Analysis and Measurement of Intrinsic Noise in Op Amp Circuits
Part IX: 1/f Noise and Zero-Drift Amplifiers

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This TechNote focuses on errors in low-frequency applications. Input offset voltage drift and 1/f noise are studied in detail. Standard topologies are compared and contrasted with the zero-drift amplifier topology, which has low offset drift and no 1/f noise.

Zero-Drift Amplifiers

Zero-drift amplifiers are operational amplifiers (op amps) that periodically self-calibrate offset, offset drift and low-frequency noise errors. The calibration frequency for different zero-drift amplifiers ranges from 10 kHz to 100 kHz (periods from 100 μs to 10 μs). A digital circuit inside the zero-drift amplifier controls the calibration; however, the amplifier acts as a normal linear op amp from the board and system level view. There are two different types of zero-drift amplifiers: auto-zero and chopper-stabilized.

The auto-zero amplifier consists of a continuous linear amplifier and a nulling amplifier. During calibration the nulling amplifier samples the offset voltage and stores it on a sampling capacitor. The linear amplifier's offset is canceled by the offset stored on the sampling capacitor. Since the calibration is done at a relatively high rate (eg 50 kHz), the offset drift and low-frequency noise are also canceled. Fig. 9.1 shows a simplified block diagram of an auto-zero circuit. Although this is an oversimplification of the correction, it is adequate from a functional board and system level designer view. See Reference [1] for a more detailed description of how this technique works.

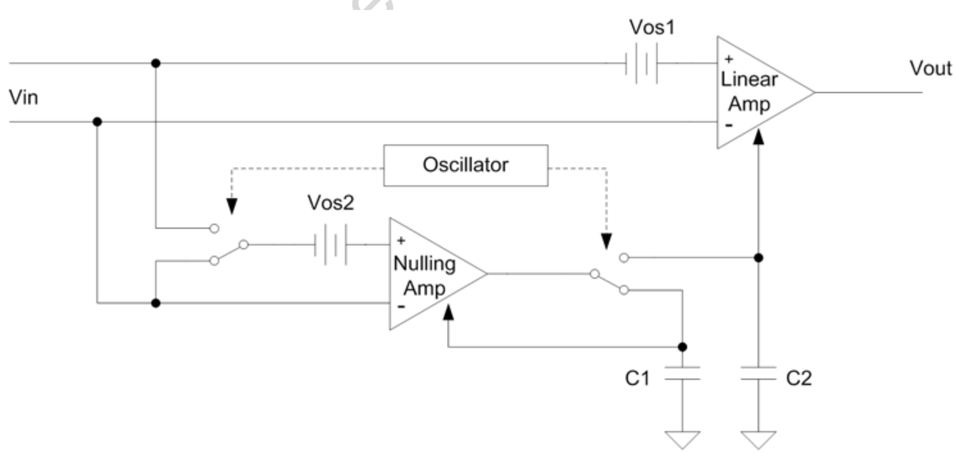


Fig. 9.1: Simplified Auto-Zero Circuit

In the chopper-stabilized circuit the input and output are synchronously inverted. Thus, the offset is inverted every other chopping cycle. This converts the offset from a constant dc value to an ac signal with an average of zero. Filtering reduces the amplitude of the ac signal created by the offset chopping (see Fig. 9.2). Modern TI chopper amplifiers use a patented switched-capacitor notch filter to eliminate the chopper signal. A more detailed description of how the chopping technique works is given in Reference [2].

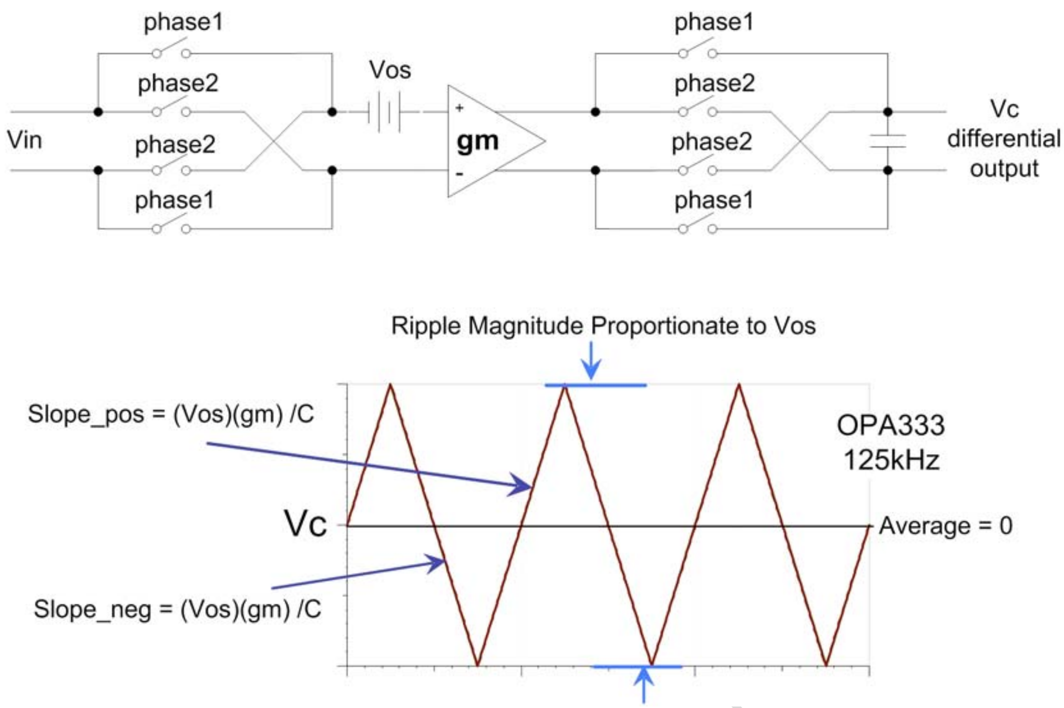


Fig. 9.2: Chopper-Stabilized Amplifier And Ac Signal Created

Both the auto-zero and chopper-stabilized amplifiers share some common characteristics and are categorized as zero-drift amplifiers. The key characteristics are low-voltage offset and low-voltage-offset drift. Although bias current and bias current drift are not calibrated during the self-calibration, they are typically low because the amplifiers are MOSFET amplifiers. Fig. 9.3 lists offset and offset drift for some common zero-drift amplifiers.

Op Amp	Offset (µV)	Offset Drift (µV/C)	Voltage Noise (nV/rtHz)
OPA333	10	0.05	55
OPA335	5	0.05	55
OPA378	50	0.25	20

Fig. 9.3: Offset And Offset Drift For Common Op Amps

Another important characteristic of zero-drift amplifiers is that they have practically no 1/f noise. Low-frequency noise can be thought of as variations in offset voltage with time. The self-calibration eliminates low-frequency noise in the same way it eliminates offset drift.

Fig. 9.4 shows a noise waveform sampled and corrected over time.

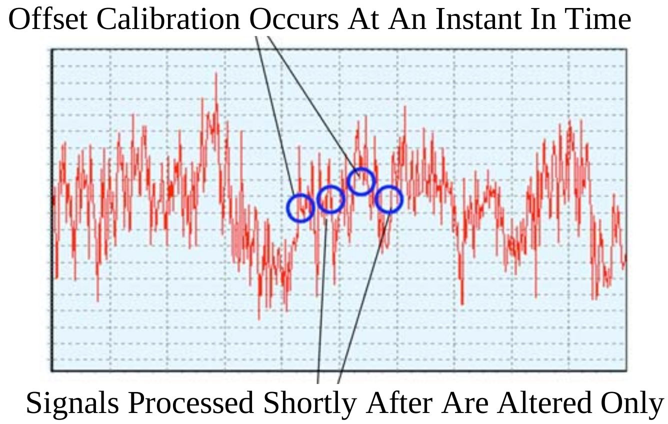


Fig. 9.4: Low-Frequency Noise Eliminated By Zero-Drift Amplifier

Zero-Drift Amplifier Spectral Density Curve

The spectral density curve for zero-drift amplifiers does not have a 1/f region. In some cases the calibration process creates signals at the calibration frequency and harmonics. These signals show up as spikes on the spectral density curve. Fig. 9.5 shows the spectral density curve for a typical auto-zero amplifier.

For most applications it is a good idea to avoid the region with the calibration signal feedthrough. This can be done with an external filter. In many cases the op amp gain bandwidth (GBW) automatically attenuates the calibration feedthrough. In Fig. 9.6 the 3-dB bandwidth is limited to 2 kHz by the GBW limitation of the amplifier; ie GBW ÷ Gain = 2 MHz ÷ 1001 = 2 kHz. In this configuration the calibration signal is significantly attenuated. Fig. 9.7 shows the noise calculation for the gain of 1001. Note that in this configuration the calculated total noise closely matches the measured total noise. For this configuration the noise is flat throughout the entire frequency of operation: ie 0 Hz to 2 kHz. The mathematics used in this calculation were covered in Part III of this series.

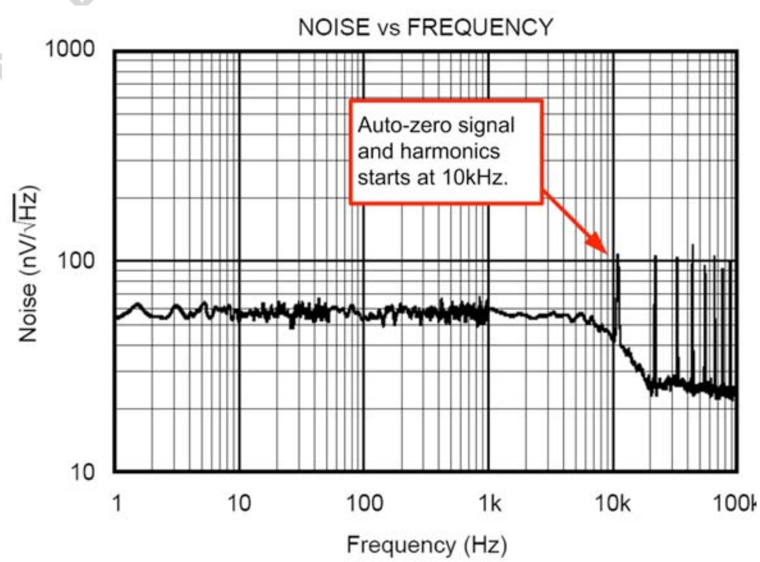


Fig. 9.5: Spectral density Curve For OPA335 (Typical Auto-Zero Amplifier)

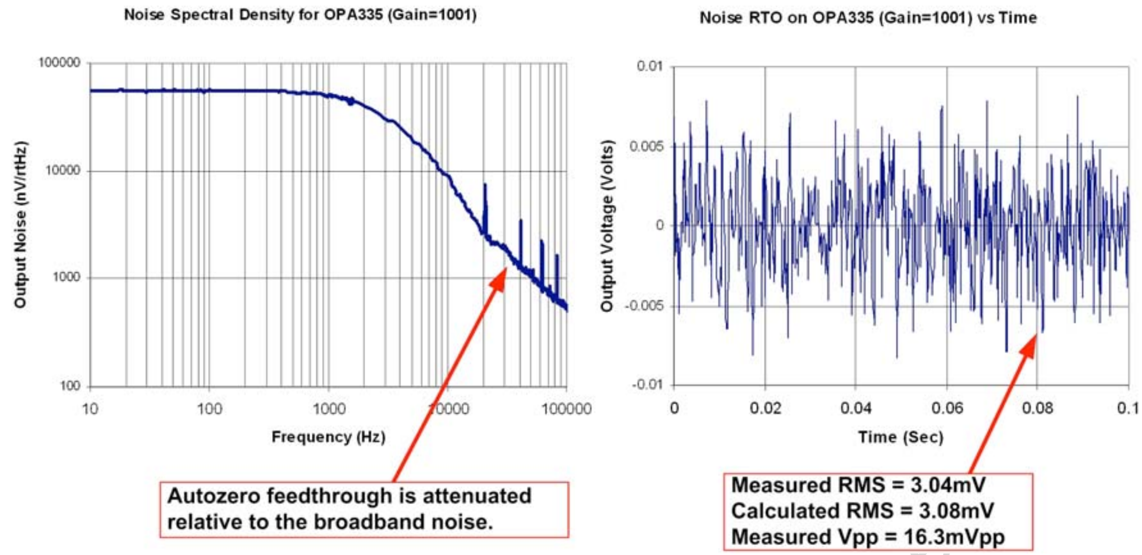


Fig. 9.6: Measured Output Noise OPA335, Gain Of 1001

$$f_H = \frac{Gain\_Bandwidth}{Gain} = \frac{2 \cdot MHz}{1001} = 1.998kHz$$

$$BW_n = f_H \cdot K_n = 1.998kHz \cdot 1.57 = 3.137kHz$$

$$V_{n\_out} = e_{BB} \cdot \sqrt{BW_N} \cdot Noise\_Gain = 55 \cdot nV \cdot \sqrt{3.137 \cdot kHz} \cdot (1001) = 3.08mV$$

Fig. 9.7: Total Noise Calculation For OPA335, Gain Of 1001

Fig. 9.8 shows the noise in gain of 101. The 3-dB bandwidth is limited to 19.8 kHz by the GBW limitation of the amplifier; ie GBW ÷ Gain = 2 MHz ÷ 101 = 19.8 kHz. Since the bandwidth is 19.8 kHz, the calibration signal at 10 kHz is not attenuated. Also note that the calculation for total noise is complicated by the fact that the spectral density curve drops from 55 nV/rtHz to 2 5 nV/rtHz at 10 kHz. The calibration signals are included in the noise signal, but cannot be easily accounted for in the calculation.

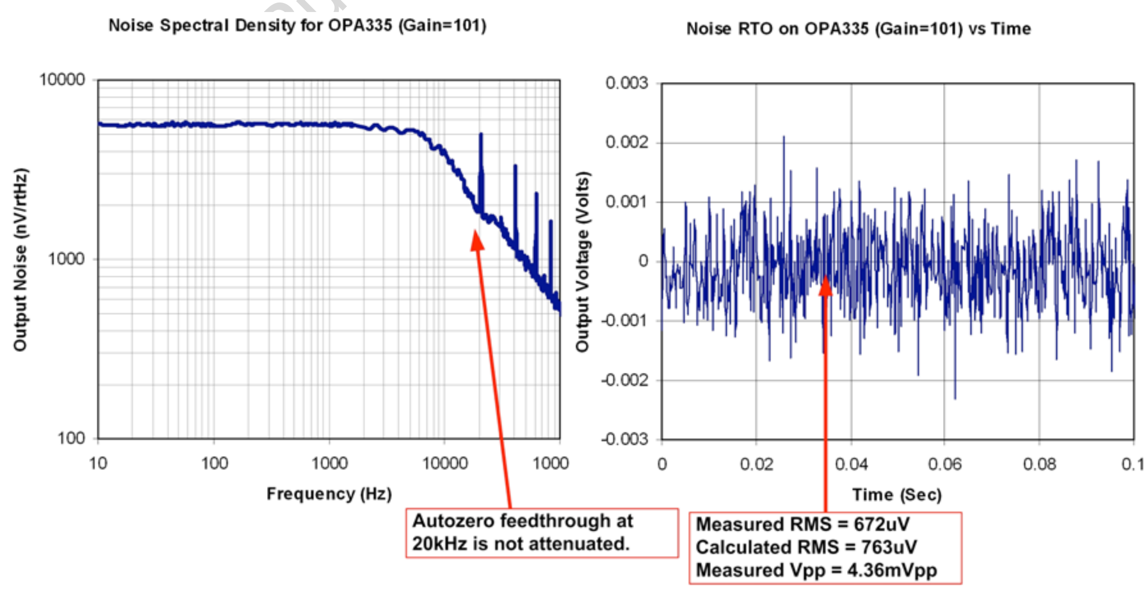
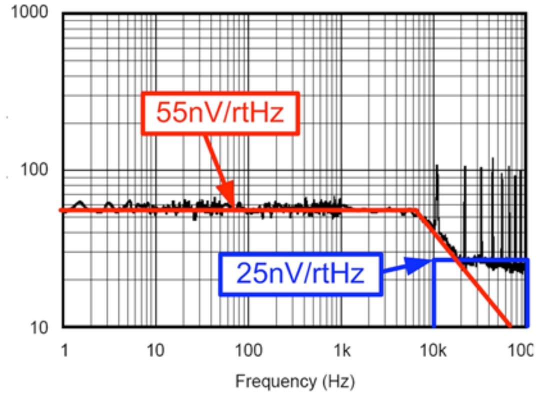


Fig. 9.8: Measured Output Noise OPA335, Gain Of 101

The total noise is calculated in Fig. 9.9.



$$BW_{n1} = f_H \cdot K_n = 10kHz \cdot 1.57 = 15.7kHz$$

$$V_{n\_out1} = e_{BB} \cdot \sqrt{BW_{n1}} \cdot Noise\_Gain = 55 \cdot nV \cdot \sqrt{15.7kHz} \cdot (101) = 696 \mu V$$

$$f_H = \frac{Gain\_Bandwidth}{Gain} = \frac{2 \cdot MHz}{101} = 19.82kHz$$

$$BW_{n2} = f_H \cdot K_n = (19.82kHz - 10kHz) \cdot 1.57 = 15.4kHz$$

$$V_{n\_out2} = e_{BB} \cdot \sqrt{BW_{n2}} \cdot Noise\_Gain = 25 \cdot nV \cdot \sqrt{(15.3kHz) \cdot 1.57} \cdot (101) = 313 \mu V$$

$$V_{n\_total} = \sqrt{V_{n1}^2 + V_{n2}^2} = \sqrt{(696 \mu V)^2 + (313 \mu V)^2} = 763 \mu V$$

Fig. 9.9: Total Noise Calculation For OPA335, Gain Of 101

Fig. 9.10 shows the noise in gain of 11. In this case the 3-dB bandwidth is limited to 182 kHz by the gain bandwidth limitation of the amplifier so the calibration is not attenuated.

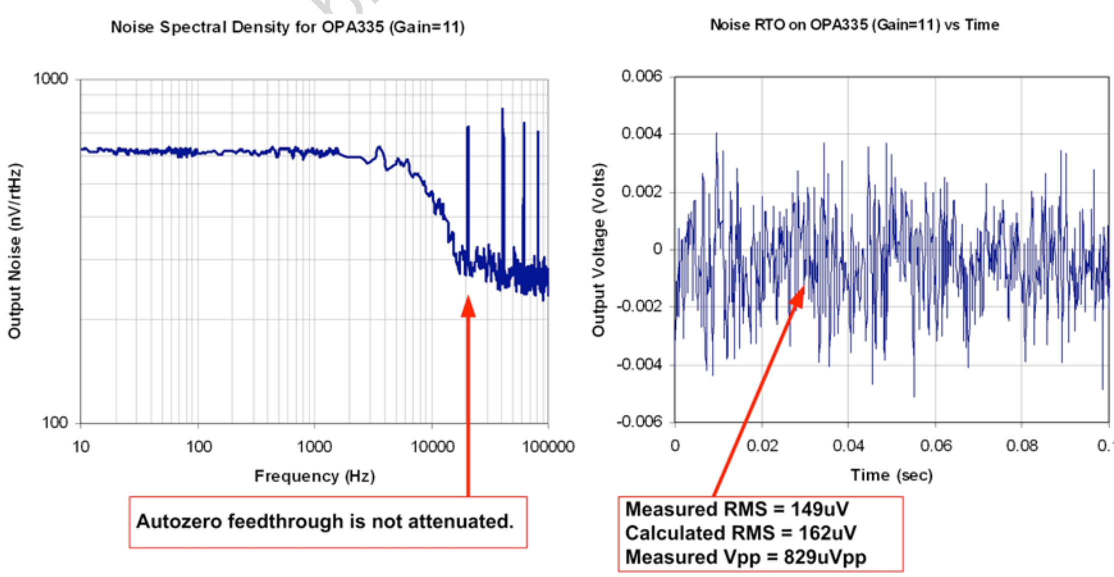


Fig. 9.10: Measured Output Noise OPA335, Gain Of 11

Fig. 9.11 compares the three gains considered in this TechNote.

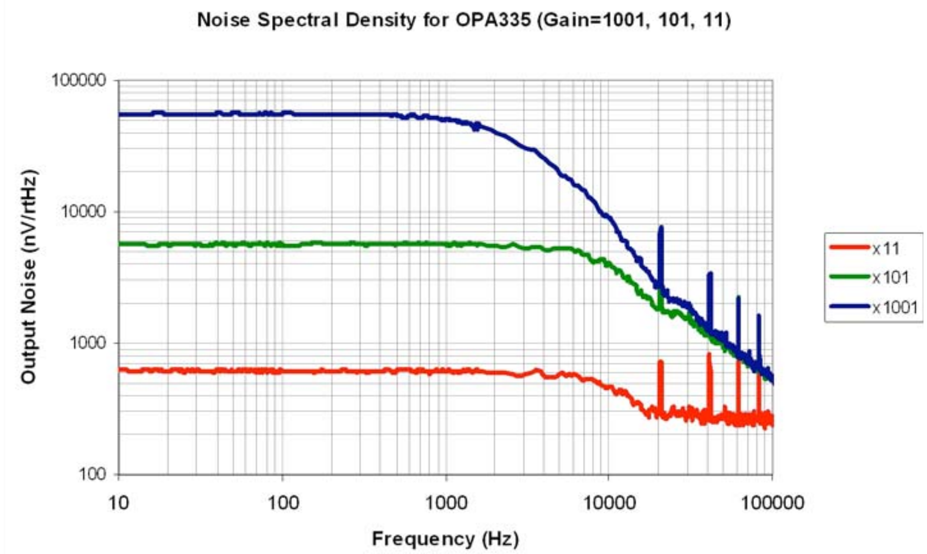


Fig. 9.11: Comparing Output Spectral Density For Different Gains

Modern Texas Instruments chopper amplifiers use filtering techniques that minimize the calibration signal feedthrough. Another trick used in modern op amp designs is to move the calibration frequency higher to increase the usable bandwidth. Fig. 9.12 shows the spectral density curve for the OPA333 chopper-stabilized amplifier where the chopper frequency is approximately 125 kHz. The GBW for the OPA333 is 350 kHz. For most gain settings the chopper frequency is outside of the bandwidth of the amplifier. For example, the bandwidth would be limited to 35 kHz for a gain of 10, and consequently the chopper signal would be substantially attenuated.

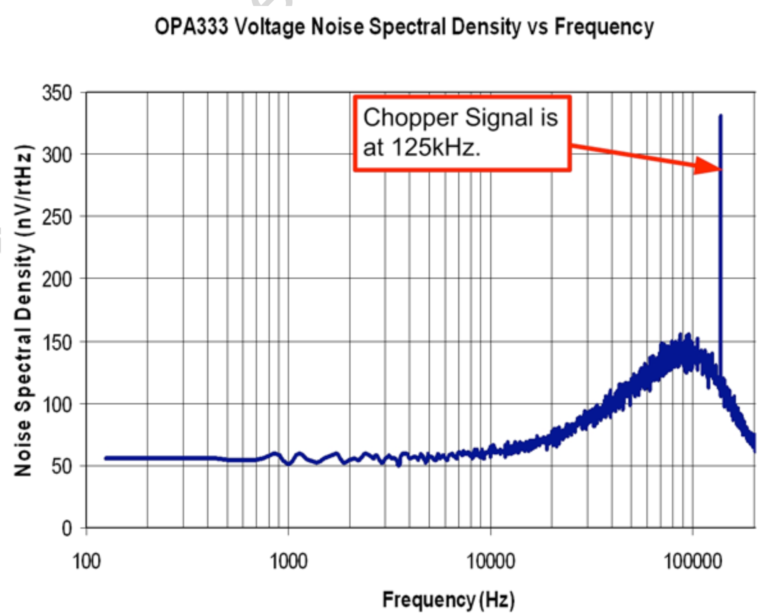


Fig. 9.12: Spectral Density Curve For OPA333: Chopper With Internal Notch Filter

Low-Frequency Noise

The spectral density curve for 1/f noise increases as frequency is decreased. In fact, noise will increase towards infinity as we approach zero frequency. This fact often leads designers to believe that the total noise should be infinite at dc, because dc is considered to be zero frequency. The best way to understand why 1/f noise does not translate to infinite noise for practical circuits is to convert frequency to time (Time = 1/Frequency).

Fig. 9.13 shows flicker noise (1/f noise) calculations for the OPA336 at different low cut frequencies. The lower cut frequency is set by the time period that the signal is observed over. Typically noise calculations use 0.1 Hz as a lower cut frequency. This corresponds to an observation period of 10 s. The same calculation can be made for any time period. Note that a frequency of 0 Hz corresponds to infinite time and, consequently, is not practically achievable. Note that extremely low frequencies correspond to years of time. Fig. 9.14 shows how noise increases for longer observation periods and is the graphical representation of the table in Fig. 9.13.

f_H	f_L	$1/f_L$ Sec.	$1/f_L$ Days	$1/f_L$ Years	Noise Calculation	Noise
10	1	1	1.1×10^{-5}	3.1×10^{-8}	$200nV\cdot\sqrt{\ln\left(\frac{10Hz}{1Hz}\right)}$	303nV
10	0.1	10	1.1×10^{-4}	3.1×10^{-7}	$200nV\cdot\sqrt{\ln\left(\frac{10Hz}{0.1Hz}\right)}$	429nV
10	1×10^{-6}	1×10^{6}	11	0.032	$200nV\cdot\sqrt{\ln\left(\frac{10Hz}{1\mu Hz}\right)}$	808nV
10	1×10^{-9}	1×10^{9}	1.1×10^{4}	32	$200nV\cdot\sqrt{\ln\left(\frac{10Hz}{1nHz}\right)}$	960nV

Fig. 9.13: Flicker Noise Calculations For OPA336

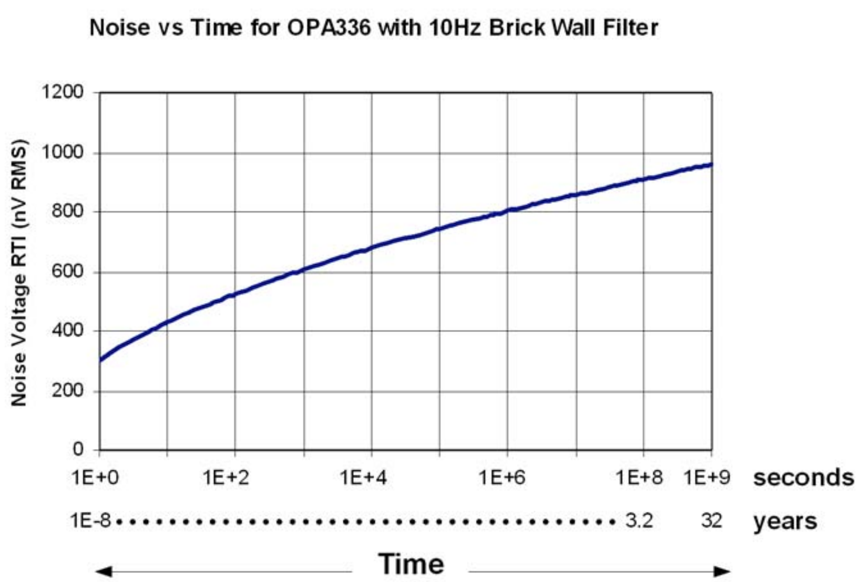


Fig. 9.14: Noise For OPA336 Vs Time (F_H = 10 Hz)

The upper cut frequency used in these calculations is 10 Hz. This is the typical upper cut frequency used in flicker noise measurements (0.1 Hz < f < 10 Hz). The noise calculation computes the noise from f_L to f_H, assuming a *brick wall filter* at f_L and f_H. This is the way 0.1 Hz to 10 Hz noise is normally specified. The term brick wall filter means that the noise drops abruptly to zero outside the specified bandwidth. If a real-world filter is used, the noise decreases gradually with the filter (eg 20 dB/decade for a first-order filter). This topic was covered in detail in Part II of this series.

The total noise from flicker (1/f noise) is equal over each decade change in frequency. For example, the total noise in the interval (0.1 Hz, 10 Hz) is the same as in the interval (0.01 Hz, 0.1 Hz); this is shown mathematically in Fig. 9.15 using the formula developed in Part I. This fact is often confusing to engineers because the area appears to be significantly larger in regions with higher flicker noise. However, keep in mind that spectral density curves are usually shown with logarithmic axes. When you look at the area of two different decade-wide intervals on logarithmic axes, they do not look equivalent. If you change to a linear axis, you see that as 1/f noise gets larger the width of the interval gets smaller. Fig. 9.16 shows the power spectral density curve on linear axes to illustrate the equivalent area of two-decade wide intervals.

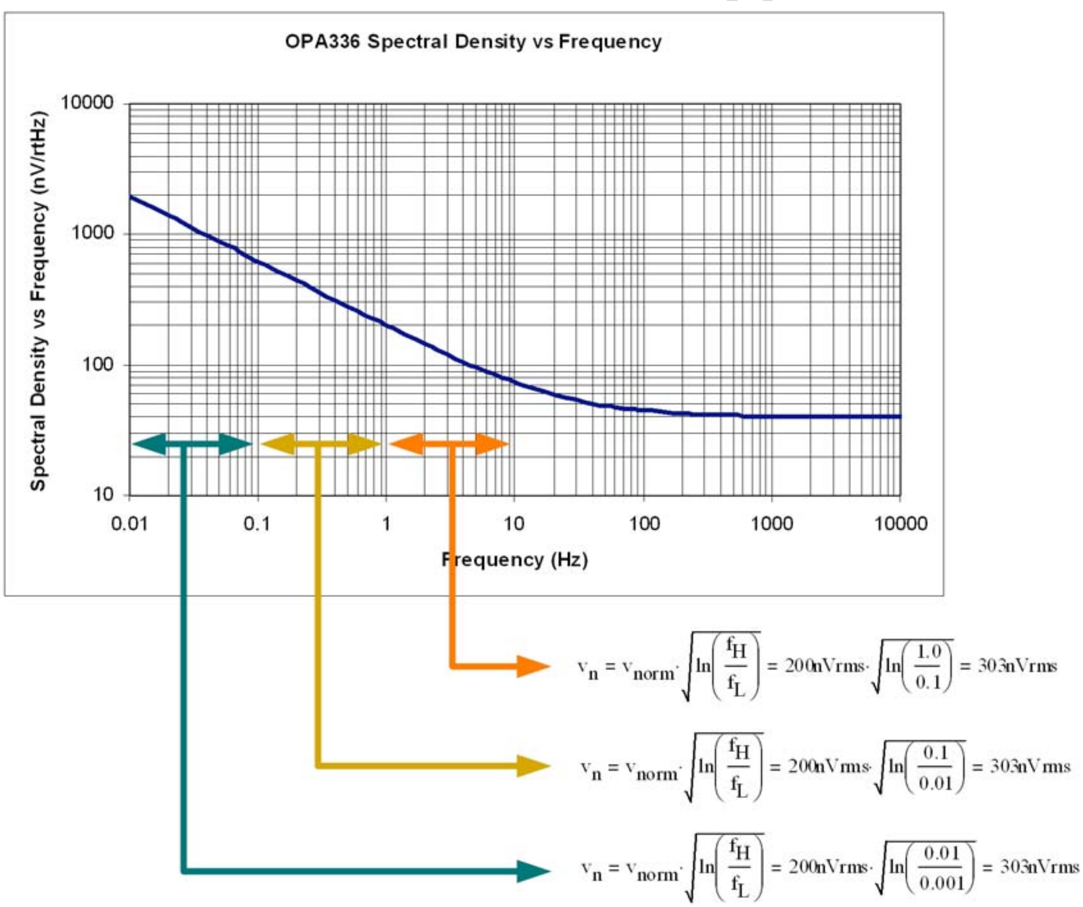


Fig. 9.15: Total Noise Is Equal In Each Decade Interval

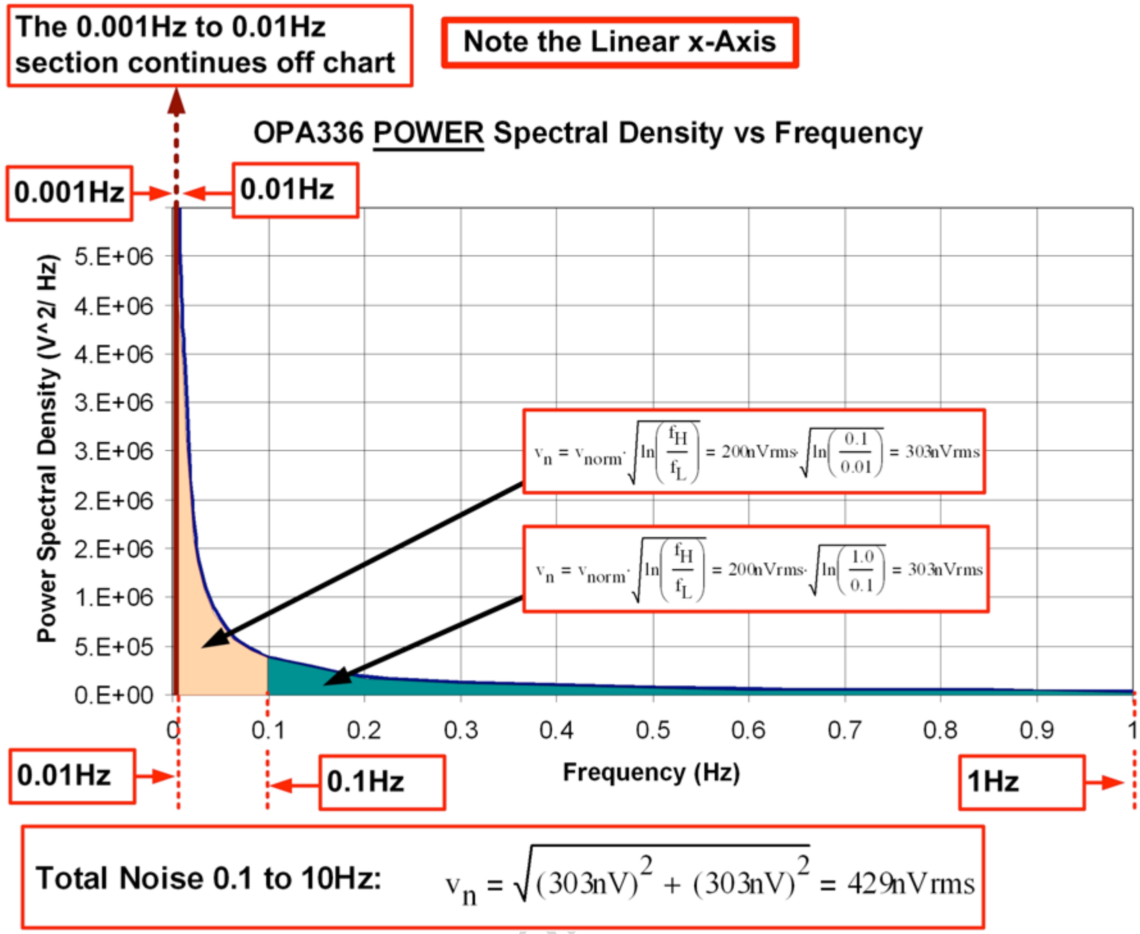


Fig. 9.16: Linear Axes Illustrate Equivalent Area Of Decade-Wide Intervals

The total noise for amplifiers with 1/f noise increases for longer observation time periods. The waveform shown in Fig. 9.17 (overleaf) shows the noise on the OPA336 over a 100,000 s interval (10 µHz). The upper cut frequency for this signal is 10 Hz. Thus, the noise bandwidth is 10 µHz to 10 Hz. The total rms noise over the entire interval is 0.74 µV. If you choose any sub-interval of time, the total rms noise will be smaller. In this example, a 10 s sub-interval is shown to have a total noise of 0.43 µV rms. The sub-interval in this example was taken from the first 10 s, but any 10-s interval will have the same total rms noise. Remember that a smaller time period corresponds to a larger lower-cut frequency, and less area under the 1/f curve.

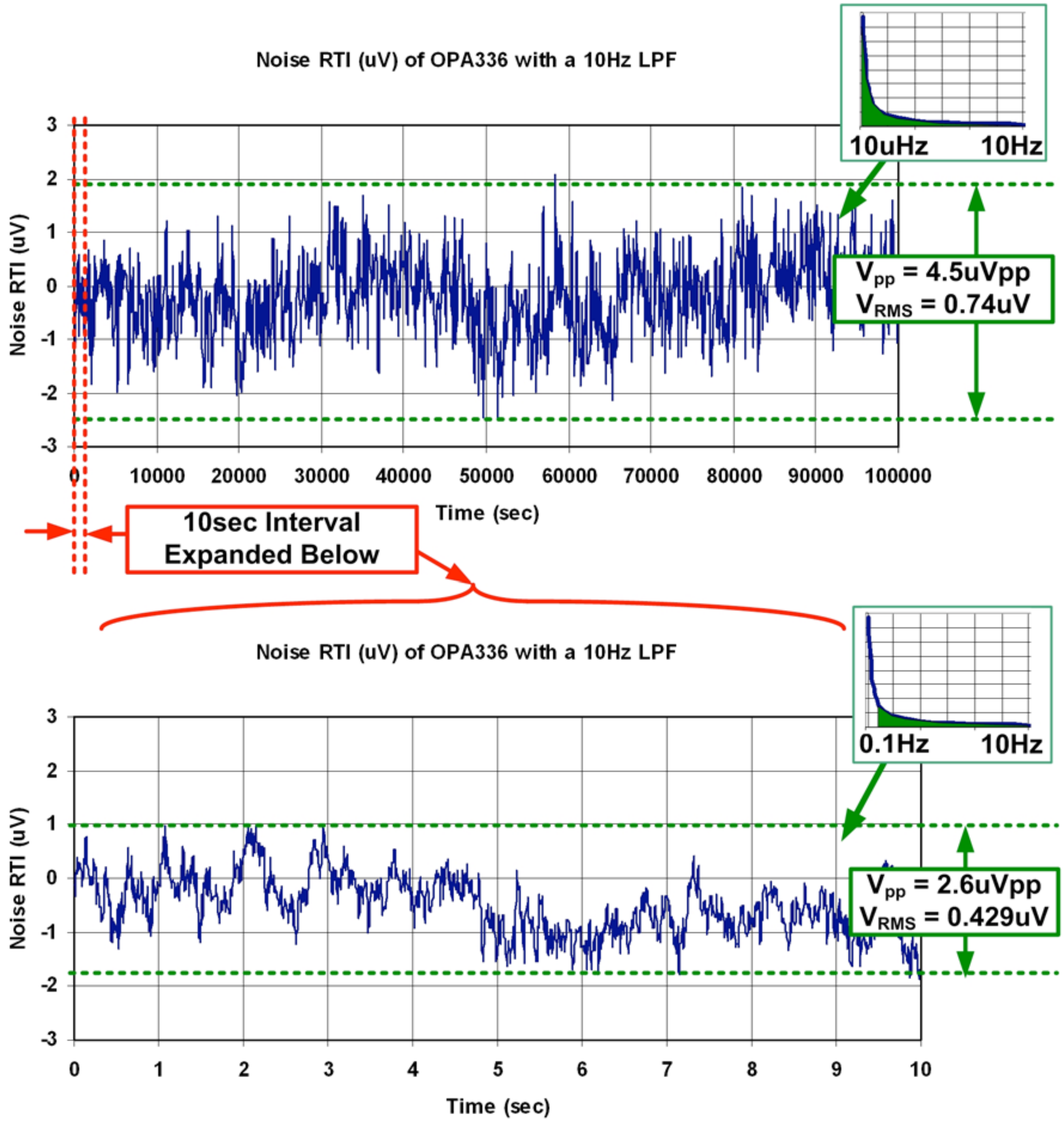


Fig. 9.17: OPA336 Noise Over A Long Time Interval

Zero-drift amplifiers do not contain flicker noise, so total noise is computed using the same method as broadband noise. Since noise spectral density is flat, it is possible to integrate noise down to 0 Hz. It is not possible to integrate down to 0 Hz with flicker noise because spectral density is infinite at 0 Hz. The total noise in each decade wide sub-interval is equal for flicker noise. With broadband noise, the total noise dramatically decreases for lower sub-intervals.

Looking at the power spectral density curve on linear axes helps illustrate how total noise decreases for lower sub intervals (see Fig. 9.18).

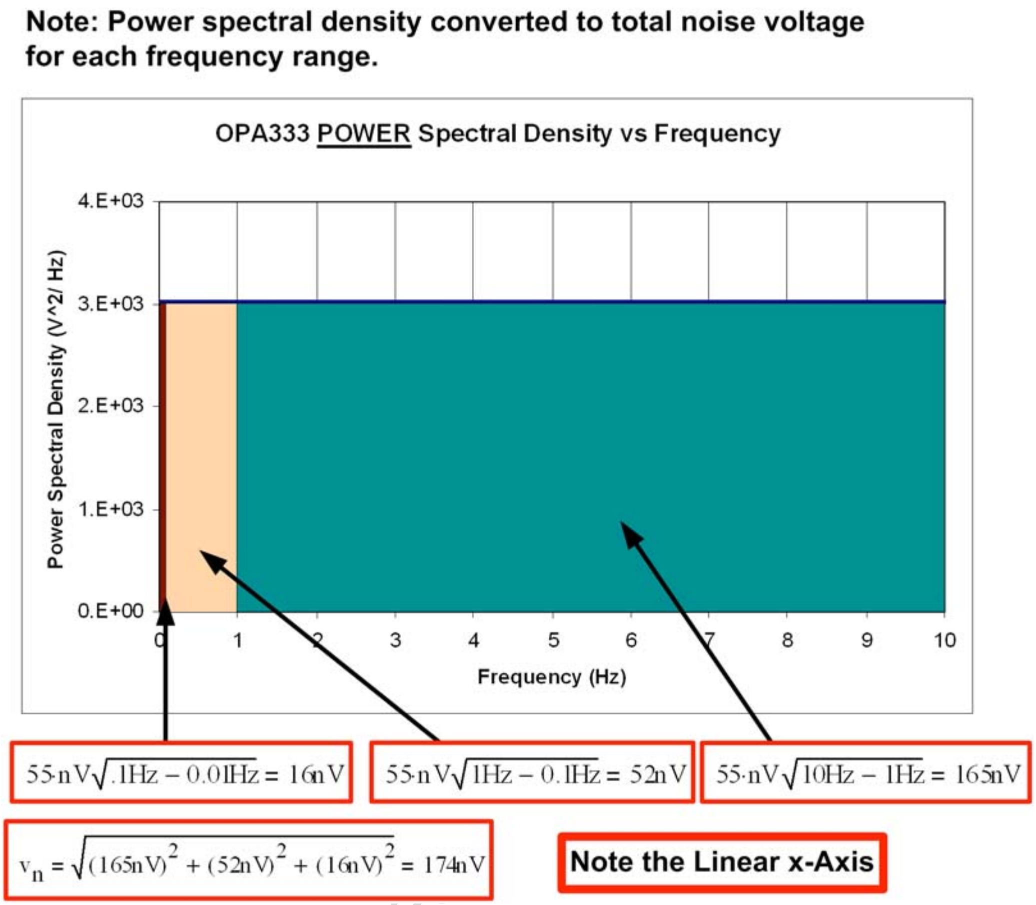


Fig. 9.18: Power Spectral Density For OPA333 On Linear Axes

Fig. 9.19 shows noise calculations for the OPA333 at different lower-cut frequencies which are set by the observation period. Note that the there is very little change in total noise with time. Because the spectral density is flat, the lower frequency sub-intervals have very little area (total noise). This is an advantage of the zero-drift topology compared to devices with flicker noise.

f_H	f_L	$1/f_L$ Sec.	$1/f_L$ Days	$1/f_L$ Years	Noise Calculation	Noise
10	1	1	$1.1\text{x}10^{-5}$	$3.1\text{x}10^{-8}$	$55 \cdot nV\sqrt{10Hz - 1Hz}$	165nV
10	0.1	10	$1.1\text{x}10^{-4}$	$3.1\text{x}10^{-7}$	$55 \cdot nV\sqrt{10Hz - 0.1Hz}$	173nV
10	$1\text{x}10^{-6}$	$1\text{x}10^{6}$	11	0.032	$55 \cdot nV\sqrt{10Hz - 1\mu Hz}$	174nV
10	$1\text{x}10^{-9}$	$1\text{x}10^{9}$	$1.1\text{x}10^{4}$	32	$55 \cdot nV\sqrt{10Hz - 1nHz}$	174nV

Fig. 9.19: OPA333 Noise Over Long Time Interval

Fig. 9.20 shows noise from the zero-drift amplifier is unchanged out to extremely long times.

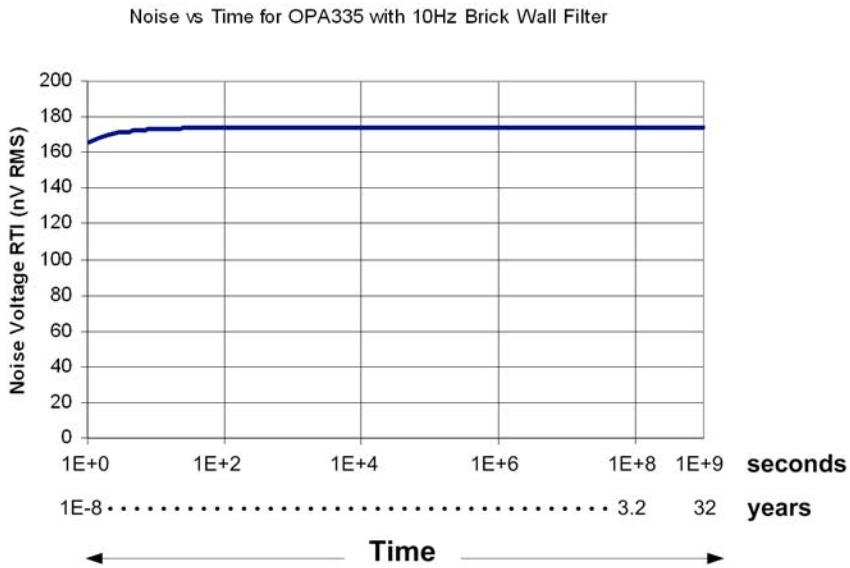


Fig. 9.20: OPA333 Noise Over Long Time Interval

The total noise for zero-drift amplifiers remains constant for different observation periods. The waveform in Fig. 9.21 (overleaf) illustrates the OPA333 over a 100,000 second interval (10 μHz). The upper cut frequency for this signal is 10 Hz. Thus, the noise bandwidth is 10 μHz to 10 Hz. The total rms noise over the entire interval is 0.173 μV. If you choose any sub-interval of time, the total rms noise will be the same.

In this example, a 10-s sub-interval is shown to have a total noise of 0.173 μV rms. The sub-interval in this example was taken from the first 10 s, but any 10-s interval will have the same total rms noise. Remember that the total noise for the two cases is very close because the area under the power spectral density curve is nearly the same.

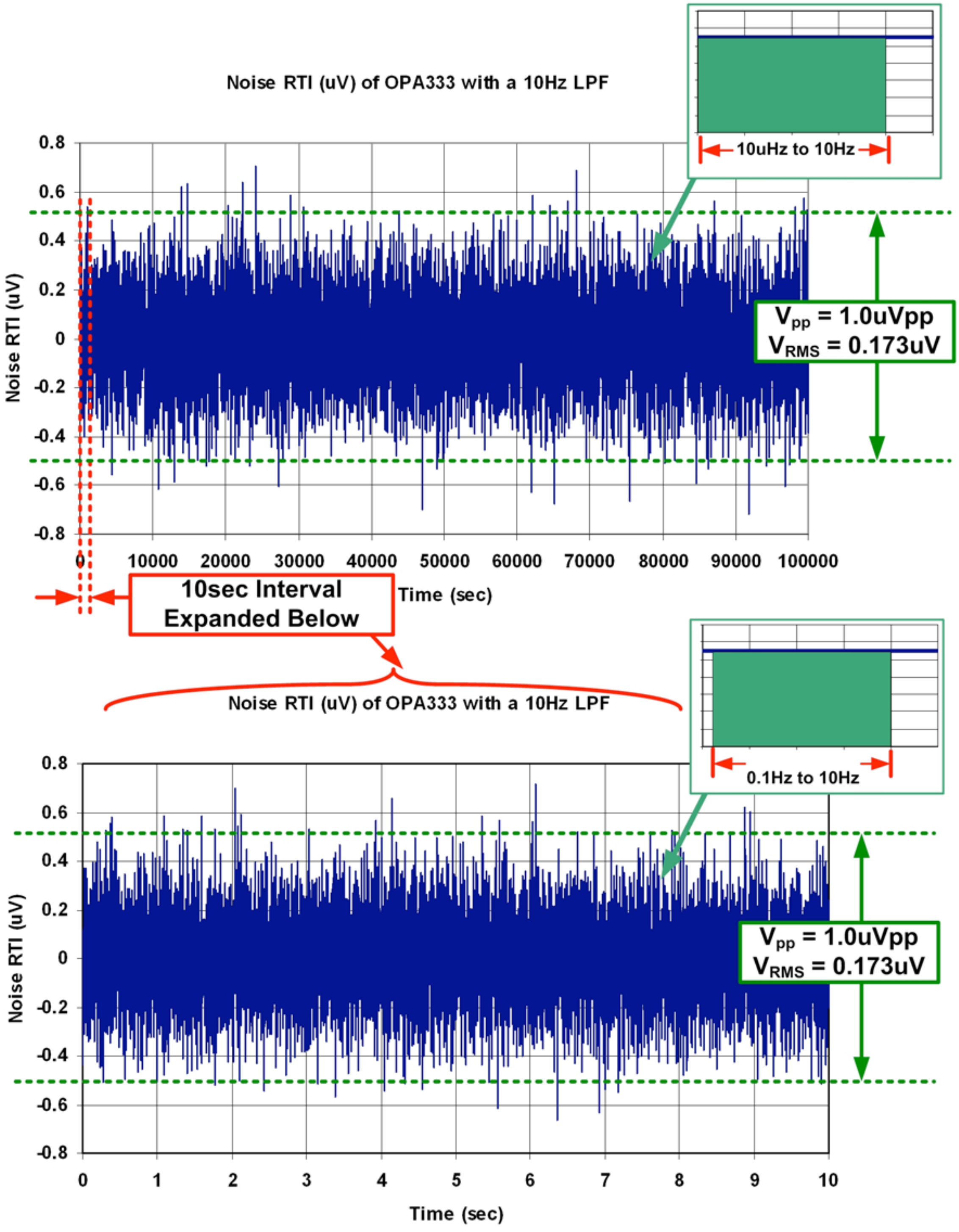


Fig. 9.21: OPA333 Noise Over Long Time Interval

Measuring Low-Frequency Noise

Most data sheets specify low-frequency noise as noise over the interval 0.1 Hz to 10 Hz. An active filter circuit for measuring this noise was covered in Part VI of this TechNote series. This article discusses low-frequency noise with extremely low-cut frequencies (10 μHz). Measuring extremely low frequencies makes it impractical to ac-couple the signal because the component values are outside of usable ranges. Fig. 9.22 shows a dc-coupled circuit that can be used to measure low-frequency noise. Depending on the noise level of the amplifier under test, the gain of this circuit should be adjusted.

One problem with this circuit is that it also amplifies the dc offset of the amplifiers. In the example in Fig. 9.22, noise is gained to a level that can be easily read on an oscilloscope (5.69 mVpp, see Fig. 9.23). Make sure that the gain of the first stage is at least 10; this makes the noise from the first stage dominate. In this example the offset is gained to 2 Vdc; this makes it impossible to measure the noise on the appropriate range. The circuit shown in Fig. 9.24 can be used to correct this offset.

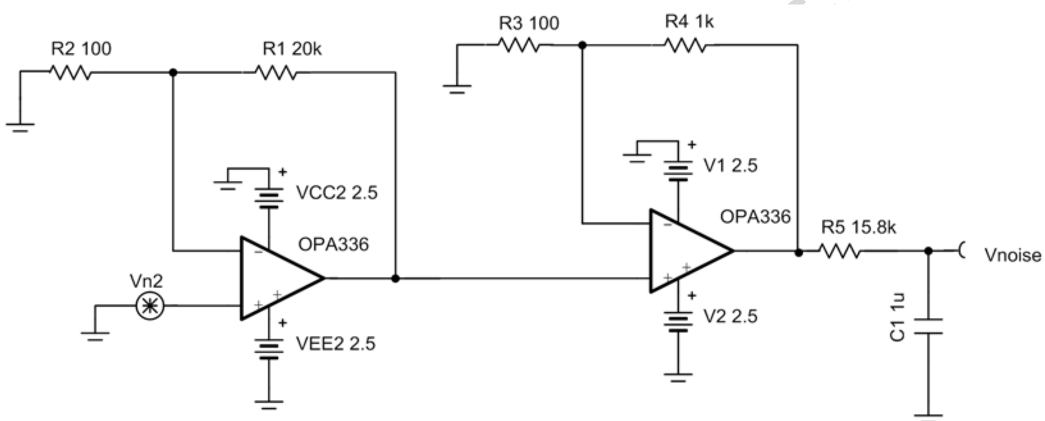


Fig. 9.22: Circuit For Measuring Noise (Dc To 10 Hz)

Total output flicker noise from 0.1Hz to 10Hz

$$V_{noise\_rms} = V_{n1Hz} \cdot \sqrt{\ln\left(\frac{f_H}{f_L}\right)} \cdot Gain$$

$$V_{noise\_rms} = 200nV \cdot \sqrt{\ln\left(\frac{10}{0.1}\right)} \cdot 201 \cdot 11 = 0.949\mu Vrms$$

$$Noise\_pp = 6 \cdot V_{noise\_rms} = (0.949\mu Vrms) \cdot 6 = 5.69mVpp$$

Output offset Voltage

$$V_{out\_offset} = V_{off} \cdot Gain1 \cdot Gain2 = (500\mu V) \cdot (201) \cdot 11 = 1.11V$$

Fig. 9.23: Noise/Offset Calculation For Fig. 9.22

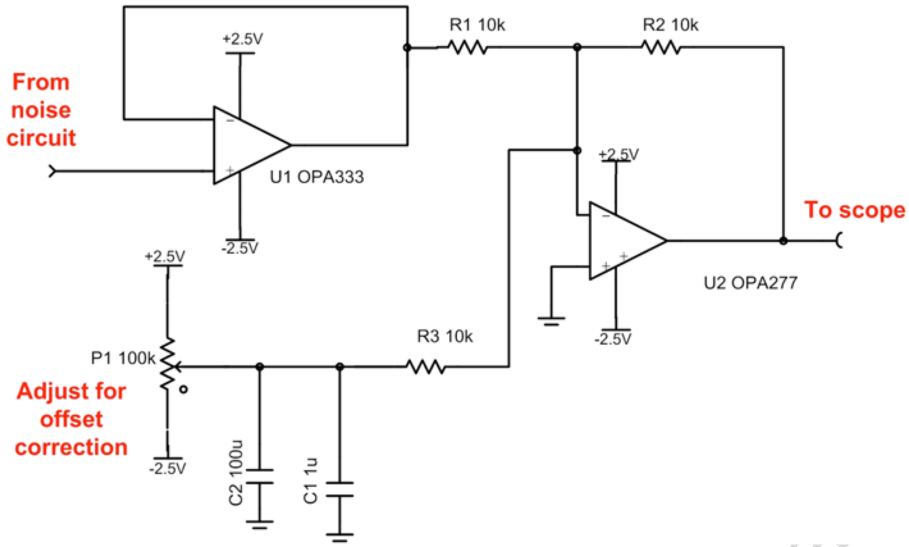


Fig. 9.24: Circuit For Canceling Output Offset

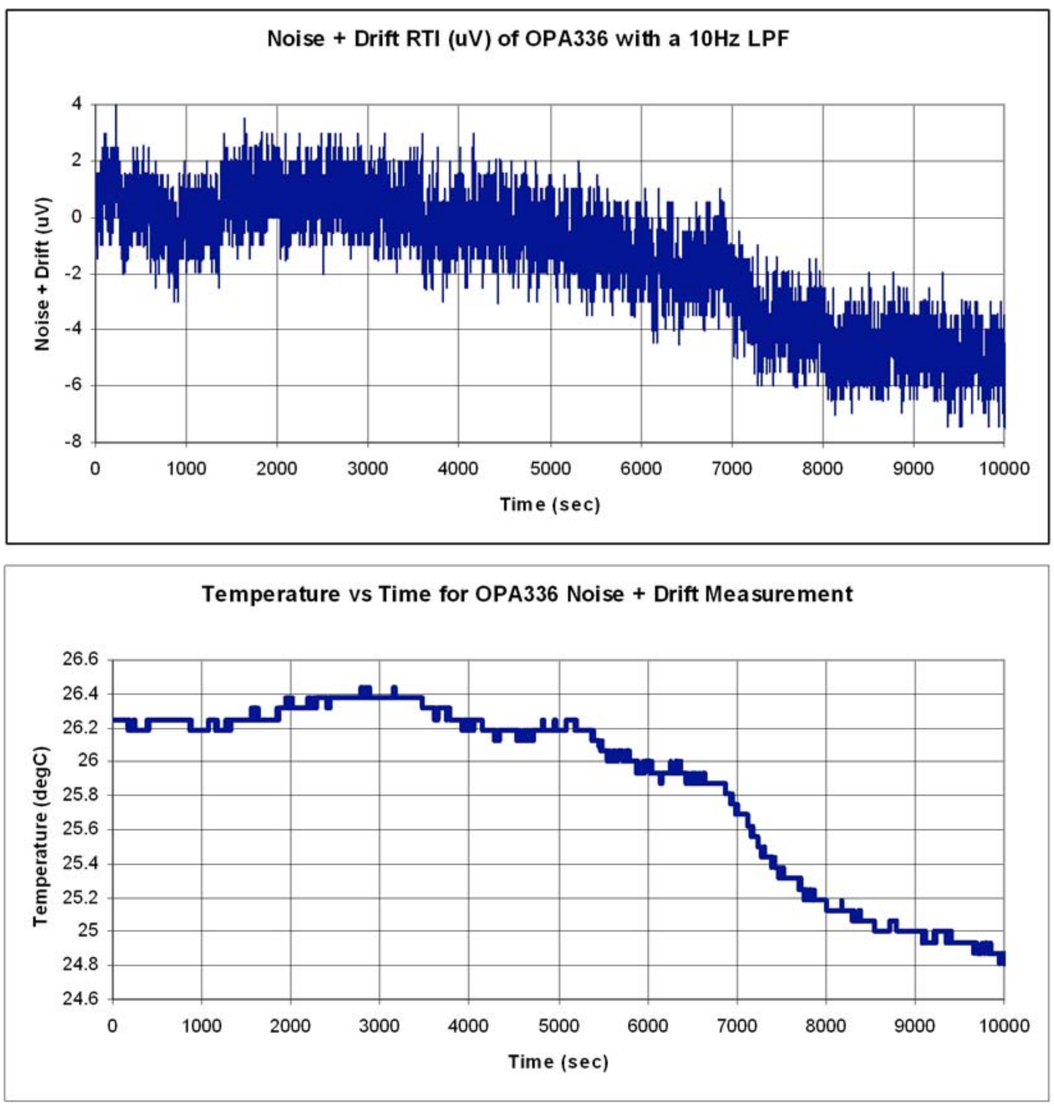


Fig. 9.25: OPA336 Noise Plus Temperature Drift At Ambient Lab Temperatures

One important consideration when making low-frequency noise measurements is to ensure that offset temperature drift does not affect the results. Offset voltage temperature drift looks very similar to 1/f noise. In fact, it is impossible to distinguish between the two error sources by analyzing the output signal. Fig. 9.25 shows the noise plus offset drift on the OPA336 over 10,000 s. No temperature control was used during this measurement (the ambient lab temperature varied 2ºC). This small fluctuation in room temperature significantly affected the output signal. Fig. 9.26 compares the results of the circuit with no temperature control to the same circuit under precision control. The precision temperature control is achieved using a thermal bath. The thermal bath used in this measurement is a chamber filled with an inert, fluorinated fluid controlled to 0.01ºC. The typical expected drift for this example is 1.5 μV x 1.6 = 2.4 μV. In this case the drift appears to be about 5 μV; ie the mean value of the signal shifts from 0 μV to –5 μV.

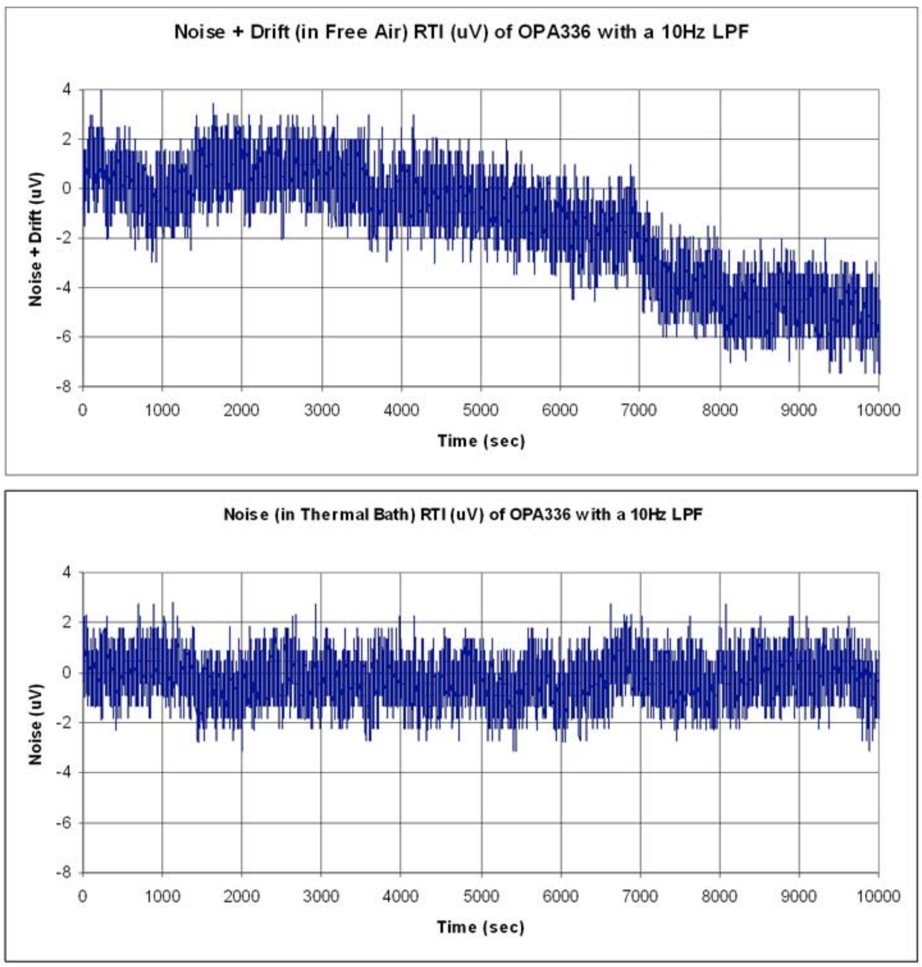


Fig. 9.26: OPA336 Noise: Tight Temperature Control Vs Ambient temperature

In the case of zero-drift amplifiers, the effect of offset temperature drift is greatly reduced. Fig. 9.27 shows noise measurements made with no temperature control (the ambient lab temperature varied 2ºC). Fig. 9.28 compares the measurement with no temperature control to the same measurement made in a thermal bath. There is no noticeable difference between the circuit with and without temperature control. The ultra-low temperature drift of the zero-drift amplifier is a key benefit of this topology.

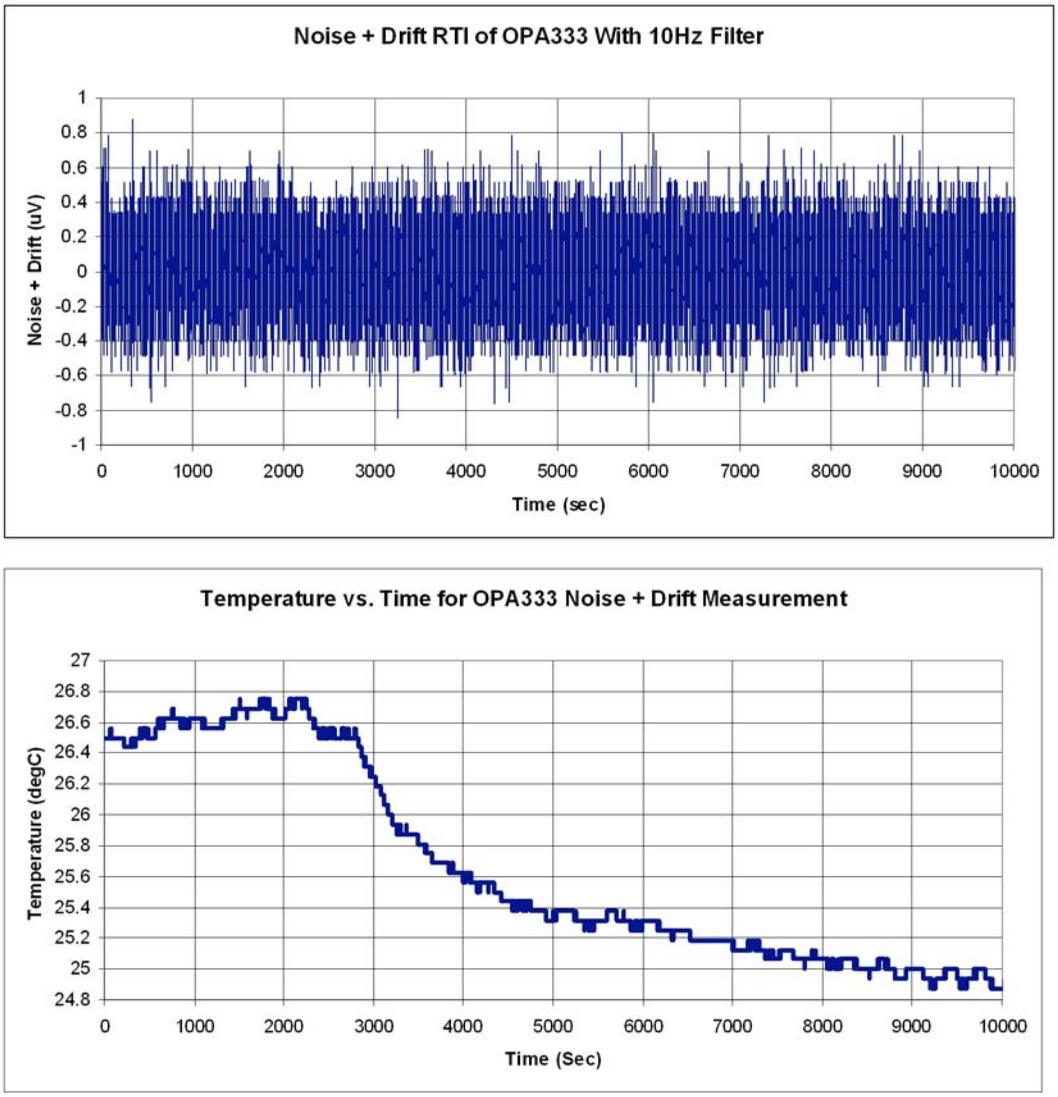


Fig. 9.27: OPA333 Noise Plus Temperature Drift At Ambient Lab Temperatures

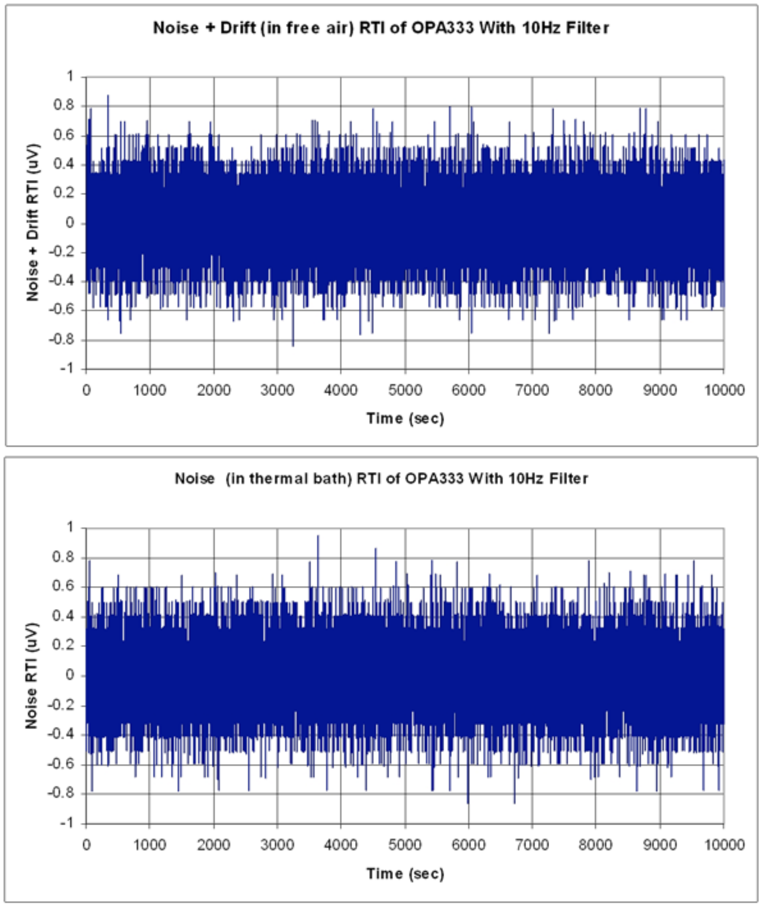


Fig. 9.28: OPA333 Noise: Tight Temperature Control Vs Ambient temperature

Another important consideration for noise measurement is to measure the rms noise (standard deviation) rather then peak-to-peak noise. Often people read noise as a peak-to-peak reading on the oscilloscope. This is good as an approximation, but is not adequate for precision results. The problem is that the number of samples can greatly affect the peak-to-peak reading. Remember that peak-to-peak can be estimated by multiplying the standard deviation by six.

Mathematically this means that there is a 99.7% chance that the noise is bounded by the peak-to-peak estimate. So, 0.3% of the noise is outside this range. If the number of samples is increased, you will see more occurrences outside the six sigma peak-to-peak estimate. The rms noise, on the other hand, remains relatively constant regardless of the number of samples. Fig. 9.29 illustrates the same signal captured on a digitizing oscilloscope with 25,000 samples and 2500 samples. The peak-to-peak measurement is significantly larger for the chart with more samples. Part I of this series gives more detail on this topic.

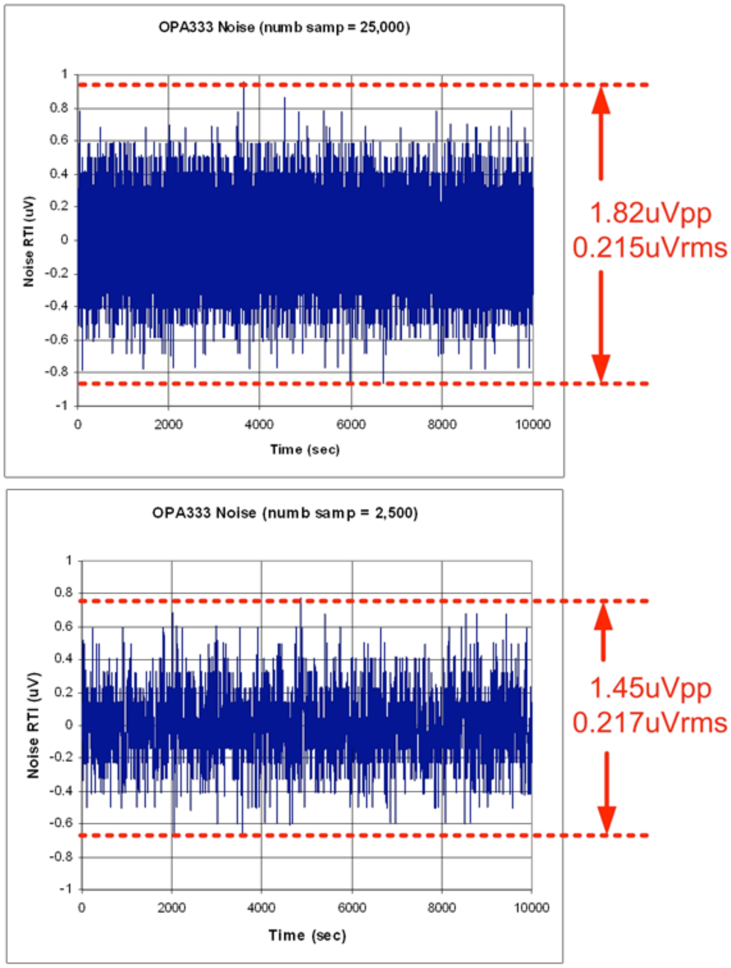


Fig. 9.29: Effect Number Of Samples Has On Peak-To-Peak Noise Measurement

One final note regarding rms noise measurements: you should be careful ***not*** to include dc into the noise measurement (average = 0). The mathematics of noise analysis assumes that the average value is zero. The easiest way to eliminate the average is to do this to find the standard deviation of the signal rather then the rms. The standard deviation is mathematically defined to be the rms with zero average (see Fig. 9.30).

Most digitizing oscilloscopes have the ability to save the results into a spreadsheet format. In the spreadsheet you can use spreadsheet mathematics to compute the standard deviation: “=STDEV(..RANGE..)” in Microsoft Excel. Some digitizing oscilloscopes provide an rms mathematical operator. However, in general, you will get more accurate results by importing the data into a spreadsheet and perform the standard deviation. This is because the signal typically has a slight dc offset even when ac-coupled. Of course, different instruments have different idiosyncrasies, so experiment with your equipment to learn its limitations.

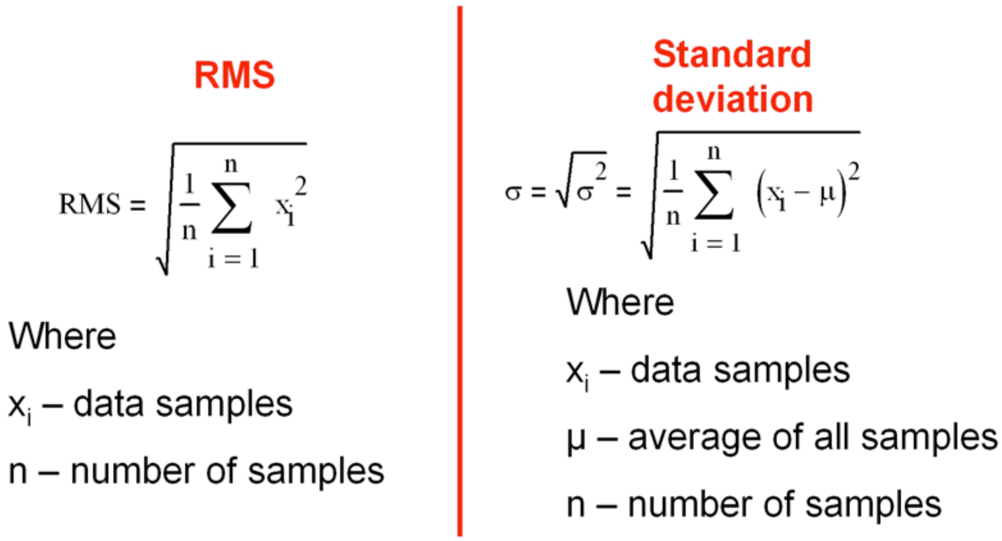


Fig. 9.30: Standard Deviation Vs Rms

Summary And Preview

In this TechNote, we discussed zero-drift amplifiers and noise calculations. Zero-drift amplifiers have low-offset, low-offset drift, and no flicker noise. Zero-drift amplifiers use an internal digital circuit to continuously correct for input voltage offset. In Part X, we will discuss reference noise and how it relates to A/D performance.

Acknowledgments

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- Rod Burt, Senior Analog IC Design Manager
- Joy Zhang, Analog IC Design Engineer
- Bruce Trump, Manager Linear Products
- Tim Green, Applications Engineering Manager

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[1] Thomas Kugelstadt, "Auto-zero amplifiers ease the design of high-precision circuits," Texas Instruments Document Number SLYT204
http://focus.ti.com/lit/an/slyt204/slyt204.pdf
[2] Thomas Kugelstadt, "New zero-drift amplifier has an IQ of 17 µA", Texas Instruments Document Number SLYT272 http://focus.ti.com/lit/an/slyt272/slyt272.pdf

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