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Measuring Oscilloscope Voltage Probe Performance

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Abstract

GaN has exceeded the limits of oscilloscope voltage measurements. Not because of the oscilloscope, but because of the probe transfer function limitations and because of the lack of de-embedding. An additional constraint is that most probes are designed and calibrated using a 50 Ω signal generator. Unfortunately, there aren't any power supplies with 50 Ω impedance. Clearly a better method of measuring and calibrating oscilloscope probes, at GaN speeds, is needed.

In this session I'll provide three methods of accurately measuring oscilloscope probe response, one in time domain and two in frequency domain, one of which uses the oscilloscope as a high frequency Vector Network Analyzer (VNA).

Author Biography

Steve Sandler has been involved with power system engineering for nearly 40 years. Steve is the founder of PICOTEST.com, a company specializing in instruments and accessories for high performance power system and distributed system testing. He frequently lectures and leads workshops internationally on the topics of Power Integrity, and Distributed Power System Design and is a Keysight Certified EDA expert.

Steve publishes articles and books related to power supply and power distribution network ('PDN') performance. His latest book, "Power Integrity: Measuring, Optimizing and Troubleshooting Power-Related Parameters in Electronics Systems" was published by McGraw-Hill in 2014. He is the recipient of both DesignCon 2017 and EDICON 2017 Best Paper Awards.

Steve is also the founder of AEi Systems, a well-established leader in worst case circuit analysis, modeling, and troubleshooting of satellite and other high reliability systems.

Introduction

A key feature of the RF Vector Network Analyzer (VNA) is that it includes the ability to calibrate and de-embed the cables and fixtures. This is essential for high frequency measurement. In contrast, most oscilloscopes do not include such a capability. The oscilloscope probe may include communication with the oscilloscope for automatic compensation of the probe. The probe itself may have adjustments, some with up to three adjustments, for optimally compensating the probe. In the ideal case, the probe compensation is adjusted to obtain a flat frequency response up to the 3 dB bandwidth of the probe.

It’s my guess that in the (near) future, the oscilloscope will be able to de-construct the probe from measured data and properly de-embed it from the measurement. This will still not be good enough. The reasons why are the subject of this paper, and two methods of measuring the probe performance are shown, one in the frequency domain and one in the time domain. This measured performance data will help in the selection of an appropriate probe as well as modeling the probe for de-embedding or optimizing the measurement.

The Probe Model

The most common oscilloscope probe is the passive high impedance probe, probably because they are included with the oscilloscope. The probe model is shown in Figure 1. The probe is comprised of a series resistor with a parallel capacitor. These connect to a coaxial cable at the tip and two adjustable compensation capacitors at the oscilloscope mount. Some more sophisticated probes can electronically adjust the compensation, but this is a decent representation of a passive high impedance probe.

The probe ships with a variety of removable probe tips. These spring-loaded probe tips provide different retractable hooks to clip onto test points or wires located on the circuit being tested. The probe also ships with ground clips and typically also ground springs. The image in Figure 1 is shown using a ground spring, with the spring-loaded tip removed. Both the ground wire and the probe tip are represented as short filaments, and as such, each of these wires appears inductive.

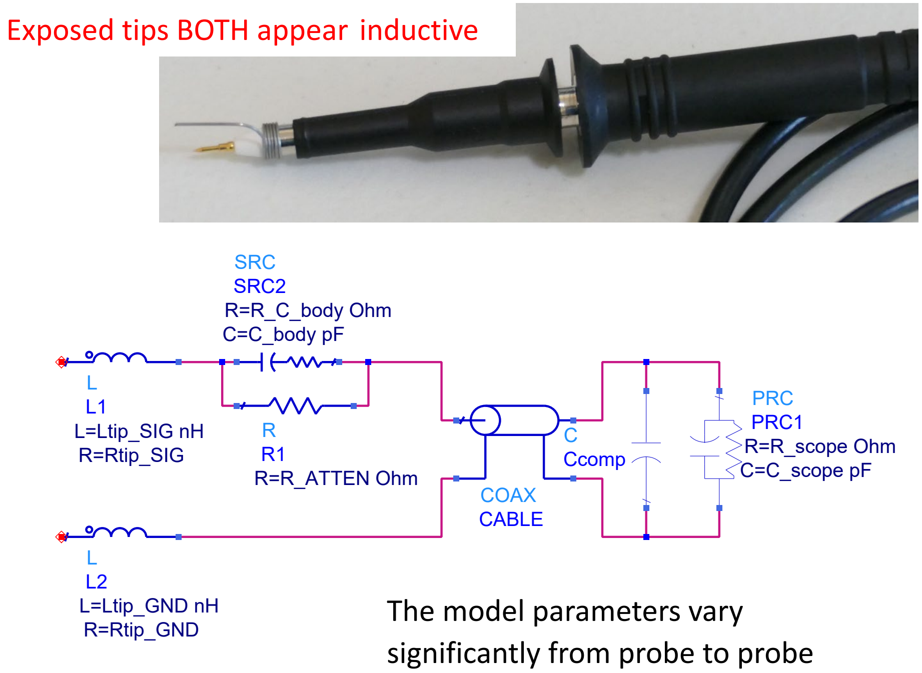

Figure 1 Typical high impedance probe with the spring-loaded tip cover removed and a short ground spring attached for a short ground connection. Electrical model below.

A simple rule is that we can directly measure the probe response so long as the magnitude of the probe impedance is much larger than the magnitude of the source impedance.

Measuring the Probe Directly

Unfortunately, it's rarely the case that the probe impedance is much greater than the source impedance at all frequenies. Consider a typical 500 MHz 10:1 probe with a specified 5.6pF tip capacitance. This is a pretty common specification. The 5.6pf capacitance at 500 MHz present a loading impedance of

$$Z_{tip} = \frac{1}{2\pi \cdot 500\ \mathrm{MHz} \cdot 5.6\ \mathrm{pF}} = 56\ \Omega \qquad (1)$$

We can directly measure the probe up to a frequency of 500 MHz *if the signal generator impedance is much lower than 56 Ω*. Unfortunately, this means that we can't directly measure the probe response using a 50 Ω signal generator, since the signal generator impedance is not much lower than the probe impedance.

There are a few additional complications. Since the probe tip wires appear inductive, they will resonate with the tip capacitance. We would prefer this resonance to be above the 500 MHz bandwidth of the probe, which limits the maximum tip inductance.

$$L_{max} = \frac{1}{(2\pi \cdot 500\ MHz)^2 \cdot 5.6\ pF} = 18\ nH\ total \qquad (2)$$

Assuming the ground and center tips are equal length and that each wire is a filament represented as 25 nH/inch, the maximum tip wire length for a 18 nH total tip inductance is 0.72 inches or 0.36 inches per lead.

At this frequency the inductance of the tips will resonate with the probe tip capacitance. The impedance at the tip will drastically fall due to the series resonance of the tip wire inductance and the probe tip capacitance, requiring an even lower source impedance to measure the probe. Adding just a bit more complication, we would like to measure the probe beyond its specified bandwidth in the event our source signal has an edge speed faster than the 500 MHz probe.

Method 1: 3 Step Process Using Correction Factor

If the source impedance isn't much lower impedance than the probe, one method is to apply a correction factor to null the attenuation of the probe, due to the generator source impedance and the probe tip impedance.

The use of a 50Ω source generator signal will result in the signal being attenuated due to the probe tip impedance. The probe tip is represented as the bottom of a voltage divider while the 50Ω source generator is represented as the top of the voltage divider. This can be seen visually in Figure 2.

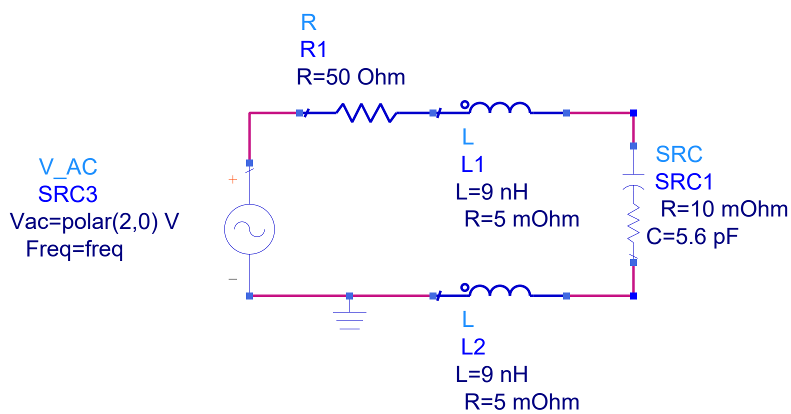

Figure 2 Considering the probe tip as inductance per lead and capacitance the 1 port reflection coefficient, S11 can be used to correct for the generator 50Ω source impedance. Here SCR3 is the source generator and R1 is the source generator 50 Ω impedance.

The signal at the probe tip, assuming a 50 Ω signal generator is:

$$V\_tip(Z11) = V\_source \cdot \frac{Z11}{Z11+50} \tag{3}$$

Substituting for Z11

$$Z11(S11) = 50 \cdot \frac{(1+S11)}{1-S11} \tag{4}$$

$$V_{tip(S11)} = V_{source} \cdot \frac{1+S11}{2} \tag{5}$$

The attenuation due to the probe tip can be cancelled by multiplying the measured value by the reciprocal of the attenuation. Note there is a factor of two missing in the correction factor. This is because the signal generator produces twice the signal voltage, expecting it to be halved by a 50 Ω load.

$$Correction = \frac{1}{1+S11} \tag{6}$$

First, the probe is simulated using a 1V zero-Ω source to obtain the correct response and probe tip impedance. These simulation results are shown in Figure 3.

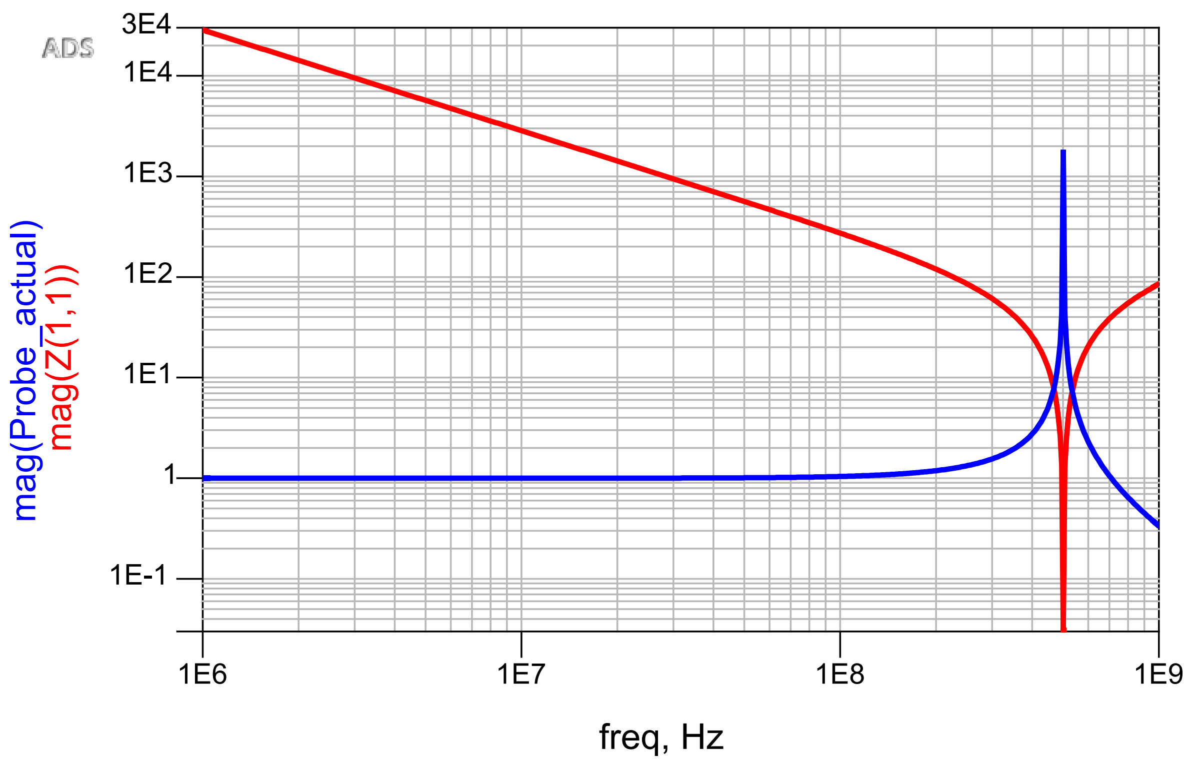

Figure 3 Simulating the probe with a zero-Ω source shows 65.3 dB peaking, which is higher than a typical probe, but serves the example well. The probe tip impedance is also shown here.

The probe tip is measured as a 1-port reflection measurement using a VNA, and the S11 data is exported. In this example, we are simulating the S11 measurement. The probe tip S11 simulation data is shown in Figure 4.

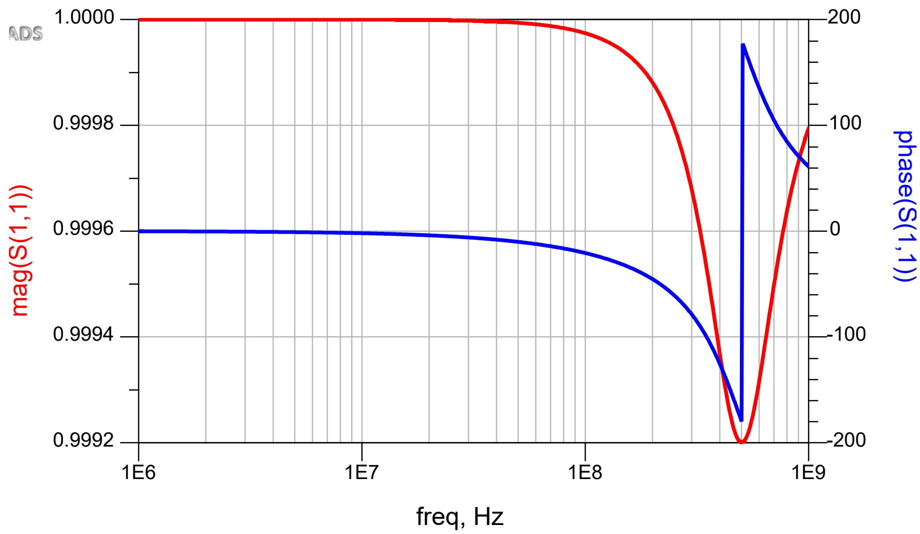

Figure 4 S11 measurement from a 1-port reflection s-parameter simulation.

The transfer function of the probe, S21, is measured using a swept 50 Ω signal generator or VNA. The results are shown in Figure 5.

Note that this measurement at low frequency shows 6 dB gain. This is because the signal source is calibrated assuming a 50 Ω load and here it is connected to a high impedance probe. Near resonance the probe shows 8 dB gain or 2 dB peaking above the 6 dB low frequency gain. This is not close to the correct result of 65.3 dB peaking shown in Figure 3.

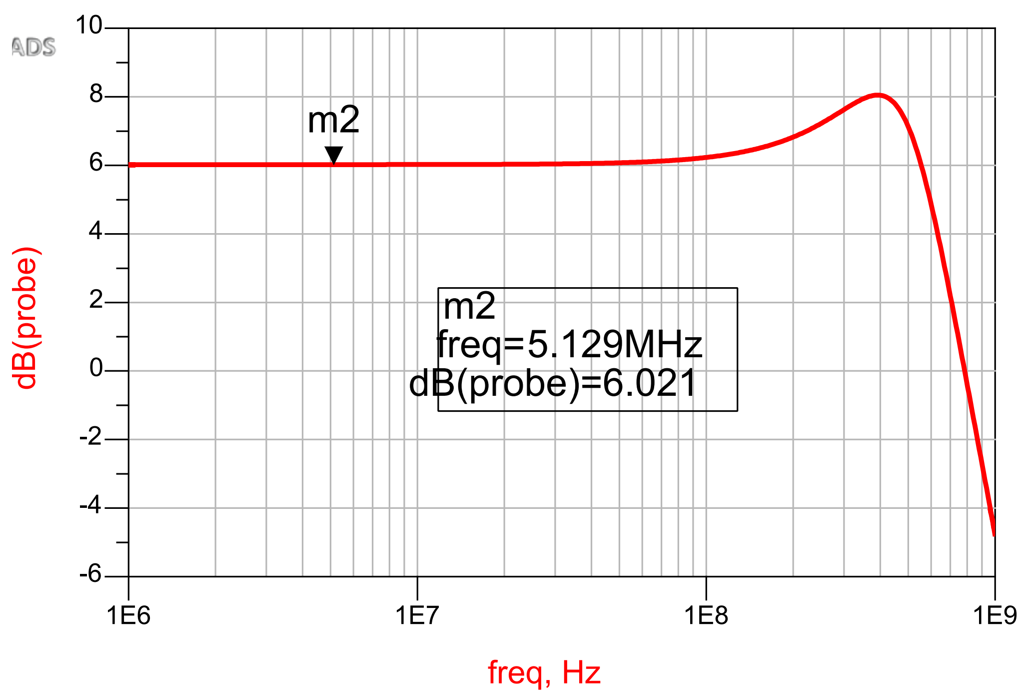

Figure 5 The simulated transfer function of the probe using the 50 Ω source generator.

The simulated transfer function is multiplied by the probe correction factor from the S11 measurement. The corrected result is shown in Figure 6.

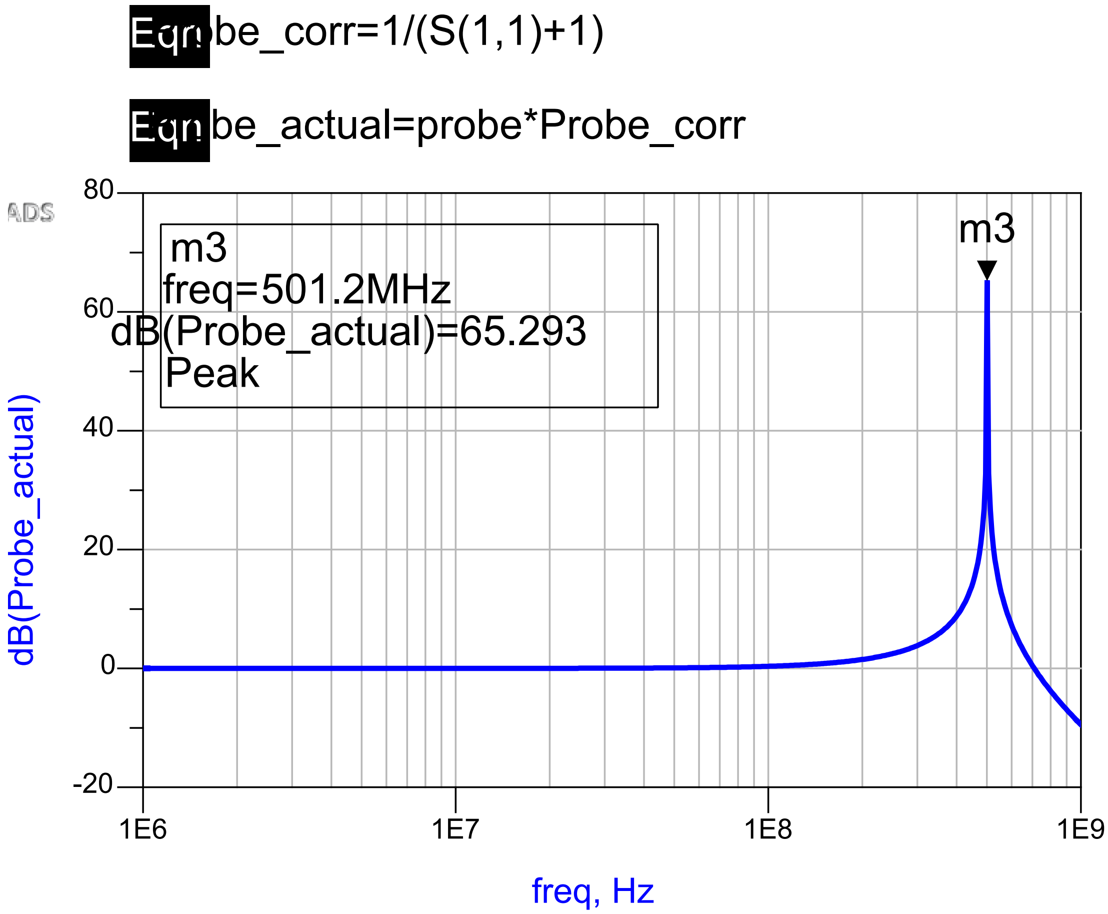

Figure 6 The product of the swept transfer function and the S11 based correction factor shows 65.3 dB peaking at the probe resonance of 500 MHz. This is in perfect agreement with the zero-Ω source measurement, validating the method.

The corrected result, shown in Figure 6, is in perfect agreement with the results in Figure 3. This method works, but is a bit cumbersome as it requires a separate S11 measurement of the tip and the correction is post-processed from the data. Also, the measurement is not made in conjunction with the oscilloscope but with a separate VNA. The results using the oscilloscope might not match the VNA results.

Method 2: Near Zero Impedance Hyper-Fast Step

In the introduction I stated that the measurement could be performed directly if the source generator impedance is much lower than the impedance of the probe tip across the entire frequency range of use.

The bandwidth of most high Z probes is less than 1 GHz, so allowing an additional octave above this frequency, the rise time of the step applied to the probe needs to be faster than,

$$T_{fall} = \frac{0.35}{2\ GHz} = 175\ ps \tag{7}$$

Also, the impedance of the step must be much less than the probe tip impedance. Again, using a 10:1, 500 MHz, 5.6pF capacitance passive probe the pulser impedance needs to be much lower than the 56 Ω probe impedance. This is possible using a low impedance GaN element with an ultra-high-speed driver. A prototype probe "pulser" is shown in Figure 7.

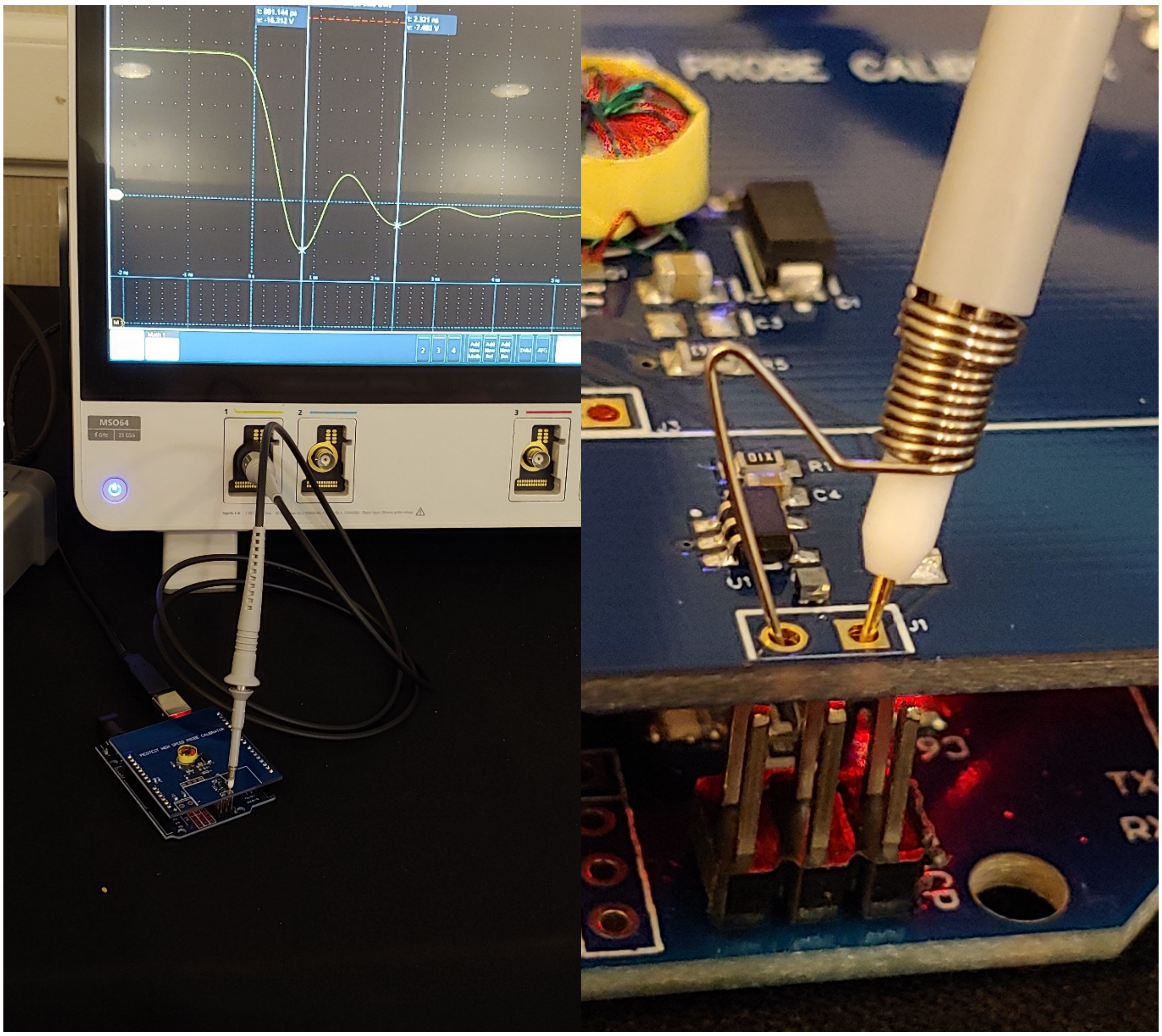

Figure 7 A hyper-fast GaN pulse generator charges the probe tip to a programmable voltage, up to 100V. The probe is then discharged using a 0.5 Ω GaN FET. The result is a time domain response as seen in the ringing here. The probe tip capacitance is measured using the rise time to charge the probe tip.

The pulser directly measures the time domain response on the falling edge and the probe capacitance on the rising edge, as shown in Figure 8.

Knowing the capacitance, the ringing frequency and the damping, it's possible to completely model the probe using inverse Laplace transformation.

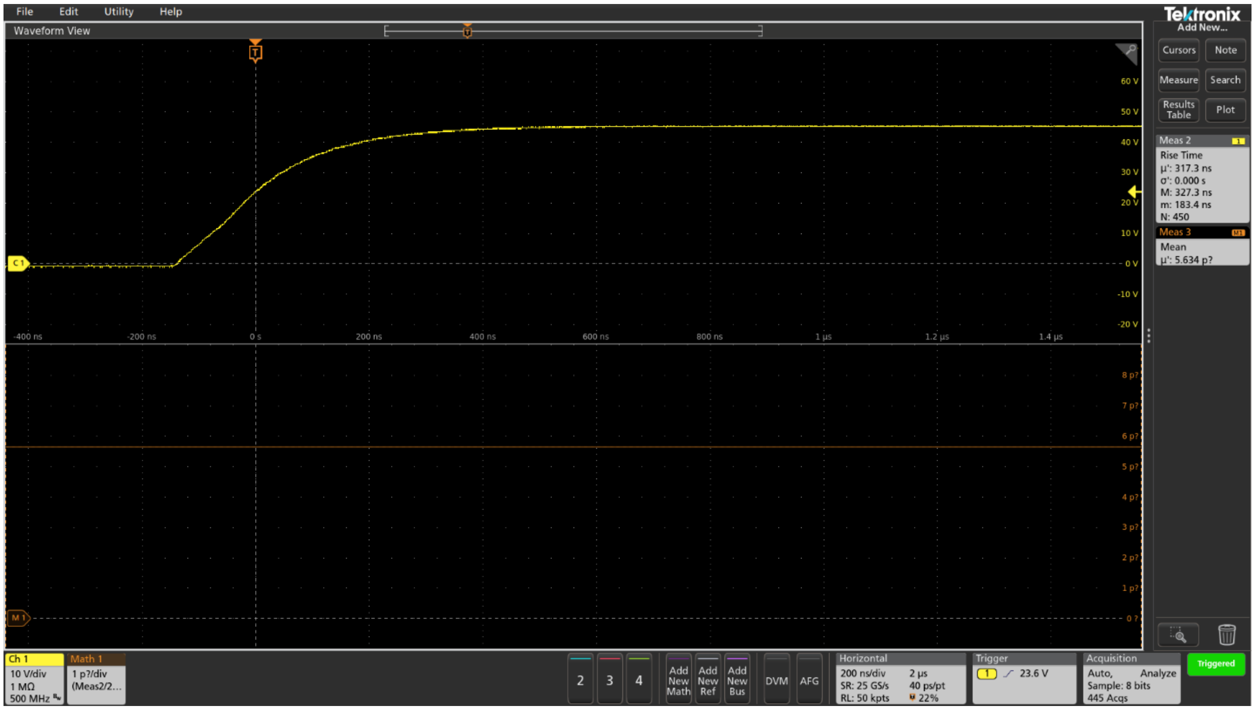

Figure 8 The rise time of the pulser is used to measure the tip capacitance. This probe measures 5.63pF, which closely matches the 5.6pF listed on the probe label.

It's still important to verify that the pulser itself isn't ringing and that the pulser meets the fall time requirement of 175 ps. This is verified by measuring a Tektronix P6150, 9 GHz, passive probe as shown in Figure 9. The fall time of the pulser, the oscilloscope and the probe combined is 179 ps. Accounting for the 9 GHz probe and the 8 GHz oscilloscope results in an actual fall time of 169 ps, besting our 175 ps goal.

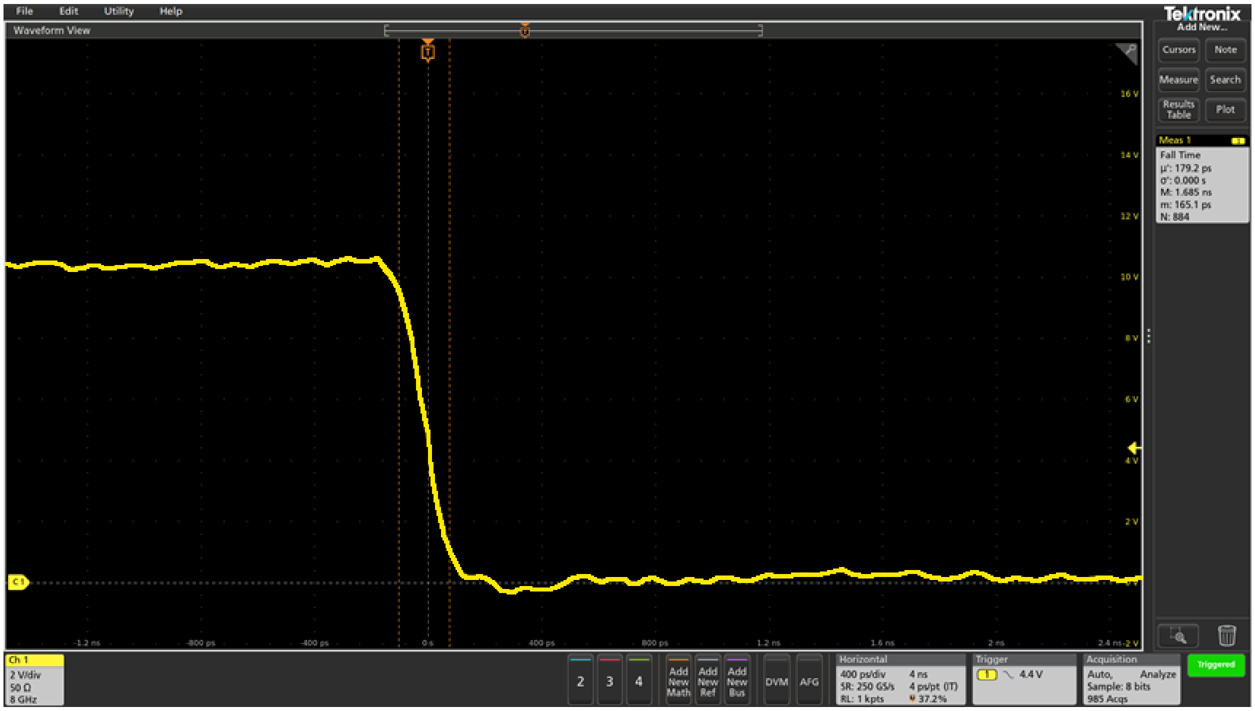

Figure 9 The measurement of the Tektronix P6150 9 GHz probe indicates a 179 ps fall time. There is no ringing, preshoot, or overshoot observed. The 179 ps accounts for the probe pulser, the probe and the oscilloscope rise times. De-embedding the scope bandwidth and the specifed probe bandwidth results in 169 ps fall time, meeitng our 175 ps goal.

This second method measures the probe together with the oscilloscope, which is simple but is limited to 2 GHz probes. Most active probes, including power rail probes, have greater than 2 GHz bandwidth, so yet another method is needed if we want to measure active probes.

Method 3: High Frequency Oscilloscope-Based VNA

In this third method, the oscilloscope is configured as a RF VNA. Applying a high bandwidth signal, the probe response can be measured, together with the oscilloscope, up to the bandwidth of the oscilloscope and probe combined.

In this example, we are using the Picotest J2151A PerfectPulse® Fast Edge Signal Generator, which has a certified 3 dB bandwidth of 10.5 GHz. The J2151A includes a 15 GHz, two-resistor power divider that is used to measure the probe tip signal using one oscilloscope channel. The other end of the probe is connected to a second oscilloscope channel, configured to match the probe. This can be either 50 Ω or 1 MΩ, as shown in Figure 10.

Since the power divider introduces a 6 dB loss, the probe factor on the probe tip channel is set to 2X and since the scope channel and the pulse generator are both 50 Ω, the Thevenin impedance at the probe tip is also 50 Ω. This will prove to be beneficial.

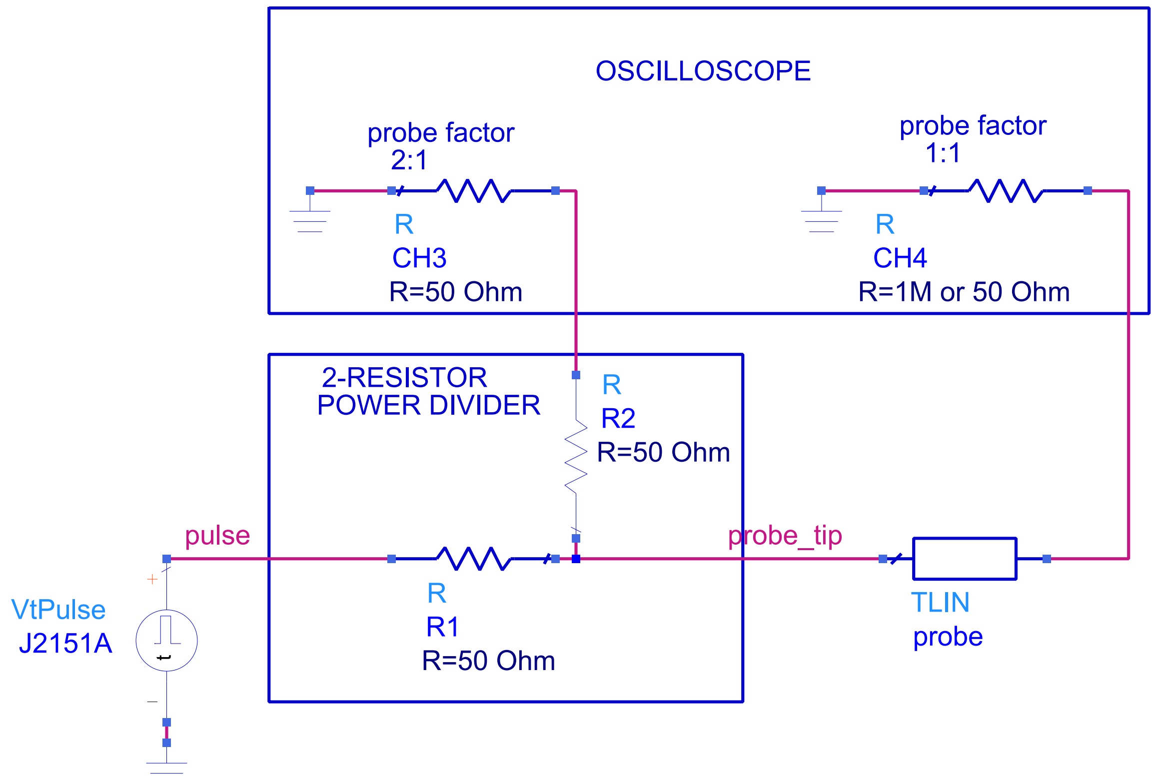

Figure 10 A 6dB power divider included with the J2151A provides a 50 Ω Thevenin impedance at the probe tip. Oscilloscope CH3 directly measures the probe tip voltage. The 50 Ω impedance allows the full oscilloscope bandwidth. The other end of the probe connects to CH4 using either a 50 Ω or 1 MΩ termination, as needed to match the probe.

A math function divides the probe output (CH4) by the probe tip (CH3) using FFT of derivatives. This results in a flat source with a bandwidth of more than 10 GHz. The Math function is shown in equation 8.

$$Math1 = dB\big(FFT(DIFF(CH4)\big) - dB(FFT(DIFF(CH3)) \qquad (8)$$

The attenuation, resulting the 50 Ω source and the probe tip capacitance, is eliminated from the measurement since the probe tip voltage is measured directly at CH3.

In general, I strongly recommend measuring something known before measuring something unknown in order to validate your test setups. A short length of 18 GHz rigid coax cable is used as the probe for the measurement in Figure 9 and the setup is shown in Figure 11. The measurement shows a flat response all the way to 10 GHz. You might be wondering why this measurement doesn't show the scope roll-off at 8 GHz. The answer is simple. Since the 8 GHz roll-off of both channels is very similar, they cancel in the division. As a result, the measurement can go above the bandwidth of the oscilloscope so long as the sample rate can support it. This requires a sample rate of at least 2x the bandwidth. A sample rate of 25 GS/s could theoretically measure to 12.5 GHz. In reality, the oscilloscope filters will limit the usable bandwidth to between 9 GHz and 10 GHz.

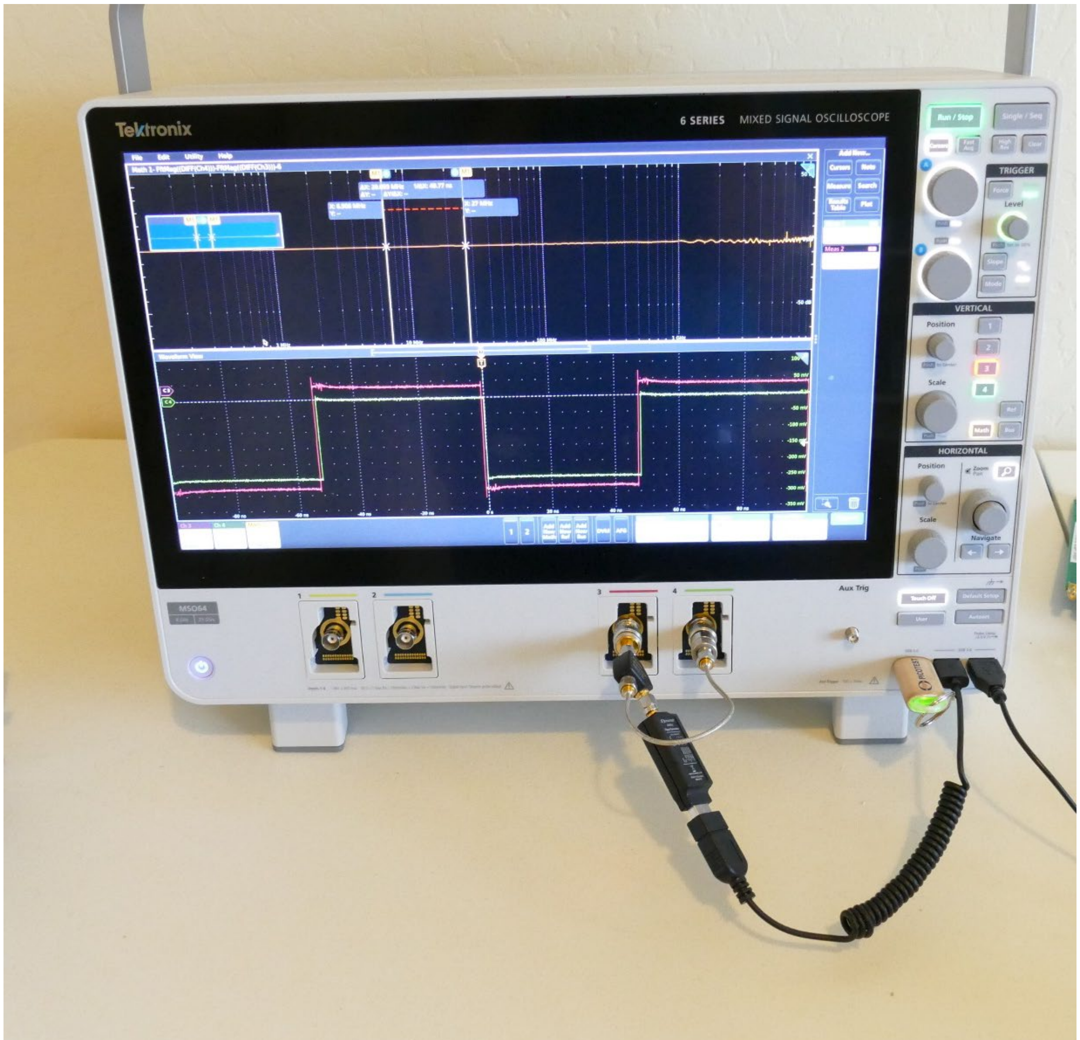

Figure 11 Replacing the probe with a short 18 GHz cable is expected to provide a flat response over the complete bandwidth of the oscilloscope, and this measurement confirms that it does. Note the measurement is also valid above the bandwidth of the signal generator since this is also applied to both channels and cancels in the division.

With the setup verified, an example measurement is shown in Figure 12 using an actual passive oscilloscope probe. A probe tip adapter connects the probe tip to the power divider using the shortest possible connection. The time domain signals are shown in the lower display window and the probe transfer function is shown in the upper window.

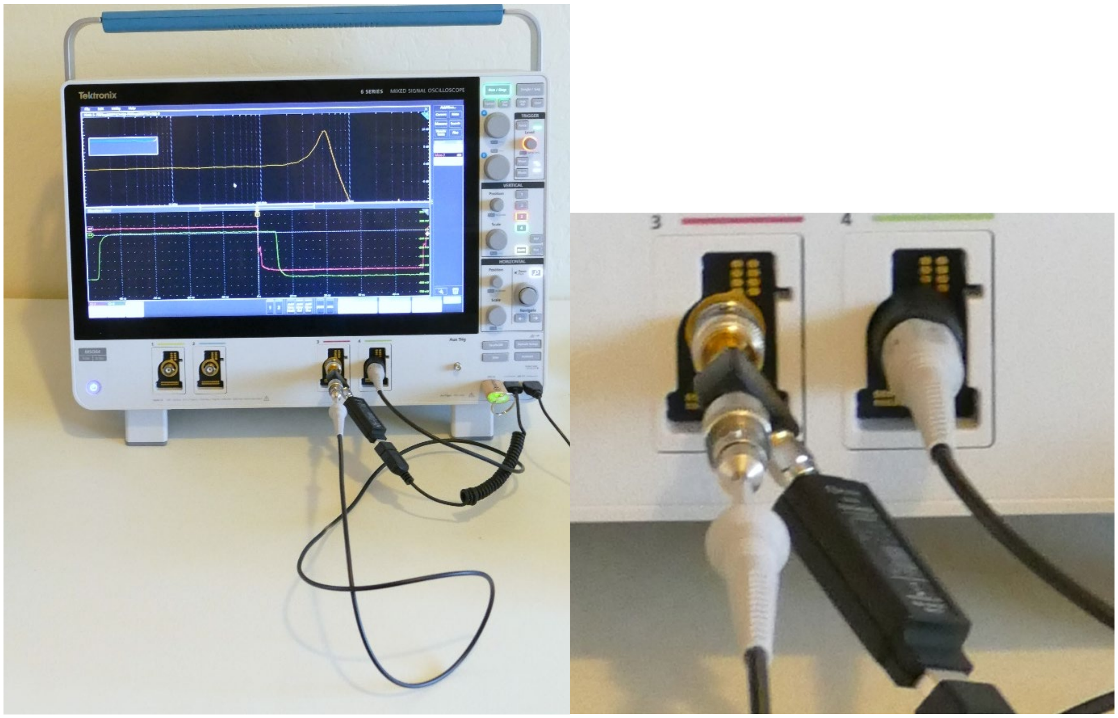

Figure 12 Measurement of a passive scope probe using the Picotest J2151A PerfectPulse TDR.

Measuring an active probe, such as a power rail probe, is quite similar except that the oscilloscope will automatically set CH4 to 50 Ω instead of 1 MΩ. The 4 GHz roll-off is clearly seen in Figure 13 and note there is no peaking. This very flat response, along with very low tip capacitance, are the main advantages of using the power rail probe to measure wideband noise.

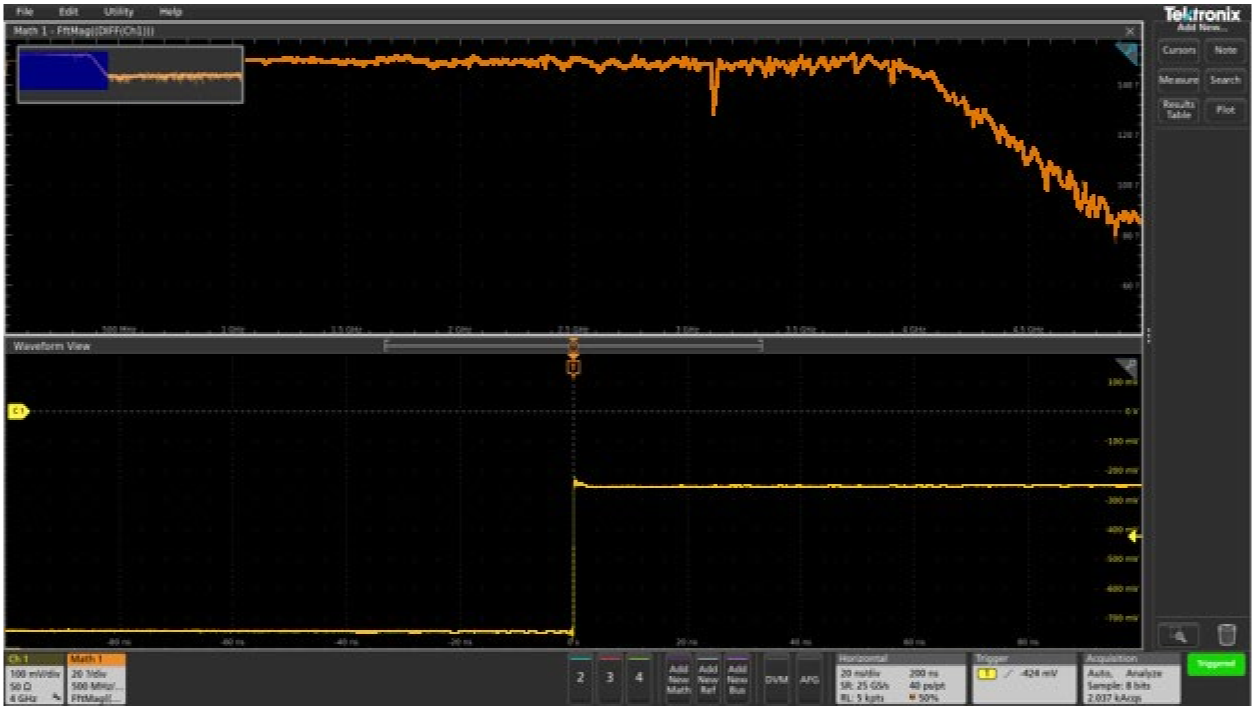

Figure 13 Using the Tek Series 6 oscilloscope as a high frequency VNA to measure power rail probe performance. The oscilloscope supports the frequency domain test and display.

It is also possible to replicate a wideband VNA using the oscilloscope, simply by rotating the power divider. This provides a 50 Ω source impedance, just as a VNA would, and the setup is shown in Figure 14. The only change is the power divider is rotated.

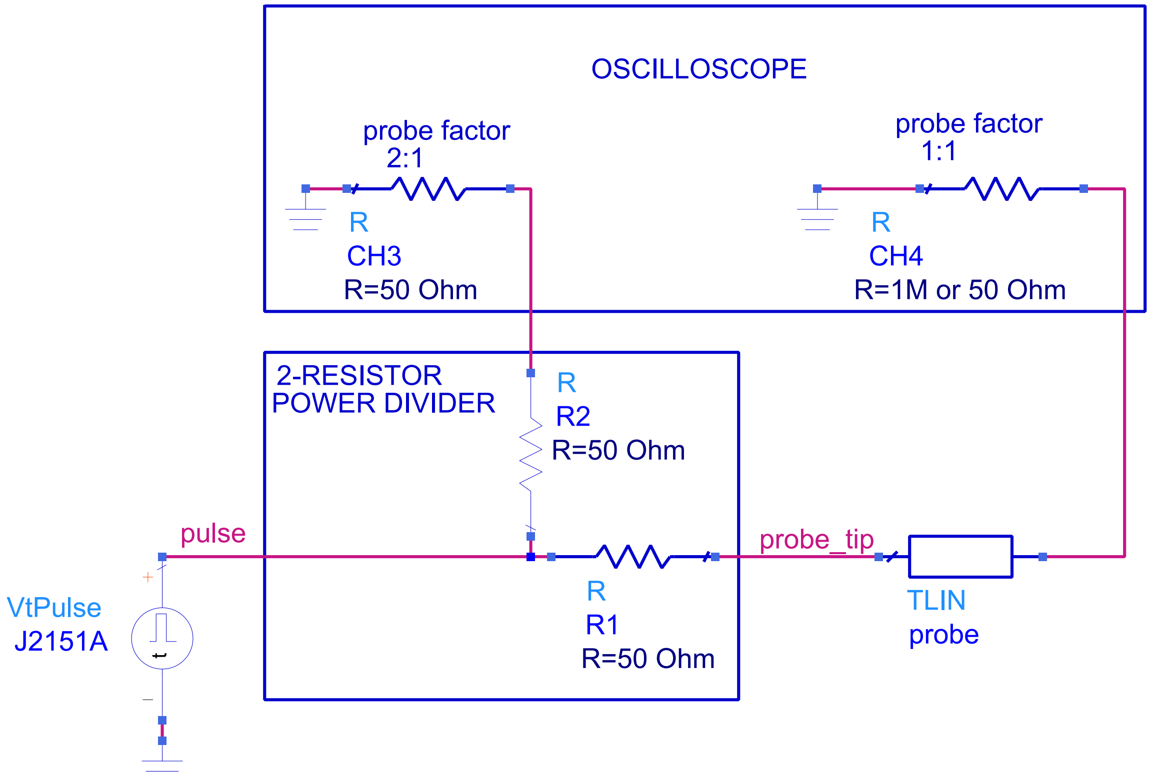

Figure 14 Rotating the J2151A's 15GHz power divider results in measuring the signal ahead of the 50 Ω source resistor. This provides a 50 Ω source signal, much like a VNA. Oscilloscope CH4 again measures the other end of the oscilloscope probe using either 50 Ω or 1M impedance as required to match the probe.

The source signal is now measured in front of the 50 Ω source resistor, R1, just as in a VNA. Using the same math function shown in equation 8, the result will be the probe transfer function with a 50 Ω source in place. While I have clearly stated this will produce an incorrect result for the probe transfer function, it does still have a benefit.

Using the results from Figure 10 and Figure 14 allows the extraction of S11 from the two S21 measurements in order to create a complete probe model. This S11 extraction also confirms the transformations since the extracted value matches the original result, shown in Figure 3.

Similarly, S11 can be transformed to Z11 using equation 4 to derive the probe tip impedance, as shown in Figure 16. The transformed Z11 result is shown in Figure 16 and is in perfect agreement with the simulation results shown in Figure 3.

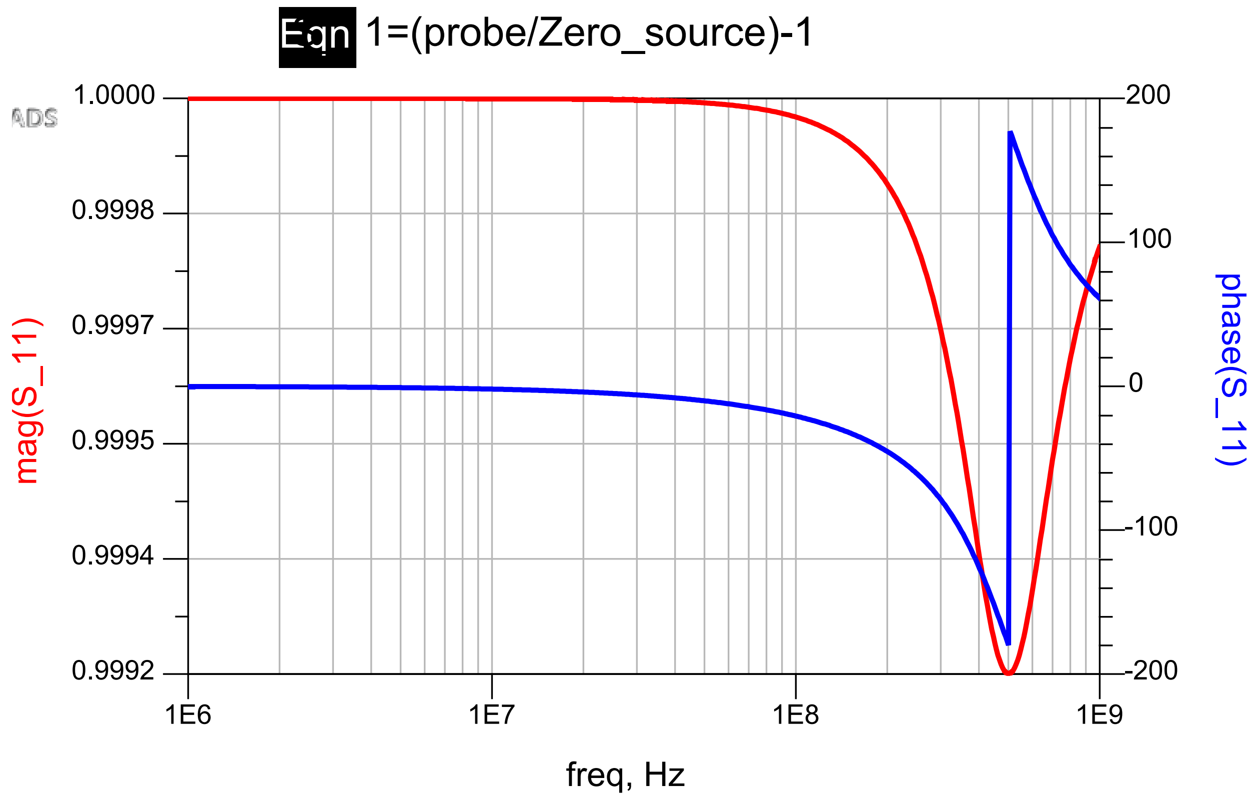

Figure 15 Having probe measurements with and without the 50 Ω source allows computation of S11 using equation 5. Note that this figure is in perfect agreement with the direct S11 seen in Figure 4.

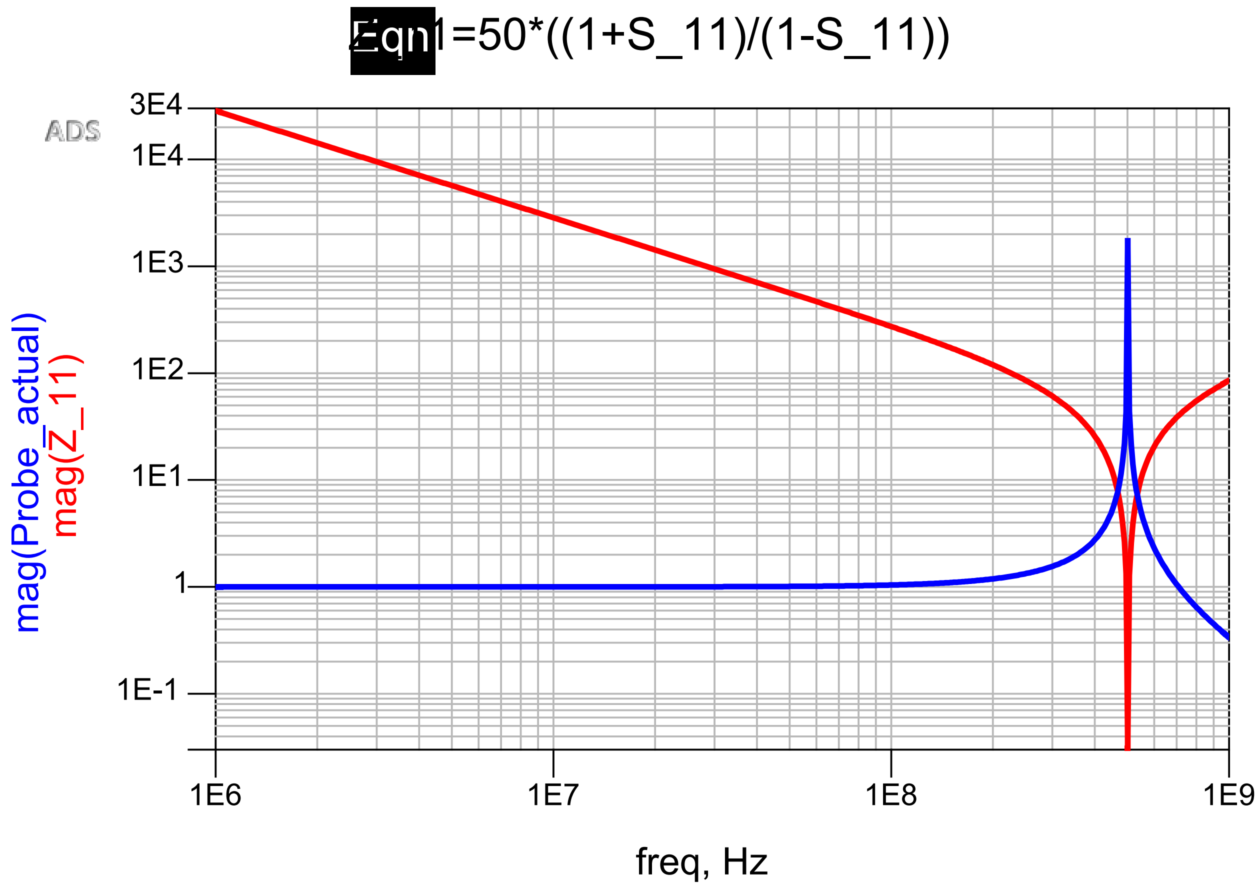

Figure 16 Using equation 4 to convert the data from S11 to Z11 perfectly matches the result in Figure 3, confirming the transform.

The results of Figure 16 confirm the transformations and the relationships between the various measurements. The results can be used to determine the probe transfer function, without the included 50 Ω series resistor, and to determine the probe tip impedance. The tip impedance is necessary to determine the interaction between the probe and the DUT, for example due to the PCB trace leading to a probe test point.

This is my preferred method since the measurement is direct and is an end-to-end measurement inclusive of both the probe and the oscilloscope. This method also supports much higher bandwidth than the time domain pulser.

A drawback of this method is that computing the tip impedance requires two measurements and post processing outside of the oscilloscope in order to develop a complete probe model.

Summary and Conclusions

Engineers spend hours troubleshooting design problems that ultimately turn out to be probe-related issues. A lot of time can be saved by measuring your probes in advance, so you know how the probe contributes to measurement inaccuracies.

I've shown three different methods of measuring the transfer function of oscilloscope probes. Each has advantages and disadvantages, though the most direct and widest bandwidth method uses the oscilloscope as a wide bandwidth VNA, as shown in Figure 10.

The second-best method uses the time domain, though it requires inverse laplace outside of the oscilloscope to determine the probe transfer functions. Combining the setup in Figure 10 and Figure 14 provides the tip impedance and transfer function all within the oscilloscope environment.

I would venture to guess that the oscilloscope will eventually include the capability to completely de-embed the probe but, like most high frequency VNAs, will de-embed cables and fixtures.

Whichever method you choose, it's important to know the transfer function of the probe to be certain that the probe is not degrading the accuracy of the measurement.

A probe that doesn't exhibit a flat response will yield incorrect results. Faster switching devices such as GaN can easily excite high frequency resonances in the probe transfer function, making it even more important to assure the probe flatness to obtain accurate measurements.

Also, keep in mind that circuit board traces, ground leads, and other connections will also contribute to the measurement accuracy. These artifacts must also be accounted for in the measurement, but in this paper, I focused only on the probe itself. The additional PCB traces will typically appear as inductive and can be derived either from an EM simulation of the PCB or using a TDR to physically measure the trace.

In case you haven't seen responses like this in your oscilloscope probes, see Figure 17. This probe has two distinct peaks, the largest at more than 23 dB. One is at 400 MHz and the other is located at above 1 GHz. Also note there is a dip in the response beginning at 100 MHz and ending at about 250 MHz.

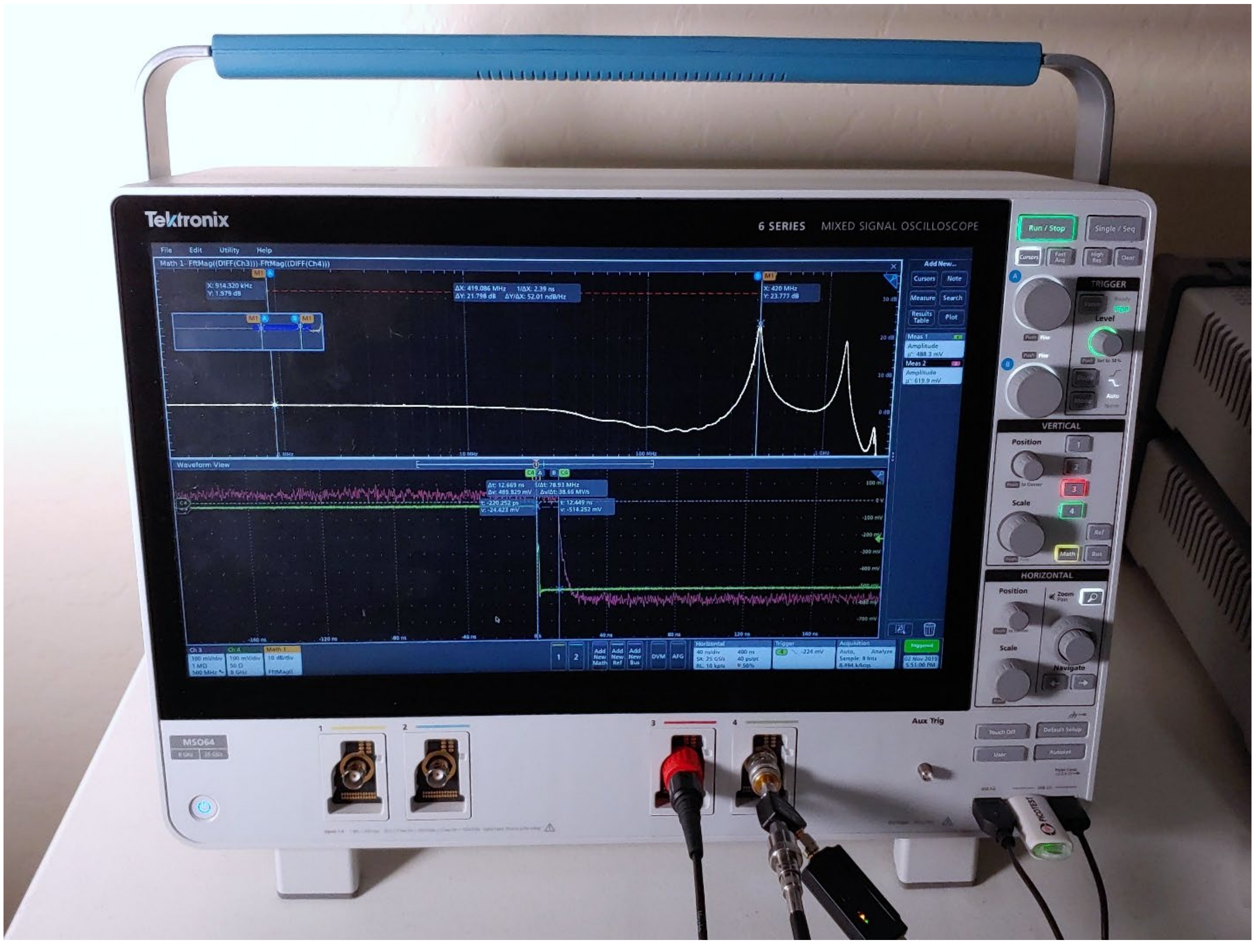

Figure 17 This probe shows two peaks with the larger peak at 23 dB. The -3 dB bandwidth is 2.3 GHz, but the in-band flatness is 23 dB. The result will be significant ringing in the oscilloscope measurement that is created by the probe resonance. It is not due to the switching waveform being measured.