room and the intensive care unit, which by their very existence indicate the importance of constant surveillance, the physiological monitor provides a continuous display of valuable data.

SIGNALS FROM THE HUMAN BODY

The human body provides many indexes of relative well-being. Excessive body temperature has long been known to accompany ailments ranging from the minor to the serious. In a similar sense, and with varying degrees of reliability, eye dilation, pulse rate, respiration rate and others provide worthwhile information regarding the viability of the human body. When considered in the time domain, certain of these physiological indicators become more critically important and can yield substantially more information than others.

For example, one or two degrees of excessive body temperature persisting for several days would be of comparatively less cause for alarm than a heart stoppage for ten seconds. Moreover, the thermal mass of the body is such that hourly sampling might provide all the information required. But the nature of the heart is such that significant information may be observed from events lasting only a few hundredths of a second.

Thus, the physiological signals can be classified according to (1) the magnitude of deviation from the norm, (2) the relative importance to the human body, (3) the rapidity with which the change can occur, and (4) the time duration of the shortest significant event within the data. Signals involving comparatively short duration cyclical events and potentially rapid change can yield considerable information when displayed in graphical form on a monitor such as the Type 410.

THE ELECTROCARDIOGRAM (ECG)

Among the key physiological indicators is the ECG; a graphical recording of the heart electrical activity. This signal is associated with the muscular contraction which produces the pumping action. Effective pumping requires coordination of the individual heart muscles, with related cyclical patterns in the electrical signal.

While the electrical signal occurs within the heart muscle tissue, it can be detected on the surface of the body. The sensing electrodes can be placed at many different sites and each pair or combination of electrodes provides a different perspective of the complex three-dimensional signal generator, the heart.

Figure 2 shows an idealized waveform representing the electrocardiogram from one of the more popular monitoring configurations which consist of a differential measurement between electrodes on the right arm and left leg with a third electrode on the right leg serving as a common-mode reference to the monitoring system.

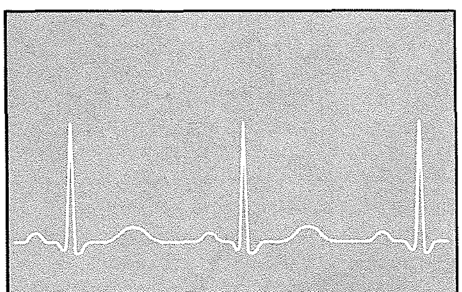


Fig 2. ECG Waveform

The information obtainable from the ECG is far too broad and technically complex to detail here, but several general uses can be mentioned: (1) Heart rate can readily be determined as can improper rhythm. (2) Heart attacks may involve dead tissue and coagulated blood in portions of the heart which can produce an abnormal ECG. (3) During certain stages of pregnancy, the fetal ECG can be detected. The presence of more than one fetus has sometimes been determined by this method. Orientation of the fetus in the womb may be determined by noting the fetal ECG polarity. (4) Victims of electrical shock may die due to heart fibrillation, a condition in which little or no blood is pumped by the heart. Fibrillation is a total loss of coordination between the various heart muscles which causes the heart to quiver rapidly rather than rhythmically contracting and expanding. Defibrillation can often be accomplished by applying a powerful electrical shock (up to 400 watt seconds in a ten-millisecond pulse) which temporarily locks the heart muscles. Within a few seconds after the intentional shock, the heart will often restart with the proper coordination. The electrical activity of the heart before and after defibrillation is readily monitored with the Type 410. Input circuitry of the instrument is protected against destruction by the defibrillator pulse so that there is no need to disconnect the monitoring electrodes during defibrillation.

THE PULSE

A normal ECG is no proof that blood is properly circulating throughout the body. Monitoring of the pulse by touch on the wrist, neck or elsewhere can show that blood is circulating, at least in that portion of the body and, in some cases, the judgement can be made that the pulse is "weak" or "strong".

The Pulse Sensor can more than replace the conventional touch method. The sensor is easily attached to the patient and will provide continuous, hands-off monitoring.

As the blood pressure rises and falls with each heart beat, the amount of blood present in any particular portion of the flesh varies slightly with the expansion and contraction of the blood vessels. This slight change can be detected from the correspondingly slight change in the translucency of the flesh. A small, low-power incandescent lamp directs light into the flesh and an adjacent photoresistor senses the light variation.

The finger tip, toe, and forehead are particularly good locations for the sensor. Contact pressure between the sensor and flesh is an important factor; excessive pressure will block blood flow and too little pressure will result in an excessive sensitivity to movement, thereby introducing interference. The Pulse Sensor is shown in Figure 3 with a removable finger adapter.

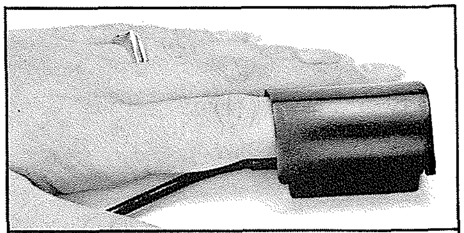


Fig 3. Pulse Sensor

This spring-loaded adapter not only holds the sensor against the finger with the proper pressure, but also excludes potentially interfering modulated light from fluorescent lamps or other sources. The adapter can be quickly attached and is self-holding on the finger.

For quick determination of heart rate, a direct reading Heart Rate Scale is provided across the top of the Type 410 graticule as shown in Figure 4. This scale is possible through the use of automatically triggered sweeps for both ECG and pulse displays,

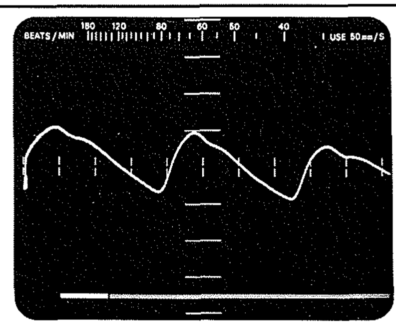


Fig 4. Pulse Sensor Waveform

and by the accurate sweep speeds of the Type 410. Three sweep speeds are provided: 25, 50, and 100 millimeters per second. The Heart Rate Scale is calibrated for use with the 50 mm/s speed.

The portion of the signal which has the greatest amplitude triggers the sweep and therefore appears at the lefthand edge of the graticule as shown in Figure 4. The corresponding portion of the next cycle appears to the right of the first at a distance determined by the time interval between the events and the horizontal sweep speed of the Type 410. From the known sweep speed and the measured distance, a simple calculation gives the pulse rate. The Heart Rate Scale is derived from this calculation and can be used with either a pulse or ECG display. (Display shows 75 beats/min.)

While the event which produces sweep triggering remains stationary at the lefthand edge of the graticule with successive sweeps, the second event changes position with any variation in heart rate. If the heart rate is uniform, the display need be watched for only two or three seconds to obtain an accurate rate indication. The scale can also be used with slightly less accuracy with the other two sweep speeds; dividing by two on 25 mm/s and multiplying by two on 100 mm/s.

THE ELECTROENCEPHALOGRAM (EEG)

In some surgical procedures, the heart is intentionally stopped and blood circulation is maintained by an external mechanical pump making it more difficult to determine the relative well being of the patient. In such cases the Type 410 can be used to monitor the EEG, the electrical activity of the brain. This complex signal, seemingly random to the layman (Figure 5), can yield valuable information through analysis of amplitude and frequency content.

The EEG is detected upon the surface of the head with electrodes similar to those used for ECG.

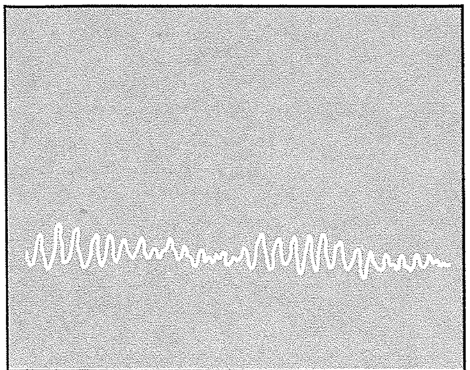


Fig 5. EEG Waveform

MONITORING CONVENIENCE

Note that all three types of signals previously discussed as applicable to the Type 410 are not only among the *most* important indicators of patient well being, but that all are available at the surface of the body.

For maximum monitoring capability and cross correlation between signals, seven electrodes and the pulse sensor may be connected to the patient as shown in Figure 6. Using only the INPUT SELECTOR switch, the user can select the EEG, ECG, or pulse waveform. With a second switch, the ECG LEAD SELECTOR, any of seven standard combinations of ECG electrodes may be chosen.

THE CLINICIAN AND HIS ENVIRONMENT

The Type 410 is of particular value to the anesthesiologist, a medical doctor specializing in anesthesiology. His activities in the operating room go far beyond the administration of anesthetics; he is responsible for monitoring patient well-being, assists the patient's breathing, monitors blood loss and replacement, monitors blood pressure, administers drugs, and in general watches for any unfavorable reaction due to the anesthetic or the surgical procedure.

Instrumentation which can provide some of the needed data can be of considerable value. To provide the anesthesiologist with continuous information, the Type 410 produces an audible "beep" coincident with the most significant event in each cycle of the ECG or pulse waveform. Most doctors and nurses, through experience, will be able to estimate heart rate quite accurately by listening to the "beep" and will most certainly be able to detect poor rhythm. Should a more qualitative determination of heart rate be desired, a quick look at the Heart Rate Scale will suffice. With the LOUDNESS control, the sound can be made audible to the entire surgical team or only to the anesthesiologist.

Several features of the Type 410 combine to insure that a display is available under nearly all circumstances. These features include the elimination of input coupling capacitors so as not to retard recovery from overdrive by high amplitude defibrillator pulses or electrocautery arcs. AC coupling for drift elimination is provided between amplifier stages and includes an overdrive scan limiter for quick recovery.

Automatic sweep triggering circuits, which require no operator controls, seek out the event of dominant amplitude in the ECG or pulse signal, regardless of polarity. If the amplitude of the dominant event should suddenly decrease, the sweep and audio "beep" temporarily stop while the trigger circuits search for lower amplitudes. However useful information continues to be available. The CRT spot will appear at the lefthand edge of the graticule and any available heart signal will cause the spot to bounce vertically. If, within two to four seconds, the triggering circuits have not found a lower amplitude signal, the audio "beep" restarts, sounding at a rapid rate to serve as an alarm.

The operating room presents several unique restrictions to the use of instrumentation. The area immediately surrounding the operating table is often crowded with people and equipment. Certain of these people must move around during the operation and their pathway must not be obstructed by equipment, patient monitoring cables or power cords. *Battery operation* of the Type 410 avoids power cords across the floor.

A suitable location for the Type 410 is available on the anesthesiologist's gas machine. This machine is usually located near the patient's head and is a wheeled cart containing gas cylinders, distribution manifolds, valves, flow gauges, etc. There is often a set of drawers in a cabinet which provides a small table top. Hoses from the gas machine connect to the face mask through which the patient breathes. By mounting the Type 410 on this machine, the patient cable parallels the hoses to the patient and therefore is not an added obstruction to traffic. The instrument is then at a convenient viewing distance for the anesthesiologist and the controls are within easy reach.

An optional mounting fixture is available for mounting the Type 410 to the side of the gas machine so that the much-needed table space is not occupied. The mount can be attached to a flat surface on the side of the drawer cabinet or to one of the vertical pipes used as structural support in some machines. The Type 410 is supported five feet above the floor by the mounting fixture in order to comply with safety regulations, and is a convenient level which permits most members of the surgical team to see the display when desired.

When a surgical operation is completed, the patient must often remain under close observation for several hours. The first stage of observation usually takes place in the recovery room adjacent to the operating room. It is sometimes undesirable to interrupt the electronic monitoring of the patient while moving from surgery to recovery room. The battery-operated Type 410 simply lifts off the mounting fixture and is easily carried along with the patient for continuous monitoring.

MEASUREMENT BARRIERS

The real test of any physiological monitor is the fidelity with which it displays the bioelectric signal. The human body is considerably less than an ideal signal source. The signals of interest are small, about one millivolt. Unless the body is grounded, it usually bears an interfering 60-Hertz signal of several volts which is electrostatically induced by power line sources such as nearby lighting fixtures and appliances. This signal will

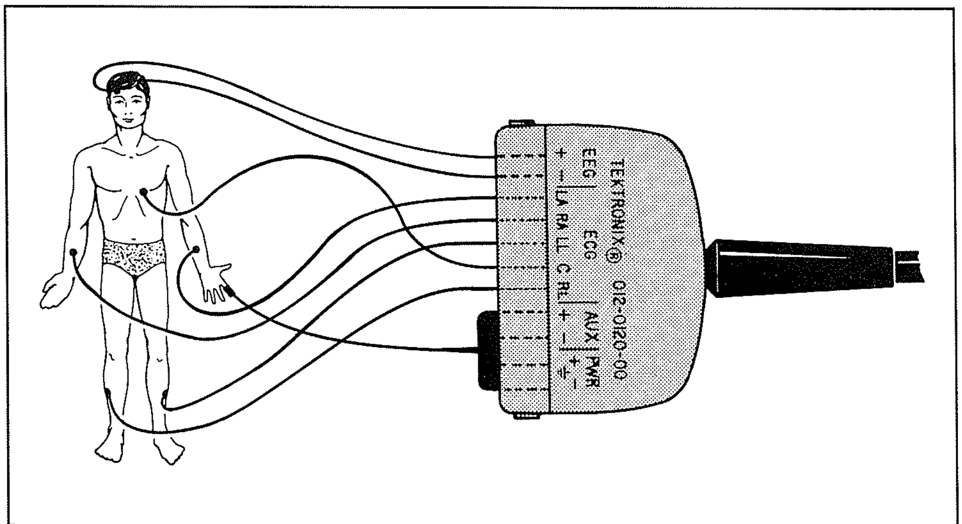


Fig 6. Electrode and Sensor Connections for Cross Correlation of Signals

be common to all active signal leads to the monitoring device and is termed common-mode signal.

The outer layer of skin is of comparatively high resistance and is therefore an undesirable element in the signal path. Moreover, when a metallic electrode is placed upon the body, the body fluids constitute an electrolyte and one-half of a battery is formed. Dissimilarities among the several electrodes on the body can cause a DC voltage to exist between them. But since they form a poor battery, the terminal voltage can vary with patient movement, perspiration, etc. This voltage variation cannot be separated from the desired bio-electric signal and therefore must be eliminated at its source.

Certain desirable characteristics of a physiological monitor can now be described. High skin resistance must be reduced and any voltage difference between electrodes must be small and stable. The monitor should be unaffected by the common-mode interference signal.

The ability of a monitoring system to reject a common-mode signal is often limited by an inability to transport the common-mode signal to the monitor by the two different paths without having the signal arrive at the monitor in dissimilar forms. If this happens, at least part of the signal is no longer in common mode, but has become a differential signal which cannot be rejected by the monitor. This problem can occur due to the skin resistance at each electrode forming an attenuator with the shunt input impedance to circuit ground within the monitor.

It is common practice to use a saline paste under each electrode to impregnate the skin, thus reducing the resistance between the highly conductive body fluids and the electrode. This can reduce skin resistance from a high of perhaps one megohm to as little as a few hundred ohms with careful preparation. However, a more practical degree of skin preparation will result in resistances ranging from one to five kilohms. Since it is highly unlikely that the resistances at the various electrode sites will match, it is probable that dissimilar attenuators will be formed with the monitor input circuitry. A few simple calculations will show that shunt impedances to circuit ground within the monitor of several hundred megohms are required to reduce this effect to an acceptable level.

The Type 410 provides excellent common-mode interference rejection capabilities by actively driving the shunt impedances in the input circuitry. This technique is called "guarding" and effectively multiplies the input impedances to several thousand times their actual values. The common-mode signal therefore arrives at the monitor in a form which permits virtually complete rejection.

The silver/silver-chloride electrodes supplied with the Type 410 eliminate virtually all of the electrochemical problems associated with ordinary electrodes. These small electrodes can be comfortably worn by the patient for many hours at a time. Electrode adapter cables are also provided with the

Type 410 which permit the use of many other standard electrode types.

POWER REQUIREMENTS

The Type 410 obtains power from a removable battery pack in the rear of the instrument. The pack contains ten rechargeable size "C" Nickel-Cadmium cells and a complete line-operated charger. Recharging is started by simply inserting the power cord into the rear of the battery pack. The battery provides eight to twelve hours of instrument operation for each recharge.

When the Type 410 is used in an Intensive Care Unit, continuous operation for days or weeks may be required. This presents no problem. With the power cord attached, the instrument can operate indefinitely because the charging current slightly exceeds the current required by the monitor.

PACKAGING CONCEPTS

The top, bottom, and sides of the Type 410 are rugged aluminum-alloy castings which provide an easily cleaned, dust-tight cabinet. The monitor weight is only 12½ pounds including the battery pack. Eighteen handle positions are provided for carrying or for tilting the instrument to the best viewing angle. The handle hub is specially shaped to fit into a cup-shaped bracket on the mounting fixture (Fig. 7).

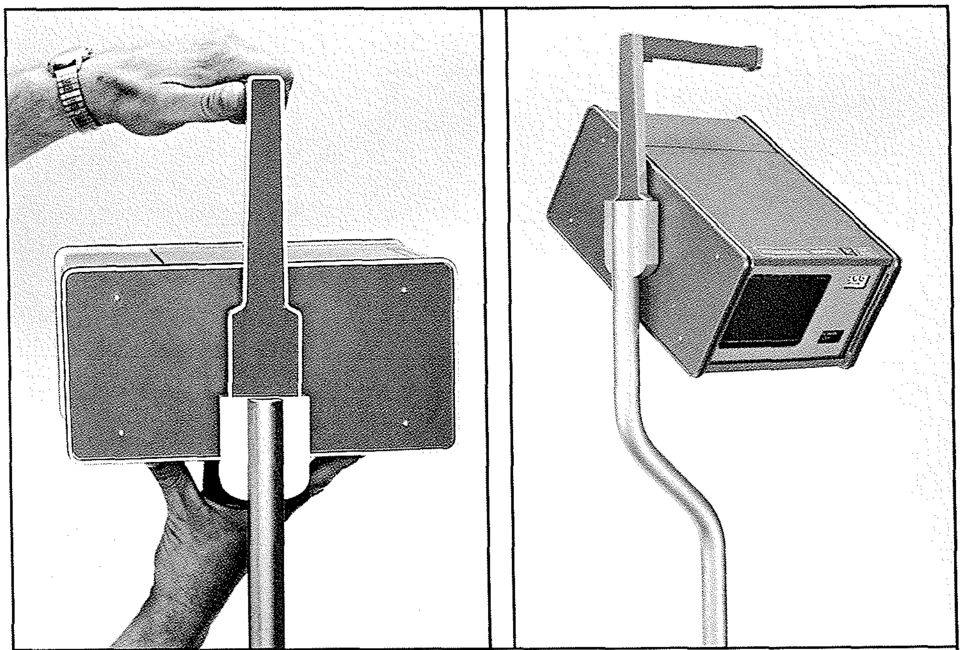


Fig 7. Attaching the Type 410 to the Mounting Stand

SUMMARY

The Type 410 Physiological Monitor was designed for patient surveillance. Only a few of the many necessary considerations have been discussed here; the physiological signals, the intended user, the environment, the fidelity of signal display, etc. The area of possible application is broad.

Output signals available from the rear of the monitor can drive a recorder to provide a permanent record as is usually required in diagnostic applications. This portable, simple-to-operate instrument will also find application in medical research with both people and animals, as well as in Veterinary Medicine.

TYPE 410 PHYSIOLOGICAL MONITOR CHARACTERISTIC SUMMARY

VERTICAL

Bandwidth

ECG and AUX	≤0.1 Hz to 250 Hz ±15%
EEG	≤0.1 Hz to 100 Hz ±15%

Calibrated Deflection Sensitivity

		Accuracy	
Display mode	Deflection Sensitivity	≤20 mV DC offset	At 100 mV DC offset
EEG	10 mm/50 µV	±5%	0 to −10%
ECG	20 mm/mV	±5%	0 to −10%
AUX	2 mm/mV	±5%	0 to −10%

Vertical Size Range ≤ X1/3 to ≥ X3

Differential Input Resistance

EEG and ECG	2 MΩ ± 15%
AUX	20 MΩ ± 15%

Differential Dynamic Range

At least 100 mV of either polarity

Common Mode Rejection Ratio

With ≤5-kΩ Source Impedance unbalance (at 60 Hz) and using properly applied electrodes.

EEG	≥ 150,000:1
ECG	≥ 150,000:1
AUX	≥ 150,000:1

Common-Mode Dynamic Range

+3 V to −3 V

Drift

≤0.5 cm/h (after 10 s warm-up)

Nondestructive Input Voltage Limits

Instrument need not be disconnected from patient during DC defibrillation or cautery

Differential Overload Recovery Time

≤4 seconds (all cases)

TRIGGER

Trigger Requirements

0.5 cm ECG display (≥ 40 beats/min)
0.5 cm blood pulse display (≥40 pulsations/minute)

Delay Before Sweep Free-runs

2 to 4 s after last trigger

HORIZONTAL & AUDIO

Sweep Speed

25, 50, 100 mm/s ±5%

Battery Check Scale

Green—Normal Operation
Yellow—Recharge needed
Operation not harmful to instrument
Red—Do not operate

Heart Rate Scale Accuracy

±5% of reading (50 mm/s range, 35 to 110 beats/min)

Audio

Audio "Beep" at heart rate with alarm activated upon loss of signal

POWER SOURCE

Line Voltage

90 V to 136 VAC
180 V to 272 VAC

Line Frequency

48 Hz to 440 Hz

Battery Operating Range

11.9 V to 15.0 V

AC Input Power

≤7 W at 115 V, 60 Hz

Battery Pack

Ten Size "C" NiCd. cells; 1.8 Ah

Charging Time

14 to 16 hours

Discharge Time

8 to 12 hours operation with maximum accessory load at +20° to +25° C

OTHER

Turn-on time

≤4 sec

Warm-up time

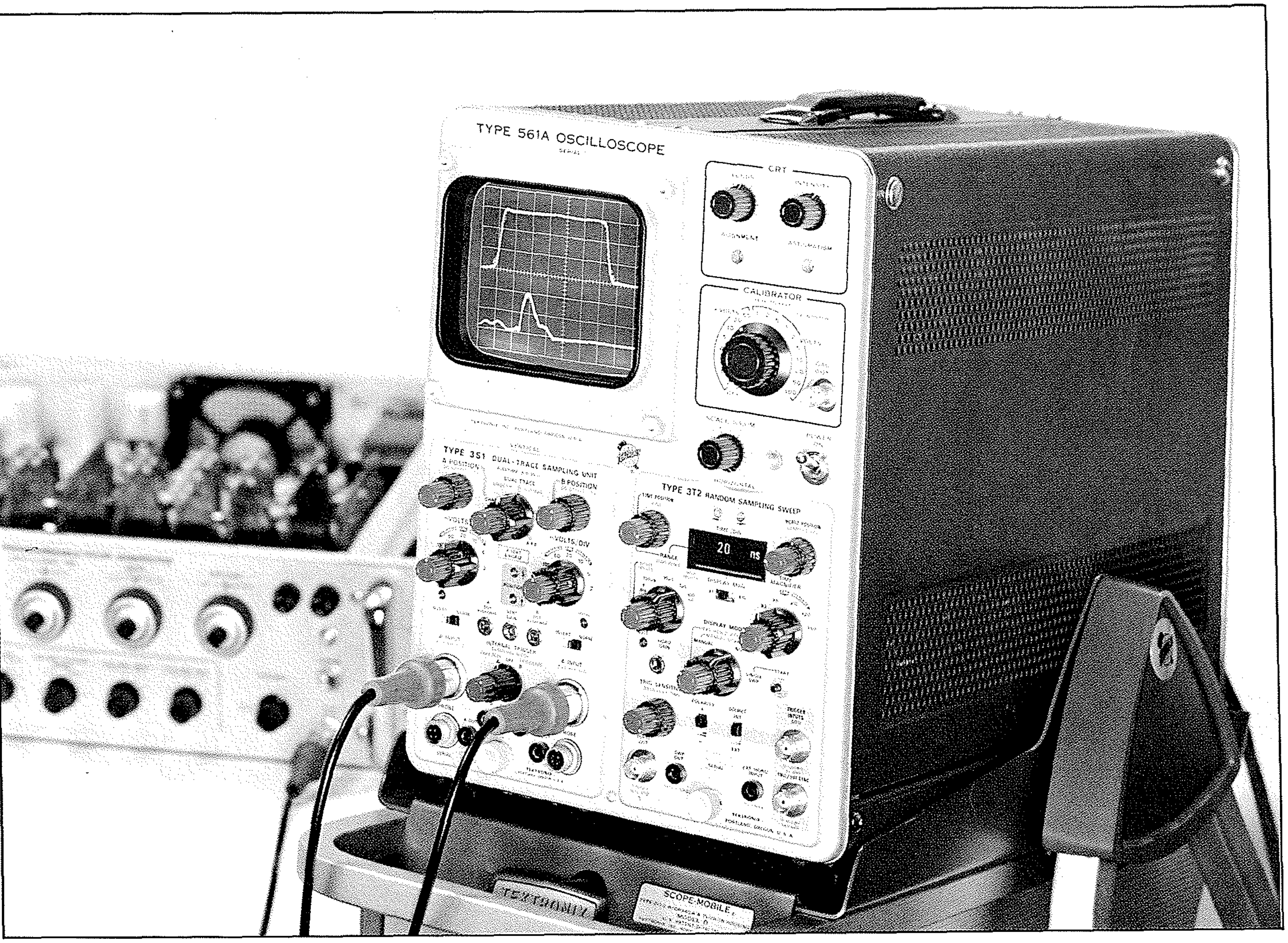
≤10 sec

CRT

5" with P-7 phosphor

RANDOM SAMPLING—
A NEW WAY OF FAST PULSE DISPLAY

by Al Zimmerman



INTRODUCTION

The Type 3T2 Random Sampling Sweep Unit provides a unique state of the art advancement in measurement capability. It permits observation of the leading edge or other portions of the signal even when used with vertical units that have *no delay lines and without a pretrigger.*

The advantages of eliminating delay line or pretrigger application are evident:

1. The inherent distortions and risetime limitations of signal delay lines are eliminated.

2. It is no longer necessary to work into the 50-Ω characteristic impedance of a delay line, so that direct sampling probes may be used for convenient high-impedance in-circuit signal pickup.

3. Trigger may occur prior to, coincident with, or after the displayed signal without sacrificing lead time in the display.

4. Signals with no convenient source of a stable pretrigger can be observed without display jitter.

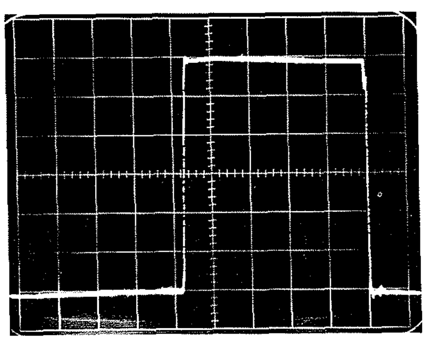


Fig 1A. 3T2 Unit in "BEFORE TRIGGER" mode: the leading edge can be observed.

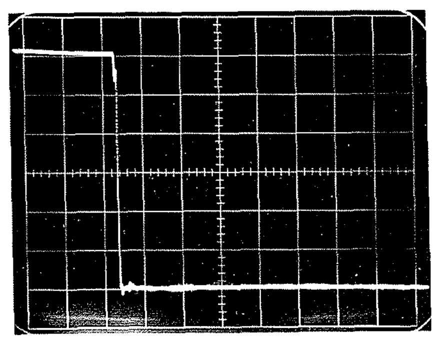


Fig 1B. 3T2 Unit in "WITH TRIGGER" mode (conventional sampling): the leading edge cannot be observed without a delay line or pretrigger.

HOW RANDOM SAMPLING WORKS

In the following explanation of the principles of the Random Sampling process, (how it is used in Tektronix Type 3T2 Plug-in Unit) an understanding of conventional sampling is advantageous.

The Random Sampling process is composed of two basic operations:

1. Originating the sample pulses randomly distributed in a time window around the part of the signal to be displayed.

2. Constructing a pulse display by deriving two analog signals, representing X and Y coordinates, from a series of those samples.

ORIGINATING THE SAMPLE PULSES

To find the right time for originating the samples a "Trigger Rate Meter" is used, which measures the trigger repetition rate. This rate meter gets its input from the trigger recognizer and holdoff circuit. On the basis of several sequential trigger rate measurements, the rate meter starts a negative going signal (slewing ramp) which generates samples within the time window. The start time of this slewing ramp is *before* the next trigger signal (ΔT) and is a forecast resulting from the previous trigger-rate measurements. So it can be seen that the start time of the slewing ramp is not in a fixed time relation to the next trigger signal, but more the "best guess" of the rate meter.

The display thus becomes a random sampling display because of the inability of the rate meter to make a perfect "guess" of when to take a sample.

The rate meter provides maximum display dot density by gathering the samples around just that section of the waveform that is used for the CRT display (see Fig. 2). This time window can be a very small portion of the total signal period. The samples which fall outside the time window do not have any contribution to the display construction and are kept as few as possible.

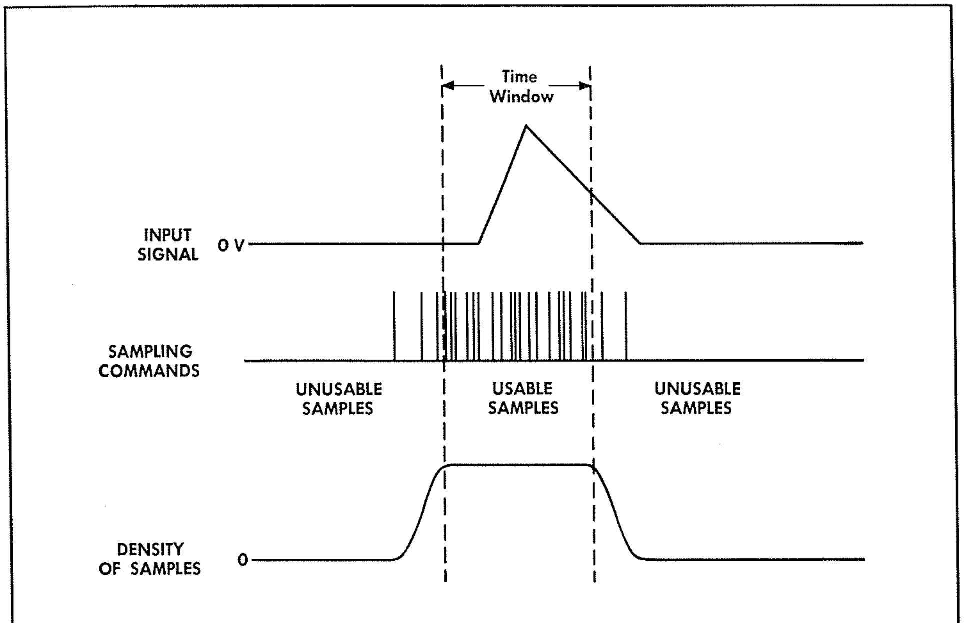


Fig 2. Distribution of samples over a time window, which covers the section of the signal to be displayed.

Error Correction

If too many samples fall outside the time window, on either one side or the other, a correction of the rate meter "guess" has to be made. The principle of that correction is based on a comparison (dot position comparator) of the horizontal signal with the staircase signal. An error signal is generated when the horizontal signal does not track along with the staircase on a basis of an average of many samples. The error signal adjusts the rate meter to make a better "guess" or forecast for the next start of the slewing ramp. Fig 3 shows the correction loop.

CONSTRUCTING A PULSE DISPLAY

In order to be displayed, a sample must have a particular time relationship to the sampled pulse. The rate meter has tried to place each sample within the time window. If a sample does occur in this time window, a dot is displayed on the CRT.

The "Y" or vertical coordinate of a sample is obtained by the same sample-and-hold process used in a conventional sampling oscilloscope. The "X", or horizontal, coordinate of the sample is obtained differently, however, and this process is illustrated in Figure 4.

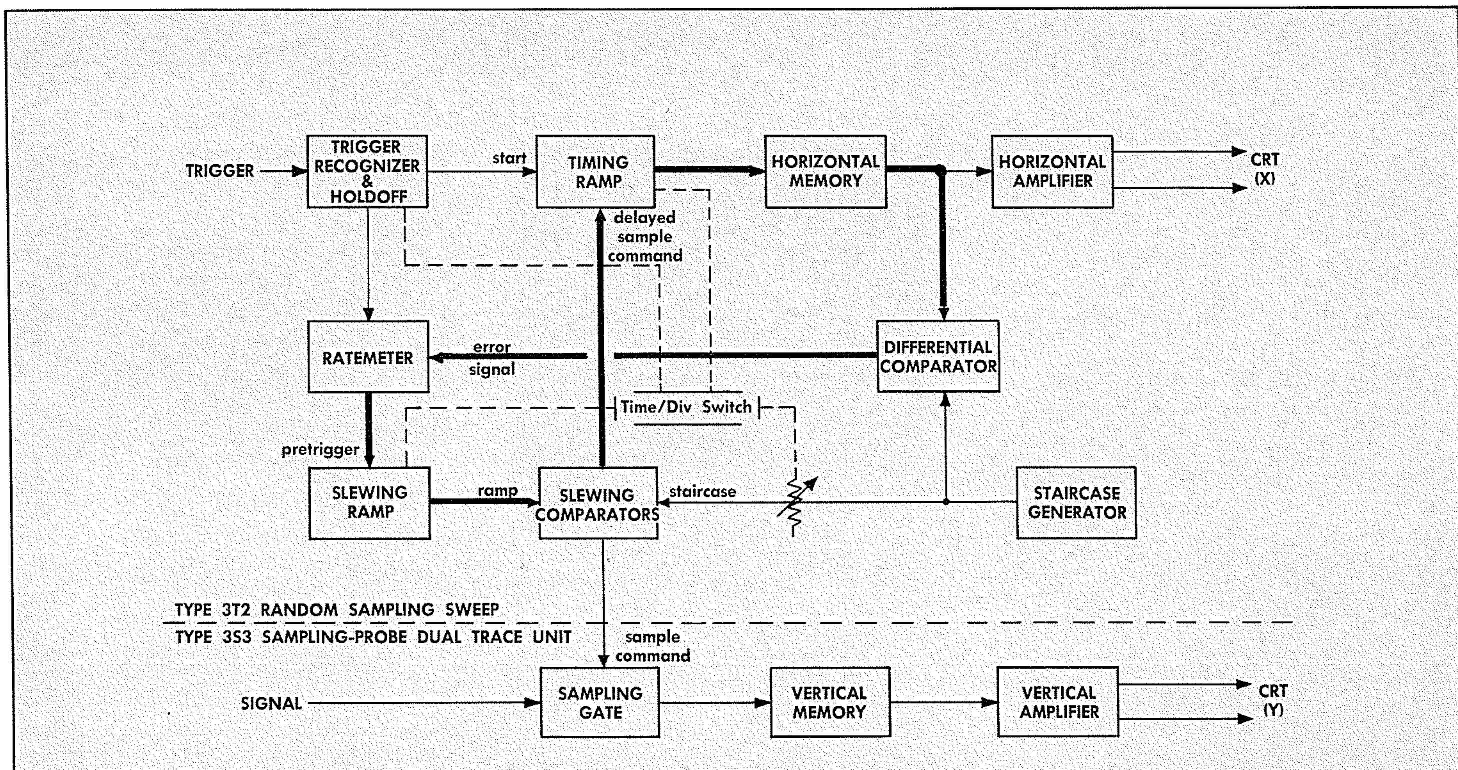


Fig 3. Block diagram of the Random Sampling process used in Tektronix Type 3T2. The dark line shows the correction process for the slewing ramp start.

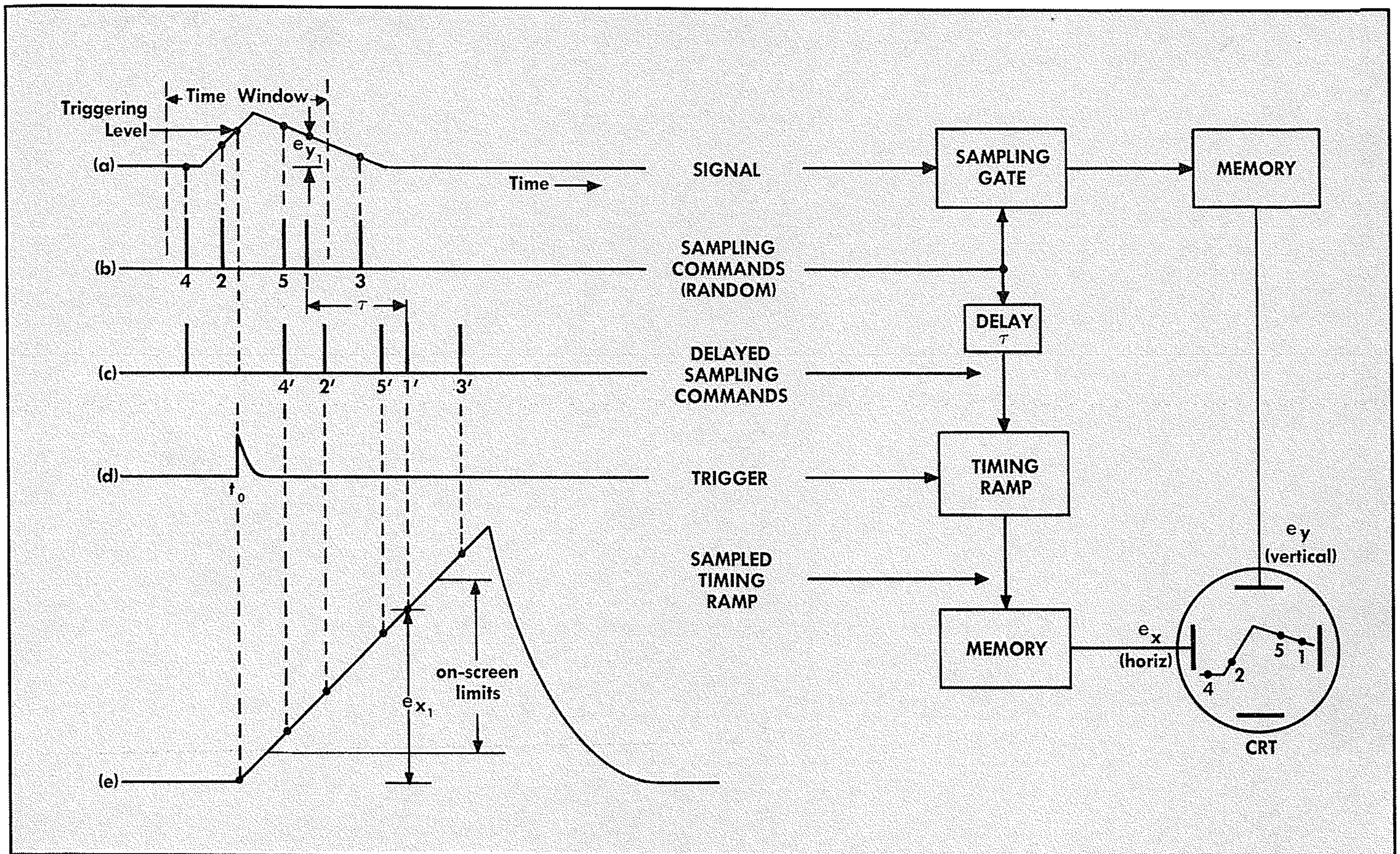


Fig 4. Derivation of "x" and "y" deflection signals.

As shown, five randomly placed samples are taken of the signal. It must be kept in mind that these five samples are taken on SUCCESSIVE repetitions of the signal. They are random samples in the "best guess" time window.

The y-component, e_y, of the first sample is held and subsequently used to position the CRT spot vertically. The sampling command which took the first sample is then delayed by a *fixed interval* τ, as indicated in Figure 4c. This delayed sampling command 1′ is used to sample a timing ramp which was started by trigger recognition along the input signal at t_0. The resulting sample e_x is held and subsequently used to position the CRT spot horizontally.

By this same process subsequent samples supply both vertical and horizontal information to deflect the CRT beam from dot to dot thus constructing a display of the signal from those samples which fall within the time window.

Some reflection will show that as the fixed interval τ is increased, more lead time will appear in the display. It should be clear that such an increase in τ for more lead time will also require a time shift of the sampling distribution to the left in Figure 4b (i.e. earlier in time) in order that the required information be collected for the display.

Figure 3 shows a complete operational block diagram of the random sampling oscilloscope including those portions which control the distribution of samples across the time window.

CHARACTERISTICS

SWEEP TIME/DIV

100 μs/div to 200 ps/div, 1-2-5 sequence, extending to 20 ps/div with X10 DISPLAY MAGNIFIER. Basic accuracy without X10 magnifier, ±3%; with magnifier, ±5%. TIME/DIV is a resultant of the combined settings of TIME POSITION RANGE, TIME MAGNIFIER, and DISPLAY MAG. The sweep rate is displayed (digitally) in the TIME/DIV "window" for all combinations of these controls.

TIME POSITION RANGE

100 ns, 1 μs, 10 μs, 100 μs, and 1 ms. TIME POSITION and FINE variable controls position start of the display through a time scale equal to TIME POSITION RANGE setting.

SAMPLES/DIV

Continuously variable adjustment of samples displayed per horizontal division from approx 5 samples/div to an immeasurable number of samples/div.

An internal switch, CALIBRATED SAMPLES/DIV, disables the front-panel SAMPLES/DIV control and converts to 100 samples/div, calibrated, for use in Digital Oscilloscopes.

START POINT

Two-position switch (concentric with TIME POSITION RANGE switch) selects either random sampling (BEFORE TRIGGER) or conventional, sequentially-stepped sampling (WITH TRIGGER).

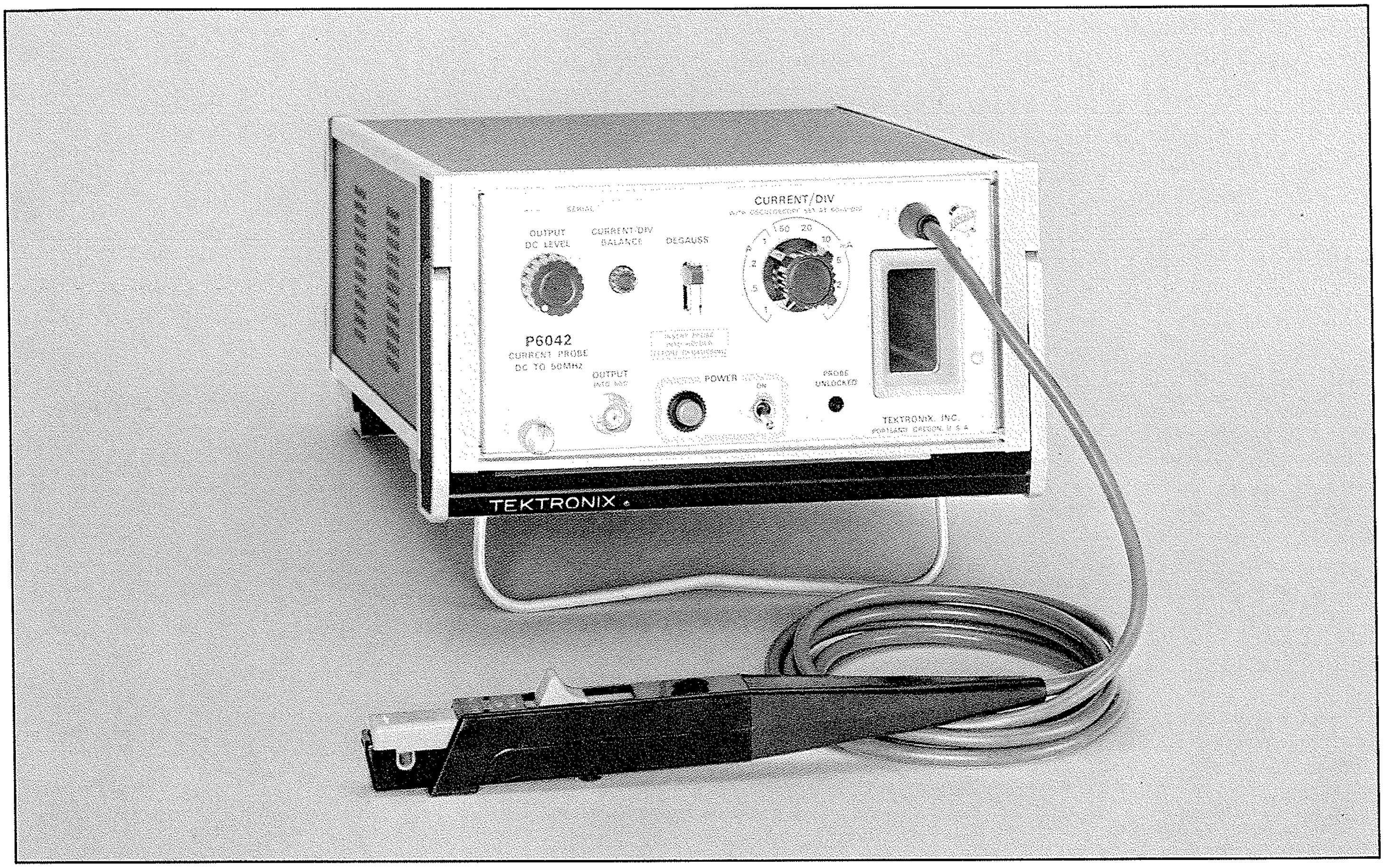
In BEFORE TRIGGER mode, the displayed "time window" may be positioned in time up to one-half times the TIME POSITION RANGE setting ahead of the trigger. This provides a base line up to 5 divisions long before the leading edge of the pulse to be viewed.

TRIGGER JITTER

Depends on signal shape, repetition rate, triggering mode. May be as low as 30 ps under optimum conditions.

P6042 DC-to-50 MHz CURRENT PROBE

by Cal Hongel



INTRODUCTION

Current probes have become increasingly useful and popular with the expanding use of semiconductor devices which are current sensing devices (current amplifiers). A new current probe has just been developed at Tektronix that provides unique measurement capabilities.

Utilizing the Hall-effect plus AC current probe technology (P6019/P6020), the P6042 DC-to-50 MHz current probe can be used simultaneously for both high-frequency and direct-current measurements. AC signals with DC components can be displayed on an oscilloscope with true waveform presentation. The probe is particularly useful for evaluating the performance of semi-conductor circuits where a wide range of parameters exist. Fast switching transients, low-frequency response, and DC level can all be displayed simultaneously (Figure 1). The P6042 can also be used to measure the sums or differences of currents in separate wires. When the probe is clipped around two wires carrying current in the same direction, the sum is displayed; around two wires carrying current in the opposite direction, the difference is displayed. For increased sensitivity the wire can be looped through the probe several times increasing the sensitivity by the number of loops.

The probe is easy to use. The conductor is simply placed into the slot of the probe and the spring loaded slide closed . . . no need to break the circuit under test. Measurements can be made only when the probe is in the locked position (push slide forward to lock). A warning light on the front panel indicates when the slide is in the unlocked position. A compartment is provided in the front panel for convenient storage of the probe when the system is not in use and an inter-lock is provided in this compartment for degaussing the probe. The probe can be degaussed only when in the compartment to prevent possible damage to the circuit under test.

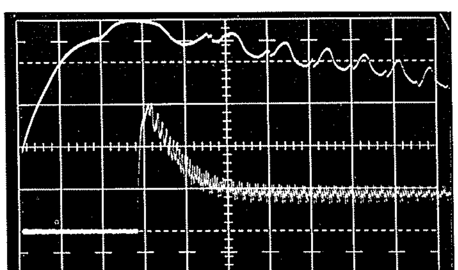


Fig 1. Double exposure photograph using the P6042 and a Type 547/1A5 Oscilloscope to display the current characteristics of a small DC motor. Lower display shows the zero current level, starting current, and running current. Current/div setting is 0.2 A/div with a sweep rate of 50 ms/cm. In the upper display, the sweep rate is increased to 5 ms/cm to show the current change as the commutator bars pass the brushes.

DESIGN CONCEPTS

The P6042 Current Probe includes a sliding-core type probe and associated amplifier as shown in Figure 2. The probe contains a stationary core around which is wound a 50-turn secondary, a moveable core which slides over the end of the stationary core and the current-carrying conductor, and a Hall voltage device. The amplifier houses the power supplies, low-frequency amplifiers, attenuators and the output amplifier.

High Frequency

High-frequency measurements are made in the same manner as in an AC current probe. The AC current probe is basically a transformer. The current-carrying conductor forms a one-turn primary winding for the transformer; the windings in the probe around the core form the 50-turn secondary winding. The relationship between the current, voltage and turns is shown below:

$$N_p I_p = N_s I_s$$

For a one-turn primary,

$$I_p = N_s I_s$$

Then for a 50-turn secondary,

$$I_s = \frac{I_p}{50}$$

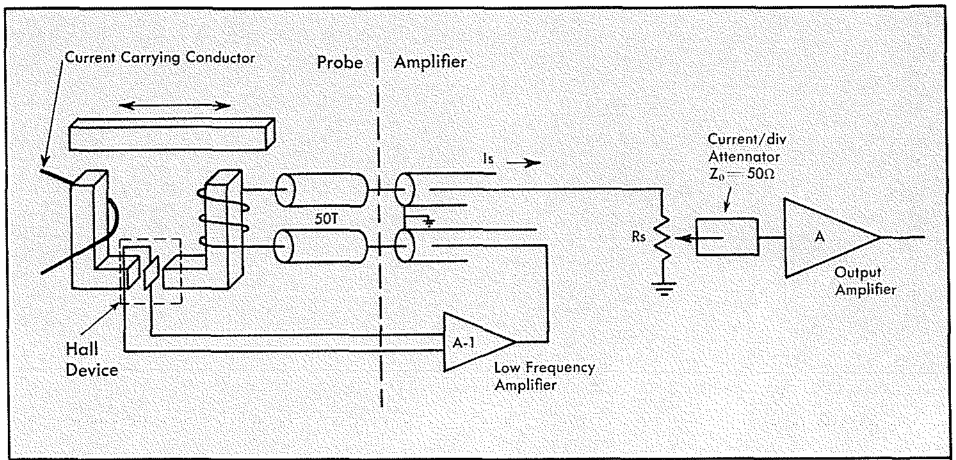


Fig 2. Block Diagram of P6042 Probe.

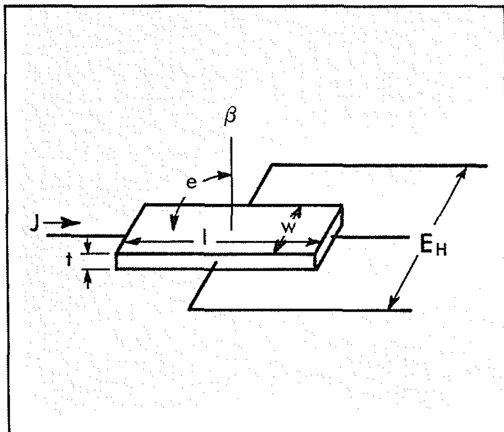


Fig 3. Hall device.

The Hall device is a thin rectangular sheet of semiconductor material sandwiched in the stationary portion of the transformer core. The Hall effect is a voltage generated across opposite edges of a current carrying conductor placed in the magnetic field.

The basis for the effect is the Lorentz force which is the deflection of charged particles moving in a magnetic field. The force is both perpendicular to the direction of the particle (current) flow and the direction of the magnetic field. Equation follows:

$E_{Hall} = wR_H\, J\, \beta \sin\phi$

E_{Hall} = voltage from Hall device
w = width of Hall element
R_H = Hall coefficient
J = current density
β = field strength

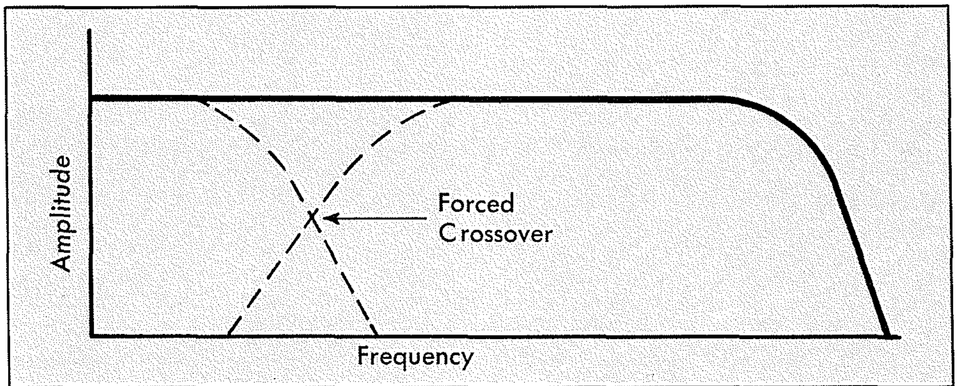


Fig 4. Forced crossover.

The secondary voltage is

$$E_s = I_s\, R_s$$

R_s is 50 Ω, so $E_s = \frac{I_p}{50}\ (50\,\Omega)$

or $E_s = I_p\ (\Omega)$

For AC signals the voltage output of the current probe into the secondary load (R_s) is 1 mV per mA of input current.

DC and Low Frequency

The heart of the DC measurement capability is a highly-sensitive Hall device developed by the Tektronix Integrated Circuit Department. The Hall device is located in a cross section of the ferrite core contained in the probe head. At the point where the AC response of the core becomes ineffective due to the low-frequency L/R time constant of the core, the back EMF of the secondary no longer cancels the flux generated in the core by the primary current. The flux remaining in the core (primarily flux due to DC and low-frequency current) passes through the Hall device generating a small voltage directly related to the applied field. Figure 3 shows the current, voltage, and flux relationship of a Hall device.

The Hall device voltage (about 50 μV per mA of applied current) is amplified by the operational amplifier (A-1) and applied to the 50-turn secondary, to cancel the remaining flux in the core. Most of the flux in the core is cancelled either by the back EMF of the secondary or by feedback from the operational amplifier. As a result, the non-linearity of the core does not affect accuracy, nor does it directly limit maximum current handling ability. At DC and low frequencies, the operational amplifier supplies an output across the secondary load (R_s) of 1 mV per mA of primary current.

The maximum input current is related to the current handling ability of the operational amplifier. To handle ± 10 A in the primary, the amplifier (A-1) must supply $10\,A \times \frac{1}{50}$ to the 50-turn secondary to cancel the flux at DC and to supply ± 200 mA across R_s.

Attenuator and Output Amplifier

The current induced in the secondary by the primary (at high frequency) and the current applied to the secondary at low frequency by amplifier (A-1), produces a voltage across the secondary load that is directly related to the input current. The adding of the low-frequency signal to the high-frequency signal is done in such a way as to force one to take over where the other leaves off (see Figure 4). This is commonly known as a forced complement system.

The sensitivity at this point is 1 mV output for a 1 mA of primary current (input current). The 50-Ω secondary load is in the form of a 50-Ω attenuator that provides attenuation of up to 1000X (1 A/div) in 10 steps with a 1-2-5 sequence. The signal from the 50-Ω attenuator is applied to a 50X DC-to-50 MHz output amplifier. The output amplifier supplies an output of 50 mV per mA of primary current or 1 mA/div with the oscilloscope deflection set at 50 mV/div.

The P6042 output amplifier has an output impedance of 50 Ω. A 50-Ω termination is supplied with the P6042 probe for use with oscilloscopes having 1-MΩ inputs.

CIRCUIT LOADING

All probes load the circuit under test in one form or another. Voltage probes have input capacitance and DC resistance. Current probes load in a different manner. They have an insertion impedance due to the secondary load being reflected into the primary and very low-capacitive loading.

Reflected Load

The secondary inductance and load resistance is reflected through the turns ratio squared and appears as a series load in the primary (current-carrying conductor). Calculations of the typical reflected loading of P6042 current probe is shown below:

$$R_p = \frac{R_s}{T^2} = \frac{50\,\Omega}{(50)^2} = 0.02\,\Omega$$

$$L_p = \frac{L_s}{T^2} = \frac{0.5\,\text{mH}}{(50)^2} = 0.2\,\mu\text{H}$$

Shield Inductance

Another factor affecting circuit loading is the reflection of the current probe shield into the current carrying conductor. The shield appears as a shorted turn around the conductor. Leakage inductance also appears in series with the primary.

Stray Capacitance

The only other factor involved with circuit loading is the stray capacitance between the probe and the conductor. This capacitance depends on the size of the current carrying conductor and its position in the hole. It is typically 1 pF and can be measured using a Type 130 LC Meter. As with voltage probes, stray capacitance can limit the risetime of the measurement ($T_{rise} = 2.2\ R_{source}\ C_{strays}$). By inserting the current probe on the ground or B+ side of the load resistor the stray capacitance loading can be reduced.

The total insertion impedance can best be represented by the graph in figure 5.

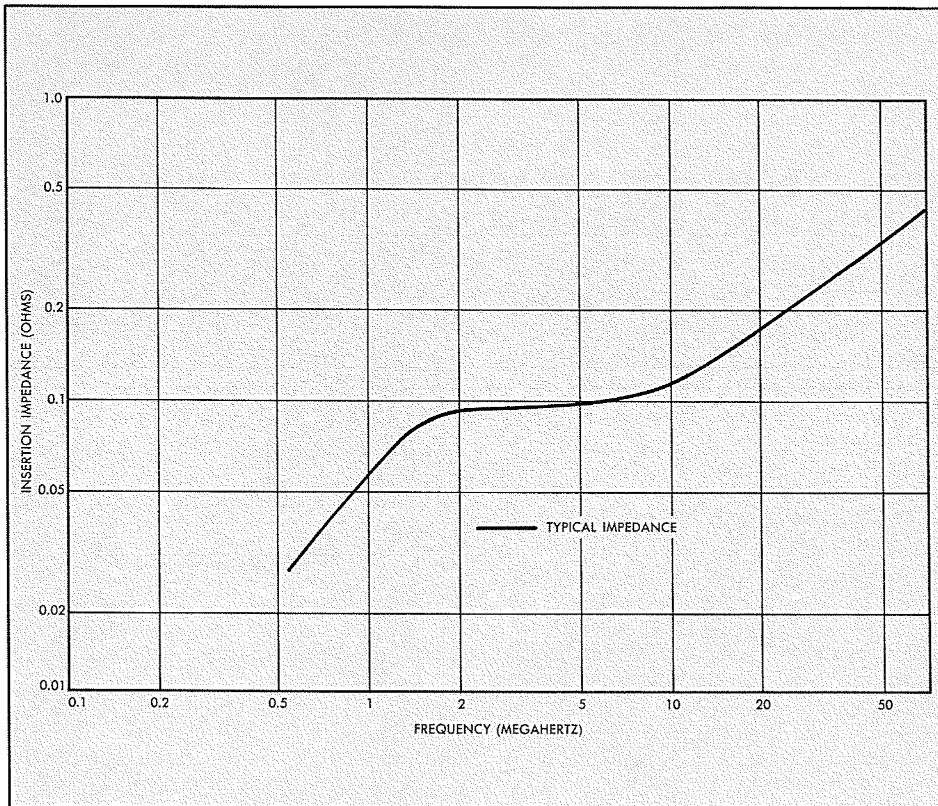


Fig 5. Insertion impedance.

PROBE DEGAUSSING

Whenever a magnetic field is applied to the transformer core in the probe with the system turned off, or if a current beyond the maximum specified level is applied, the core may become magnetized. A portion of this magnetic flux is likely to remain in the current probe core causing measurement errors. To remove this flux the probe is placed in the storage compartment and the degaussing switch is depressed. The degaussing switch connects the 50-turn secondary winding to an oscillator as shown in Figure 6. The oscillator produces a 10-kHz exponentially decreasing sinewave which initially saturates the core. The decaying current eliminates stored flux due to core hysteresis.

An interlock switch for the degaussing oscillator is provided in the probe storage compartment. The switch eliminates any possibility of introducing transformed current from the oscillator into the test circuit. The compartment, accessible from the front panel, provides convenient storage for the probe when not in use.

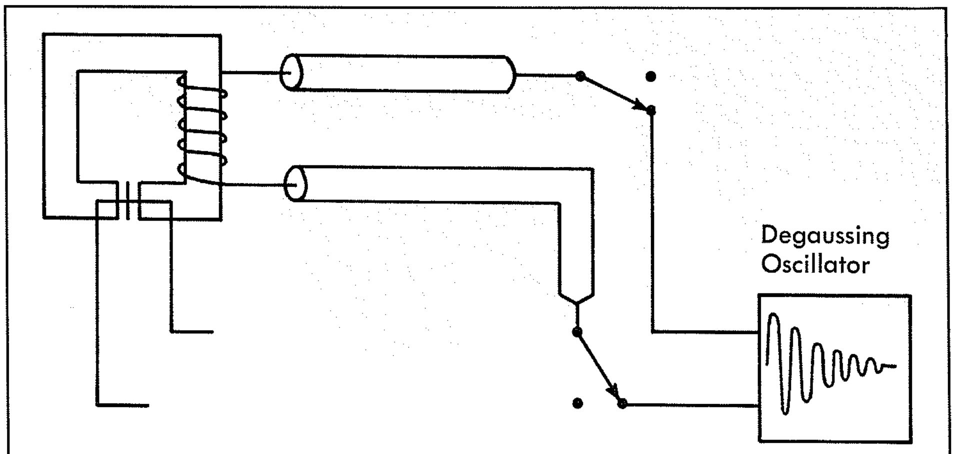


Fig 6. Probe degaussing.

CHARACTERISTICS

Probe and Amplifier

SENSITIVITY is 1 mA/div to 1 A/div in 10 calibrated steps, 1-2-5 sequence, accurate within 3% (with an oscilloscope deflection factor of 50 mV/div).

BANDWIDTH is DC to 50 MHz at 3-dB down.

RISETIME is 7 ns or less.

DYNAMIC RANGE is + and − 10 divisions of display.

NOISE (periodic and random deviation) is 0.5 mA or less plus 0.2 or less major divisions of display. Random trace shift is 1.5 mA or less.

THERMAL DRIFT is 2 mA/°C or less, plus 0.2 or less major division of display per °C.

MAXIMUM INPUT CURRENT is 10 A (DC plus Peak AC).*

*Peak-to-peak current derating is necessary for CW frequencies higher than 2 MHz. At 50 MHz, the maximum allowable current is 2 A.

MAXIMUM INPUT VOLTAGE is 600 V (DC plus Peak AC).

OUTPUT IMPEDANCE is 50 Ω through a BNC-type connector. A 50-Ω termination is supplied with the probe for use with 1-megohm systems.

AMPLIFIER POWER REQUIREMENT is approximately 10 W, 50 Hz to 400 Hz. Quick-change line-voltage selector permits operation from 90 V to 136 V or 180 V to 272 V.

DIMENSIONS AND WEIGHT of the amplifier are 4½ in. (11.4 cm) high by 7½ in. (19.2 cm) wide by 9¾ in. (24.8 cm) deep; 6½ lbs. (3.1 kg).

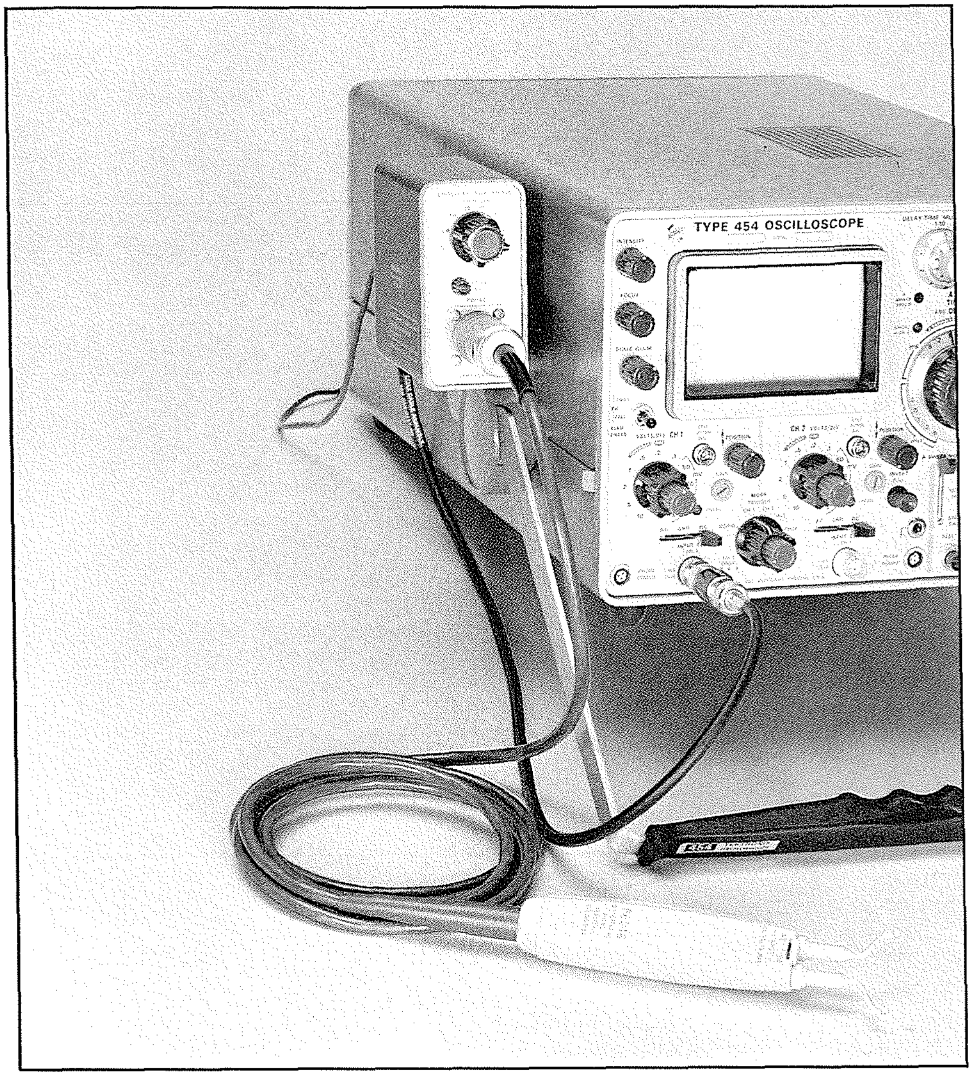
PROBE CABLE is 6 feet long, permanently connected between the probe head and amplifier.

P6042 DC CURRENT PROBE PACKAGE (010-0207-00)

Includes: 50-Ω BNC cable (012-0057-01); 50-Ω BNC termination (011-0049-00); 3-inch ground lead (175-0263-00); 5-inch ground lead (175-0124-00); two alligator clips (344-0046-00); 3-wire to 2-wire adapter (103-0013-00); instruction manual (070-0629-00).

P6046 DC-to-100 MHz DIFFERENTIAL PROBE AND AMPLIFIER

by Glenn Bateman



INTRODUCTION

The P6046 Differential Probe and P6046 Amplifier Unit provides new measurement capabilities when used with all Tektronix oscilloscopes. With this new probe system, the differential-signal processing takes place in the probe itself, resulting in high common-mode signal rejection at higher frequencies. Differential probe-tip signal processing minimizes the measurement errors caused by differences in probes, cable lengths, and input attenuators. In addition, the wide-band capability of the P6046 Probe and Amplifier provides DC-to-100 MHz single-ended measurement performance.

The P6046 probe circuitry utilizes 13 semiconductors including dual FET's for the balanced input. A switch on the probe selects AC or DC input coupling. Accessories include a plug-on 10X attenuator for increasing the differential input voltage range, and a ground tip for applications requiring single-ended input. Unique swivel tips provide variable spacing to accommodate the varying distance between test points.

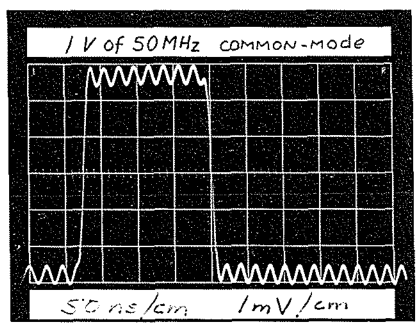


Fig 1. High-frequency differential measurement.

The P6046 Amplifier mounts conveniently on the side of the oscilloscope and features a calibrated 1 mV/div to 200 mV/div (2 V/div with 10X attenuator) deflection factor (oscilloscope deflection factor set at 10 mV/div). Output impedance of the amplifier is 50 Ω. A 50-Ω cable and termination is supplied with the amplifier for use with 1-MΩ systems.

DIFFERENTIAL AMPLIFIERS

The primary use of differential amplifiers is to measure the signal difference between two points that need not be referenced to ground.

An oscilloscope differential amplifier is a device that amplifies and displays the voltage difference that exists at every instant between signals applied to its two inputs. For example, two pulses that differ in both amplitude and coincidence that are applied to a differential amplifier will cause the oscilloscope display to be a complex waveform that represents the instantaneous difference between the two pulses. On the other hand, two signals that are identical in every respect will cause no output on the CRT screen (limitations to this statement will be described under Common-Mode Rejection).

The amount of difference signal due to common-mode signal that one can expect from a particular differential amplifier is specified by the common-mode rejection ratio (CMRR). This ratio and associated terms are defined as follows:

Common-Mode Signal—The instantaneous algebraic average of two signals applied to a balanced circuit, all signals referred to a common reference.

Common-Mode Rejection—The ability of a differential amplifier to reject common-mode signals.

Common-Mode Rejection Ratio (CMRR)—The ratio of the deflection factor for a common-mode signal to the deflection factor for a differential signal.

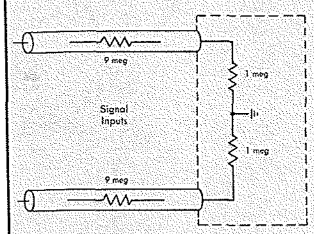
Differential Signal—The instantaneous, algebraic difference between two signals.

Measurements made with a differential amplifier should contain an allowance for the output voltage that is due to a common-mode signal. For example, if an amplifier with a CMRR of 1,000:1 is used to measure the difference between two similar five-volt signals, the output seen on the oscilloscope screen is the result of two voltages: (1) the actual difference between the input signals, and (2) the difference voltage that results from the common-mode signal. Because of this combination, the actual difference voltage cannot be exactly measured. Therefore, the voltage measured on the CRT screen should include a tolerance that is equal to the computed, or measured output voltage due to the common-mode signal.

In the above example, CMRR of 1,000:1 with a common-mode signal of 5 V, if a difference signal of 0.015 V is measured on the CRT, it should be recorded as 0.015 V ±0.005 V.

MEASUREMENT PROBLEMS

The major difficulty in making differential measurements is in connecting the signal source to the measuring device. Measurement errors can be caused by differences in probes, cable lengths, and input attenuators.



Probe	Tolerance	CMRR (Apparent)
A	+1%	50:1
B	−1%	50:1
C	+2%	25:1
D	−2%	25:1
E	+3%	16:1
F	−3%	16:1

Fig 2. Simplified input circuit and table that shows the change in CMRR (apparent) due to X10 probes that have a 1, 2, and 3% difference in their attenuation value.

Probes

Attenuator probes extend the usable voltage range of a differential amplifier by reducing the input signals to a level that is below the maximum common-mode linear dynamic range. In doing this, however, the probes may cause a reduction in the apparent CMRR due to component value differences within the probes. For example, Figure 2 illustrates the change in CMRR (apparent) due to X10 probes that have a 1, 2, and 3% difference in their attenuation value. Bear in mind that the reduction in apparent CMRR can also be caused by different values of the signal source resistance.

Cable Length

Probes and cables of different lengths may introduce enough signal delay between them to cause a difference voltage at the input to the amplifier. At 50 MHz, an 0.1-inch difference in cable length will reduce the CMRR from 1,000:1 to 250:1. Also an inductance difference due to a ⅛-inch difference of lead length at 50 MHz can reduce the CMRR from 1,000:1 to 400:1. Processing the differential signal in the probe reduces these problems.

Input Attenuators

To minimize measurement errors due to input attenuators, the P6046 Probe and Amplifier provides attenuation by reducing the differential gain and common-mode gain within the amplifier. High-frequency common-mode rejection is difficult to obtain when using input attenuators. This is due to the stray capacitance that is distributed along the resistor length resulting in an infinite number of RC time constants that cannot be compensated for over a wide frequency range.

The dual 10X attenuator head included with the P6046 is calibrated with the P6046 to provide maximum CMRR. The attenuator head is keyed with the probe so that the + and − inputs are always matched. The attenuator head should be used only when necessary as it will reduce the CMRR at 50 MHz from 1,000:1 to ≈50:1.

Source Impedance

As the signal source impedance increases, the common-mode measurement problems increase. If the source impedance of the two signals to be measured is different, the CMRR will change due to the different ratios between the source impedance and the input impedance. At high frequencies an increase of source impedance will magnify the problems of CMRR measurements due to a mismatch of stray capacity.

DESIGN CONCEPT

The design objective was to develop a system that would overcome most of the differential measurement problems of high frequency differential amplifiers. The solution was to process the differential signal at the signal source, thereby eliminating most of the problems caused by probes, cable length and input attenuators. It was also necessary to obtain good common-mode rejection for high-frequencies of reasonable voltage levels. These capabilities had to be built into a reasonably small and convenient-to-use probe.

P6046 Probe

In order to obtain high common-mode rejection ratios at higher frequency, the design is a departure from conventional input systems using emitter followers, bootstrapped for all frequencies into the differential comparator. This approach is limited at high frequencies by the bootstrap system.

The input comparator of the P6046 rejects the common-mode signals directly without using an emitter-follower input stage. The thermal time constants of the dual FET's limits the low-frequency CMRR. To eliminate these problems, the input comparator is bootstrapped only for low frequencies (DC-to-100 kHz).

The input-comparator FET's operate with a gain of 0.4. This low gain permits a larger differential dynamic range and a wider bandwidth. An amplifier in the probe restores the probe gain to unity. The probe has one differential attenuator circuit that is controlled from the amplifier and reduces the differential gain and common-mode gain to 1/10.

P6046 Amplifier

Gain changing and converting from a differential-signal to single-ended operation is accomplished in the P6046 Amplifier. Gain changing in the P6046 Amplifier eliminates the differential measurement problems associated with input attenuators.

The P6046 Amplifier has a gain of 10 and features a calibrated 1 mV/div to 200 mV/div deflection factor (with oscilloscope deflection factor set at 10 mV/div).

The output impedance of the amplifier is 50 Ω. A 50-Ω termination is supplied with the P6046 Amplifier for use with oscilloscopes having 1-MΩ input impedances.

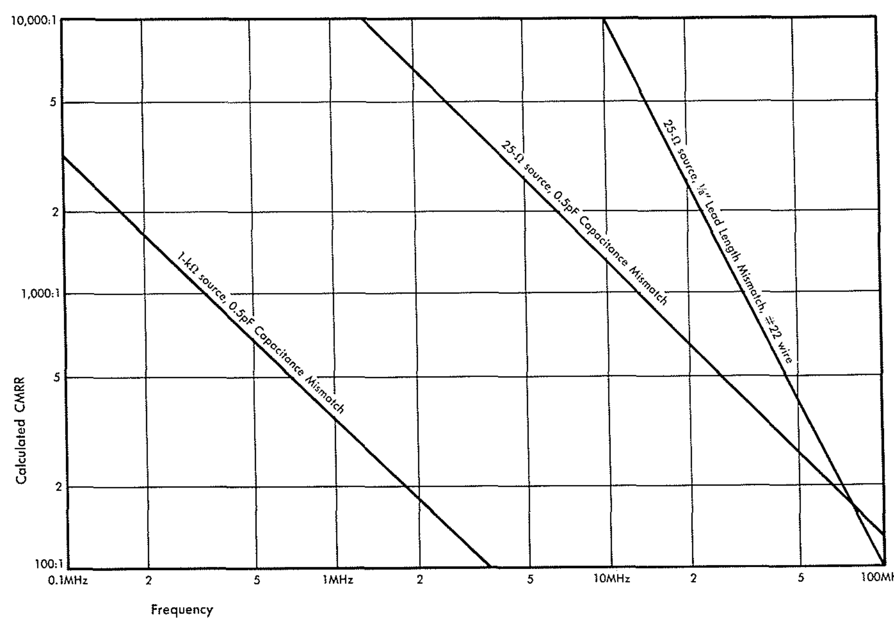


Fig 3. CMRR vs frequency with capacitance mismatch and different source impedances.

MECHANICAL DESIGN

Mechanically, the P6046 probe is made as rugged as possible without sacrificing performance or usability. Thirteen semiconductors, including dual FET's for the balanced input, are housed in the P6046 probe.

The body of the probe is made of high-impact plastic, plating grade. The inside of the body is plated with a low-resistance material that provides an excellent ground plane and electrostatic shield from outside radiation.

Several probe tips have been designed for the P6046 probe. The probe-tip input connectors are mounted on ½-inch centers and are designed to mate with coaxial connectors permanently mounted on circuit boards. The permanent connectors provide excellent ground and signal connection and should be used whenever possible.

When it is not convenient to use the permanent coaxial connectors, a number of special tips are included. For making measurements from test points that are not spaced ½-inch apart, swivel tips are included that provide variable spacing from 3/16 inch to 1½ inches. See Figure 4.

Also included is a ground tip that shorts one of the input tips to the coaxial ground for single-ended measurements, hooked tips for hanging the probe into circuits, and sleeve-type adapters for insulating the tip's coaxial ground. The dual 10X attenuator head included with the P6046 probe has the same tip configuration as the probe.

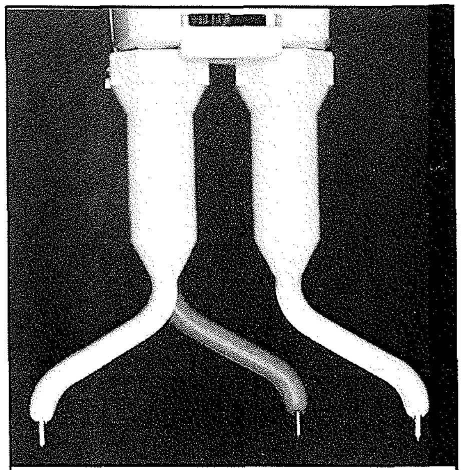


Fig. 4. P6046 Swivel tips.

CHARACTERISTICS

Probe and Amplifier

DEFLECTION FACTOR is 1 mV/div to 200 mV/div in 8 calibrated steps, 1-2-5 sequence, accurate within 3% (with an oscilloscope deflection factor of 10 mV/div).

BANDWIDTH is DC-to-100 MHz at 3-dB down.

RISETIME is 3.5 ns or less.

COMMON-MODE REJECTION RATIOS with deflection factors of 1 mV/div to 20 mV/div are 10,000:1 at DC, 1,000:1 at 50 MHz, and typically 100:1 at 100 MHz.

COMMON-MODE LINEAR DYNAMIC RANGE is ±5 V (DC + peak AC), ± 50 V with 10X attenuator.

INPUT RC is 1 MΩ paralleled by approximately 10 pF.

INPUT COUPLING is AC or DC, selected by a switch on the probe. Low-frequency response AC-coupled is 3-dB down at 20 Hz, at 2 Hz with 10X attenuator.

NOISE (periodic and random deviation) referred to the input is 280 μV or less.

MAXIMUM INPUT VOLTAGE is ±25 V (DC + peak AC), ±250 V with 10X attenuator.

OUTPUT IMPEDANCE is 50 Ω through a BNC-type connector. A 50-Ω termination is supplied with the probe for use with 1-megohm systems.

LINEAR OUTPUT is ± 10 div with the oscilloscope set at 10 mV/div.

PROBE CABLE is 6-feet long, terminated with a special nine-pin connector.

P6046 DIFFERENTIAL PROBE AND AMPLIFIER (010-0106-00)
Includes P6046 Probe (010-0214-00); Amplifier for P6046; 50-Ω BNC cable (012-0076-01); 50-Ω BNC termination (011-0049-00); dual 10X attenuator head (010-0361-00); four swivel-tip assemblies (010-0362-00); special ground tips (010-0363-00); 5-inch ground lead (175-0124-00); 12-inch ground lead (175-0125-00); two alligator clips (344-0046-00); two hook tips (206-0114-00); two test jacks (131-0258-00); two insulating tubes (166-0404-00); two ground clips (214-0283-00); carrying case (016-0111-00); two instruction manuals (070-0756-00).

THE TYPE 1A5 DIFFERENTIAL PLUG-IN with the Type 530, 540, 550, 580-Series Oscilloscopes can use the P6046 Differential Probe without the P6046 Amplifier. The P6046 probe extends the Type 1A5 differential measurement capabilities to 50 MHz, CMRR is 1,000:1 at 50 MHz.

P6046 PROBE PACKAGE (010-0213-00)

SERVICE SCOPE

USEFUL INFORMATION FOR

USERS OF TEKTRONIX INSTRUMENTS

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P.O. Box 500
Beaverton, Oregon, U.S.A. 97005

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FRANK GREENWOOD
DEPARTMENT OF TRANSPORT
T & E SYSTEMS LAB.
BOX 4028, STATION E
OTTAWA, ONTARIO, CANADA

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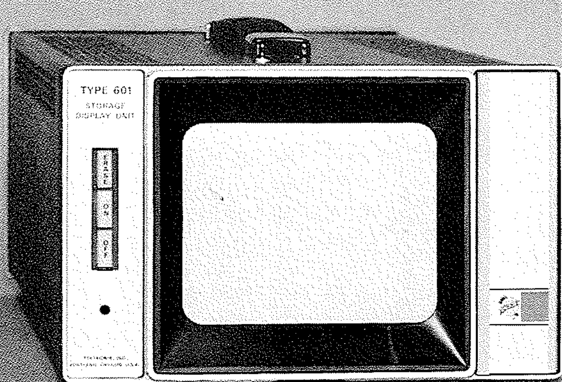
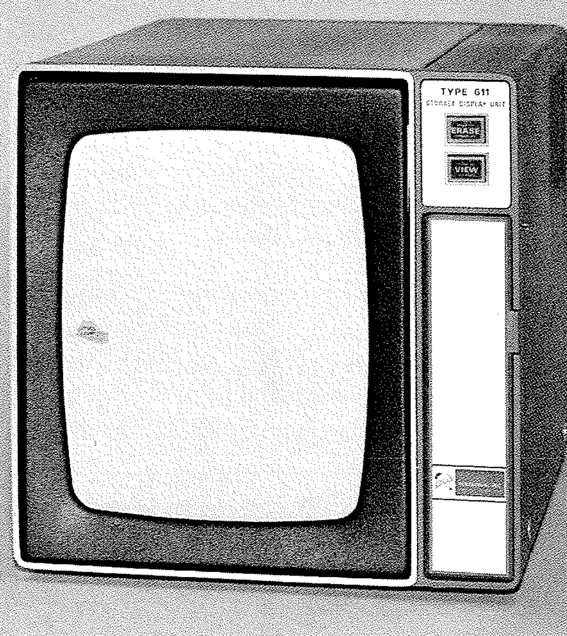
SERVICE SCOPE

NUMBER 47

DECEMBER 1967

A NEW VECTORSCOPE, Page 2





STORAGE DISPLAY INSTRUMENTS, Page 6

A NEW VECTORSCOPE

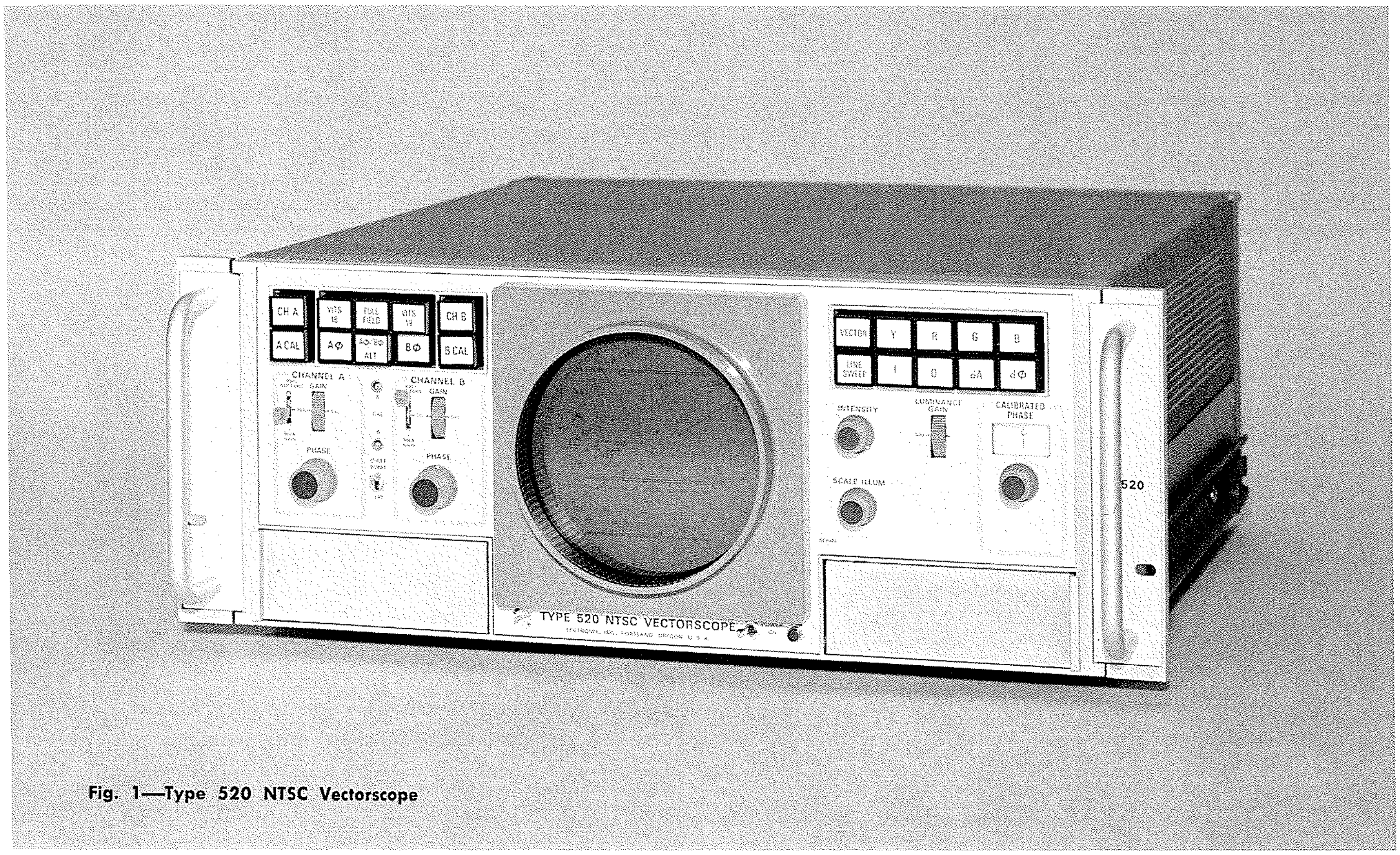


Fig. 1—Type 520 NTSC Vectorscope

INTRODUCTION

Color TV has caused Television Broadcast facilities to expand at a rapid rate. The resulting increase in the quantity and complexity of studio equipment has created a need for more versatile, easier to use, measuring equipment. Oscilloscopes are extensively used in the TV Broadcast industry as program monitors to measure picture levels, as troubleshooting devices to isolate equipment malfunctions, and as a measurement tool to determine waveform compliance with FCC Standards.

The studio operator now routinely verifies or adjusts a variety of parameters that were not required with black and white equipment; burst phase, I, Q, burst level, color bar phase, luminance to chrominance level and others.

In addition to the variety of color system setup requirements, greater emphasis is being placed on in-service monitoring of picture quality. Signal distortion limits which define acceptable picture quality are well established. Suitable test signals have been developed to check distortion; i.e. NTSC Color Bar Test Signal, Linearity stair-step, etc. Using these standard test signals as reference values, quantitative measurements of picture quality are readily made. However, certain signal distortions—particularly differential phase may be difficult to identify or diagnose.

NEW VECTORSCOPE

A new Vectorscope was developed, to assist the Broadcast Engineer and Technician in making these color measurements. The new Tektronix 520 NTSC Vectorscope (Fig 1) was designed with emphasis as an "operating" instrument rather than a special test oscilloscope used only to identify or solve special video problems.

The new oscilloscope is designated as a "vectorscope", however, the addition of a luminance channel has extended the measurement versatility to include most of the routine checks required in a color camera chain.

While measurement versatility is important when considering the purchase of any piece of equipment, ease of operation and long term reliability are equally important. Through special design efforts the Type 520 Vectorscope provides the user with stable drift-free displays at the touch of a button, without sacrificing measurement resolution or accuracy.

PUSH-BUTTON OPERATING CONVENIENCE

The push-button operating controls are arranged into two groups—one for signal selection and one for measurement display mode. All controls unnecessary for specific measurement are automatically disconnected from use to eliminate front-panel confusion. Controls such as the CRT display focus and positioning controls which require only periodic adjustment are located behind front-panel doors.

Fig 2 shows the typical vector display of a 75% saturated color bar test signal. Note the sharply defined spots which permit increased phase-angle resolution, particularly useful for detecting phase jitter or very small phase errors.

Fig 3 illustrates the comparison of two signals on a time-shared basis. Channel A is displayed for two scanning lines and channel B displayed on the next successive two scanning lines. The time-shared signals appear as if they were being displayed on a dual-beam oscilloscope. The time difference between the two input signals normally causes a phase-angle displacement between the two displays (Note Fig 3). This condition is normal and is due to subcarrier reference in the instrument being locked to only one of the signals. By adding a second phase-control knob to the front panel and time sharing its operation with the first phase control, the burst of each input signal can be independently rotated to the X axis permitting an overlay of the vectors for direct comparison. For example, Fig 4 illustrates the input and output waveforms of a distribution amplifier, with intentional distortion, applied to Channel A and Channel B inputs respectively. Differential phase and differential gain can easily be seen to exist on the yellow, cyan and to a lesser extent, the green, red and magenta vectors. Only slight differential phase is evidenced by the blue vector. While this illustration is somewhat severe, note that the differential gain is more severe than the differential phase—a valuable clue when determining the cause of the distortion, such as clamping failure in an amplifier.

DIFFERENTIAL GAIN AND DIFFERENTIAL PHASE

One of the more familiar applications of the vectorscope is the measurement of differential phase and differential gain. The Type 520 Vectorscope provides these measurement capabilities quickly and accurately at the touch of a button. Fig 5 illustrates the Differential Gain operating mode using a modulated stairstep waveform. A different graticule is used to measure differential gain than was used to make the measurement in Fig 4. The IRE Graticule and the parallax-free vector graticule are in place but each graticule is selectively illuminated when the button for the desired measurement is depressed. The one percent error on the tenth step (counting from left to right) is easily observed.

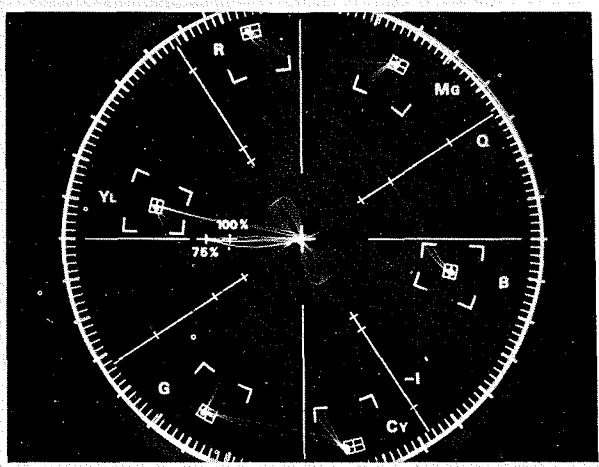


Fig 2—Vector display of 75% saturated color bar test signal

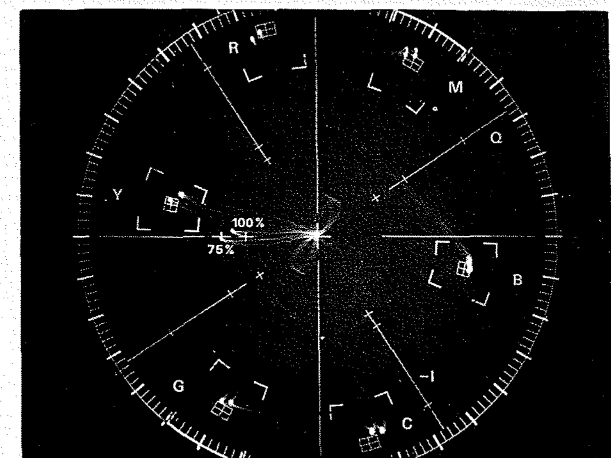


Fig 3—Vector display with channel A and B time-shared

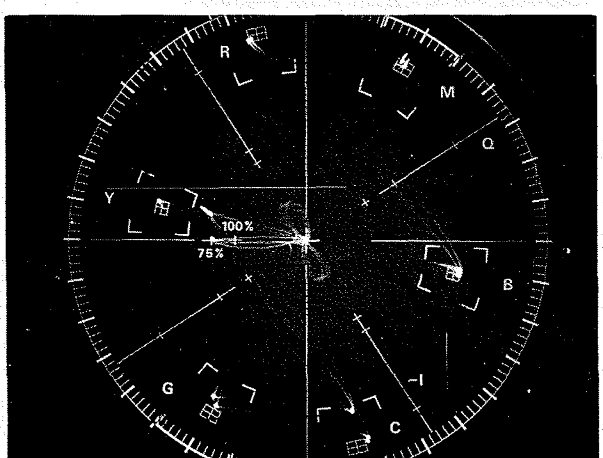


Fig 4—Vector display of input and output waveforms from a distribution amplifier

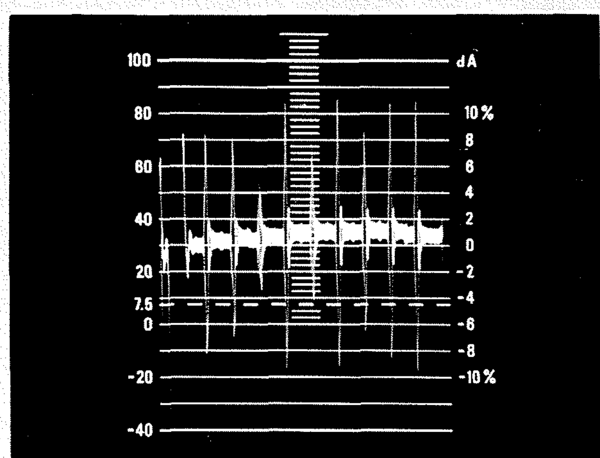


Fig 5—Differential Gain display using a modulated stairstep waveform

When the d ϕ (differential phase) button is depressed, the display is automatically repositioned to the center of the CRT and the graticule illumination removed. The measurment is then taken from the calibrated phase shift dial and the CRT display now serves only as a null indicator. Since considerable amplifier gain is required to resolve phase differences of 0.2 degree or less, small amounts of noise existing on either the applied signal or in the vectorscope will make display interpretation difficult. Display interpretation is simplified by alternately inverting the display to produce the mirror image observed in Fig 6. Any noise on the display (as evidence by a wide trace) can then be averaged out by simply adjusting the overlayed traces for minimum trace width. In Fig 6 each step has a phase transient in the middle amounting to about 0.09°.

NEW MEASUREMENT CAPABILITY

The Type 520 Vectorscope utilizes the luminance portion of the composite color signal to permit measurement of the transformed red, green and blue picture values which normally appear at the picture tube of a color monitor. The Y (or M) luminance signal and the chrominance signal are transformed at the receiver into red, green and blue picture components. Currently, the only means by which the transformation can be verified is by observing the display from the color picture tube itself. Subjective evaluations of picture quality made from color bars viewed directly on a color monitor are influenced by several variables:

1) Phosphor light output efficiency is reduced with age or usage.
2) Color response of the human eye varies from viewer to viewer making consistent readings difficult.
3) Picture monitor characteristcis may vary.
4) Small error differences between the luminance and chrominance waveform amplitudes are almost impossible to detect without using picture comparison techniques.

MEASURING Y, R, G, and B

Measurement of the transformed signal is made by selecting one of four buttons labeled Y, R, G, B. These buttons correspond to luminance, red, green and blue video displays.

When saturated colors (75% or 100%) are displayed on a picture monitor, the monitor electron guns are either on or

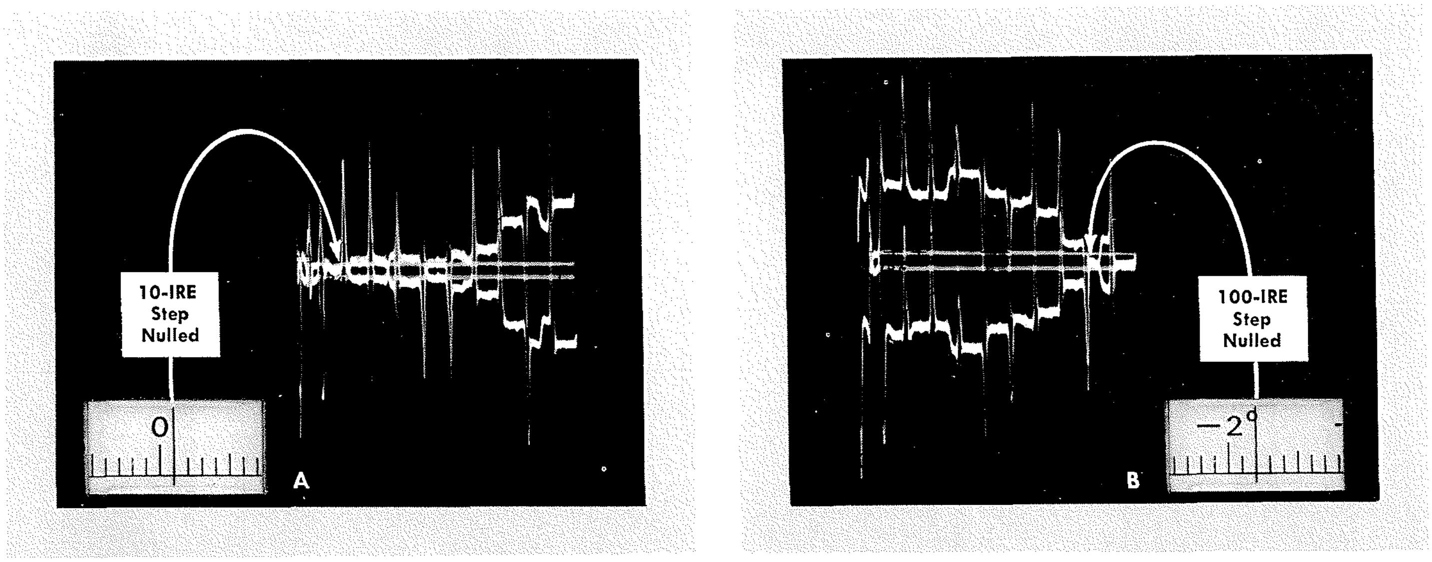


Fig 6. Differential Phase measurement using a modulated stair-step signal. Dial reading A to dial reading B indicates 2.1° differential phase.

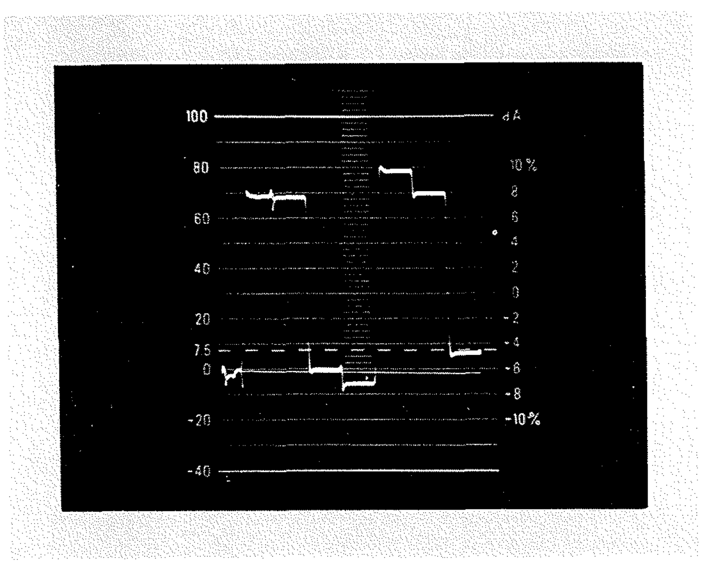


Fig 7—Red values of NTSC Color Bar Test Signal with magenta and red bars oversaturated

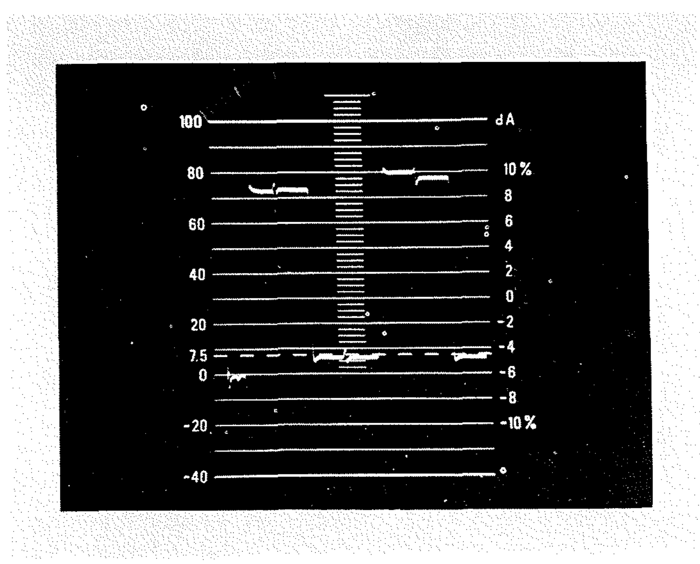


Fig 8—Red values of the NTSC Color Bar Test Signal with luminance amplifier nonlinearity in the white region

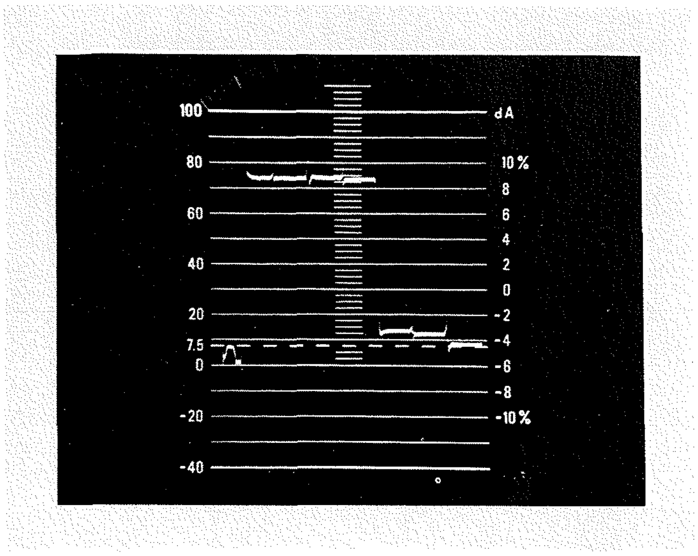


Fig 9—Green values of NTSC Color Bar Test Signal

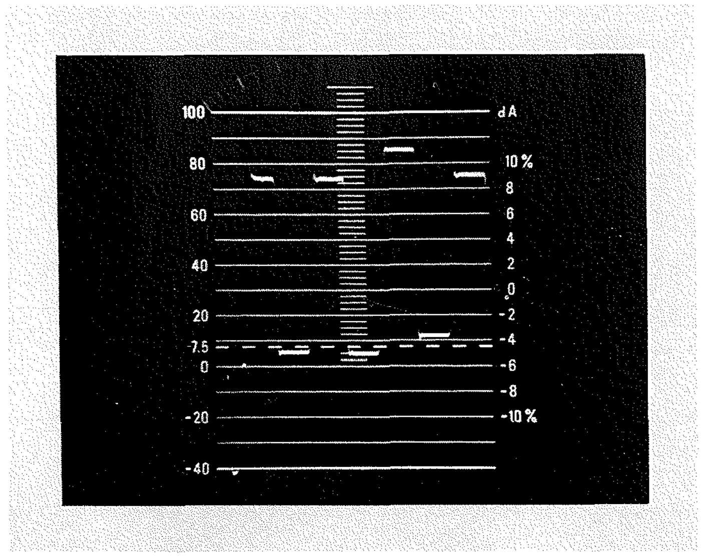


Fig 10—Blue values of NTSC Color Bar Test Signal

off as primary and complementary colors are reproduced. During the primary color "Red", for instance, the red gun is on and green and blue guns are off. The complementary color of "Red" is "Cyan", which is reproduced by the green and blue guns with the red gun held off.

In Fig 7 the red (R) "image" is displayed on the vectorscope using the "standard" decoded color bar signal. The colors are arranged from the left of the display to the right in order of descending luminance—grey, yellow, cyan, green, magenta, red, and blue. Note that the magenta and red bars are over-saturated, however, the error is not easily detected by the eye on the picture monitor because the error is not too large. Observing the vector display of the same signal in Fig 2 indicates that the chrominance portions of the encoded color bars are correct. Therefore the luminance levels for magenta and red must be incorrect—too high in this case. In this illustration, since only two color bars are affected, the error is not due to non-linear luminance gain but simply because the luminance pedestal levels of the color bar generator are incorrectly adjusted.

Fig 8 shows the same display except that the luminance amplifier is non linear in the white region. The *effect* however, is not apparent on the grey and yellow bars but on the magenta, red and blue bars and is due to the luminance amplifier gain having been adjusted with a white pedestal.

Fig 9 illustrates the green (G) display of the same waveform. Note the luminance distortion previously observed affects the green more seriously. Green should be off during the last three bars. While the slight presence of green during the red bar will cause the displayed red to appear orange to the eye (because red is a primary color) the magenta error would be more difficult to detect in the reproduced picture.

Fig 10 shows the blue picture display which is not as seriously affected by luminance errors.

SUMMARY OF 520 NTSC VECTORSCOPE CHARACTERISTICS

Push-button controls provide new operating convenience and permit rapid selection of displays for quick analysis of television color signal characteristics. Amplitude calibrated displays of chroma and luminance are assured with internal calibration test signals to verify amplifier accuracy. The luminance component of the composite color signal is derived for displaying separately or in combination with the red (R), green (G), or blue (B) components.

Two 0° to 360° phase-shifters provide independent phase control of channel A and B. Phase differences caused by unequal signal paths are easily cancelled. A precision calibrated phase shifter with a range of 30°, spread over 30 inches of dial length provides excellent resolution for making small phase angle measurements. Video cable lengths can be accurately matched for time delay at the color subcarrier frequency to less than 0.5° phase difference. Differential gain and differential phase measurement capabilities are provided with accuracies within 1% for gain and 0.2° for phase.

A digital line selector permits the display of a single line Vertical Interval Test Signal from a selected line of either field 1 or field 2.

A parallax-free vector graticule, or IRE graticule, is automatically selected and edge-lighted concurrent with operating mode selection. All silicon solid-state design provides long-term reliability and cool, quiet operation.

The Type 520 NTSC Vectorscope is available in electrically identical cabinet or rackmount models.

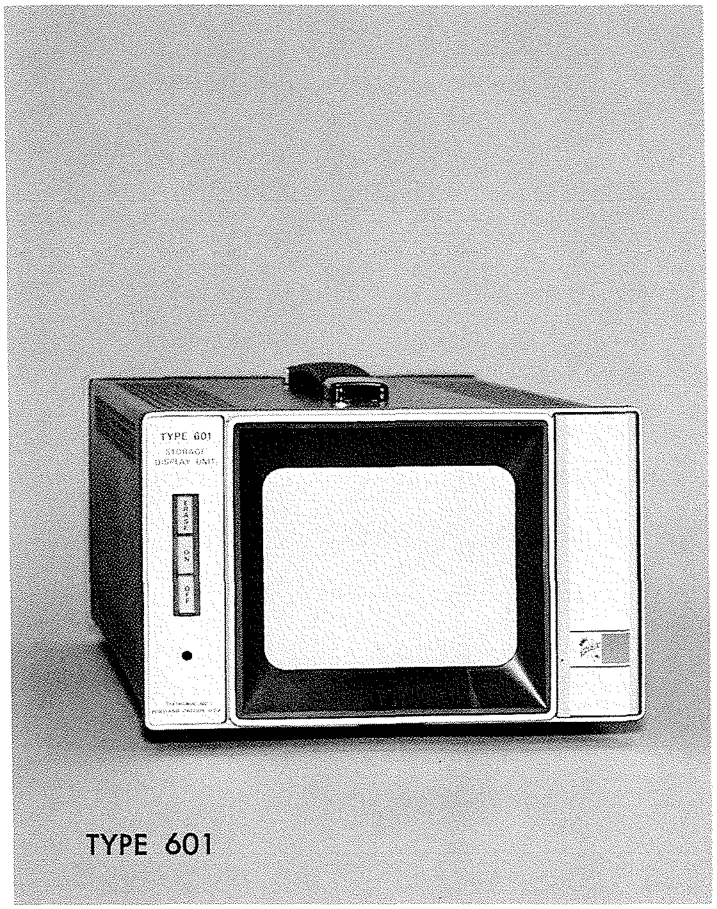
A more complete description of this instrument is found in the Tektronix New Products Catalog Supplement recently distributed.

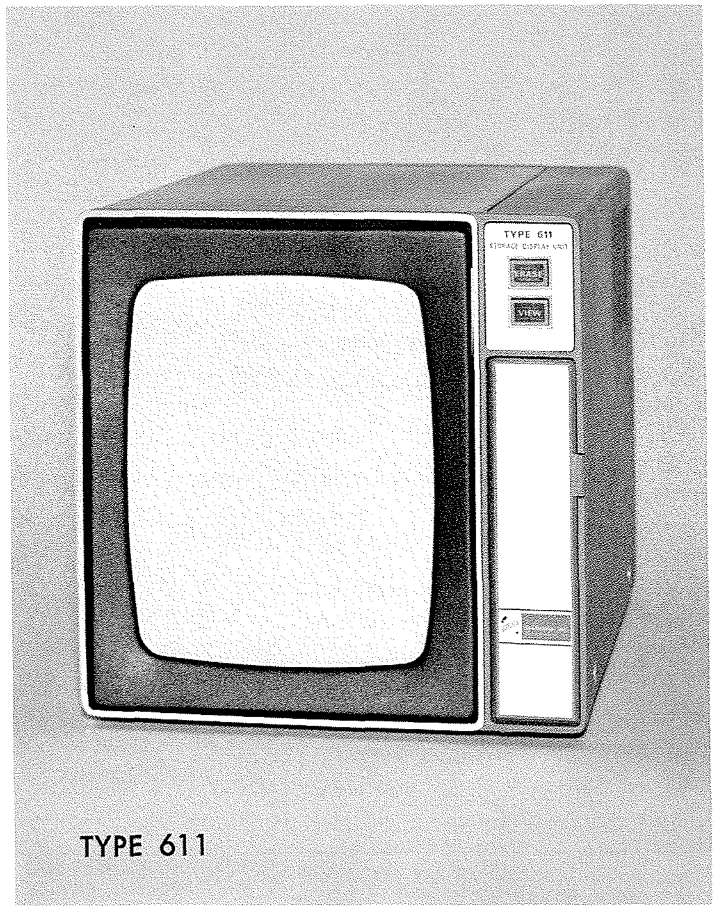
TYPE 520 NTSC VECTORSCOPE $1850

TYPE R520 RACKMOUNT $1850

U.S. Sales Price FOB Beaverton, Oregon

STORAGE DISPLAY INSTRUMENTS





INTRODUCTION

The Type 564 Storage Oscilloscope, a measuring instrument, served as an excellent exploratory tool to determine the advantages of bistable-storage cathode-ray tubes as computer readout devices. Several groups experimented with the bistable storage tube in this application and the results show that a bistable storage tube when used with the appropriate periphal equipment provides high-resolution, non-refreshed, alpha-numeric and graphics displays without flicker or fade.

A sequence of events occurred as the computer market developed that contributed toward the development of the bistable storage tube as a computer display device.

(1) Computer usage was being discouraged by man-to-machine interface problems, that is, a problem is submitted through a programmer, a misunderstanding is found after a period of time, the problem is resubmitted, etc.

(2) Larger and faster computers were developed to help offset computation costs.

(3) Techniques to improve computer time utilization were developed.

(4) Computer time-sharing appeared to be a solution to efficient use of computer time but because of input-output limitations, many parallel or time-shared users are required in order to keep the computer busy.

(5) Time-sharing a central computer requires remote terminals convenient to the users.

(6) The cost per remote terminal for time-sharing application must be sensibly low.

(7) A major economic consideration of remote terminals is local memory cost, especially if arbitrary format alpha-numeric and graphic capabilities are required. It is not economically wise to provide display refreshing from the computer memory, and even with a buffer memory the communication link bandwidth may be too narrow to allow refreshing a display at above flicker rates.

(8) For applications where flexible format is required and large amounts of data are to be presented, the Tektronix simplified direct-view bistable-storage CRT provides an economic solution to the memory/display problem.

A NEED FOR NEW INSTRUMENTS

The interested groups who experimented with the Type 564 Storage Oscilloscope as a computer remote-terminal readout device were encouraged by the results obtained and indicated the need for an instrument optimized for computer display rather than measuring applications. Producing an instrument specifically for computer readout purposes required different design objectives than those for measuring devices.

(1) Writing-speed parameters could be traded off for more uniform and smaller spot size.

(2) Plug-ins replaced with built-in amplifiers resulting in a more compact unit.

(3) The Z axis modified for "on-off" operation.

(4) The CRT target modified for improved isolated stroke or dot appearance (a key contribution).

NEW DISPLAY DEVICES

The recently announced Types 601 and 611 Storage Display Units were designed to be used as integral parts of computer remote terminals. When driven by the appropriate periphal equipment these units will present non-refreshed displays of alpha-numerics and graphics without flicker or fade.

The Type 601 and 611 are intended for *individual* use, *not* group viewing. The high resolution of the 601 and 611, require the viewer to sit fairly close to the instruments in order to resolve the displayed information.

5-INCH STORAGE DISPLAY UNIT

The Type 601 Storage Display Unit features a new, Tektronix developed, 5-inch bistable-storage display tube, providing clear, non-fading presentations. Resolution in an 8-cm x 10-cm display area is 100 stored line pairs in the vertical axis and 125 stored line pairs in the horizontal axis providing an information capacity of about 400 alpha-numerics. The information storage rate is 100-thousand dots per second and time required to erase the stored information is 200 ms. All solid-state modular circuit design insures long-term stable performance.

The operating functions are remotely programmable by simply grounding program lines at a rear-panel connector. Access to X, Y and Z inputs is through rear-panel BNC connectors or a remote program connector.

11-INCH STORAGE DISPLAY UNIT

The Type 611 Storage Display Unit features an 11-inch magnetically deflected, bistable-storage display tube developed by Tektronix. This new storage tube offers high information density and excellent resolution on a 21-cm x 16.3-cm display screen. The information capacity of the Type 611 is about 4000 alpha-numerics. Dot settling time is 3.5 μs/cm plus 5 μs and dot writing time is 20 μs. The time required to erase and return to ready-to-write status is 0.5 seconds.

The operating functions are remotely programmable through a rear-panel connector with access to X, Y and Z inputs through rear BNC connectors or the remote program connector. A "Write-Through Cursor" feature permits positioning the writing beam to any point on the display area without storing the cursor or destroying previously stored information. Write through for alpha-numerics and graphics can be done by shortening the unblank pulse duration from the normal value of 9 μs (Type 601) or 20 μs (Type 611). This mode of operation is useful for manual graphics, with the aid of equipment like the Rand Tablet or with an SRI Mouse. An internal test signal provides a quick check of focus, storage and general performance status of the instrument.

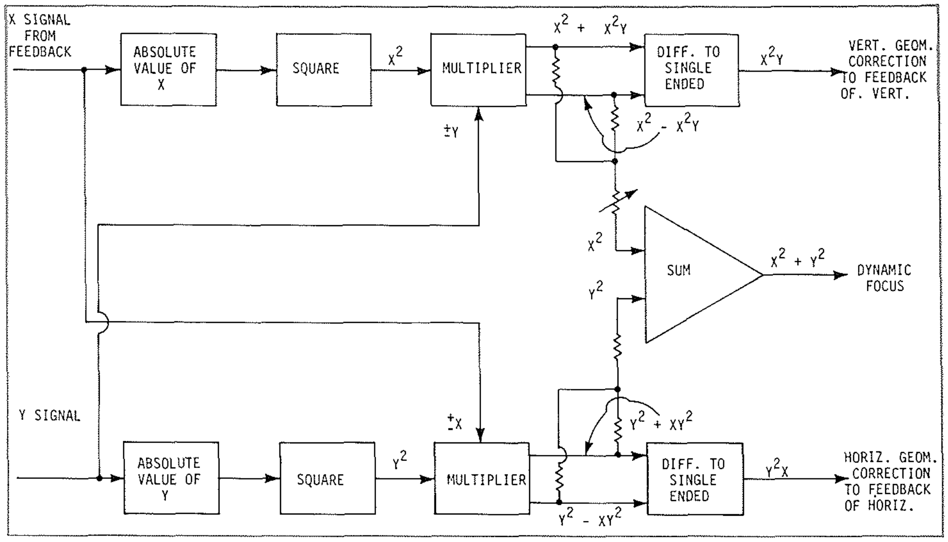


Fig 1—Block diagram of Type 611 pincushion and dynamic focus correction.

SOME DESIGN CONSIDERATIONS

Compatibility between the Types 601 and 611 is maintained with regard to the input connectors and selection of common functions, such as erase. However there are differences which should be kept in mind. The Type 601 has ± 6 cm continuously variable position controls for X & Y, while the Type 611 has three position switches that permit the operator to select one of nine beam resting positions. Variable controls provide a ±10% range for small adjustments of each position. The limited variable range was chosen because of the more stringent drift requirements of the Type 611. Both units have internal gain calibration adjustments to set the full screen deflection voltage within 2% of 1 volt. Both units have provision for other less-sensitive deflection factors.

Trace alignment of the two instruments is different. The Type 611, using an electromagnetrically-deflected tube, has an external deflection yoke which may be rotated to align the traces; orthogonality is a function of how well the yoke was manufactured. The larger screen requirement of the Type 611 requires magnetic deflection through an angle of 70°, in order to keep the length of the instrument reasonable (the Type 611 is about 20% longer than the Type 601). The wide magnetic deflection angle of the Type 611 CRT, together with the flat faceplate, requires correction to the deflection geometry, linearity and focus. Without going into the mathematical details, it can be said that both pincushion and dynamic focus require squared deflection terms. Figure 1 shows the block diagram. The squaring circuit is a single FET. The multiplier is a differential pair driven from a current source. Thus with comparatively simple circuitry, the circuit generates the required X^2Y and Y^2X for pincushion correction, and $X^2 + Y^2$ for focus correction. The dynamic focus summing circuit gets its input from the multipliers, rather than the squaring circuits directly, because of the signal levels involved; that is, the output of the multiplier is at a more convenient level than the squaring circuit. This combination of corrections appears to be new, and unexpectedly simple.

The Type 601 with an electrostatically deflected tube has a *unique* method of correction; instead of the usual rotation coil, signals are independently mixed from the X and/or Y amplifiers into the Y and/or X amplifier, thus introducing tilt and/or slant as necessary to correct trace alignment and/or orthogonality. Because of this cross-mixing, the use of the Type 601 as a waveform monitor should be restricted to applications involving bandwidths below 100 kHz. The smaller deflection angle and lower resolution requirement of the Type 601 make dynamic scan or focus corrections unnecessary.

SUMMARY

The first instrument to use the Tektronix developed bistable storage tube was the Type 564 Storage Oscilloscope, introduced in the spring of 1962. Since that time, the Type 564 has found extensive use in a multiplicity of applications including information display. Early experiments with the Type 564 as an information display device proved the validity of the concept and helped define new storage tube requirements; the Type 601 and 611 Storge Display Units are the first display instruments to employ these new storage tubes.

A more complete description of these new instruments is found in the Tektronix New Products Catalog Supplement recently distributed.

Type 601 Storage Display Unit .. $1050

Type 611 Storage Display Unit .. $2500

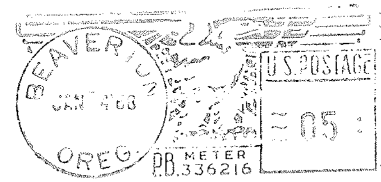
U.S. Sales Prices FOB Beaverton, Oregon

SERVICE SCOPE

USEFUL INFORMATION FOR

USERS OF TEKTRONIX INSTRUMENTS

Tektronix, Inc.
P.O. Box 500
Beaverton, Oregon, U.S.A. 97005



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