Analysis and Measurement of Intrinsic Noise in Op Amp Circuits
Part XII: Photodiode Amplifier Noise (Continued)

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This TechNote covers a sample calculation of the photodiode circuit from Part XI http://www.en-genius.net/includes/files/avt_120511.pdf The sample calculation uses a real world photodiode and operational amplifier (op amp). Part XI also shows the measurement results for this example circuit. In this example measurement of a post-amplifier is required to increase the noise amplitude to a level that the equipment can accurately measure.

Photodiode Example Specifications

Fig. 12.1 shows some typical specifications for a photodiode. The photodiode used in this example is intended for lower frequency applications like smoke detectors, dimmers, and bar code scanners.

Symbol	Characteristic	Condition	Min	Typ	Max	Units
Rsh	Shunt resistance	Vr = 10 mV	100	150		MΩ
Cj	Junction capacitance	Vr = 0 V		70		pF
		Vr = 10 V		10	25	
I_D	Dark current	Vr = 0 V		2	30	nA
V_{BR}	Breakdown voltage	I = 10 µA	30	75		V

Fig. 12.1: Photodiode Specifications

Photodiode Current Noise Calculations

Fig. 12.2 calculates the total current noise spectral density from the photodiode. Part of this current noise is from the photodiode shunt resistance, and part is from shot noise. One shot noise term is zero because the diode has no photocurrent.

EQUATION	DESCRIPTION
$k_b := 1.38 \times 10^{-23} \frac{J}{K}$	Boltzmann constant
$T_n := 298K$	Temperature in Kelvin (25°C)
$R_{sh} := 150 \times 10^6 \Omega$	Shunt Resistance in photodiode
$i_j = \sqrt{\frac{4k_b \times T_n}{R_{sh}}} = 10.472 \times 10^{-15} \frac{A}{\sqrt{Hz}}$	Thermal (Johnson noise)
$q := 1.602 \times 10^{-19} C$	One electron charge
$I_D := 2 \times 10^{-9} A$	Dark current in photodiode
$i_{sL} = \sqrt{2q \times I_D} = 25.314 \times 10^{-15} \times \frac{A}{\sqrt{Hz}}$	Shot noise (dark)
$I_L := 0\ A$	Photo current in photodiode (our measurements are dark)
$i_{sL} := \sqrt{2q \times I_L} =$	Shot noise (w. Light)
$i_{n\_diode} = \sqrt{i_j^2 + i_{sD}^2 + i_{sL}^2} = 27.4 \times 10^{-15} \times \frac{A}{\sqrt{Hz}}$	Total diode current noise

Fig. 12.2: Photodiode Current Noise Spectral Density

Op Amp Specifications

The OPA827 is the amplifier we choose to use for our example photodiode amplifier circuit. The table shown in Fig. 12.3 is a subset of the OPA827 data sheet. The key parameters are highlighted in yellow.

Parameter	Standard Grade OPA827			Unit
	Min	Typ	Max	
Noise				
Input voltage noise f = 0.1 Hz to 10 Hz e_n		250		nVpp
Input voltage noise density f = 1 kHz e_n		4		nV/√Hz
f = 10 kHz e_n		3.8		nV/√Hz
Input current noise density f = 1 kHz i_n		2.2		fA/√Hz
Input Impedance				
Differential		10^{13} || 9		Ω || pF
Common-mode		10^{13} || 9		Ω || pF
Frequency response				
Gain-bandwidth product GBW		22		MHz

Fig. 12.3: Specifications From OPA827 Data Sheet

Op Amp Voltage Noise Calculations

Fig. 12.4 shows calculations of the key poles and zeros in the voltage noise gain curve. The poles and zeros separate the different regions of the noise gain curve. The poles and zeros depend on photodiode capacitance, feedback resistance R_f, feedback capacitance C_f, and the unity gain bandwidth of the op amp used.

EQUATION	DESCRIPTION
$R_f := 100\ k\Omega$	Feedback resistance
$C_f := 4\ pF$	Feedback capacitor
$C_j := 70\ pF$	Photodiode junction capacitance (from photodiode manufacturer)
$C_{opa} := 18\ pF$	Opamp input capacitance (OPA827 data sheet)
$C_i := C_j + C_{opa}$	Total input capacitance
$f_c := 22\ MHz$	Unity gain bandwidth (OPA827 data sheet)
$f_p = \frac{1}{2\pi R_f \times C_f} = 398\ Hz$	
$f_z = \frac{1}{2\pi R_f \times (C_i + C_f)} = 17.3\ kHz$	
$f_i = \frac{C_f}{C_i + C_f} \times f_c = 957\ kHz$	

Fig. 12.4: Poles And Zeros In Noise Gain Curve

Fig. 12.5 calculates the 1/f noise voltage corner frequency which separates the 1/f region from the broadband region on the op amp voltage noise spectral density curve.

EQUATION	DESCRIPTION
$e_{nif} := 3.8 \frac{nV}{\sqrt{Hz}}$	Broadband noise spectral density (OPA827 data sheet)
$f_L := 0.1 Hz$	Lower bound on frequency (1/f region) (arbitrary lower bound of frequency)
$e_{at\_f} := 60 \frac{nV}{\sqrt{Hz}}$	Flicker noise measured at f_L (OPA827 data sheet noise curve)
$e_{fnorm} = e_{at\_f} \times \sqrt{f_L} = 10\ nV$	
$f_f = \frac{e_{fnorm}^{2}}{e_{nif}^{2}} = 24.9\ Hz$	

Fig. 12.5: 1/f Noise Corner

Fig. 12.6 shows the total voltage noise calculation. Each term E_{noe1} through E_{noe5} corresponds to a different region in the noise output voltage curve. That, the equations, and the delineation between the different regions are provided in Part XI. The total noise is calculated by the root sum of the squares of each region term. Note that Eqs. 3, 4, and 5 dominate the total noise. In this case the 1/f noise performance is not a significant issue.

EQUATION	DESCRIPTION
$E_{noe1} = \sqrt{e_{nif}^{2} \times f_f \times 1n\left(\frac{f_f}{f_L}\right)} = 44.9\ nV$	
$E_{noe2} = \sqrt{e_{nif}^{2} \times (f_z - f_f)} = 499\ nV$	
$E_{noe3} = \sqrt{\left(\frac{e_{nif}}{f_z}\right)^{2} \times \frac{f_p^{3} - f_z^{3}}{3}} = 31.8\ \mu V$	
$E_{noe4} = \sqrt{\left(e_{nif} \times \frac{C_i + C_f}{C_f}\right)^{2} (f_i - f_p)} = 65.3\ \mu V$	
$E_{noe5} = \sqrt{\frac{(e_{nif} \times f_c)^{2}}{f_i}} = 85.4\ \mu V$	
$E_{noe} = \sqrt{E_{noe1}^{2} + E_{noe2} +^{2} E_{noe3} +^{2} E_{noe4} +^{2} E_{noe5} +^{2}} = 112\ \mu V$	

Fig. 12.6: Total Rms Output Noise From Op Amp Voltage Noise Source

Thermal (Resistor) Noise Calculations

Fig. 12.7 shows the calculation for thermal (resistor) noise from the feedback resistor. The feedback capacitor limits this noise by limiting the bandwidth (f_p).

EQUATION	DESCRIPTION
$R_f = 100\ k\Omega$	Feedback resistance
$k_b = 1.38 \times 10^{-23} \frac{J}{K}$	Boltzmann constant
$T_n = 298\ K$	Temperature in Kelvin (25°C)
$f_p = 397.887 \times 10^3\ Hz$	Transconductance bandwidth
$K_n = 1.57$	Noise current from OPA827 data sheet
$BW_n = K_n \times f_p$	Noise bandwidth (brick wall filter)
$e_{n\_r} = \sqrt{4k_b \times T_n \times R_f \times BW_n} = 33\ \mu V$	Thermal noise at output

Fig. 12.7: Total Rms Output Noise From Feedback Resistor Thermal Noise

Op Amp Current Noise Calculations

Fig. 12.8 shows the hand calculation for the transimpedance amplifiers current noise. The current noise is translated into a voltage noise by the feedback resistor. The feedback capacitor limits this noise by limiting the bandwidth (f_p).

EQUATION	DESCRIPTION
$R_f = 100 \times 10^3 \Omega$	Feedback resistance
$f_p = 397.887 \times 10^3 \text{Hz}$	Transconductance bandwidth
$i_{n\_opa} = 2.2 \times 10^{-15} \frac{A}{\sqrt{Hz}}$	Noise current from OPA827 data sheet
$i_{n\_diode} = 27.395 \times 10^{-15} \frac{A}{\sqrt{Hz}}$	Noise current from diode (calculated)
$i_{n\_total} = \sqrt{i_{n\_opa}{}^2 + i_{n\_diode}{}^2}$	Total noise current
$K_n = 1.57$	Noise bandwidth factor first order filter
$BW_n = K_n \times f_p = 624 \text{ kHz}$	Noise bandwidth (brick wall filter)
$E_{noI} = i_{n\_total} \times R_f \times \sqrt{BW_n} = 2.17 \ \mu V$	Current noise at output

Fig. 12.8: Total Rms Output Noise From Op Amp Current Noise

Total Noise For Example Transimpedance Amplifier

Fig. 12.9 is the total noise including all the noise components considered in this analysis (ie the op amp voltage noise, resistor noise, and current noise). In this case the op amp voltage noise is the dominant noise source.

EQUATION	DESCRIPTION
$E_{noe} = 112\ \mu V$	Op amp voltage noise
$E_{noR} = 32\ \mu V$	Resistor noise
$E_{noI} = 2.17\ \mu V$	Op amp voltage noise
$E_{no} = \sqrt{E_{noR}^2 + E_{noI}^2 + E_{noe}^2} = 117\ \mu V$	Total output noise for OPA827 transimpedance amp

Fig. 12.9: Total Rms Output Noise Including All Components

SPICE Analysis Of Example Circuit

The circuit required to perform as TINA SPICE analysis for the example circuit is shown in Fig. 12.10. Note that the full model for the diode is not required. In this analysis we only use the diode junction capacitance Cj = 70 pF.

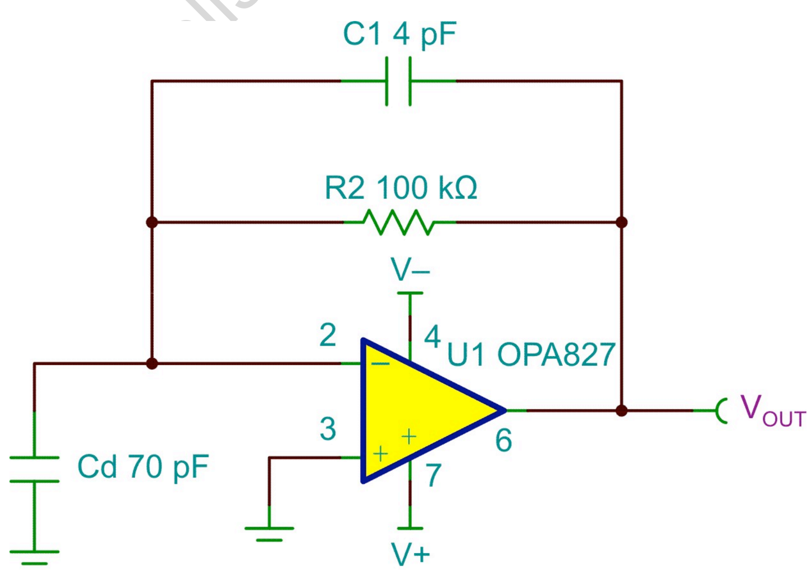
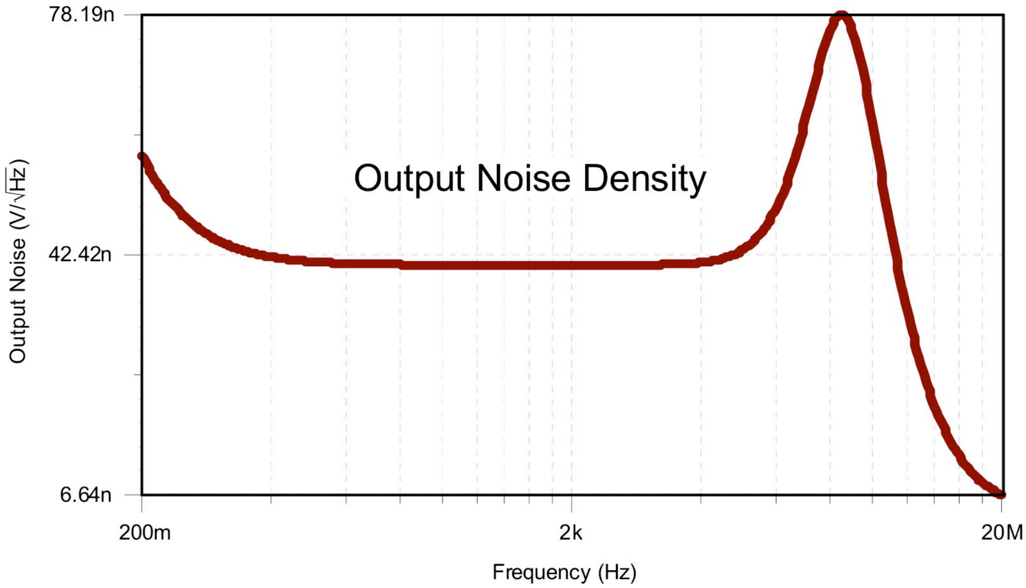


Fig. 12.10: Example Circuit For TINA SPICE Simulation

Fig. 12.11 shows the results of a TINA SPICE noise simulation for the circuit shown in Fig. 12.10. The output noise voltage spectral density curve shows noise peaking of 78 nV/√Hz. The noise density curve can be generated in TINA SPICE using the "output noise" option on a "noise analysis." The total noise curve is the integration of the power spectral density. The total noise converges to 109 µVrms; this corresponds well to the total noise of 117 µV from the hand calculations.



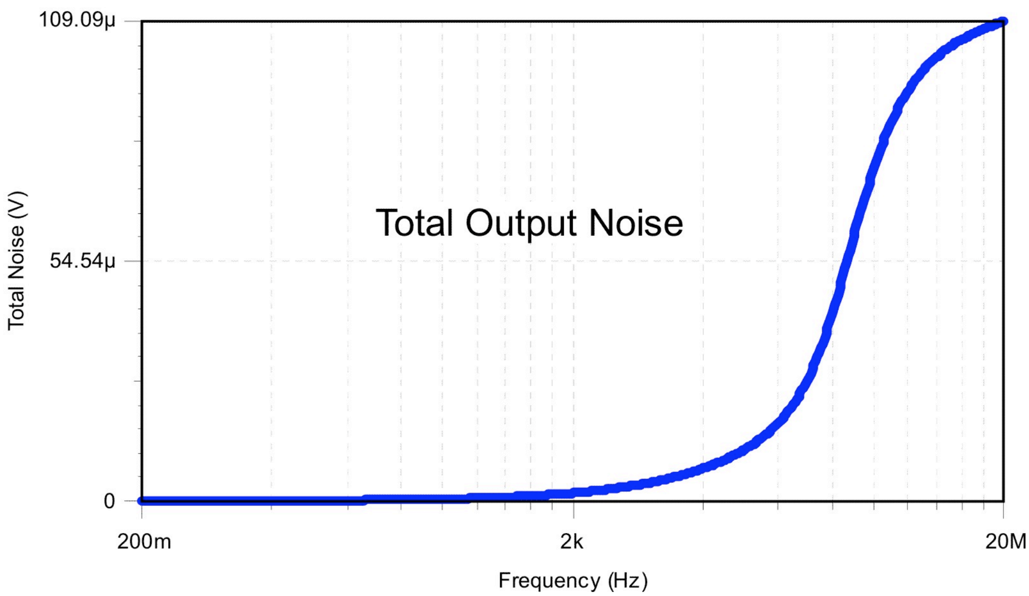


Fig. 12.11: TINA SPICE Results For Example Circuit

Fig. 12.12 (overleaf) shows the affect of adjusting the feedback capacitor on bandwidth, spectral density, and total noise. The figure compares the 4 pF capacitor used in the previous calculations with a 2 pF capacitor. Generally, decreasing the feedback capacitor increases the I-to-V bandwidth. In the example shown the effect on the I-to-V bandwidth is not obvious, but there is noticeable gain peaking for the 2 pF case. The gain peaking is indicative of marginal stability. The noise spectral density curve shows significant noise peaking for the 2 pF capacitor. This additional noise peaking corresponds to attritional total noise.

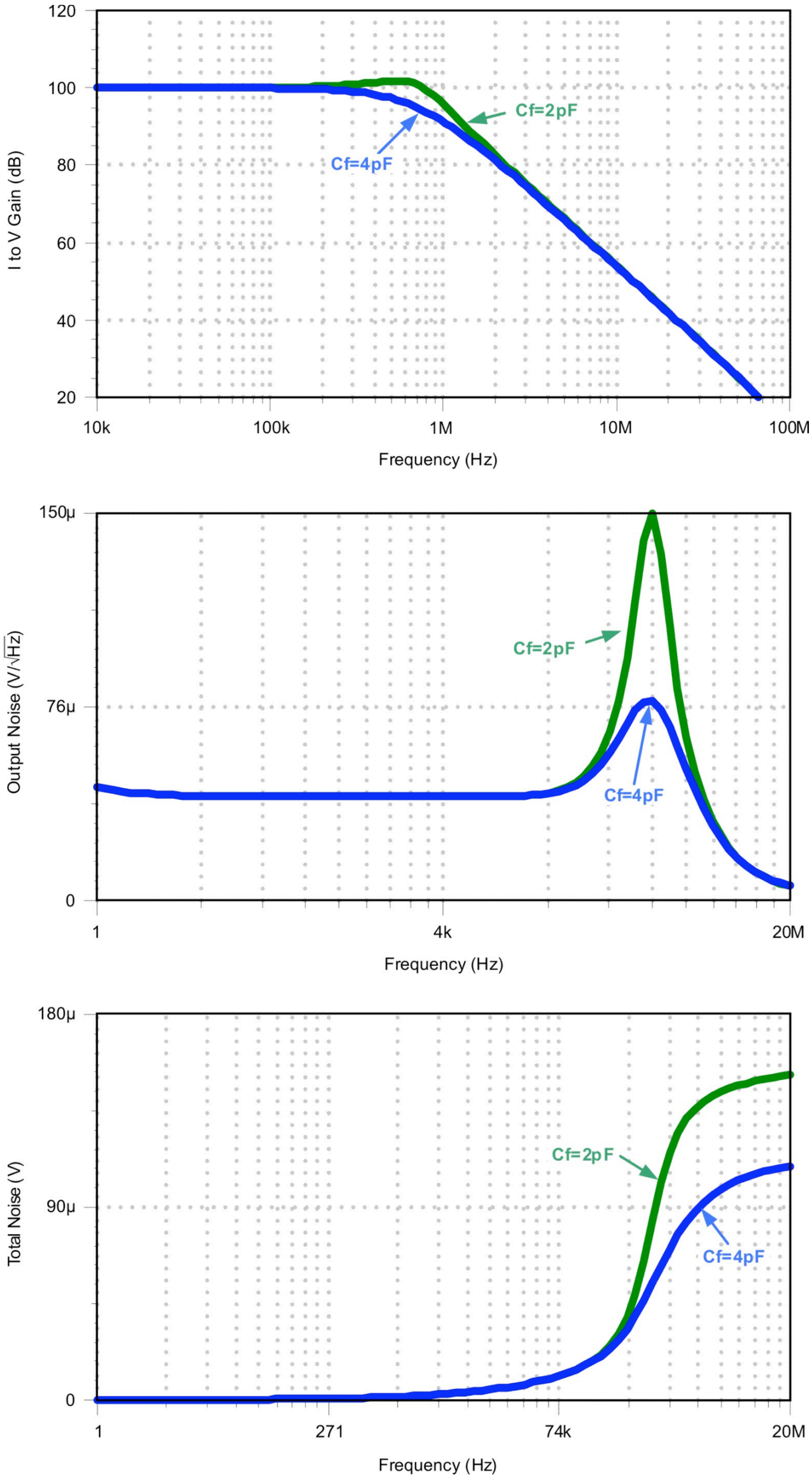


Fig. 12.12: Effect Of Changing Feedback Capacitance On Noise

Measuring The Noise For The Example Transimpedance Amplifier

Fig. 12.13 shows the simplest approach to measuring noise. In this case the amplifier output is directly connected to the measurement instrument. To determine if this approach can be used to successfully measure noise, you need to compare the expected noise level to the measurement noise floor.

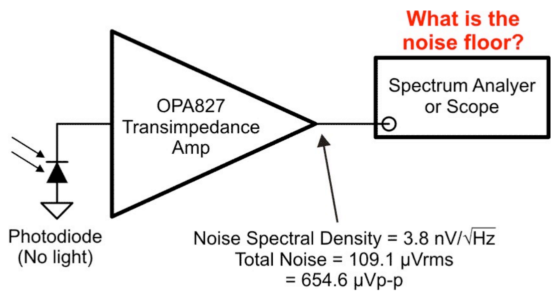


Fig. 12.13: Check Equipment Capability For Measuring Example Circuit Noise

Fig. 12.14 gives the specifications for two common examples of equipment used for noise measurements. The table also lists the noise signals from the transimpedance amplifier for comparison to the noise floor. The noise floor is close to the amplitude of the noise signal. Ideally the noise floor would be significantly smaller then the noise signal. Without a post-amplifier the accuracy of the noise measurement is marginal.

	Scope	Spectrum Analyzer	Noise Signal
Noise floor	48 µV rms	10 nV/√Hz	109 µV rms 3.8 nV/√Hz
Range	1 mV/div		
Input impedance	1 MΩ	50 Ω	

Fig. 12.14: Noise Floor For Common Test Equipment

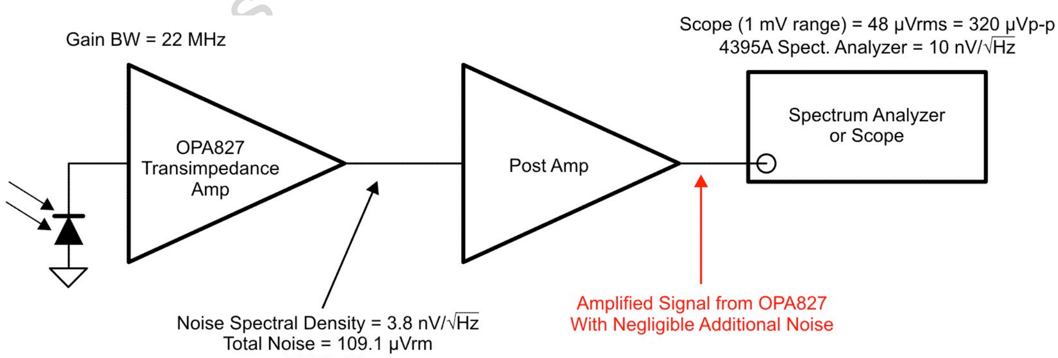


Fig. 12.15: Low-Noise Post-Amplifier Increases Noise To Measurable Level

Fig. 12.15 shows how a post-amplifier can be used to improve the noise floor of the measurement circuit. The post-amplifier increases the amplitude of the noise signal from the transimpedance amplifier so that the equipment is capable of measuring the noise. In order for this technique to be effective, the post-amplifier must have an input noise that is significantly smaller then the output noise of the transimpedance amplifier. It is also important to make sure that the bandwidth of the post-amplifier is wider then the amplifier under test. In cases where the amplifier under test is low noise, it can be challenging to select the post-amplifier. For this example circuit we will investigate using the OPA847 as a post-amplifier.

Fig. 12.16 shows the affect of the gain on the noise signal from the transimpedance amplifier. The point is to ensure that the gain sufficiently increases the amplitude of the noise signal so that it can be easily measured by the test equipment. In Part I, we learned that noise signals three times greater then the noise floor dominate. In this example the noise signal is amplified so it is substantially larger than the noise floor to further minimize error. Note that most equipment has poor resolution near the noise floor, so operating significantly above the noise floor improves accuracy.

	Scope	Spectrum Analyzer	Noise Signal	Noise Signal x 150
Noise floor	48 µV rms		109 µV rms	16.35 mV rms
		10 nV/√Hz	3.8 nV/√Hz	570 nV/√Hz
Range	1 mV/div			
Input impedance	1 MΩ	50 Ω		

Figure 12.16: Amplify Noise Signal To Improve Noise Measurement Accuracy

Fig. 12.17 shows the bandwidth calculation for the post-amplifier. It is important to note that the post-amplifier bandwidth must be greater then the transimpedance amplifier bandwidth for accurate noise measurements. On the other hand, using a bandwidth significantly wider than the amplifier under test adds unnecessary noise. In our example the post-amplifier bandwidth is relatively close to the transimpedance amplifier bandwidth (ie post-amplifier 26 MHz, and transimpedance amplifier 22 MHz).

EQUATION	DESCRIPTION
$Gain\_Bandwidth\_Product = 3900\text{ MHz}$	For OPA847 gain > 50
$Post\_Amp\_Bandwidth = \frac{Gain\_Bandwidth\_Product}{Gain} = \frac{3900\text{ MHz}}{150} = 26\text{ MHz}$	
$Transimpedance\_BW = 22\text{ MHz}$	Unity gain bandwidth of the transimpedance amplifier is lower than the post

Fig. 12.17: Noise floor For Common Test Equipment

The OPA827 is a low-noise, wide-bandwidth device. Fig. 12.18 shows the proposed post-amplifier. The gain of the amplifier is set to 150. Note that the parallel combination of the feedback network was intentionally selected as a low resistance so that it does not contribute significant thermal noise. The 50 Ω output resistor is for impedance matching to 50 Ω input impedance equipment. The output capacitor allows for ac coupling to the test equipment.

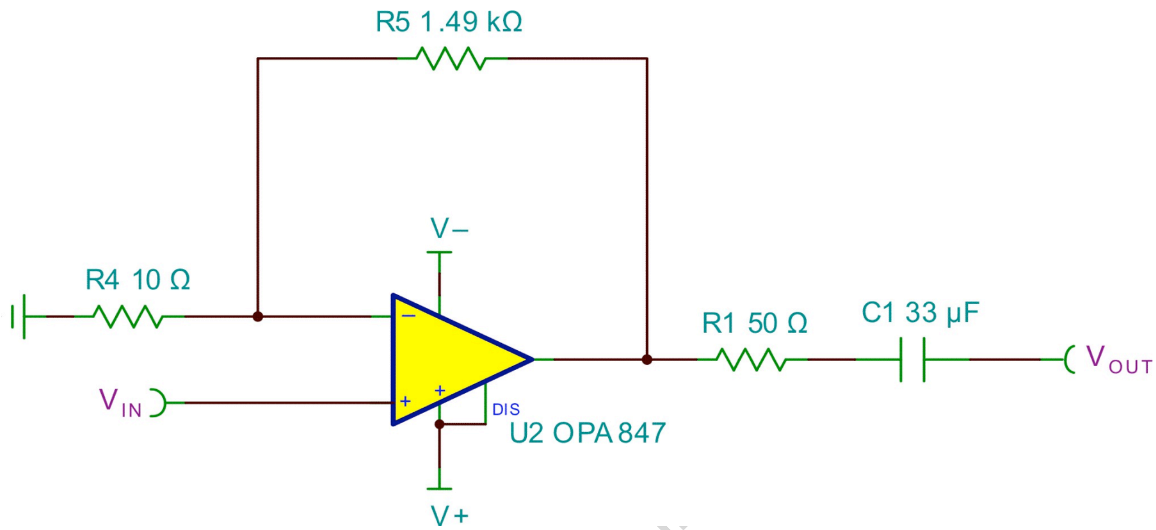


Fig. 12.18: Low-Noise Wide Bandwidth Post-Amplifier

Fig. 12.19 shows the total noise simulation results for the post-amplifier. The data from this simulation is used in Fig. 12.20 (overleaf) to show that the post-amplifier noise is small when compared to the transimpedance amplifier.

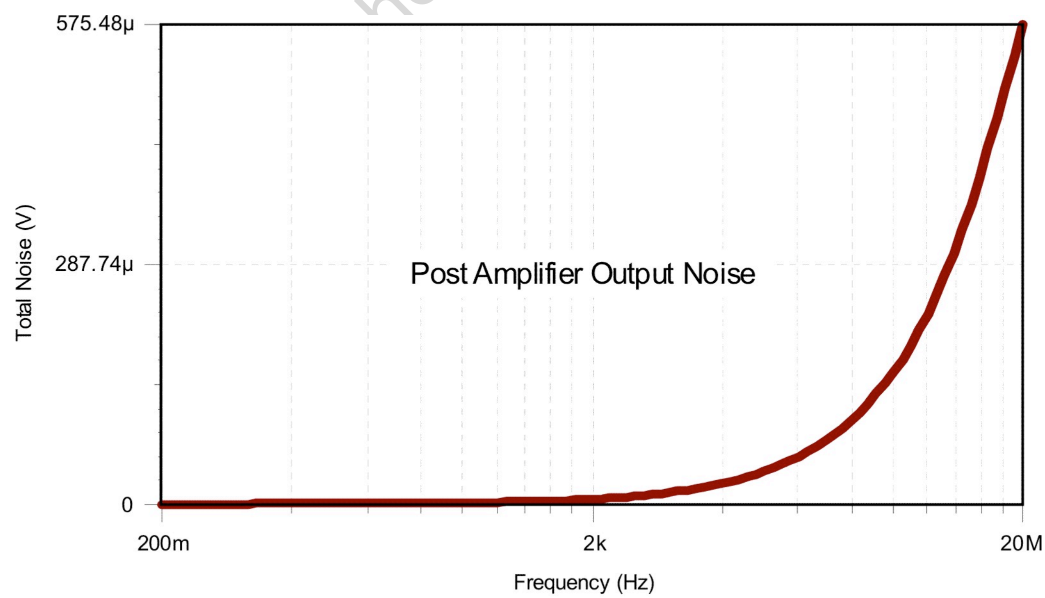


Fig. 12.19: Post-Amplifier Simulated Noise

The noise output from the transimpedance amplifier adds to the input noise of the post-amplifier. The table in Fig. 12.20 illustrates that the noise referred to the input of the post-amplifier is small, compared to the noise at the output of the transimpedance amplifier. Thus the post-amplifier increases the noise amplitude from the transimpedance amplifier (150x), but does not add significant additional noise.

Transimpedance amp output noise from simulation (Fig. 12.11)	**Post-amp input noise (RTI) RTO / 150**	**Post-amp noise RTO (150 x RTI) from simulation** (Fig. 12.19)
109 μV rms	3.83 μV rms	575.5 μV rms
3.8 nV/√Hz	0.85 nV/√Hz	127.5 nV rms

Fig. 12.20: Comparison Post-Amp Input Noise With Transimpedance Output Noise

Fig. 12.21 shows the hardware setup used for the transimpedance amplifier and the post-amplifier. The printed-circuit board on the left side is the photodiode amplifier. Note that the photodiode is covered with black tape to ensure that light does not affect the noise measurements.

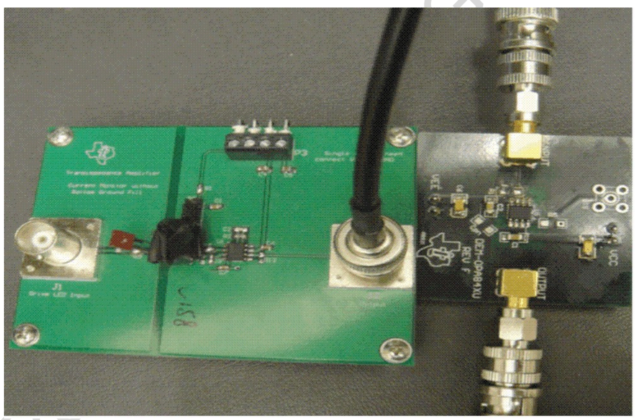


Fig. 12.21: OPA847EVM (Right) Used For Post-Amplifier

Fig. 12.22 shows the TINA SPICE schematic for the photodiode amplifier and the post-amplifier. Simulating this circuit allows us to confirm that the post-amplifier increases the amplitude of the transimpedance amplifier noise, but does not contribute significant additional noise.

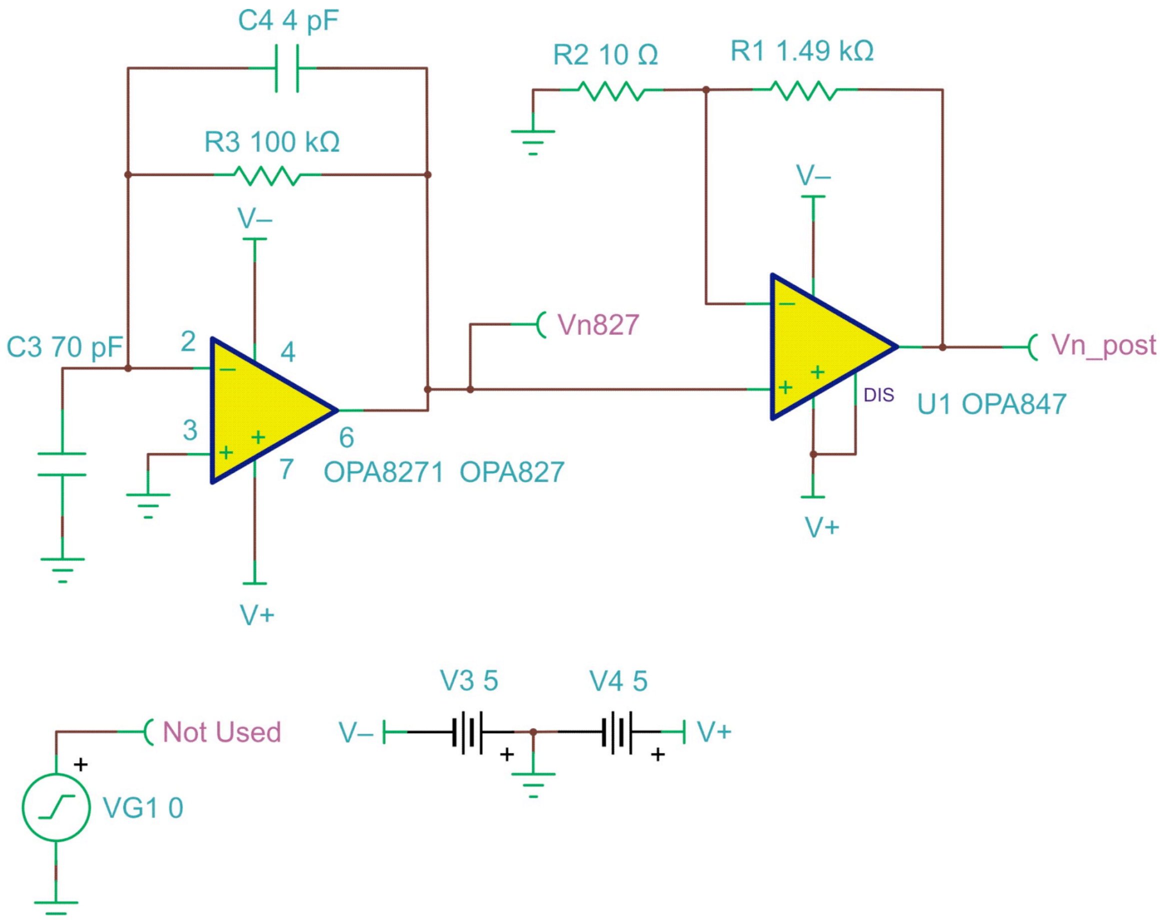


Fig. 12.22: Transimpedance Amplifier And Post-Amplifier

Fig. 12.23 shows the simulation results of the circuit in Fig. 12.22. The important point is that the post-amplifier amplifies the noise by a factor of 150, but does not add significant noise.

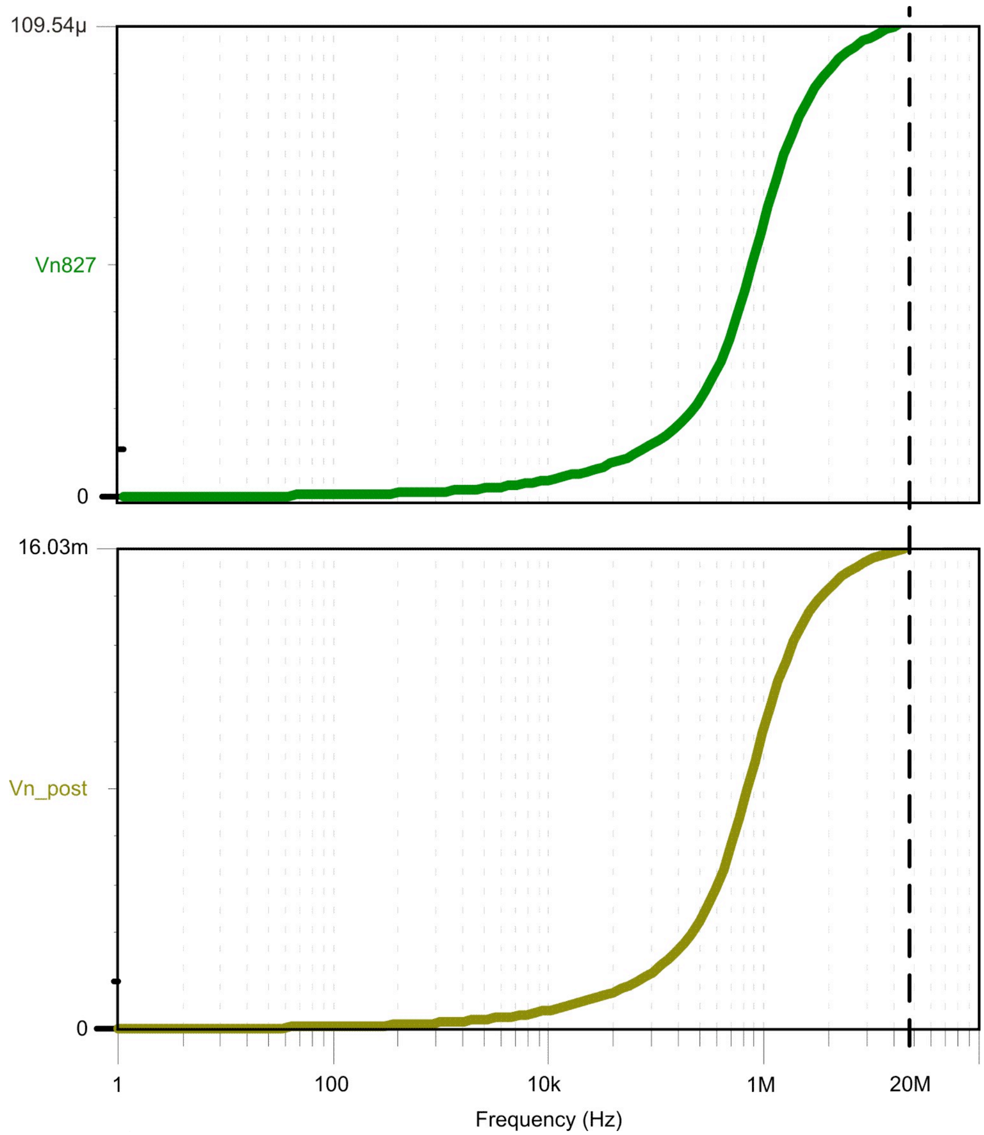


Fig. 12.23: Simulation Result For Transimpedance Amp With Post-Amp

Vn OPA827 transimpedance amp (Fig. 12.23 Top)	Vn OPA847 post-amp (Fig. 12.23 bottom)	Vn post/gain 16 mV/150
109.4 µV rms	16 mV rms	106.8 µV rms

Fig. 12.24: Simulation Results Show Post-Amp Noise Is Insignificant

Fig. 12.25 shows the connections to the spectrum analyzer and oscilloscope. Note that the spectrum analyzer has a 50 Ω input impedance. This is common on high-frequency test equipment. Note that the 50 Ω impedance creates a voltage divider with the output impedance of the post-amplifier.

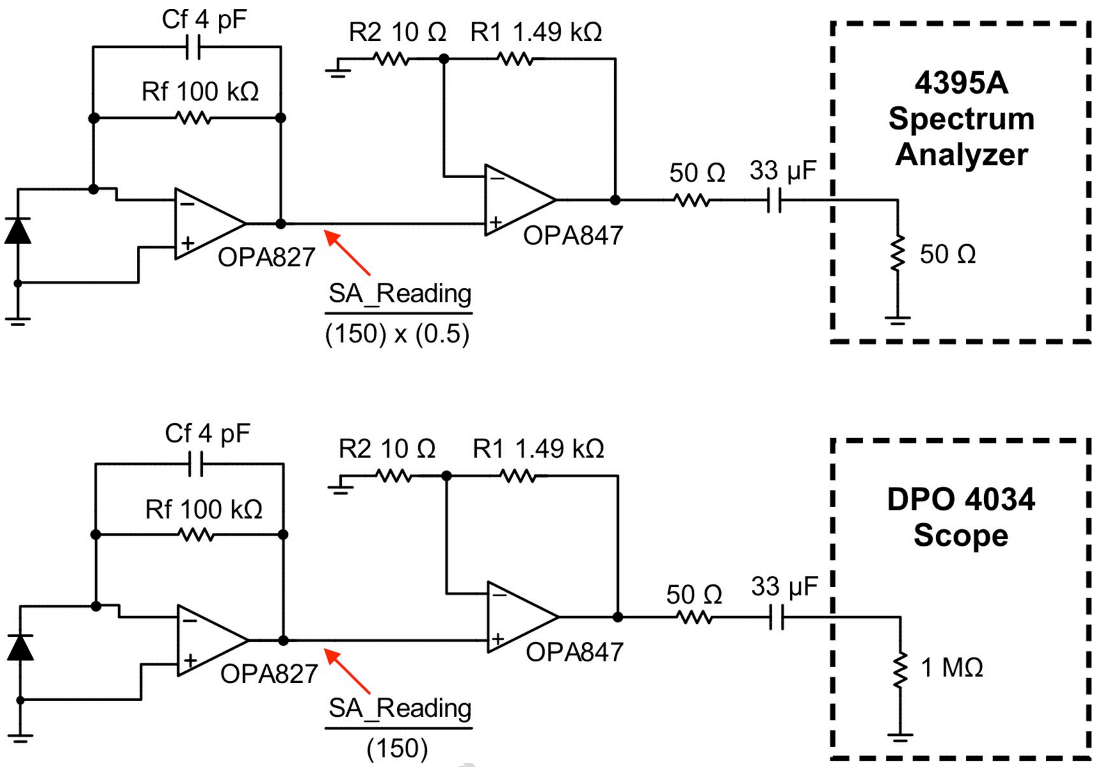


Fig. 12.25: Connections To Test Equipment

Fig. 12.26 shows the oscilloscope noise measurement for the photodiode and post-amplifier circuit.

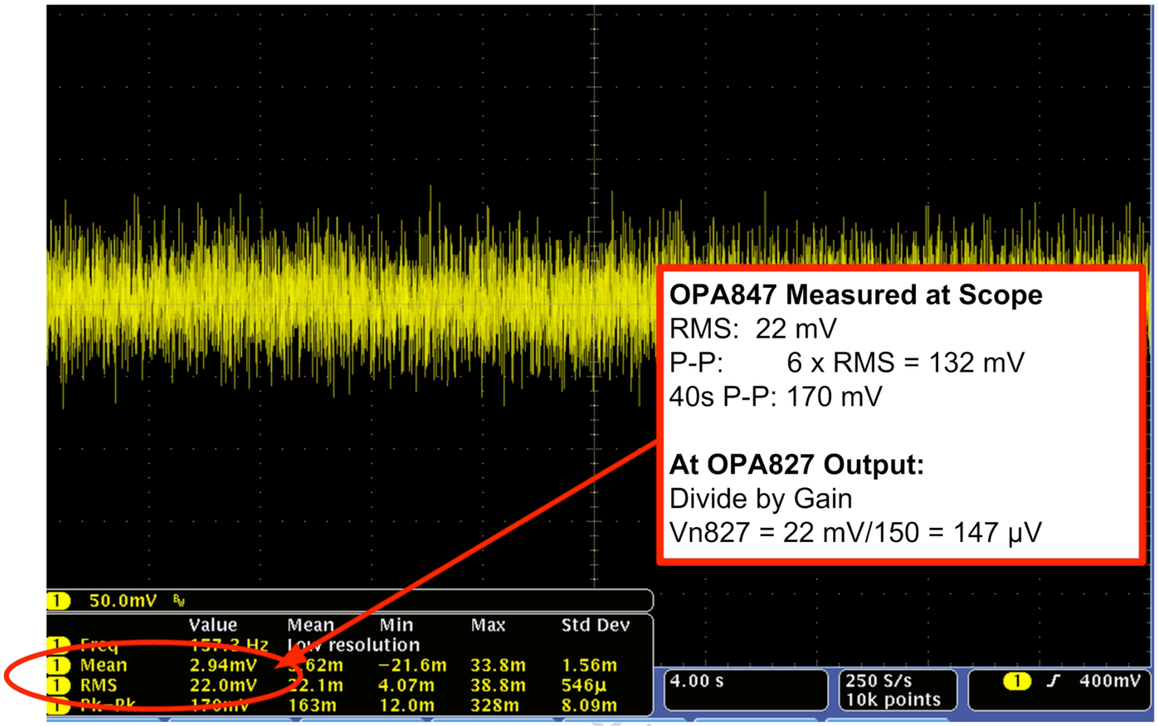


Fig. 12.26: Oscilloscope Measurement Of Photodiode Noise Circuit

Fig. 12.27 summarizes the results for the calculation, simulation, and measurement of this circuit. In this case the measured results are about 30% larger than the calculated and simulated results. This difference could be the result of component tolerances, or parasitic capacitance.

Calculated (rms)	Simulated (rms)	Measured (rms)
116.7 µV	109.1 µV	147 µV

Fig. 12.27: Summary Of Calculated, Simulated, And Measured Results

Fig. 28 shows a comparison of the spectrum analyzer measurement and the simulated output voltage spectral density (displayed with a linear frequency). The TINA SPICE spectral density results are shown with both linear and logarithmic axes.

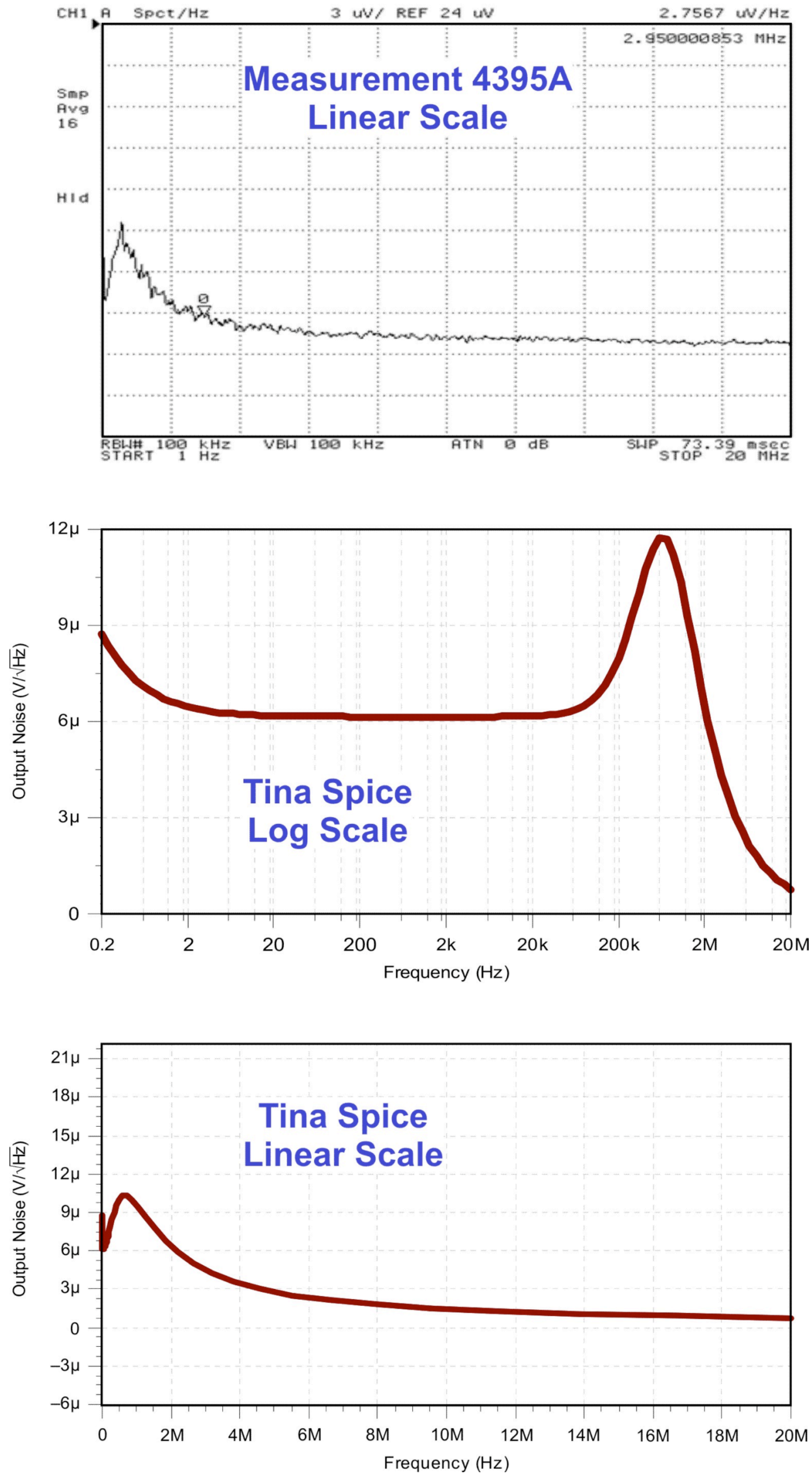


Fig. 12.28: Spectral Density Measurement Compared To Simulation Results

Fig. 12.29 shows the measured vs. simulated spectral density with key points on the curve identified. The measured spectral density curve compares well with the simulations. Any difference is likely the result of stray capacitance, and component tolerance.

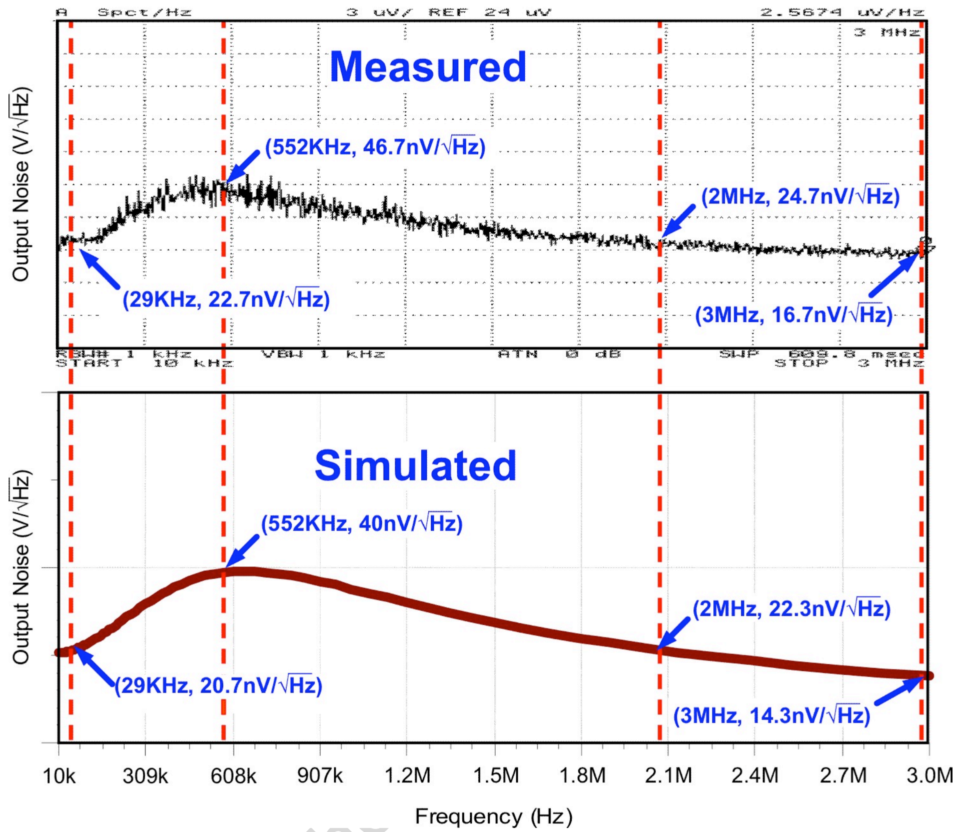


Fig. 12.29: Measured Results Vs Simulations

Summary

This TechNote uses the theory developed in Part XI to analyze an example circuit. Both hand analysis and SPICE analysis are used to predict the total output noise. The TechNote also covers methods for measuring total noise including the use of a post-amplifier to expand the capability of the test equipment.

References

- Jerald Graeme, Photodiode Amplifiers: OP AMP Solutions, McGraw-Hill Professional
- Tim Green, Operational Amplifier Stability, http://www.en-genius.net
- http://www.hamamatsu.com, Photodiode Technical Information

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