Service Scope

USEFUL INFORMATION FOR USERS OF TEKTRONIX INSTRUMENTS

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NOISE—SOME BASIC DATA

At any given state of the art, any increase in the gain or bandwidth of an amplifier inevitably means more *noise*. Some of the limitations are absolute, based on the nature of current-flow; some are only relative to the quietness of available components.

Noise limitations apply not only to oscilloscope inputs, of course, but also to the circuits to be examined with the oscilloscope.

Random (as opposed to systematic) noise is almost always specified in rms terms. Being random, it is capable of analysis only in statistical terms. Instantaneous amplitudes (determining peak-to-peak values) under a given set of circumstances are distributed on a probability curve. The probability of any given rms noise level having all of its peaks between two limits varies with the time-limits set. (If you stand around long enough, you may get struck by lightning.)

It is possible to approximate the rms/peak-to-peak ratio for wideband noise. Normally, 90% or more of the peaks will fall within 3X the rms value of the noise. On an oscilloscope, this represents the main "body" of a noisy trace observed at a relatively slow time/cm rate.

Noise power adds directly; noise voltage vectorially. One milliwatt of noise plus one milliwatt of noise is two milliwatts of noise. One millivolt of noise plus one millivolt of noise is 1.414 millivolts of noise (square root of the sum of the squares).

Noise generally may be broken down into two types: Broadband "white" noise (so-called because of the analogy to white light) in which the power is evenly distributed throughout the frequency spectrum, and low-frequency noise (referred to by some as "pink" noise) in which the power varies inversely with the frequency. All resistances, tubes and semiconductors exhibit both types of noise to some extent.

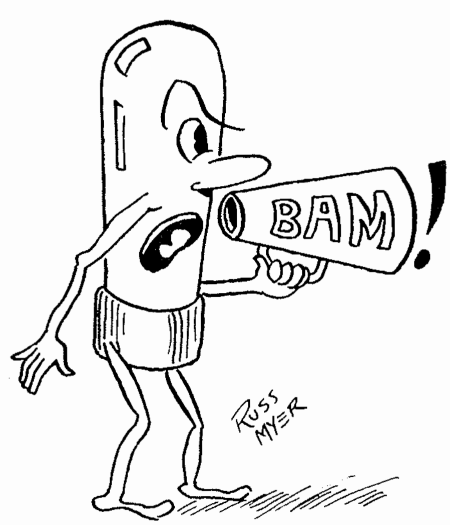
The noise discussed here is more or less inherent in electronic components—not that due to manufacturing defects or to deterioration or damage (as, gas in tubes or moisture in transistors), or that due to external interference (e.g., atmospheric noise or RFI).

RESISTANCE NOISE

ABSOLUTE: At any given temperature, any resistance or resistive component of an impedance generates random wideband noise due to thermal agitation of electrons. The noise *power* generated is proportional to resistance, temperature and bandwidth of the circuit; the observed value may be limited by the bandwidth of the measuring instrument.

RELATIVE: Current flowing in a resistor (particularly a carbon resistor) produces a low-frequency noise called "excess noise" or "current noise." The amount varies widely with the type and construction of the resistor. This low-frequency noise is generally measurable only in the region below 100 kc, and the noise power varies inversely with frequency.

TUBE NOISE



WIDEBAND TUBE NOISE: This "shot noise"—so-called because the effect of pouring electrons from cathode to plate is analogous to the noise of pouring buckshot into a barrel—is a combination of the effects of cathode temperature and resistance and the fact that current flow, being a flow of many discrete charges, is subject to random fluctuation: the number of electrons reaching the plate at any given instant for any given average current is a matter of statistical probability. The flow of grid current is likewise subject to random fluctuation; the resulting output noise and frequency distortion depends on the impedance into which the grid "looks."

"FLICKER": While the wideband noise in tubes is fairly predictable from tube parameters, low-frequency noise, commonly called flicker or 1/f noise, is highly unpredictable. Not much is known about controlling it in tube manufacture. The power distribution of flicker noise varies inversely with frequency, and is predominant over broadband noise below 1-10 kc. It is usually quite a serious limitation to vacuum-tube amplifier performance below 50 cps. The ultimate low-frequency flicker we call drift.

TRANSISTOR NOISE

ABSOLUTE: Electrons crossing any semiconductor "barrier" generate wideband noise. The noise is proportional to the current and to the circuit bandwidth.

RELATIVE: Low-frequency flicker noise is generated in transistors as well as in vacuum tubes, but transistor manufacturers seem to have had more luck in controlling it. The 6 db per octave rise of low-frequency noise over wideband noise may start as low as 1 kc in commonly available types today.

The following information should aid in determining noise values stemming from the causes under discussion here.

RESISTANCE NOISE

Thermal or "Johnson" noise *power* is proportional to temperature, resistance and bandwidth. The rms noise *voltage* is proportional, then, to the *square root* of these factors:

$$E_{rms} = 2\sqrt{1.38 \times 10^{-23}\ TRf}$$

where T is the temperature in degrees Kelvin (absolute), R is the resistance or resistive component of an impedance, and f is the effective bandwidth of the system. The constant shown is Boltzman's constant expressed in meter-kilogram-second units (joules/degree Kelvin).

A simpler formula for use at room temperature (25° C = 77° F) is:

$$E_{rms} = \sqrt{1.65 \times 10^{-20}\ Rf}$$

The values obtained will be essentially the same (±10%) from −15° C to +65° C (5° to 150° F).

Technically, the effective bandwidth f is narrow, sharply defined, and may be located anywhere in the frequency spectrum. Practically, it is convenient to use −3 db bandwidths in calculation, on the assumption that the total of the (attenuated) noise outside the −3 db points is about equal to the rolled-off noise inside the −3 db points which is weighted at 100%.

A one-megohm resistance at room temperature generates about 130 μv of rms noise over a 1 mc bandwidth. However, when the resistance is shunted by a capacitance of, say 10 pf, the noise output above 16 kc will be rolled off at 6 db/octave. We can then approximate the noise level by calling f 16 kc, giving us a value of 16 μv rms for the noise voltage, or about 50 μv peak-to-peak over a short span of time—say, a few hundred msec.

A resistance of 50 ohms generates about 28 μv rms over a one Gc bandwidth.

Because resistor noise is typical of all broadband noise, white noise levels are often specified as "equivalent noise resistance," which allows specification of noise in terms applicable to any bandwidth. The temperature assumed for "equivalent noise resistance," is 25° C, and for any particular bandwidth, the rms voltage can be calculated from the formula above.

CURRENT NOISE

Composition and deposited carbon resistors—to a greater extent than wirewound or the better grade of metal-film resistors—generate low-frequency noise proportional to the applied voltage. This noise is governed primarily by *current density;* a noisier-than-normal resistor is assumed to have localized bottlenecks of high current density. In most resistors, this current noise, "excess noise" or "1/f noise" as it is variously called, is about equal to thermal noise at 100 kc or so and is negligible above about 1 mc. The noise power of current varies inversely with frequency.

The absolute value for composition resistors is usually specified in terms established by the National Bureau of Standards—microvolts rms per applied volt for one frequency decade. Typically, the values will fall between 0.1 μv and 10 μv per applied volt in a frequency decade, but poor manufacturing techniques and quality control can produce much larger figures. For estimating purposes, 1 μv/v in a decade, or 2.24 μv rms per applied volt for the 5 decades from 1 cps to 100 kc, can be used for resistors of good quality. Note that for several decades, we multiply by the square root of the number of decades, since it's the *power* not the voltage, which varies as 1/f.

TUBE NOISE

WIDEBAND NOISE (SHOT-EFFECT)

TRIODES: The various factors affecting the wideband noise in a triode (grid grounded) are approximated by the formula:

$$I_{rms} = 2\sqrt{1.38 \times 10^{-23}\ T_cKG_mf}$$

Where T_c is the cathode temperature (in degrees Kelvin), K is a tube merit factor between 0.64 and 1.28, f is the bandwidth, and I_{rms} is the noise *current* in the plate circuit. Assuming a cathode temperature of 1000° K and a merit factor of 1.0, the formula becomes:

$$I_{rms} = \sqrt{5.5 \times 10^{-20}\ G_mf}$$

Relating this to the input, we get

$$E_{rms} = 2.34 \times 10^{-10}\sqrt{\frac{f}{G_m}}$$

For a 6DJ8 operated at a current of 10 ma and 100 v on the plate (giving a G_m of about 9,000), the grounded-grid equivalent input noise over a 1 mc bandwidth would be about 2.5 μv rms.

A quicker approximation for the grounded-grid noise level in triodes gives the equivalent wideband "noise resistance" at the grid:

$$R_{eg} = \frac{3}{G_m} \quad \text{(some sources say } 2.5/G_m\text{)}$$

For the 6DJ8 in the case above, R_{eg} becomes 333 ohms. To convert equivalent noise resistance into volts, insert the resistance and desired bandwidth figures into the simplified formula for 25° C thermal noise. The figure comes out about 2.4 μv rms for the example above, close to the value obtained before.

If—as is usually the case in high-impedance input stages—the grid is not grounded, grid current developing a voltage across the input resistance adds another noise factor. The noise component of the grid current amounts to:

$$I_{rms} = \sqrt{2 \times 1.6 \times 10^{-19}\ I_g\ f}$$

where the constant 1.6×10^{-19} is the charge (coulombs) of an electron, I_g is the steady-state grid-current, and f is the bandwidth. If the grid looks into an impedance of R and C in parallel, grid-current noise for any small bandwidth Δf at a center frequency F is:

$$E_{rms} = R\sqrt{\frac{3.2 \times 10^{-19}\ I_g\Delta f}{1 + R^2\ (2\pi FC)^2}}$$

The total noise can be approximated as was done for thermal noise by equating f with the −3 db bandwidth $1/(2\pi RC)$, and ignoring the term in the denominator, which approaches the value of 1.0 below −3 db frequency. Now,

$$E_{rms} \approx \sqrt{\frac{3.2 \times 10^{-19}\ I_gR}{2\pi C}}$$

Taking a 6DJ8 with a 10 nanoamp grid current, a grid resistor of 1 megohm and shunt capacitance amounting to 50 pf, we obtain a value of approximately 3.2 μv for rms noise due to grid current. This noise will be primarily in the dc-to-3 kc region. Of course, if we connect a low-impedance signal source to the grid, this noise will, to a great extent, disappear.

PENTODES: In pentodes, the wideband shot-effect noise in the plate circuit is complicated by the random variation in the division of cathode current between screen and plate—so-called "partition noise." In this case, it is easiest to calculate "equivalent noise resistance" first, and go on from there to total noise for a given bandwidth.

$$R_{eg} = \frac{I_b}{I_k}\left(\frac{3}{G_m} + \frac{20\ I_s}{G_m^2}\right)$$

where I_b is plate current, I_k is cathode current, I_s is screen current and R_{eg} is the equivalent noise "resistance" at the grid.

If we consider the Tektronix Type 502's input stage (6AU6's) as a typical pentode application for low-noise operation ($I_k \approx 720\ \mu a$, $I_b \approx 430\ \mu a$, $I_s \approx 430\ \mu a$, G_m about 1100), expectable wideband noise resistance would be about 4500 ohms per side. Or, adding push-pull noise components vectorially, about 3.8 μv rms over 100 kc passband of the oscilloscope. Needless to say, the Type 502's actual noise performance is not this good, primarily because of low-frequency noise which almost completely masks the broadband noise.

LOW-FREQUENCY (FLICKER) NOISE

Because researchers into noise have been occupied primarily with getting answers for the communications industry—which is mostly concerned with tuned RF amplifiers when working with microvolt signals—not much has been done about identifying the causes and cures for low-frequency flicker noise in tubes. This noise is most serious at frequencies below 1 kc and in tubes with oxide coated cathodes. Flicker noise, like current noise in resistors, varies inversely with frequency, and is quite serious in high-sensitivity dc-coupled amplifiers. It is believed to be related to variations in the conductivity of the cathode coating, to thermal agitation and migration of cathode material and areas of emission activity, with consequent shifts in the configuration of the space charge, and to interface resistance between the cathode coating and sleeve, among other hypothetical causes. One investigator, noting excessive noise in a direct-heated cathode with very small filament diameter, concluded that at least some of the noise was due to high-velocity gas ions, speeding from plate to cathode, being captured *in orbit* around the cathode (like little satellites), crashing into freshly emitted electrons and generally creating a nuisance. This seems quite plausible—*collision* of gas ions with the cathode is a common source of noise. Some researchers have found the flicker noise to vary as the square of the cathode current. However, variations among tube types and even among samples of the same tube types are so great that no consistent theory has been developed to explain all the phenomena. Flicker noise in vacuum-tube circuits operated down to 10 cps or below will commonly be three to four times the value of the broadband shot noise and other contributing factors (plate and cathode resistors, etc.). For instance, resistor noise and broadband tube noise account for about 5 μv rms in the front end of the Type 502—corresponding to perhaps 15 μv peak-to-peak.

The observed peak-to-peak value is about 40 μv, with 20 μv seen in exceptional cases.

Since there is no standardized method of measuring or specifying low-frequency tube noise, it's pretty much up to the user to select circuits and tube types and then hope the tube manufacturer keeps his product consistent.

In general, because wideband equivalent noise varies inversely with G_m and both low and high-frequency noise increases with increasing current, the best candidate for a low-noise tube type is one which offers the best transconductance at the lowest cathode current.

TRANSISTOR NOISE

WIDEBAND SHOT NOISE: A fixed minimum wideband noise value for any semiconductor carrying a given value of current is:

$$I_{rms} = \sqrt{3.2 \times 10^{-19}\ I_f} \; *$$

where I is the dc collector current and f is the bandwidth. The constant this time is twice the charge of an electron (1.6×10^{-19} coulomb). In a transistor operated at 1 ma collector current, then, the minimum wideband noise over a 1 mc bandwidth would be about 18 nanoamperes rms, at the collector. With a 1 k collector load, 18 na becomes about 50-60 μv peak-to-peak.

To convert to equivalent input-noise current, divide the output noise current by beta. As is evident from the above (all other things being equal), the only way to avoid this limitation for low-noise performance is to seek transistor designs which offer highest values of beta for a given collector current, but without increased leakage or other noise-source problems.

LOW-FREQUENCY NOISE: As with vacuum tubes, the low-frequency flicker noise in transistors is not mathematically predictable. It frequently is unspecified, even for so-called "low-noise" transistor types. A few years ago, it was exceptional for a transistor's low-frequency noise to be less than broadband noise below 10 kc. Today, "turnover" points as low as 100 cps may be obtained. Below the turnover point, the flicker noise increases at 6 db/octave. Turnover is that point below which low-frequency noise exceeds the broadband value.

TRANSISTOR SPECIFICATIONS: Even though circuit considerations have a great effect on transistor noise, transistor manufacturers have made more effort to assign numbers to noise levels than have tube manufacturers. However, in the absence of industry standards, the methods of measurement and specification are not uniform, and numbers are often hard to interpret.

Aside from the basic collector current noise mentioned above, noise-current in the collector circuit increases with increasing collector voltage (leakage current noise at the reverse-biased collector-base junction), and with increasing emitter current (surface phenomena at the forward-biased base-emitter junction). Base driving impedance affects these two noise "generators" oppositely; for any given set of voltage and current conditions, there is an optimum base driving impedance for lowest noise, generally between 300 ohms and 3 k. Transistor noise specifications are often based on very low voltage and current settings, plus optimum driving impedance. A collector voltage of 2 v, current of 500 μa and perhaps 1 k driving impedance are typical numbers for "spec" noise levels.

A commonly used spec is "Noise Figure" (NF). This is defined as the ratio of the signal-to-noise ration at the collector to the "available" signal-to-noise ratio at the base, and is normally expressed in db. Usually, it does *not* include flicker noise.

The "available" signal-to-noise is noise of that optimum driving resistance. If a transistor exhibits lowest noise when driven by a very high impedance, its noise figure may be very good but its actual noise contribution quite high. It's important to know the "R_{opt}" when evaluating a specification.

The noise figure NF is calculated as

$10 \log \dfrac{R_{opt} + R_{equiv}}{R_{opt}}$ where R_{opt} is the optimum driving impedance and R_{equiv} is the equivalent input noise resistance of the transistor itself. So to find out the actual transistor noise level R_{equiv}, we work this formula:

$$R_{equiv} = R_{opt}\ (\text{antilog}\ \frac{NF}{10})\ - R_{opt}$$

or

$$R_{equiv} = R_{opt}\ (\text{antilog}\ \frac{NF}{10} - 1)$$

Here is an example (from the General Electric handbook, 6th edition). General Electric Type 2N123. At 5 v and 1 ma, NF is 1.94 db with a driving resistance of 720 ohms.

$$\begin{aligned} R_{equiv} &= 720\ (\text{antilog}\ 0.194 - 1) \\ &= 720\ (1.56 - 1) \\ &= 420\ \Omega \end{aligned}$$

For a 1 mc bandwidth, then, the transistor will contribute about 2.55 μv of rms noise. However, the base driving resistance brings up the total equivalent input noise to about 4.3 μv rms. A lower value driving impedance might (depending on the transistor) provide lower total input noise.

Noise figures specified in "microvolts per square root cycle" ($\mu v/\sqrt{f}$) or "nanoamperes per square root cycle" ($nA/\sqrt{f}$) may refer to measurements taken on a Quan-Tech transistor noise analyzer over a *one cycle* bandwidth centered at 100 cps, 1 kc or 10 kc, and must be multiplied by the square root of the intended bandwidth before becoming meaningful. Even so, they are not too useful, referring only to open-circuit and short-circuit base conditions. Additional calculation (μv/nA) yields the R_{opt} driving impedance and the noise figure for the conditions specified.

Noise figures specified in "db below 1 μv" or the equivalent are of little use without the conditions being specified. One abridged specification sheet, for instance, describes a 2N207B transistor as having a noise level of "2db below 1μv." The full specification sheet reveals that this performance was measured over a 2700 cycle bandwidth of 300-3000 cps at a collector current of 500 μa.

Noise specifications at best are only a general guide, and in-circuit evaluation with transistors, as with tubes, is the only way as yet to evaluate the limits of achievable performance, especially with regard to low-frequency noise.

In conclusion then, the very nature of an oscilloscope—a "search" tool capable of responding to random or unpredictable waveforms—demands that it respond to noise in the circuit being "searched". Thus, the signal to noise ratio in the circuit being investigated imposes one absolute limitation on usable sensitivity and bandwidth. Only to the extent that one can *predict* the nature of the signal he wishes to measure and also delineate the characteristic of rejectable, non-significant signals, can substantial improvements in sensitivity and bandwidth be made at any given state of the art. (Compare the cost and complexity of obtaining a gain of 10^6 at 1 mc by means of a dc-coupled amplifier and by means of a little ac-dc radio, and then consider their comparative signal-to-noise ratios.)

Advancement of the state of the oscilloscope art depends upon improvements in the performance of components and an abilty to discover and apply those circuit techniques which allow an approach to the absolute limitations imposed by the nature of electron flow. Two possible techniques that may help to overcome these limitations are cyrogenics to reduce thermal noise and micro-circuitry to reduce noise associated with current (the smaller the L's and C's the less current required to achieve a given bandwidth). At this time, however, material gains in sensitivity with wide bandwidth and at high impedance by these techniques appear to be still far in the future.

* This same relationship is true of tubes in "plate saturation" when the noise-modifying space charge is depleted, and in general of any current flowing across a "barrier."

SINGLE-SHOT MULTIVIBRATOR CIRCUITS

One of the characteristics common to most single-shot multivibrator circuits is their sensitivity to the rate of rise of the trigger signal as well as to the amplitude of the signal. For this reason, the single-shot multivibrator tends to become increasingly difficult to trigger as the rate of rise of the trigger signal decreases. (Trigger risetime becomes slower for constant amplitude triggers.)

Finally, there is often a rise rate which is so slow that the circuit cannot be triggered even with triggers of very high amplitude. In vacuum tube single-shot multivibrators, this effect is produced by either the input coupling time constant (Hi-pass) or the timing network itself failing to couple sufficient signal to initiate regeneration.

In tunnel-diode single-shot multivibrators it is usually an L/R network that determines the timing (duration) of the multi. It is also this network which "robs" trigger current away from the tunnel diode if the rate of rise of the trigger is too small.

To avoid this problem, both in tube-transistor circuits or tunnel-diode circuits, a Schmitt trigger circuit is sometimes used.

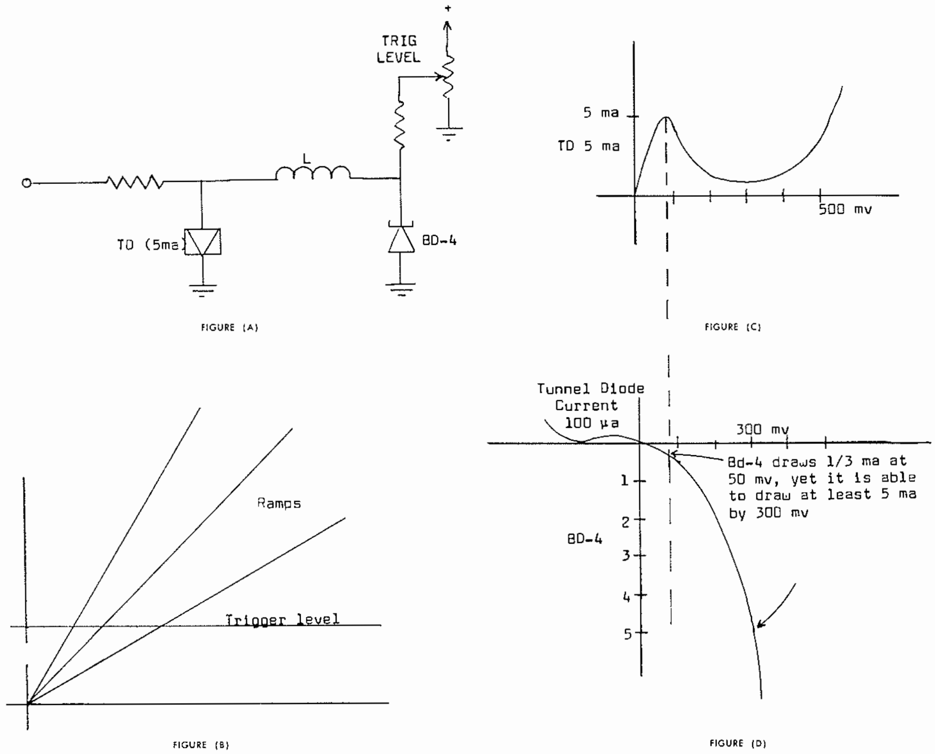


FIGURE 1

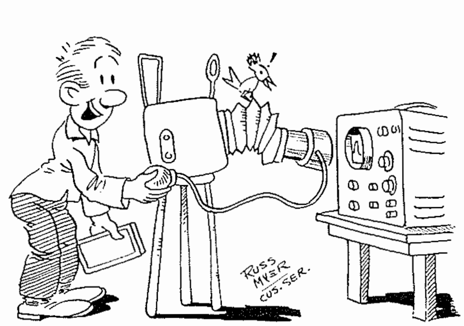
These, however, are not generally as sensitive as the single-shot multivibrator. Another (and we believe, better) solution that is useful in tunnel-diode applications, makes use of a "Back-Diode" to hold the timing circuit (L/R) disconnected and then connect it to perform its normal function after the regeneration of the main tunnel diode has occurred — the normal function here being that of switching the multi after a certain time interval.

Figure 1 shows a typical application. The requirement here is to pick off from ramps, having a wide variety of slopes, a pulse corresponding to the time the ramp crosses a certain voltage.

Since the back-diode (BD-4) takes no more than 0.33 ma (out of 5) at the trigger voltage (50 mv), it will not affect the firing level by any more than this over a wide range of trigger slopes. The ability to guarantee switching depends upon the current drawn at the valley of the tunnel diode (TD) by the BD. The BD-4 draws at least 5 ma by this voltage, and since switching would be assured even if it were only to draw 1 ma (Iv), the circuit is safely mono-stable.

This circuit has proven useful in several applications involving triggers of varying slopes. The combination of a 20 ma tunnel diode with a 1 ma Back-Diode (BD-1) is also useful.

OPTIMUM WRITING-RATE TECHNIQUES FOR OSCILLOSCOPE PHOTOGRAPHY



A competent technician should be able to provide satisfactory Polaroid* pictures of single-shot traces on an oscilloscope at the maximum speed and amplitudes for which the trace can be resolved.

There are a number of critical factors in an oscilloscope recording system which are under the control of an operator. It is necessary to employ to the optimum each of these factors for best results. P31 or P2 phosphor can be used successfully; P11 is not as effective because the phosphor cannot be prefogged. The trace must be properly focused; keep critical portion of display centered on the crt, where the system writing rate is highest. Display a pattern traversing the screen at an angle of about 45°, at a repetition rate (60 cycles or less) just rapid enough to permit focusing while observing the trace through the viewing channel or a light-tight viewing hood. Adjust the intensity to just below the point where a stationary spot appears, then focus for the finest trace. Reduce the intensity as necessary to maintain good focus.

A Tektronix C-19 camera with the fastest lens (f/1.5) and a 2-to-1 image reduction is preferred. Use the widest lens aperture but be sure the *camera* is precisely focused; at widest aperture opening the depth of the field is less than a millimeter. It is sometimes worthwhile to take a shot of a slow trace to prove out the optical focus.

If Type 47 Polaroid film is used, it should be prefogged to where the background is dark grey, rather than black. To prefog Type 47 film: Swing the camera away from the scope, tape a sheet of bond paper across the front of the camera, and shine a 60-watt lamp toward the paper from a distance of three feet. Expose the film at f/16 for 1/50 second. If the fog level is too high, increase the distance from the lamp, or decrease the time to 1/100 second. It would be well to try one or two shots to get an optimum degree of fog. Use the same technique with Type 410, but reduce the exposure to 1/100 second.

Polascope* Type 410 should be used if available. After loading the camera, develop an unexposed picture to determine the condition of the film. If the film is fresh, and has been properly stored, the print will be a definite black. Fresh film should be prefogged as directed above. If the film is not fresh, the unexposed print will be mottled grey, and prefogging will not provide further gain, however, the speed of a stale film may be as fast as that of prefogged fresh film.

For high speed traces you will need to get additional light gain by "prefogging" the P2 phosphor. This procedure provides an excitation bias for the phosphor. With the camera in place on the scope and everything ready for the exposure, open the viewing-tunnel door and shine a 60-watt lamp into the viewing tunnel in a manner that will expose the phosphor area to be occupied by the trace. The lamp should be about three feet away from the tunnel, and held for a few seconds. Then close the viewing-tunnel door, wait for about 15 seconds (for P2), open the shutter and trigger the scope in the usual manner. If you use P31 phosphor, wait only about two seconds before taking the shot.

In general, use:

Smallest f stop (widest aperture opening).
Low amplitude display; one or two cm.
A 45° trace to focus beam.
Fresh film.
Type 410 film.
Prefogged film.
20-second development.
Centered display.
High intensity, but sharp trace.
Trace carefully focused at low repetition rate.
Precise camera focus.

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*Polaroid and Polascope are registered trade marks of the Polaroid Corporation.

? LARGER INPUT CAPACITORS FOR THE TYPE 503 AND TYPE 504 OSCILLOSCOPES ?

The Type 503 and Type 504 Oscilloscopes use 0.022 µf capacitors in their input circuits. From time to time we receive inquiries about the installation of 0.1 µf capacitors in these circuits. While 0.1 µf capacitors in these positions may be an advantage in some cases, people making such a request should consider the information that follows. It may help them to reach the right decision.

Usually the basis for such a request is extension of low-frequency measurement accuracy. We offer here a timely reminder: A 10X probe will accomplish almost the same purpose as will the 0.1 µf input capacitors! Simply using a 10X probe extends the low-frequency 3 db point down by a factor of 10. The Type 503's 22 msec time constant becomes 220 msec (−3 db at 0.7 cps) when you attach a P6000, P6006, or P6017 probe. A word of caution though; don't assume this also applies to a 100X probe! The P6002, 100X probe, for instance, because of the divider circuit used, does not extend the time constant by 100, but only about 10%.

The 0.022 µf capacitors used in the Type 503 and Type 504 Oscilloscopes offer these advantages:

1. Lower leakage: For any given style of capacitor, the leakage specification is given in "megohm-microfarads", which says that as capacitance goes up, leakage resistance goes down. An input capacitor leakage resistance of 100,000 megohms will cause a trace displacement of 1 mv (in the Type 503, up to 1 cm) per 100 v applied. The 0.022 µf capacitor gives us a leakage, lower by a factor of 4, than the leakage we would get in the same capacitor in the 0.1 µf size.
2. Amplifier protection: AC-coupling is normally used when measuring small signals riding on high DC voltages. When the input is connected to a high DC voltage, the amplifier receives a severe overload signal, the duration of which is determined by the input coupling time constant. The shorter this time constant, the better reliability we get out of the amplifier.
3. Greater operator convenience: With a 0.022 µf input capacitor, when you overdrive the amplifier the trace will return to the crt screen in less than a quarter of the time required with an 0.1 µf input capacitor.

In terms of low-frequency measurement accuracy Table 1 compares the error introduced by the capacitor for 0.1 µf and 0.022 µf inputs, with and without 10X probes. Since capacitor values are typically ± 5% to ± 20%, the same order of variation should be expected in the frequencies shown in the table.

The lower section of the table shows the pulse or square wave width for a given amount of tilt for the same four cases.

If you really need 0.1 µf input capacitors, Tektronix-made Mylars are probably the best bet in what we have available. These capacitors carry a nominal 10% tolerance. Tektronix part number is 285-556. To preserve the original balance specifications in the Type 503, pairs should be selected for 5% match. For even better differential performance at low frequencies, these capacitors are available already matched in pairs to within 1% of each other under Tektronix part number 295-054 pair.

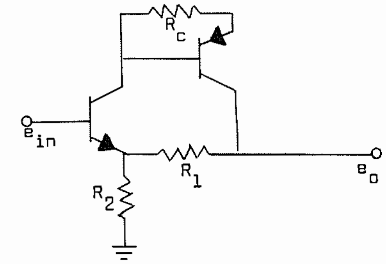
	0.022 µf		0.1 µf	
Measurement Error due to Capacitor	From Signal Source < 1 k	With 10X Probe	From Signal Source < 1 k	With 10X Probe
1%	50 cps	5 cps	11 cps	1.1 cps
2%	35 cps	3.5 cps	7.8 cps	.78 cps
3%	30 cps	3 cps	6.4 cps	.64 cps
5%	22 cps	2.2 cps	4.8 cps	.48 cps
10%	15 cps	1.5 cps	3.3 cps	.33 cps
20%	10 cps	1 cps	2.1 cps	.21 cps
30%	7.2 cps	0.72 cps	1.6 cps	.16 cps
	(above) Frequency for given error (below) Pulse width for given tilt			
Tilt				
10%	2.2 msec	22 msec	10 msec	100 msec
5%	1.1 msec	11 msec	5 msec	50 msec
2%	0.4 msec	4.4 msec	2 msec	20 msec
1%	0.2 msec	2.2 msec	1 msec	10 msec

TABLE 1

OUR APOLOGIES

In the last (June '63) issue of Service Scope several errors occurred that are serious enough to warrant our calling them to the attention of our readers.

In the article "Transistors in Degenerative Feedback Combinations," on page one, paragraph three, the phrase "three-port devices" should read "three-terminal devices." On page two, Figure 2(a) the statement "Z_o is less than R, A is greater than R_o/Z_o except in (a)" applies to all four diagrams in Figure 2. Therefore, it should be included as part of the caption for Figure 2. Next, in part (b), Figure 2, "$e_o = -i_{in}^2 R$" should read "$e_o = -i_{in} R$" and "$P_o = i_{in} R\left[\frac{R}{Z_o} + 1\right]$" should read "$P_o = i_{in}^2 R\left[\frac{R}{Z_o} + 1\right]$". Also in Figure 2, but in part (c), "$e_o = i_{in}^2 R$" should read "$e_o = i_{in} R$." Then on page two but in Table 11, under the column headed "Transfer" and in the "Impedance" box, "Low output Z" should read "High output Z." Turning now to page three, Table III, the first circuit shown here contains a "funny" looking transistor—one with two emitters. Correct this by removing the emitter on the lower leg of the upper transistor and placing it on the lower leg of the lower or "e_{in}" transistor as shown in the diagram below.



Finally, in the article "High Rep-Rate Bursts from Multiples of Type 111 Pulse-Rate Generators," we neglected to identify the resistor to ground in Figure 2, page seven. This resistor should be identified with a lower case "r".

Fairness compels me to confess that I must bear the responsibility for the errors which marred these two fine articles. The errors were not present in the authors' original manuscripts.

Please accept my sincere apologies,

The Editor.

A MEASUREMENT TECHNIQUE USING A Z UNIT WITH 1000 MEGOHM INPUT

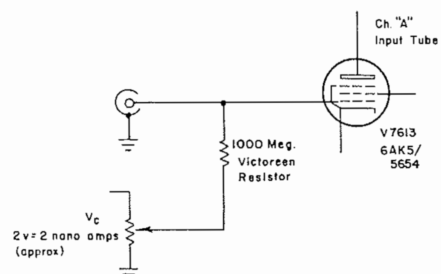
Charlie Rhodes, Tektronix project Engineer, contributes a technique for making measurements not ordinarily possible with an oscilloscope because of circuit loading by the usual 1 or 10 meg input resistor. This technique requires the use of a modified Type Z Plug-In Unit in a Type 530, Type 530A, Type 540, Type 540A, Type 550, or Type 580-Series Oscilloscope.

The Type Z Unit is modified to give an input resistance of 1000 megohms by installing a 1000 megohm resistor between the A input grid and the slider on the Z Unit's Comparison Voltage potentiometer. Disconnect everything between the A input grid and the A channel UHF input connector. Use a stiff piece of wire and route it in the air to bring input signals directly to the grid (pin 1) of V7613, a 6AK5/5654 tube. Set the VAR. ATTEN. control to A ONLY.

The input resistance is now, of course, 1000 megohms. The usual 2 nanoamps of grid current are supplied by adjusting Comparison Voltage to about -2 volts; the trace is on screen and quite stable.

A current of 50 picoamps through 1000 megohms resistance equals 50 mv or 1 cm deflection. Input sensitivity is 50 picoamps/cm. Extremely high values of leakage resistance can, therefore, be measured dynamically.

Capacitor-type inputs such as sonar transducers, strain gauges, etc., are potential applications.



Schematic showing the Type Z Unit input modified to give on input resistance of 1000 megohms.

USED INSTRUMENTS FOR SALE

1 Type 531A Oscilloscope, s/n 9199 and a Type CA Plug-In Unit, s/n 30886. F. C. Shidel, 4620 Ethel Avenue, Sherman Oaks, California.

2 Type 525 Television Waveform Monitors, s/n's 1204 and 1216. Price: $750.00 each. Information on these instruments can be obtained through Dean Butts, Tektronix, Inc., 11681 San Vicente Boulevard, Los Angeles 49, California. Phone: GR 3-1105 or BR 2-1563.

1 Type 502 Oscilloscope, s/n 6890. Kaiser Foundation Hospital, 4900 Sunset Boulevard, Los Angeles 27, California. This instrument has seen very little, if any, use.

1 Type 560 Oscilloscope, s/n 229 and 2 Plug-In Units—a Type 60, s/n 372, and a Type 67, s/n 189. These instruments are in new condition. Contact: David Hammel, 5 Devon Court, Riverton, New Jersey. Phone: Area Code 609, 829-1561.

1 Type 561A Oscilloscope, s/n 7177, approximately 4 months old. Will discount 15% from the purchase price of $399.50. Contact: Dr. von der Groeben, Stanford Medical Center, Department of Cardiology, Palo Alto, California.

1 Type 545A Oscilloscope and 1 Type CA Plug-In Unit. Instruments are about two years old. Louis G. Fields, Starling Corporation, 2047 Sawtelle Boulevard, Los Angeles 25, California. Phone BR 2-7131.

2 Plug-In Units for Type 530, Type 540, Type 550, or Type 580-Series Oscilloscopes—one 53/54K, s/n 5455, and one 53/54G, s/n 1923. Thorobred Photo Service, Inc., 7618 Sepulveda Boulevard, Van Nuys, California.

1 Type 502 Oscilloscope, s/n 004162, (Purchased in 1962). Boris Stefanov, 5628 Harold Way, #8, Los Angeles 28, California. Phone: NO 3-8011.

1 Type 541A Oscilloscope, s/n 7099, and 1 Type 53/54G Wide Band Differential Plug-In Unit, s/n 2969. Contact: R. Rechter, 11611 Chenault Street, #219, Los Angeles 49, California. Phone 648-4132 or evenings 472-1418.

1 Type G Plug-In Unit (only 8 months old.) Industrial Dynamics Company, 3423 S. LaCienega Boulevard, Los Angeles 16, California. Attn: Ed Wagner. Phone VE 7-3330.

1 Type 541A Oscilloscope, s/n 7763; 1 Type CA Plug-In Unit, s/n 2199; 1 Type 53/54K Plug-In Unit, s/n 988; and 1 Scopemobile (model not given). Contact: Philips Applied Research, 1640 21st Street, Santa Monica, California.

Richard D. Brew and Company, Incorporated offer the following Tektronix equipment:

1 Type 180 Time-Mark Generator, s/n 756

1 Type 180S1 Time-Mark Generator, s/n 1033

2 Type 121 Wide-Band Amplifiers, s/n 2701 and 2703

1 Type 127 Preamplifier Power Supply, s/n 413

1 Type 551 Dual-Beam Oscilloscope, s/n 369

1 Type 511AD Oscilloscope, s/n 1547

All this equipment is in good working condition and will meet Tektronix manual specifications. Contact Samuel A. Oliva, Electronic Division, Richard D. Brew and Company Incorporated, Concord, New Hampshire. Phone: Area Code 603, 225-6605.

1 Type 512, s/n 1691, in very good condition. Electronic Engineering Company, 1601 Chestnut Avenue, Santa Ana, California. Attn: A. Harman, Purchasing 1 Agent. Phone: KI 7-5501.

1 Type 317 Oscilloscope— $650.00, and 1 Type 105 Square-Wave Generator—$325.00. Both instruments were purchased in 1960, but never used. J. George Rakonitz, 565 Willow Road, Menlo Park, California.

1 Type 517 Oscilloscope, s/n 738. Wyle Laboratory, 128 Maryland Avenue, El Segundo, California, Attn: Ray Prasta.

1 Type 570 Electron Tube Curve Tracer, s/n 5231. F. Andrews, Canadian Marconi Company, 90 Trenton Avenue, Montreal, Quebec, Canada. Phone RE 8-9441.

1 Type 532 Oscilloscope, s/n 5100; 1 Type 53G Plug-In Unit, s/n 100; and 1 cart. S. P. Dobisz, N.J.E. Corporation, 20 Boright Avenue, Kenilworth, New Jersey.

1 Type 511A Oscilloscope. Make us an offer! Chief Engineer WPIX-TV, 220 E. 42nd Street, New York, 17, New York. Phone: MU 2-6500.

4 Type FM122 Low-Level Preamplifiers, s/n's 6923, 6924, 6925, and 6926; and 1 Type FM125 Power Supply, s/n 1076. These instruments are practically new. They have seen only about one hour service and are in "original-equipment" condition except for holes drilled in the back panel to accommodate input and output connectors. Please direct your inquiries to John West, Tektronix, Inc., 442 Marrett Road, Lexington, Massachusetts. Telephone number is VOlunteer 2-7570.

MISSING INSTRUMENTS

The "grey market" for oscilloscopes has evidently sailed out of the doldrums and is *stealing* along at a good clip. Since the July issue of Service Scope, in which we had only one "lost" instrument to report, we have received notices of six presumably stolen instruments.

Our Long Island Field Office reports a C-12 Oscilloscope Camera and carrying case disappeared from the Presbyterian Hospital. Mr. Sheridan, Chief of Security of the hospital, gives the serial number of this camera as 1474. The camera and case belong to the College of Physicians and Surgeons of Columbia University Radiology Research Lab and was purchased on an Atomic Energy Commission grant.

Information regarding the whereabouts of this camera should be telephoned to Mr. Sheridan at 212-579-2145 or Dr. William Gross at 212-579-3545.

Field Engineer Bill Lewis with our Chicago Field Office lost two oscilloscopes plus plug-ins to car prowlers in the Des Plaines, Illinois, area. While Bill was assisting a customer to repair an instrument, thieves damaged a vent window, unlocked the car door and removed a Type 535A Oscilloscope, s/n 27138, with a Type CA Plug-In Unit, s/n 46850; and a Type 561A Oscilloscope, s/n 5984, with two Plug-In Units; a Type 3A75, s/n 415, and a Type 3B3, s/n 147.

Information regarding these instruments should be relayed to your Tektronix Field Engineer or local field office.

A Type 545 Oscilloscope, s/n 35888, along with a few other instruments totaling $4,000 was removed from the laboratory of the Puget Sound Bridge and Dry Dock in Seattle, Washington. This loss occurred around the last of January of this year, but the information did not reach your editor until just recently. Ernie Hiser, Supervisor with the Puget Sound Bridge and Dry Dock company would appreciate hearing from anyone with information regarding these missing instruments.

A Type 317 Oscilloscope, s/n 1848, was apparently stolen from a motel in Kankakee, Indiana. This instrument is the property of the Shell Oil Co., 8500 North Michigan Road, Indianapolis 8, Indiana. Mr. George Axmann, telephone number AX 1-7440, ext. 62 is the man to contact if you have information regarding this instrument.

Western Scientific of 1200 W. Olympic Boulevard in Los Angeles suffered the loss of two Tektronix instruments recently. A Type 107 Square Wave Generator, s/n 2298, and a Type 180A, s/n 9164, were apparently stolen out of one of their trucks. Western Scientific will appreciate any assistance our readers can give them in helping to locate these instruments.

A very brief message from our Lathrup Village Field Office states succinctly that a Type 310A, s/n 017915, disappeared from the Toledo Scale Company's premises in Pomona, California. Despite the terseness of the message, we are sure the Toledo Scale people will appreciate any information you have that will help them locate their oscilloscope.

USED INSTRUMENTS WANTED

1 Type 310 Oscilloscope. Please contact Mr. Griffin, Filmotype Corporation, 7500 McCormick Blvd., Skokie, Illinois. Phone: OR 5-7210, Area Code 312.

1 Type 514/AD or Type 531 Oscilloscope. Frank Stabile, 1560 Brande Avenue, Anaheim, California. Phone: PR 4-5934.

1 Type 515A or Type 317 Oscilloscope. William Skidmore, 10756 Willworth Avenue, Los Angeles 24, California. Phone: GRanite 3-0403.

1 Type 515A Oscilloscope. Joe DeMichael, 12 New Haven Avenue, Derby, Connecticut. Phone: RE 5-5253.

1 Type 502 Oscilloscope. Oliver W. Osborne, American Geophysical & Instrument Co., 16440 S. Western Avenue, Gardena, California. Phone: FAculty 1-2634.

1 Type 515 Oscilloscope. Tom Burroughs, 557 Riford Road, Glen Ellyn, Illinois. Phone: 727-3441.

1 Type 515 or 514AD Oscilloscope. Instrument need not be in working condition but should be in good mechanical condition and electrically repairable. Contact: Chuck Keating, 23 S. E. 81st Street, Portland, Oregon, Phone: ALpine 3-9780.

Service Scope

USEFUL INFORMATION FOR

USERS OF TEKTRONIX INSTRUMENTS

Tektronix, Inc.
P. O. Box 500
Beaverton, Oregon