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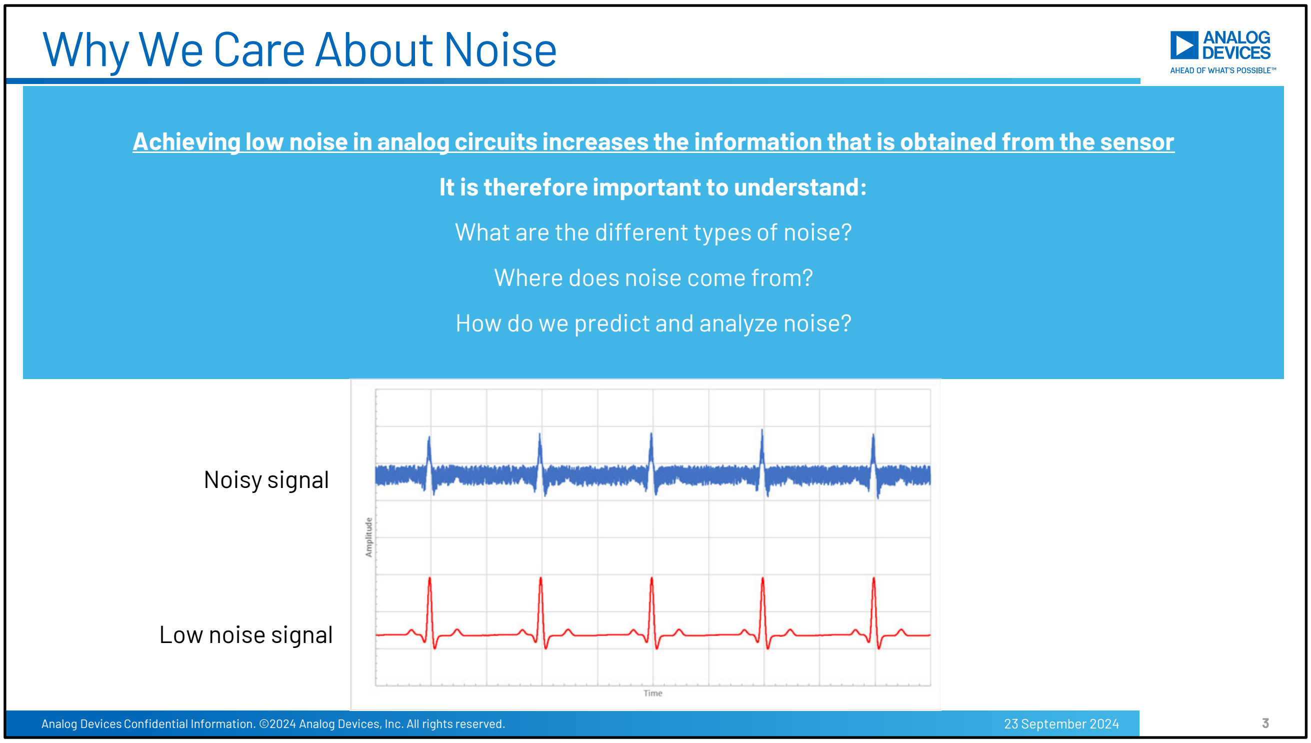
Demystifying Noise

Ed Mullins, Principal Applications Engineer

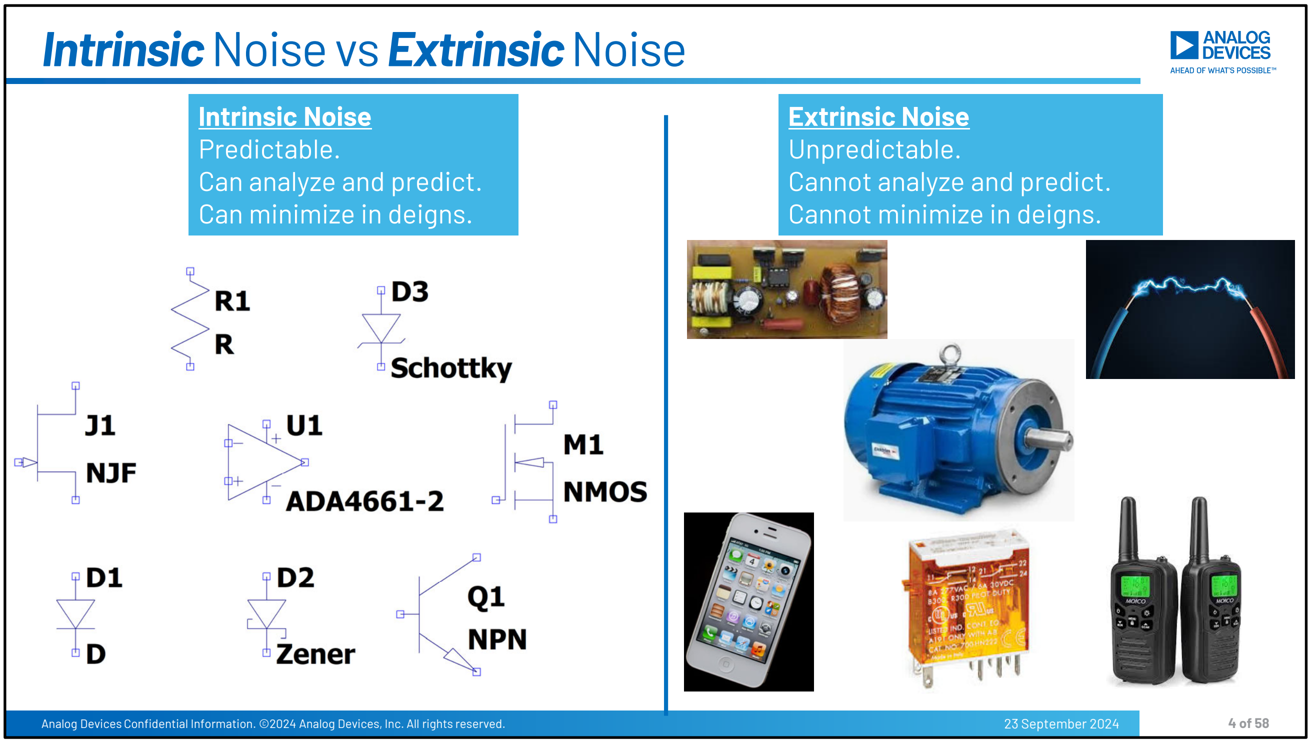
Why We Care About Noise

It can be difficult to carry on a conversation in a noisy environment such as a bar or a restaurant or near a waterfall

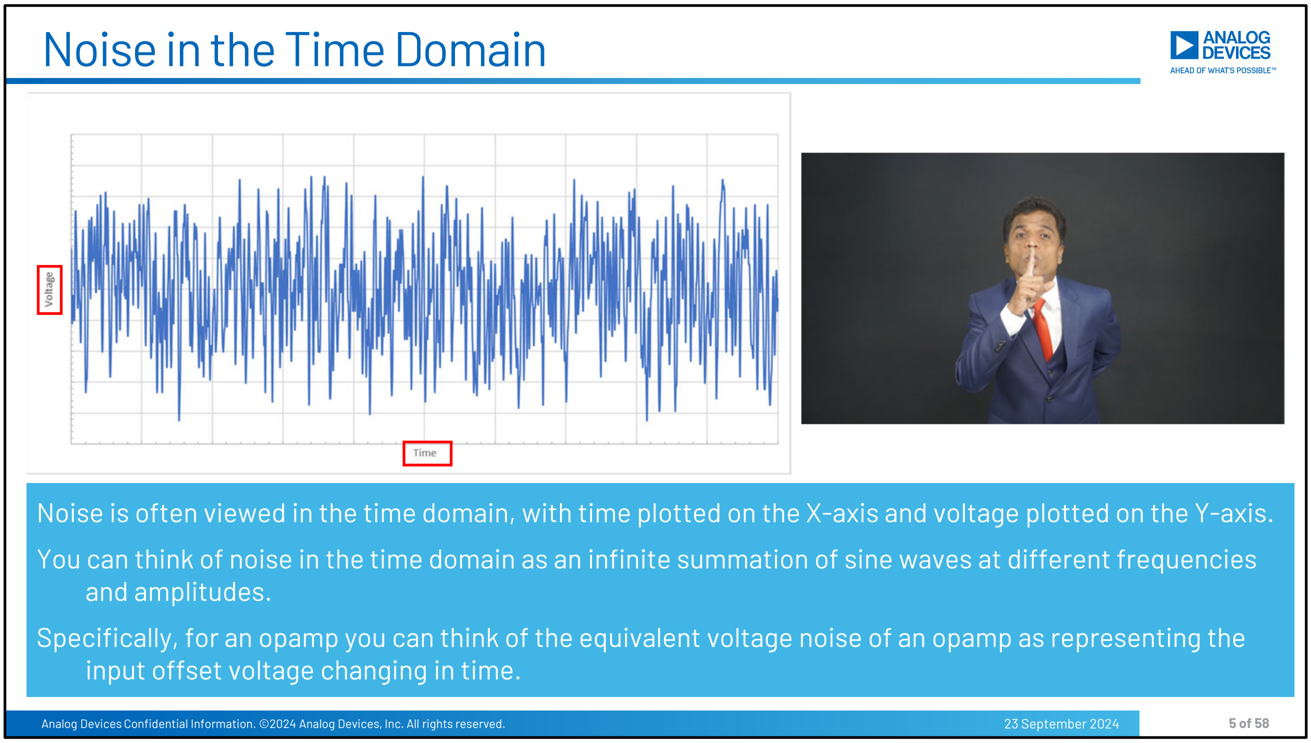
Similarly, it can be difficult to measure small signals in the presence of noise



- Data within a signal can be easily masked by noise. Start your design with the goal of obtaining the lowest overall system noise and select the best signal chain components for your project to maximize performance. Understanding where noise originates from and how it makes its way into the ADC are key to minimizing the overall noise in the design, resulting in maximum and repeatable and dependable performance.



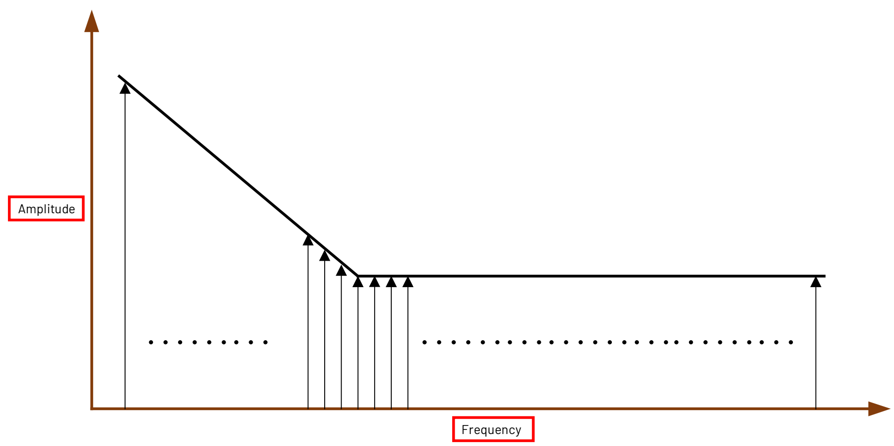
- We are going to focus on predicting, analyzing and minimizing INTRINSIC noise in our application circuits.
- INTRINSIC noise is that internal to the op amp and internal to support components such as resistors, transistors, diodes and other semiconductor devices.
- EXTRINSIC noise is that generated external to the op amp and its support components.
- Sources of EXTRINSIC noise include power electronic devices, control circuits, arcing equipment, switching power supplies, arcs generated by lightning and electrical storms, electromagnetic interference due to switches and relay contacts, operating inductive loads, such as motors, coils, solenoids and relays, electromagnetic interference from currents in cables, and frequency interference from equipment like radio transmitters and cell phones.



- Think of the offset voltage of an opamp changing in time…if you had a perfect noiseless voltmeter and measured the opamp offset voltage 1000 times you would have 1000 different readings due to the opamps voltage noise.
- This shows the erratic and unpredictable behavior of noise.

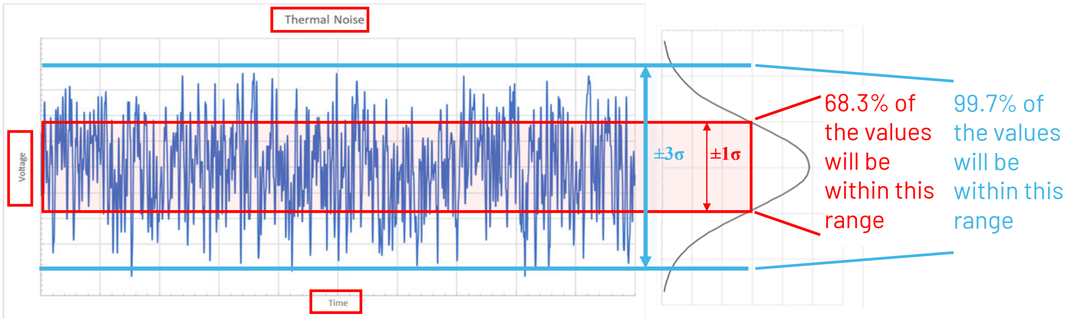
Noise in the Frequency Domain

Noise in the frequency domain can be thought of as an infinite number of impulses across the frequency spectrum.



- While I like to think of noise in the time domain, many of you may prefer to think of noise in the frequency domain.

Random Nature of Noise



Number of Standard Deviations	Chance of Measuring Voltage (%)
2 *sigma* (+/-1 sigma)	68.3
3 *sigma* (+/-1.5 sigma)	86.6
4 *sigma* (+/-2 sigma)	95.4
5 *sigma* (+/-2.5 sigma)	98.8
6 *sigma* (+/-3 sigma)	99.7
6.6 *sigma* (+/-3.31 sigma)	99.9

Noise is random in nature and generally has a Gaussian distribution. Statistical methods are used to analyze and combine noise signals.

- Noise is random in nature and generally has a Gaussian distribution.
- Examining the ratio of peak-to-peak to the RMS is similar to the relationship between max and min to the std deviation.
- If you know either the peak-to-peak value of the noise or the RMS value, the other can be determined with a given probability.
- For example, if you want to cover 99.9% of all voltage values, you will use a multiplication factor of 6.6
- (Here, RMS is not 0.707 of a sinewave – it's statistical analysis).

Rules of Thumb

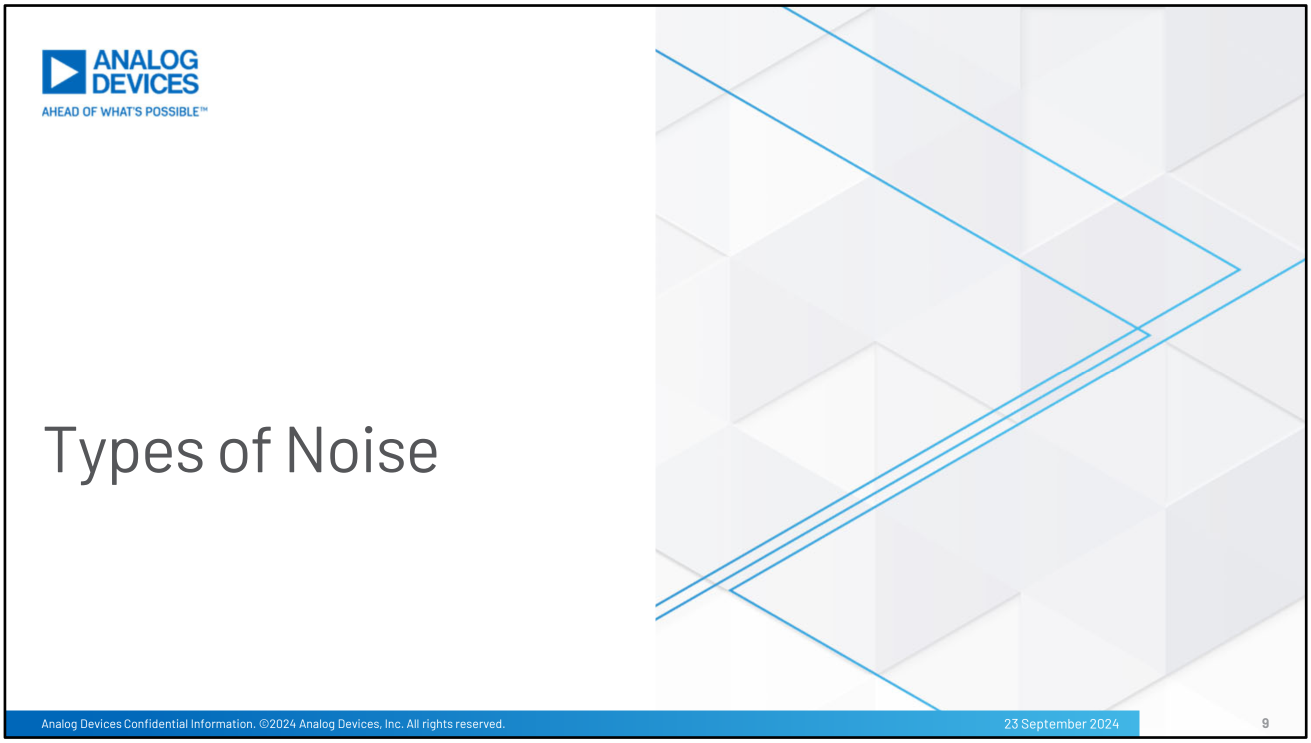
Because noise has a Gaussian distribution you can estimate the relationship between RMS and Peak-to-Peak values as:

$$V_{PP} = 6 \times V_{RMS} \text{ for 99.7\% of the population (use 6.6 for 99.9\%)}$$

Because noise sources are uncorrelated, they are combined (added) as the square-root of the sum of the squares:

$$V_{n_{total}} = \sqrt{\left(V_{n_1}\right)^2 + \left(V_{n_2}\right)^2 \ldots + \left(V_{n_n}\right)^2}$$

- Noise sources are random in nature and generally, uncorrelated…in this way there are no phase relationships between various noise signals in a system and as such noise sources can be combined into a total noise by using the root-sum-squares method of combining signals.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Shot Noise

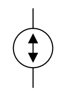
- DC Current flow through PN junctions will occur as individual carriers with charge *q*.
- The electrons will transition across the PN junction randomly in time and in the aggregate will result in an average DC current.
- If examined closely on a sensitive oscilloscope a DC current will look like a bunch of random current pulses and this is modelled as a noise current source.

$$I_{SHOT} = \sqrt{2qI_{DC}\Delta f}$$

$q =$ electronic charge, 1.602×10^{-19} Coulombs

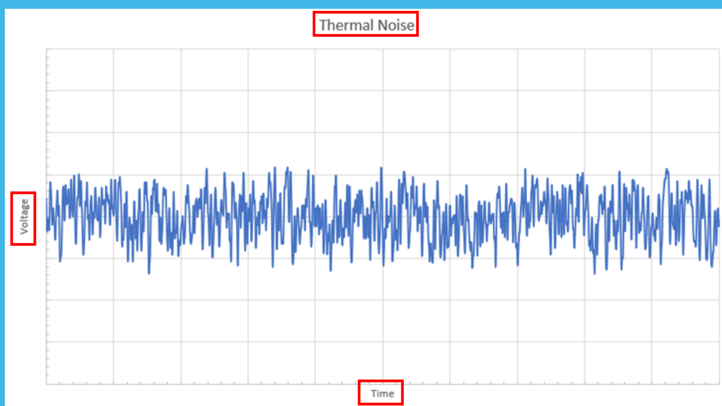
I_{DC} = DC current in amps

Δf = Noise bandwidth in Hertz


Shot Noise

Thermal Noise

- Thermal noise in conductors is due to the random thermal motion of the electrons.
- Thermal noise is directly proportional to temperature.
- In a resistor, thermal noise can be represented by either a voltage source in series with the resistor or as a current source in parallel with the resistor.



$$V_N = \sqrt{4kTR\Delta f}$$

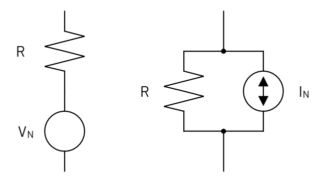
$$I_N = \sqrt{4kT\frac{1}{R}\Delta f} = \frac{V_N}{R}$$

k = Boltzmann's constant, $1.38 \times 10^{-23}\ \frac{\text{W-s}}{K}$

T = Temperature K

R = Resistance Ω

Δf = Noise bandwidth in Hertz



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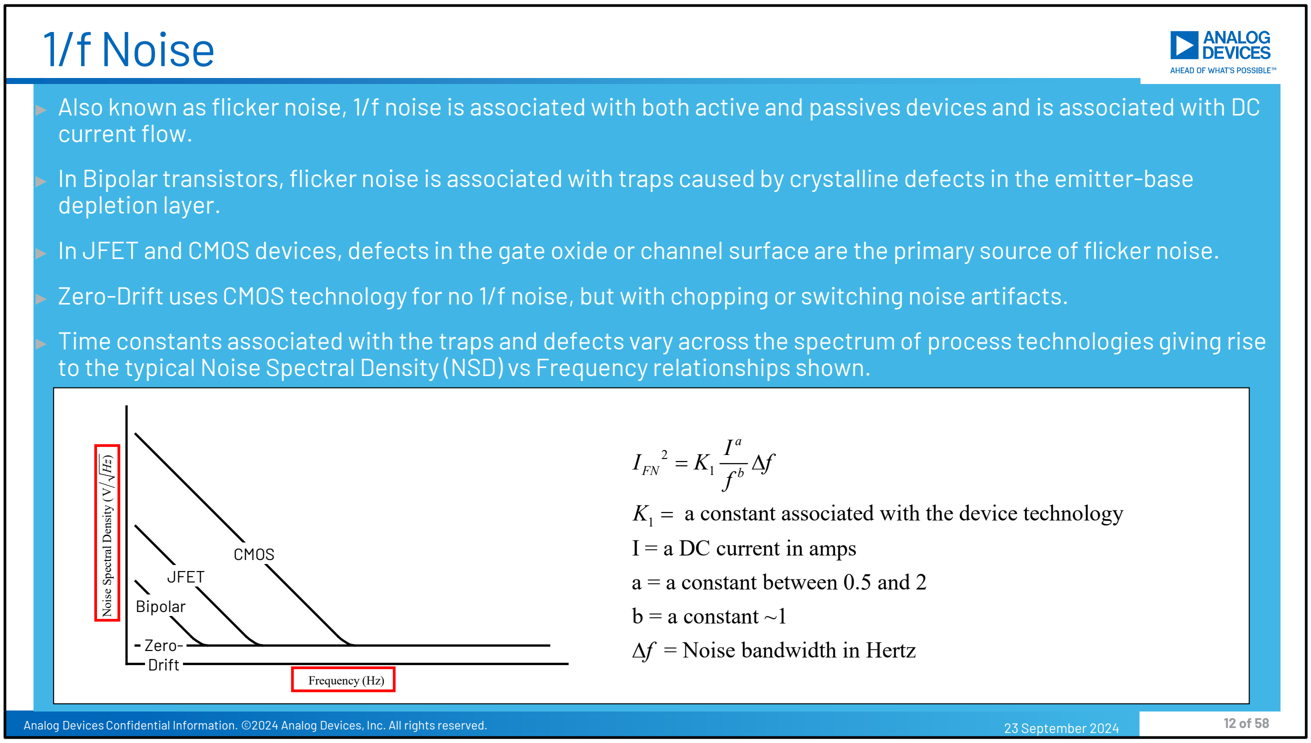
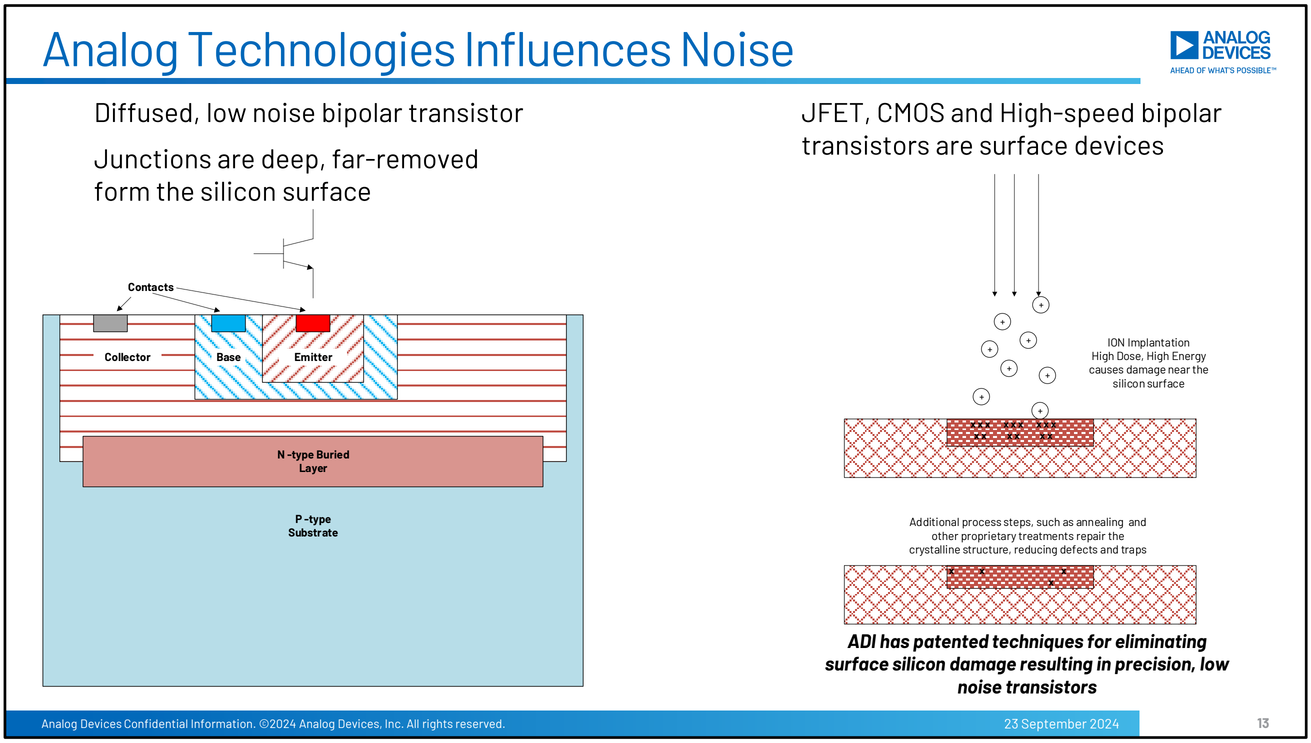
- I generally like to model this as a voltage source....



- Historically Precision bipolar amplifiers have offered the lowest flicker corner of all amplifier technologies. This is particularly true for older BJT technologies.
  - More recent bipolar technologies are constructed using ion implantation…for these reasons the noise corner for modern high-speed, bipolar technologies is much higher than their older precision counterparts.
- JFET's historically has the next lowest flicker corner frequency and combines low offset, low noise and low input bias current.
- CMOS has the highest corner frequencies of the three technologies. Not all CMOS process are created equally but ADI does use special wafer processing techniques to mitigate low frequency noise.
  - A more recent development of Zero-drift amplifiers allows for CMOS amplifiers to achieve a flat NSD vs frequency, eliminating the 1/f component, but comes with the addition of chopping or switching noise and the possibility for intermodulation effects.
- Finally, the amount of noise in your circuit depends on the design and the resistors you use. For example, a CMOS amp that has a higher 1/f noise can still produce lower noise if you pick the right resistor ratio.



- To further examine why JFET and CMOS technologies have higher 1/f noise corner frequencies we need to have a high-level understanding of how these technologies are fabricated. Older, low-noise bipolar technologies have deep, diffused junctions. Because these PN junctions are deep in the silicon (relative to JFET and CMOS transistors) they are largely free from defect sites associated with wafer processing steps. More modern, polysilicon emitter, high-speed bipolar transistors tend to have much higher 1/f noise than their deeply diffused precision bipolar counterparts.
- A JFET transistor is a surface device constructed from a conducting channel sandwiched between two gate layers, above and below the channel layer. An Nch JFET will have a P-gate and vice-versa. All of the layers in a JFET (top gate, bottom gate and channel) are constructed by a process called ion implantation. Ion implantation is a method by which dopant species (P or N types) are ionized and "injected" with high energy into the silicon surface with a certain dose and energy. As these ion collide with the silicon lattice, defects are created. These defects, if left unattended, will result in very high leakage currents and traps that create noise. Analog devices has long used patented approaches in production to fix or repair these damage locations resulting in low noise transistors.
- CMOS transistors are similar in that they are also constructed via an ion implantation process…further challenging CMOS form a 1/f noise perspective is the very thin gate oxide region. Any defects or charged particles trapped in the oxide region will result in noise with the 1/f characteristic. Analog devices also has specific process steps and mask steps to address these types of defects, reducing them from standard CMOS technologies.
- As an example, the 1/f corner frequency for a CMOS device built on Analog Devices technologies may be at frequencies of 1kHz or so, compare that to competitors who use low-cost, digital CMOS process nodes, without any special analog treatments, and have corner frequencies extending into the MHz region. Such amplifiers are not suitable for any

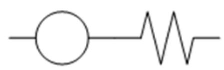
precision applications.

Noise in Resistors

Modeling Thermal Noise in Resistors

The noise voltage of a resistor is modelled as a voltage source placed in series with a noiseless resistor:

$$V_n = \sqrt{4kTR\Delta f}$$



Where:

k = Boltzmann's constant, $1.38 \times 10^{-23}\ \frac{\text{W-s}}{K}$

T = Temperature °K

Δf = Noise bandwidth in Hertz

Rule of Thumb
At room temp, a **1kΩ** resistor has a noise voltage of 4nV/SQRT-Hz, you can use this simple Rule of Thumb to determine the noise for any other resistor with the following equation:

$$V_{n_{R2}} = 4\,nV/\sqrt{Hz} \times \sqrt{\frac{R_2}{1k}} \Longrightarrow \text{Example: a 100k resistor has a noise density of } 40\,nV/\sqrt{Hz}$$

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- Thermal noise is due to the random motion of electrons in a conductor.
- Any conductor that is not at absolute zero, will have sufficient thermal energy to produce this random motion of electrons – resulting in noise that is a function of temperature and resistance.
- There's a simple Rule of Thumb to determine the noise for any other value of resistor…does anyone know the equation?

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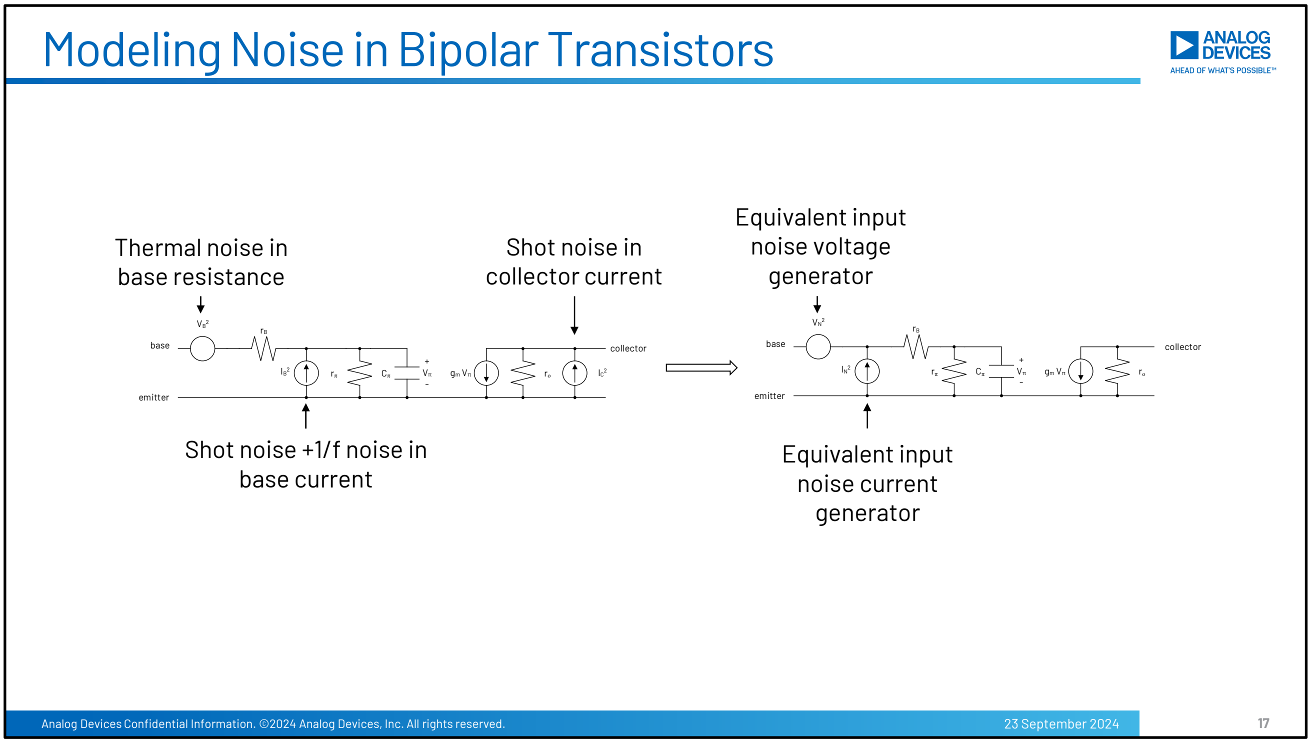
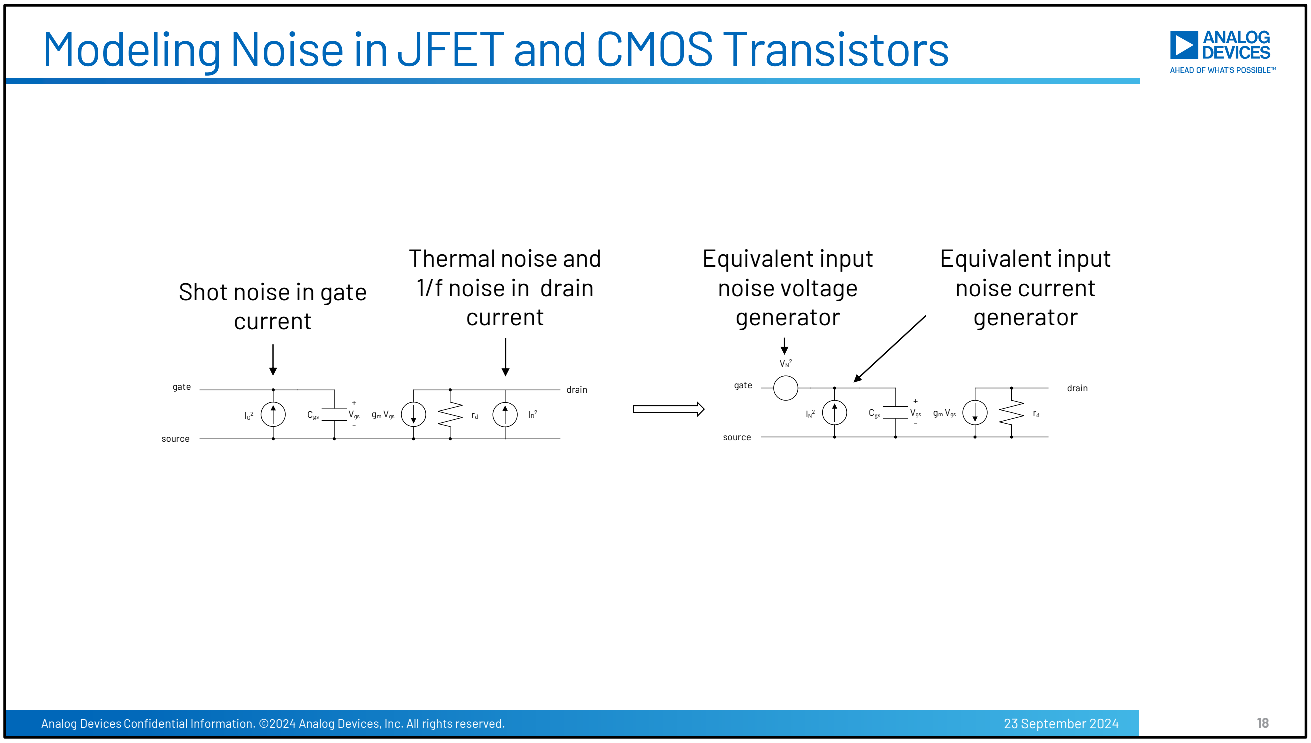
Noise in Transistors



- Bipolar transistors have shot noise associated with their DC collector current, shot noise and 1/f noise associated with their base current and thermal noise associated with the physical resistance in the base region. All these sources can be combined into simplified input referred equivalent noise voltage and noise current generators. It is worth noting that higher collector current results in lower input referred noise voltage. This is since the transconductance of the transistor is directly proportional to the DC collector current and the shot noise in the collector is proportional to the square root of the DC collector current. Higher quiescent current amplifiers tend to have lower noise voltage.



- JFET and CMOS transistors have thermal noise and 1/f noise associated with their DC drain current, shot noise associated with their gate current. All these sources can be combined into simplified input referred equivalent noise voltage and noise current generators. It is worth noting that higher drain current results in lower input referred noise voltage.

Summary (so far)

- Noise, if not mitigated can mask out small details in analog signals
- Noise is random in nature, having a Gaussian distribution
- You can think of noise as the input offset voltage of an opamp changing in time (similar for input bias current)
- There are common types of noise in every opamp circuit
  - Thermal noise from resistors, JFETS and CMOS devices
  - Shot noise from bipolar transistors and input leakage currents
  - Flicker noise from bipolar, JFET and CMOS transistors
  - Popcorn noise is generally eliminated for amplifiers and rarely needs to be considered
- The lowest noise amplifiers have historically been Bipolar, followed by JFET then CMOS This is particularly true for the low frequency region dominated by 1/f noise
- Zero-Drift amplifiers are built on CMOS technologies and eliminate 1/f noise and as such only the broadband noise spectral density needs to be considered in any noise analysis
- Ultimately overall system noise performance can be limited by the 1/f noise in very low frequency systems

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Noise in Amplifiers

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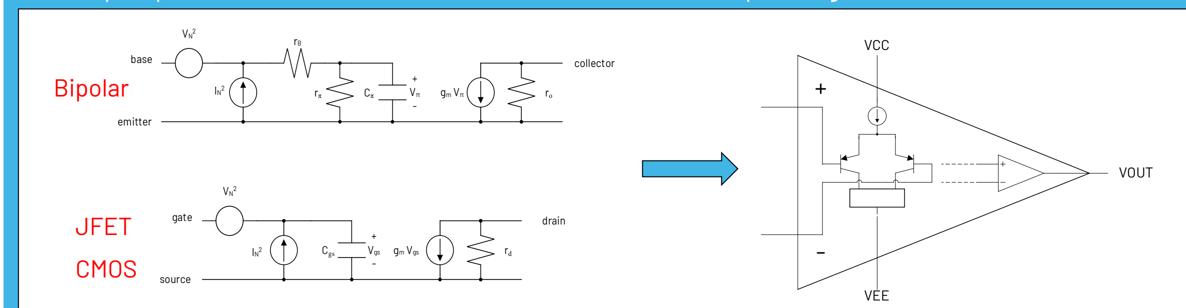
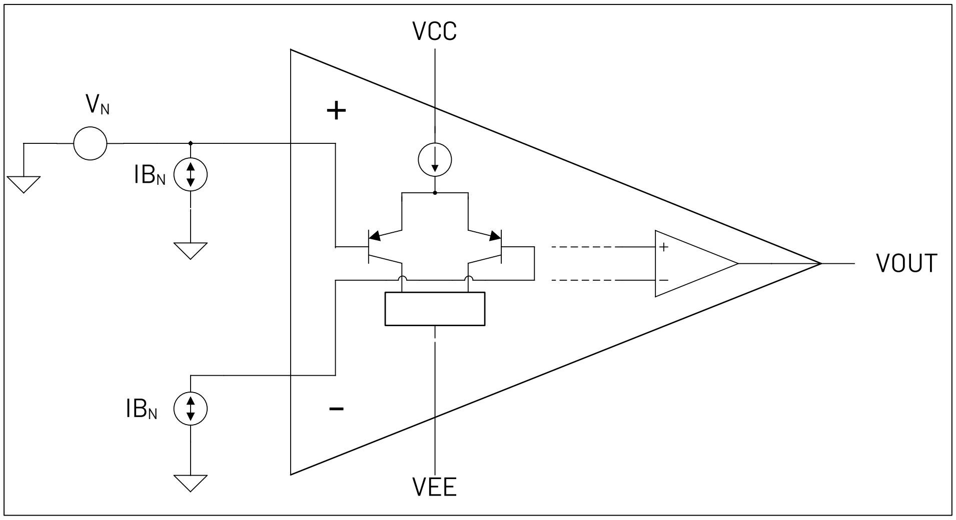
Sources of Noise in Operational Amplifiers

- Operational amplifiers are made from transistors, resistors and capacitors.
- Transistors will be either Bipolar, JFET or CMOS.
- Each of these noise sources can be represented by a noiseless element with its corresponding noise sources (voltage and/or current) placed at the input.
- It stands to reason then an amplifier will likewise consist of many internal noise sources than can be modelled as voltage and current sources, lumped together, at the amplifier inputs.
- In an opamp the dominant sources of noise are found in the input stage.

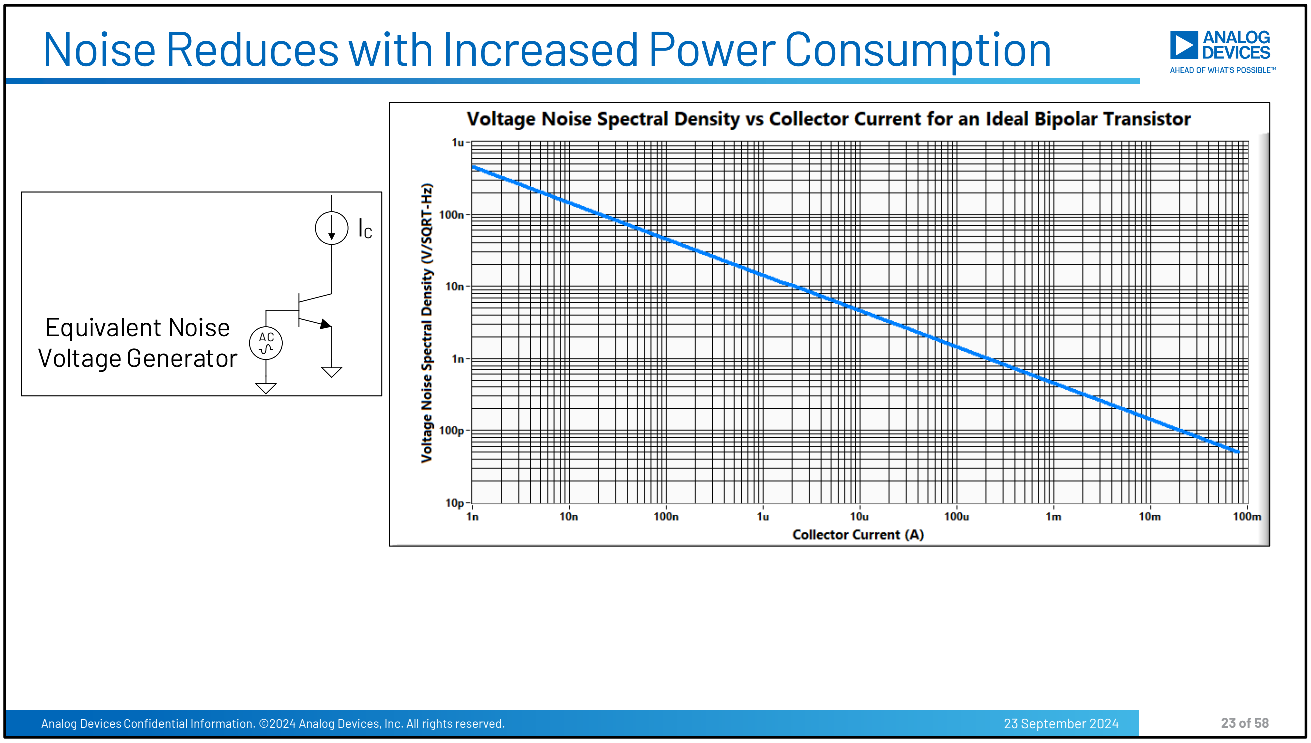


Opamp Noise Model

- The opamp is modelled with an equivalent noise voltage source connected in series with the non-inverting input and two equivalent noise current sources, one at each input.



- Always model the Vn of op amp at non-inverting input.
- So, the triangle of the op amp is now noiseless (noise modelled outside).



- Examining the voltage noise spectral density vs collector current for a single bipolar transistor reveals the fact that input referred voltage noise spectral density decreases with increasing current (power). This is also true for JFET and CMOS transistors, albeit they have a slightly different slope than does the bipolar device.
- The key point is, since amplifiers are made up from transistors, it stands to reason that lower power amplifiers have higher input referred noise voltage than do higher power amplifiers.
- (As the input diff stage biasing current increases, you have a higher operating current to noise ratio).

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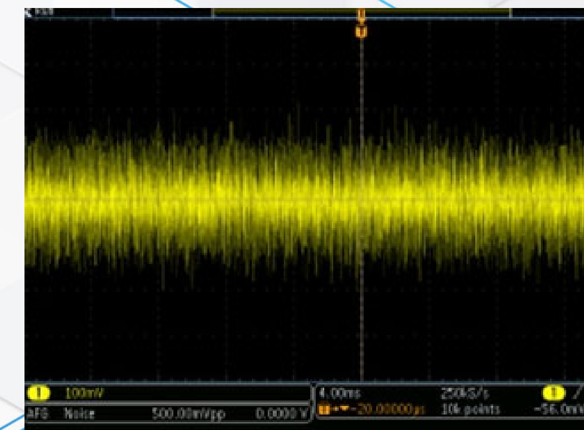
Integrated Noise



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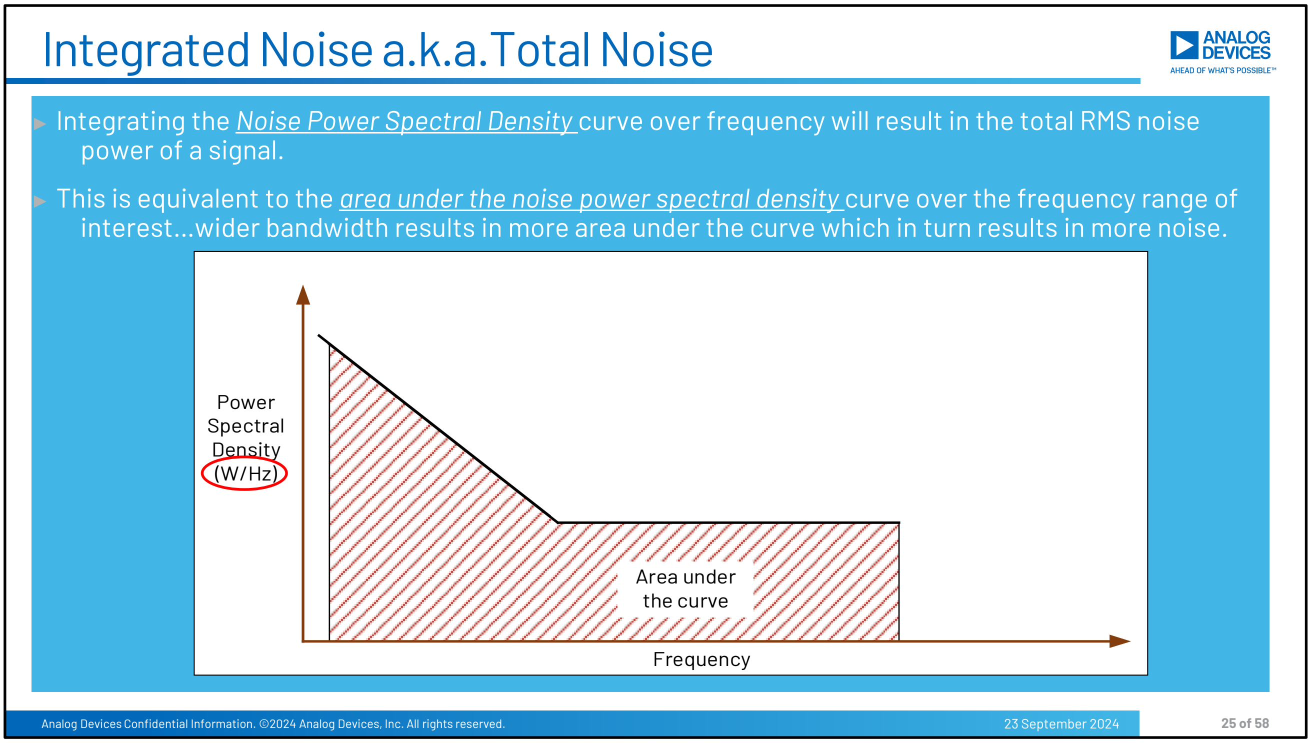
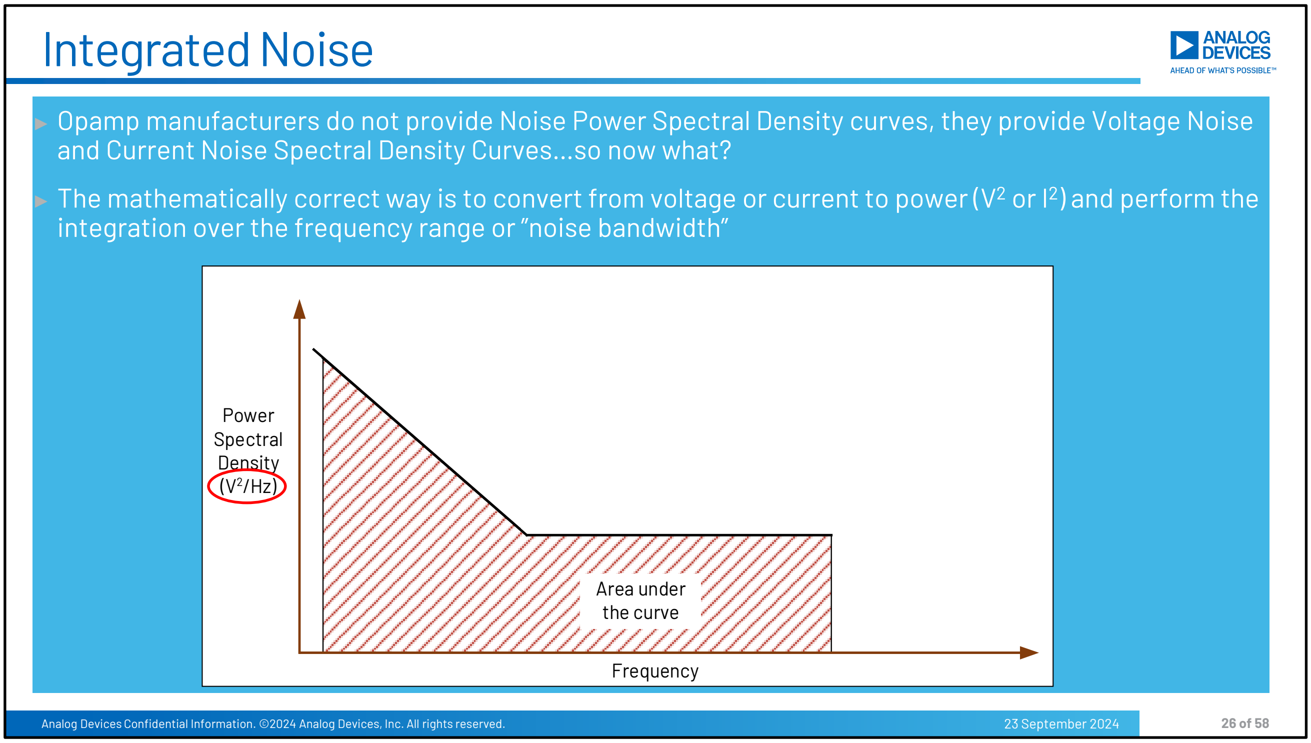


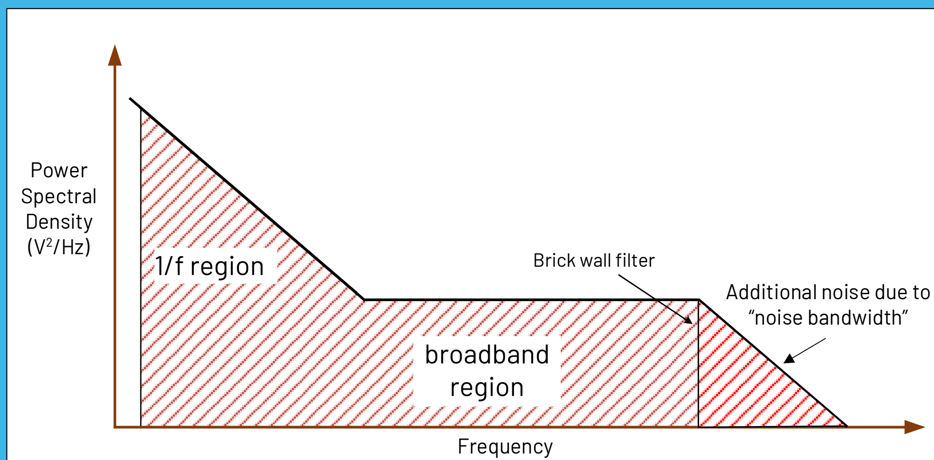
- Integrated noise, or total noise, is essentially the amount of "fuzz" you might see on an oscilloscope trace of a noisy signal.



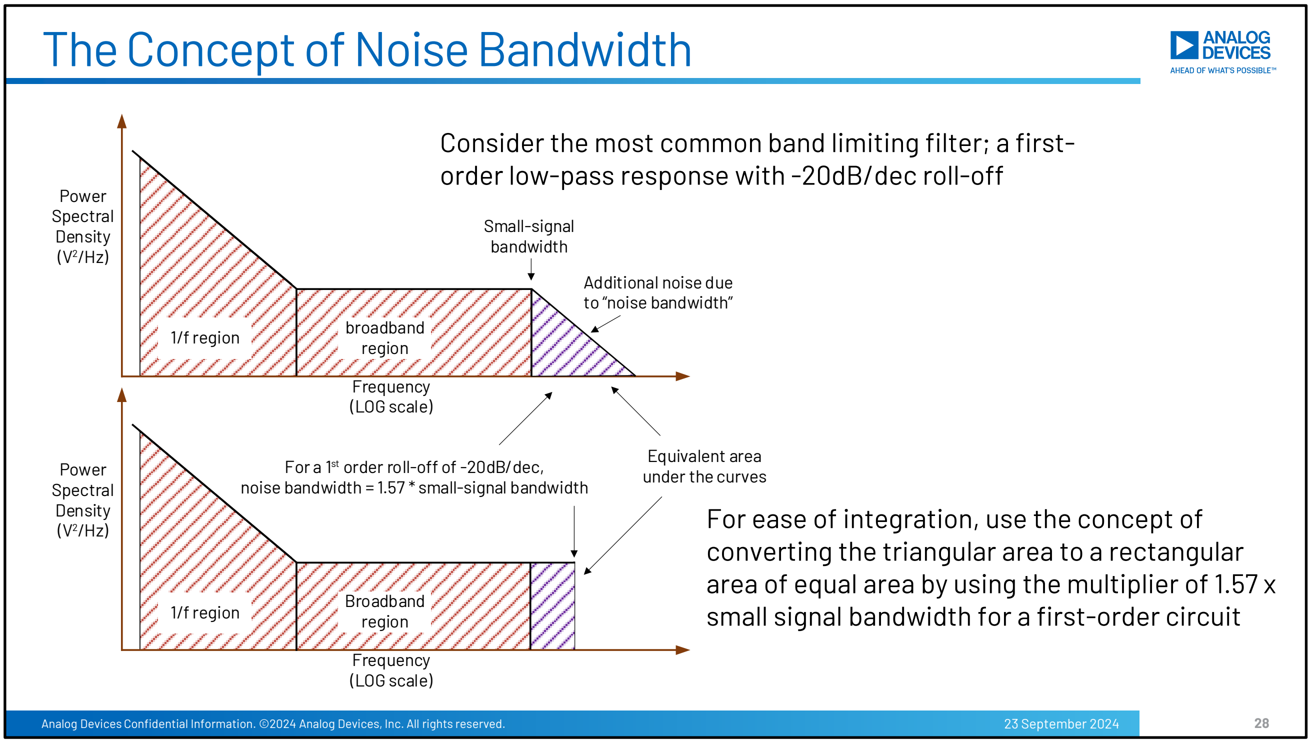
- Op amp datasheets don't give us the noise power spectral density curve. Instead, they give us the noise voltage or noise current spectral density curve which has units of V/sqrt(Hz) or A/sqrt(Hz).
- V^2/Hz is prop to W/Hz.
- The correct way is to convert from voltage or current to power (V^2 or I^2) and perform the integration.
- I don't know about you but I don't like the idea of doing a complex integration so there are a set of formulas to help you easily determine the total noise of your circuit from the graphs given in the datasheet.

The Concept of Noise Bandwidth

- Circuit Bandwidths in practice do not extend to infinity nor do we have ideal "brick wall filters".
- Circuit bandwidth is defined as the -3dB frequency where magnitude is -3dB down from its low frequency value.



- The small signal bandwidth of a system is the point where the amplitude reaches -3dB of its low frequency value…you may think of this small signal bandwidth as the "brick wall" filter of a circuit.
- But, in reality, this is not a brick wall filter and will have a slope. This additional area under the curve gives rise to the concept of Noise Bandwidth.



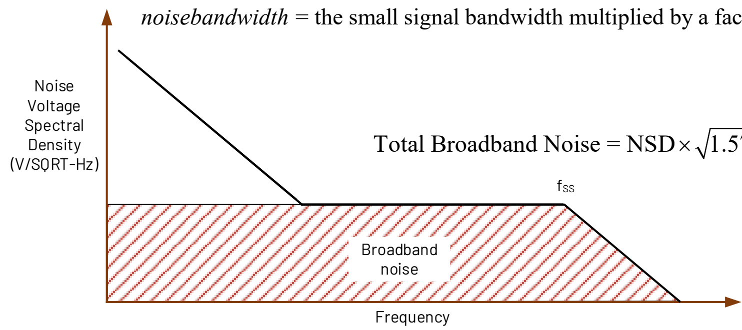
- The small signal bandwidth of a system is associated at the point where the amplitude reaches -3dB of its low frequency value…you can think of this small signal bandwidth as the "brick wall" filter of a circuit. But because the noise voltage is associated with the area under the curve the additional area due to having a non brick wall filter gives rise to the concept of noise bandwidth.
- Using this simple transformation from a triangular shape to a rectangular shape, allows us to estimate the total broadband noise voltage in a system by simply integrating over the rectangular shape, which is a very simple calculation

Computing the Total Broadband Noise

$$\text{Total Broadband Noise} = \text{NSD} \times \sqrt{noisebandwidth}$$

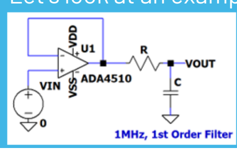
NSD = Voltage Noise Spectral Density given in units of $\left(\frac{V}{\sqrt{Hz}}\right)$ referred to the output

$noisebandwidth$ = the small signal bandwidth multiplied by a factor related to the roll-off of the gain vs frequency curve



$$\text{Total Broadband Noise} = \text{NSD} \times \sqrt{1.57 \times \text{small-signal bandwidth}}$$ (for a 1st order system)

- Let's look at an example with a -20dB/dec roll-off (a first order filter)



$NSD = 5\,nv/\sqrt{Hz}$

Small-signal bandwidth = 1MHz

$$\text{Total Broadband Noise} = 5x10^{-9} \times \sqrt{1.57 \times 1x10^{6}} = 6.3\mu V_{RMS} = 41.3\mu V_{PP}$$

- To compute the "total broadband noise" rather than convert noise voltage spectral density to noise power spectral density we can take the noise voltage spectral density (with units of Vrms/SQRT-Hz) and simply multiply by the SQRT of the noise bandwidth. The noise bandwidth is the small signal bandwidth multiplied by 1.57 (for a first order system). These units will result in Vrms, to convert to peak-to-peak multiply by 6.6.

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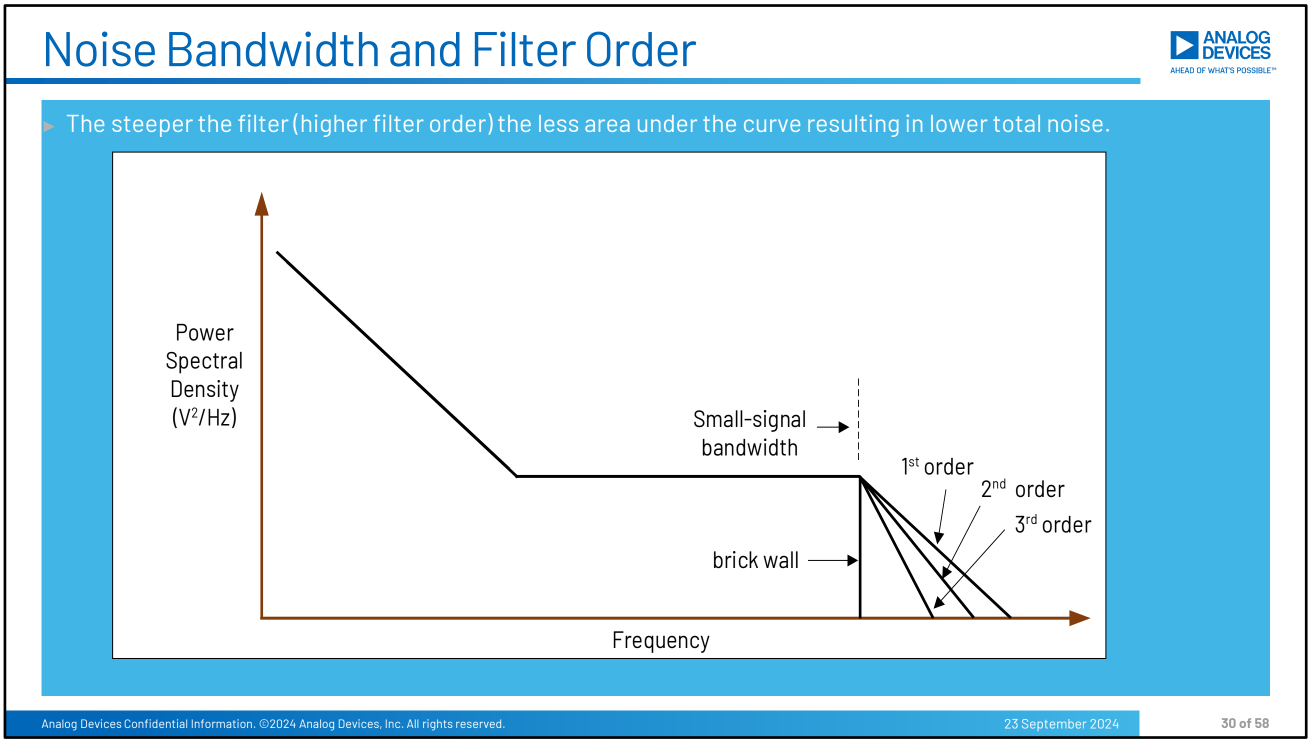

- The steeper the filter (higher filter order), the less area under the curve, resulting in lower total noise.

Noise Bandwidth

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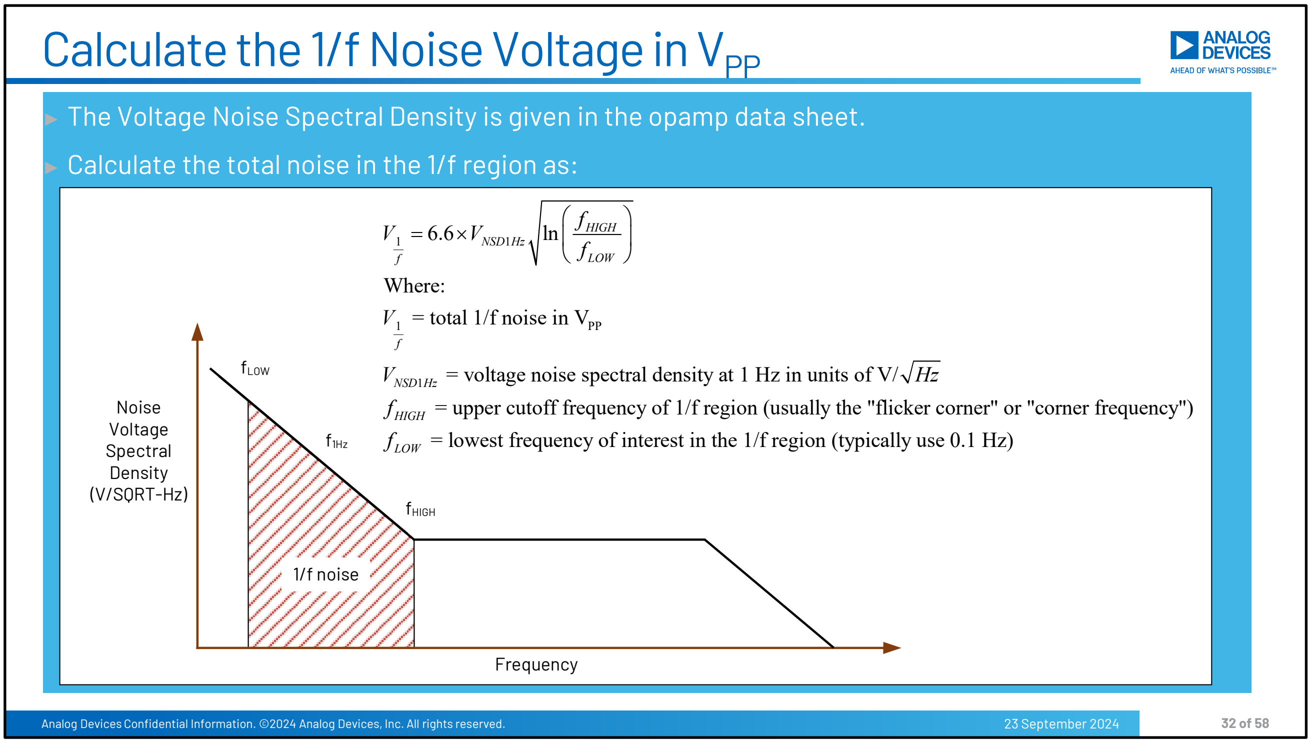
- The concept of "noise bandwidth" accounts for the additional area under the noise power spectral density curve
- The steeper the filter, the closer to an ideal "brick wall filter" a system will behave
- Conveniently if the order of the LPF is known, a scaling factor can be used to relate the small-signal bandwidth and noise bandwidth of the system

$$\text{noise bandwidth} = \text{small signal bandwidth} \times \text{noise bandwidth ratio}$$

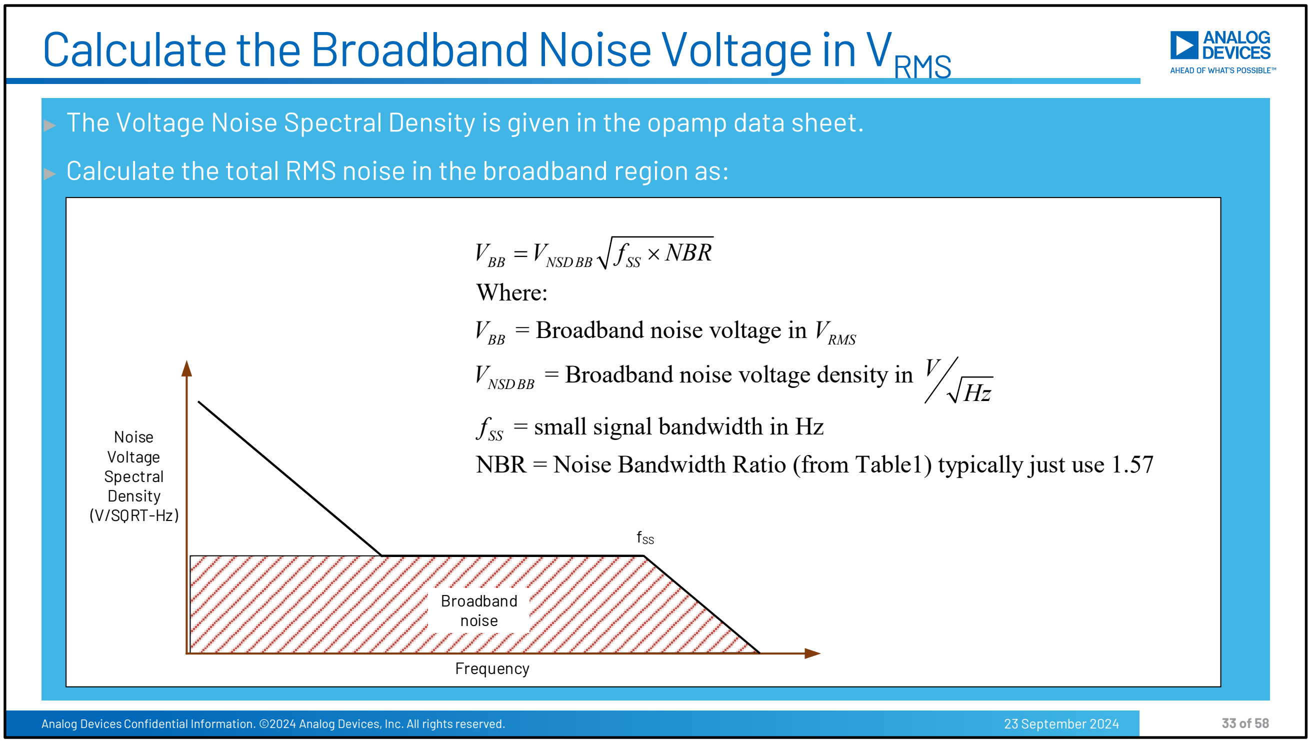
Number of Poles in Filter	Noise Bandwidth Ratio
1	1.57
2	1.22
3	1.16
4	1.13
5	1.12

Table 1

- The concept of "noise bandwidth" accounts for the additional area under the noise power spectral density curve
- The steeper the filter, the closer to an ideal "brick wall filter" a system will behave
- Conveniently if the order of the LPF is known, a scaling factor can be used to relate the small-signal bandwidth and noise bandwidth of the system



- The Voltage Noise Spectral Density is given in the opamp data sheet.
- Calculate the total RMS noise in the 1/f region as per the equation shown.

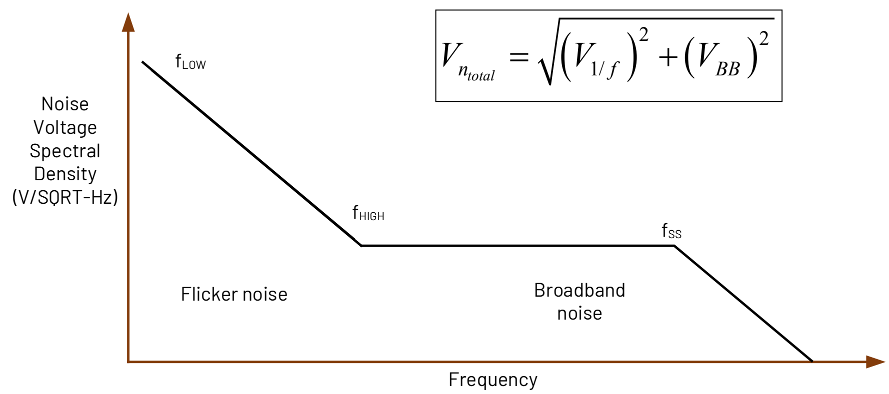


- Does anyone see a problem here?
- The amount of "double counting" is completely negligible in most practical examples of noise analysis.
- Therefore, when analyzing broadband noise, you only need to consider the upper frequency when determining the noise bandwidth of a system or circuit.
- Notice the units calculated here are in Vrms whereas the units for the 1/f region are typically given in Vpp...this is just a convention. Remember to convert to the same units when combining different noise sources.
  - To convert from Vrms to Vpp multiply by 6.6.
  - To convert from Vpp to Vrms divide by 6.6.

Computing the Total Noise

- To compute the total RMS noise in a system, add the 1/f noise to the broadband noise by the Root means squares method.
- Always make sure you have the same units for each term in the equation.

$$V_{n_{total}} = \sqrt{\left(V_{1/f}\right)^2 + \left(V_{BB}\right)^2}$$



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- To determine the total noise, you must add the 1/f noise total with the broadband noise total as the square root of the sum of the squares.

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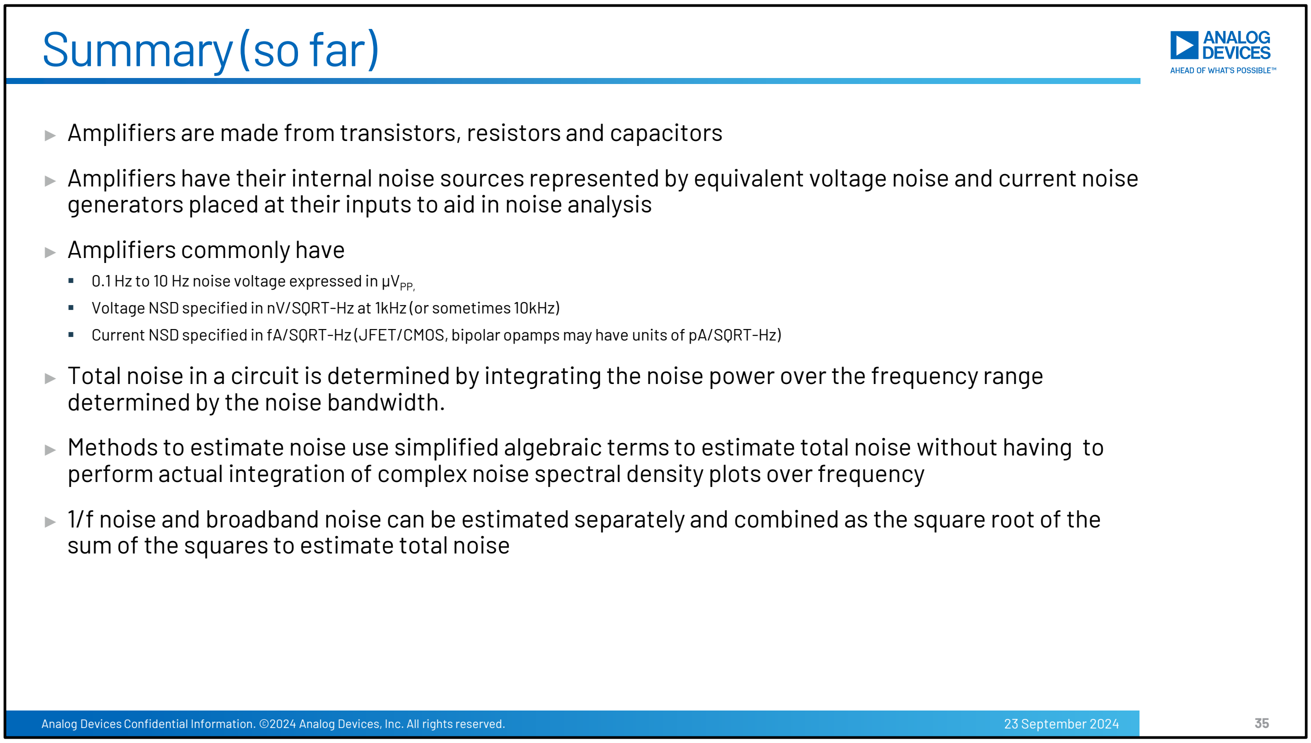

- Amplifiers are made from transistors, resistors and capacitors
- Amplifiers have their internal noise sources represented by equivalent voltage noise and current noise generators placed at their inputs to aid in noise analysis
- Amplifiers commonly have
  - 0.1 Hz to 10 Hz noise voltage expressed in $\mu V_{PP,}$
  - Voltage NSD specified in nV/SQRT-Hz at 1kHz (or sometimes 10kHz)
  - Current NSD specified in fA/SQRT-Hz (JFET/CMOS, bipolar opamps may have units of pA/SQRT-Hz)
- Total noise in a circuit is determined by integrating the noise power over the frequency range determined by the noise bandwidth.
- Methods to estimate noise use simplified algebraic terms to estimate total noise without having to perform actual integration of complex noise spectral density plots over frequency
- 1/f noise and broadband noise can be estimated separately and combined as the square root of the sum of the squares to estimate total noise

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Estimating Noise in Amplifier Circuits

Estimating Noise in Amplifier Circuits

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- Estimating noise in an amplifier circuit can be a daunting task...there is no doubt attention to detail is warranted...but breaking the process down into a several simple steps will ease the process
  1) Determine all the sources of noise in the circuit
  2) Refer each noise source to the output (RTO)
  3) Combine all output referred noise sources into a single noise value RTO (because noise sources are random and uncorrelated their noise terms add as square root of the sum of the squares)

$$V_{n_{total}} = \sqrt{\left(V_{n_1}\right)^2 + \left(V_{n_2}\right)^2 \ldots + \left(V_{n_n}\right)^2}$$

  4) Convert from units of V/SQRT-Hz to Vrms or Vpp by integrating over the noise bandwidth
  5) Refer the combined RTO noise value back to the input if interested to know RTI noise based upon the noise gain of the amplifier

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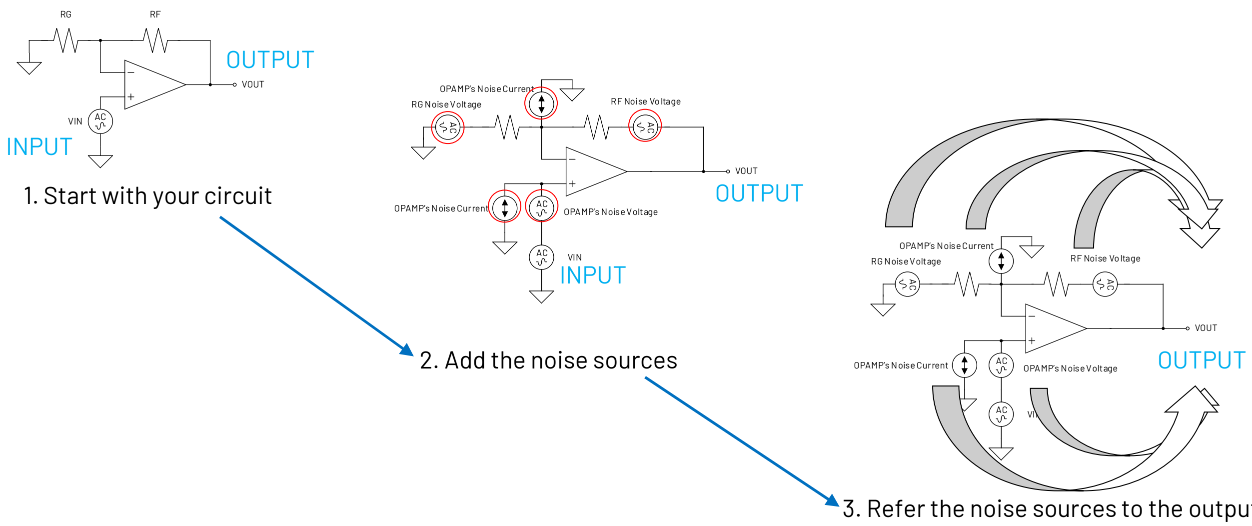
Example for a Simple Non-Inverting Amplifier



Hand noise analysis will quickly identify which noise sources to focus on reducing, for the lowest noise circuit.

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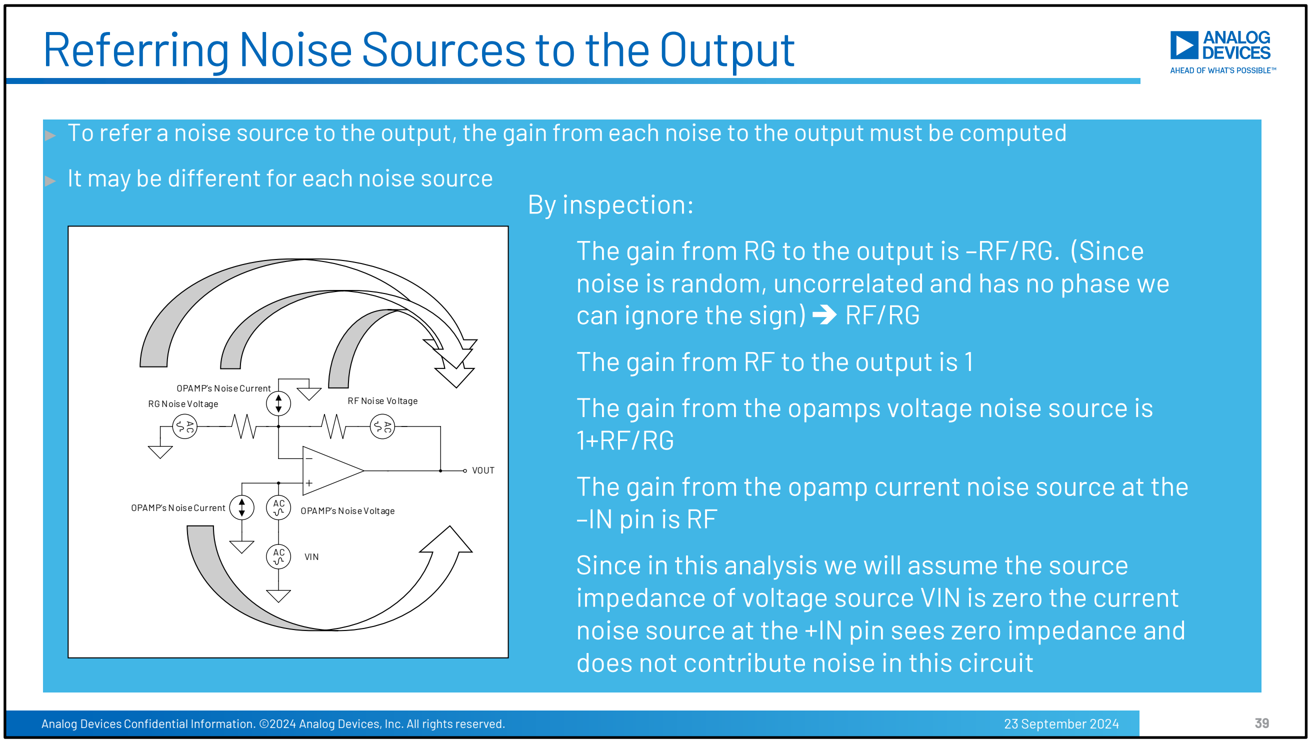
- Hand analysis will help you understand what noise source to swap out if you're trying to reduce noise in your circuit.



- To refer a noise source to the output, the gain from each noise to the output must be computed
- It may be different for each noise source

- By inspection:

- The gain from RG to the output is –RF/RG. (Since noise is random, uncorrelated and has no phase we can ignore the sign) ➔ RF/RG

- The gain from RF to the output is 1

- The gain from the opamps voltage noise source is 1+RF/RG

- The gain from the opamp current noise source at the –IN pin is RF

- Since in this analysis we will assume the source impedance of voltage source VIN is zero the current noise source at the +IN pin sees zero impedance and does not contribute noise in this circuit

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Referring Noise Sources to the Output

The RTO noise from R_G is given by:

$$V_{n\_RG} \times gain = V_{n\_RG} \times \frac{R_F}{R_G}$$

The RTO noise from R_F is given by:

$$V_{n\_RF} \times gain = V_{n\_RF} \times 1$$

The RTO noise from the opamp voltage noise is given by:

$$V_{n\_OPAMP\_VOLTAGE} \times gain = V_{n\_OPAMP\_VOLTAGE} \times \left(1 + \frac{R_F}{R_G}\right)$$

The RTO noise from the opamp current noise is given by:

$$I_{n\_OPAMP\_CURRENT} \times gain = I_{n\_OPAMP\_CURRENT} \times R_F$$

Summing all the individual RTO noise sources:

$$V_{n_{RTO}} = \sqrt{\left(V_{n\_RG} \times \frac{R_F}{R_G}\right)^2 + \left(V_{n\_RF}\right)^2 + \left(V_{n\_OPAMP\_VOLTAGE} \times \left(1 + \frac{R_F}{R_G}\right)\right)^2 + \left(V_{n\_OPAMP\_CURRENT} \times R_F\right)^2}$$

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- Summarizing the analysis from the previous slide we can write the following equations for our example circuit.

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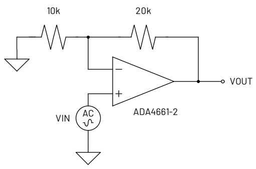
Let's Add Some Actual Values to Our Circuit



Let's determine the noise voltage from each resistor using a simple rule of thumb where:

$$1k = 4nV/\sqrt{Hz} \rightarrow \begin{array}{l} 10k = 4nV/\sqrt{Hz} \times \sqrt{\frac{10k}{1k}} = 4nV/\sqrt{Hz} \times \sqrt{10} = 12.65\,nV/\sqrt{Hz} \\ 20k = 4nV/\sqrt{Hz} \times \sqrt{\frac{20k}{1k}} = 4nV/\sqrt{Hz} \times \sqrt{20} = 17.89\,nV/\sqrt{Hz} \end{array}$$

From the ADA4661-2 data sheet the noise is given as:

NOISE PERFORMANCE				
Total Harmonic Distortion Plus Noise	THD + N	$A_V = 1$, $V_{IN} = 0.44$ V rms at 1 kHz		
Bandwidth = 80 kHz			0.002	%
Bandwidth = 500 kHz			0.003	%
Peak-to-Peak Noise	e_n p-p	f = 0.1 Hz to 10 Hz	3	µV p-p
Voltage Noise Density	e_n	f = 1 kHz	18	nV/√Hz
		f = 10 kHz	14	nV/√Hz
Current Noise Density	i_n	f = 1 kHz	360	fA/√Hz

$$\rightarrow \begin{array}{l} \text{ADA4661-2 Voltage Noise} = 14\,nV/\sqrt{Hz} \\ \text{ADA4661-2 Current Noise} = 360\,fA/\sqrt{Hz} \end{array}$$

Summing all the individual RTO noise sources:

$$V_{n_{RTO}} = \sqrt{\left(12.65e^{-9} \times \frac{20k}{10k}\right)^2 + \left(17.89e^{-9}\right)^2 + \left(14e^{-9} \times \left(1 + \frac{20k}{10k}\right)\right)^2 + \left(360e^{-15} \times 20k\right)^2}$$

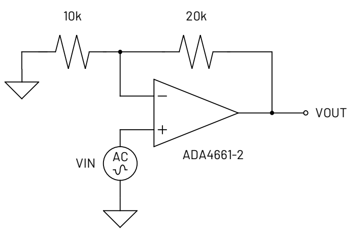
$$V_{n_{RTO}} = \sqrt{\left(25.3e^{-9}\right)^2 + \left(17.89e^{-9}\right)^2 + \left(42e^{-9}\right)^2 + \left(7.2e^{-9}\right)^2} = 52.68\,nV/\sqrt{Hz}$$

To compute the RTI noise voltage divide by the signal gain:

$$V_{n_{RTI}} = \frac{52.68e^{-9}}{3} = 17.56\,nV/\sqrt{Hz}$$

- Now, time for an example

Let's Estimate the Total Broadband Noise in the Circuit



The unity gain frequency of the ADA4661 is 4MHz.

In the gain of 3 shown, the small signal bandwidth is 1.33MHz (this is that "brick wall" filter value).

Since there are no additional filters in the circuit the roll off is only from the opamp and is first order.

Compute the noise bandwidth as 1.33MHz*1.57 = 2.1MHz.

Using the estimated RTI noise voltage of 17.56nV/SQRT-Hz, a noise gain = 3V/V and 2.1MHz noise bandwidth we estimate the broadband RMS noise to be:

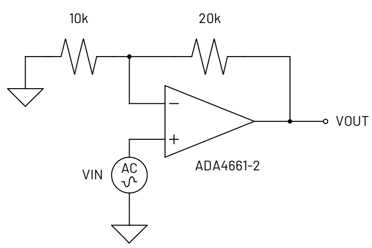
$$V_{NOISE\_TOTAL} = e_{BB} \times noisegain \times \sqrt{noisebandwidth}$$

$$V_{NOISE\_TOTAL} = \boxed{17.56e^{-9}} \times 3 \times \sqrt{2.1e^{6}} = 76\mu V_{RMS}$$

RTO: RMS Broadband Total Noise Calculation

Let's Estimate the Total 1/f Noise in the Circuit

From the data sheet curve, we can estimate the 1Hz 1/f noise at 300nV/SQRT-Hz



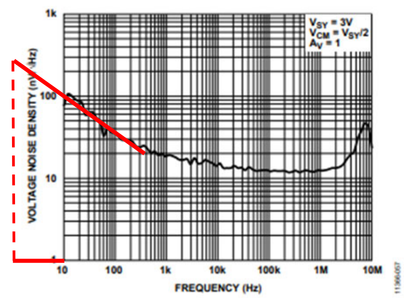


Figure 56. Voltage Noise Density vs. Frequency

We can also estimate the 1/f corner frequency of 10kHz

Using:

$$V_{\frac{1}{f}} = e_{1Hz}\sqrt{\ln\left(\frac{f_{HIGH}}{f_{LOW}}\right)} \times noisegain$$

We can estimate the **1/f noise, RTO**, as 300e-9*SQRT(LN(10e3/1)) *3 = **2.7µV$_{RMS}$**

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- In the case of the ADA4661 the NSD plot does not extend as low as 1 Hz…we can simply estimate the noise at 1 Hz by extending the curve as indicated.

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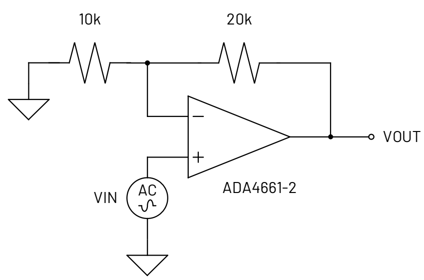
Let's Estimate the RTO Total Noise in the Circuit



Combining the 1/f noise and the broadband noise we estimate the total noise as:

$$V_{n_{total}} = \sqrt{\underbrace{(76e^{-6})^2}_{\text{Broadband}} + \underbrace{(2.7e^{-6})^2}_{\text{1/f}}} = 76\mu V_{RMS}$$

And converting to peak-to-peak (rms X 6) we estimate 456μV_{PP}.

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- Because noise adds as the square root of the sum of the squares you can see that in this example the total noise is dominated by the broadband noise. In fact, we could have completely ignored the 1/f noise calculation and ended up with the same result. This is due to the large noise bandwidth of this example.
- For circuits that are bandlimited to much lower frequencies, ignoring the 1/f noise is not recommended.

Summary (so far)

- To analyze noise, identify all the sources of noise, refer them to the output and combine them appropriately to estimate the RTO NSD
- Determine the RTI NSD by referring the RTO noise to the circuit input. This requires determination of the noise gain.
- The noise gain is generally indicating the gain from the non-inverting terminal of an amplifier to the output but could be whatever you determine you want to use as your input. You simply must be consistent within your analysis
- Determine the noise bandwidth of your circuit
- Multiply RTI NSD by Noise gain by the square root of the noise bandwidth to estimate rms noise at the output
- Multiply by 6 or 6.6 to estimate the peak-to-peak noise voltage
- Understanding the dominate noise sources and the nature of the noise gain and the nature of the noise bandwidth can help understand how to optimize the circuits noise performance

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Simulating Noise in Amplifier Circuits

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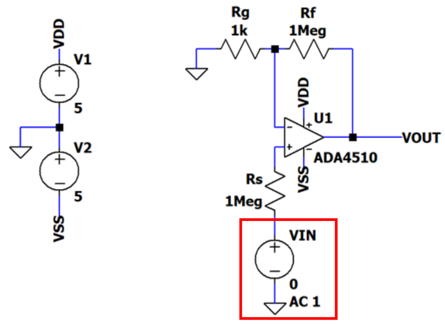
Simulating Noise in LTSpice

- Add the .noise command as a SPICE directive with the listed arguments:
  - Output: The **node** where you want to examine the ***output referred noise***
  - Input: The **source** where you want to examine the ***input referred noise***
  - Type of sweep: octave, decade, linear or list
  - Number of points: per octave, per decade, etc
  - Start Frequency: Lowest frequency in the sweep in Hz
  - Stop Frequency: Highest frequency in the sweep in Hz



SPICE directive | Input source | Points/dec

.noise V(VOUT) VIN dec 101 1 10MEG

Output node | Decade sweep | Start frequency | Stop frequency

Must have an AC noise source somewhere in the circuit for Noise Analysis to run.

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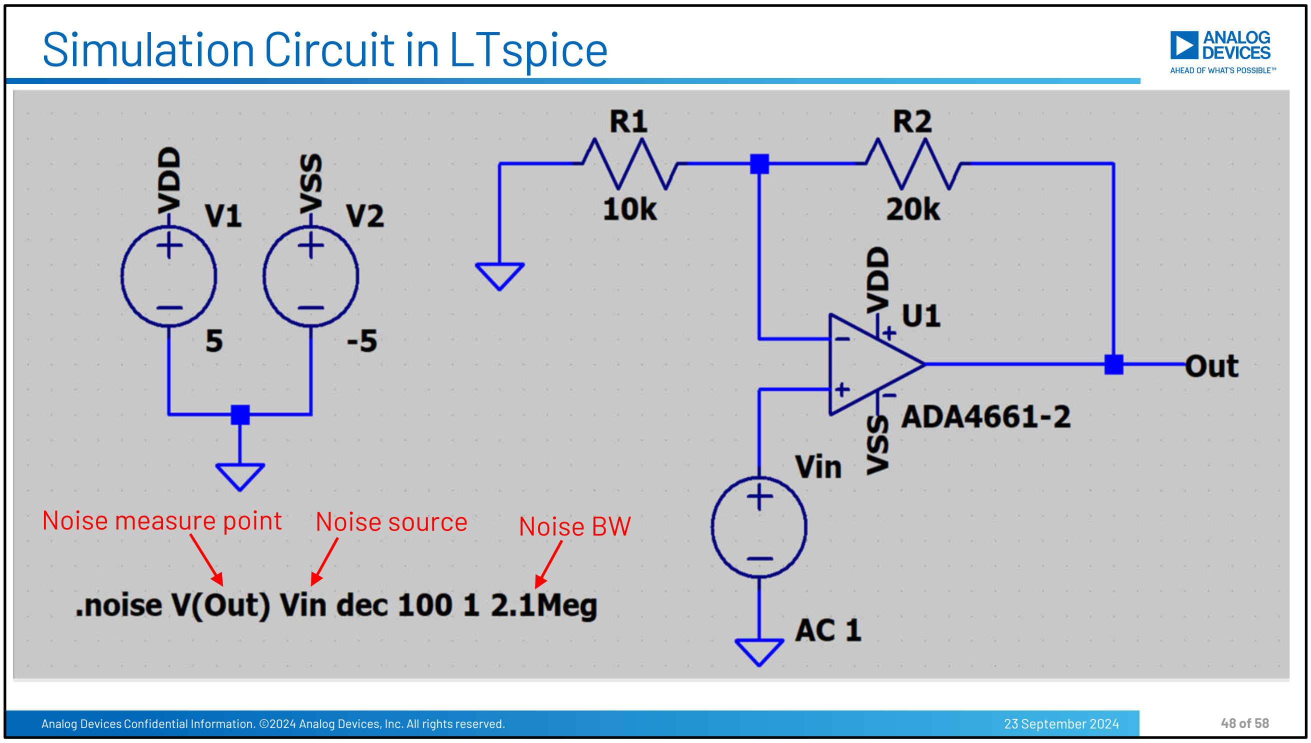
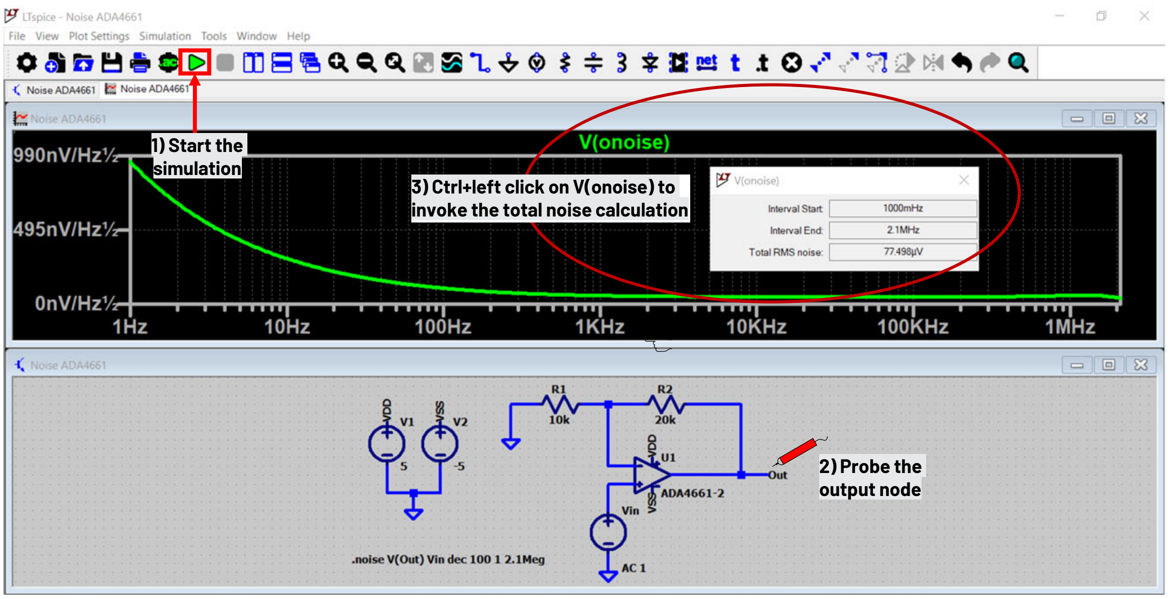
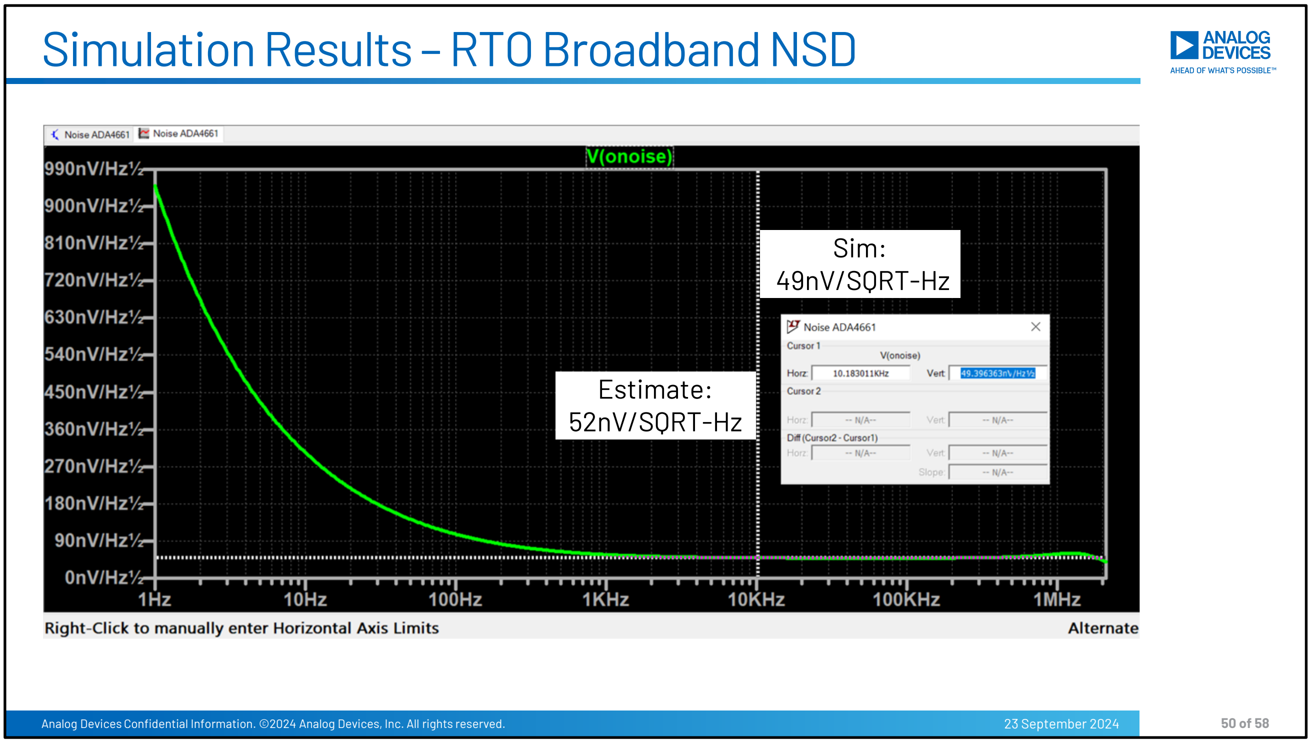
- You need to have a source of noise – here we used a voltage source.



- Simply draw your schematic as you normally would in LTspice…you do not need to add any additional noise voltage or noise current sources to your circuit. LTspice will model the amplifier noise and noise from resistors automatically. Add the .noise statement to invoke a noise simulation.
- You must identify the output node for the circuit and place an AC source at the location you want to determine the RTI noise. In this example the output of the circuit is labeled Out and is described in the .noise command line as V(Out) and we decide to place a voltage source at the opamp non-inverting input to tell LTspice this is the node to where we want to refer our RTI noise. This is indicated by Vin included in the .noise command. We also want to specify to perform a decade analysis with 100 points per decade from 1 Hz to 2.1MHz.
- I've limited the sweep to the noise BW which we calculated by hand – matches almost perfectly (or you could also go up to 10Meg or and use an RC filter to cut it off & control the BW).

Simulating Noise in LTSpice





- After running the noise analysis probe the output node to see the NSD for the RTO noise.
- Left click on V(onoise) and place a cursor over any area in the broadband region.
- Compare to the hand analysis...49 nV/SQRT-Hz simulated vs 52 nV/SQRT-Hz estimated by hand.

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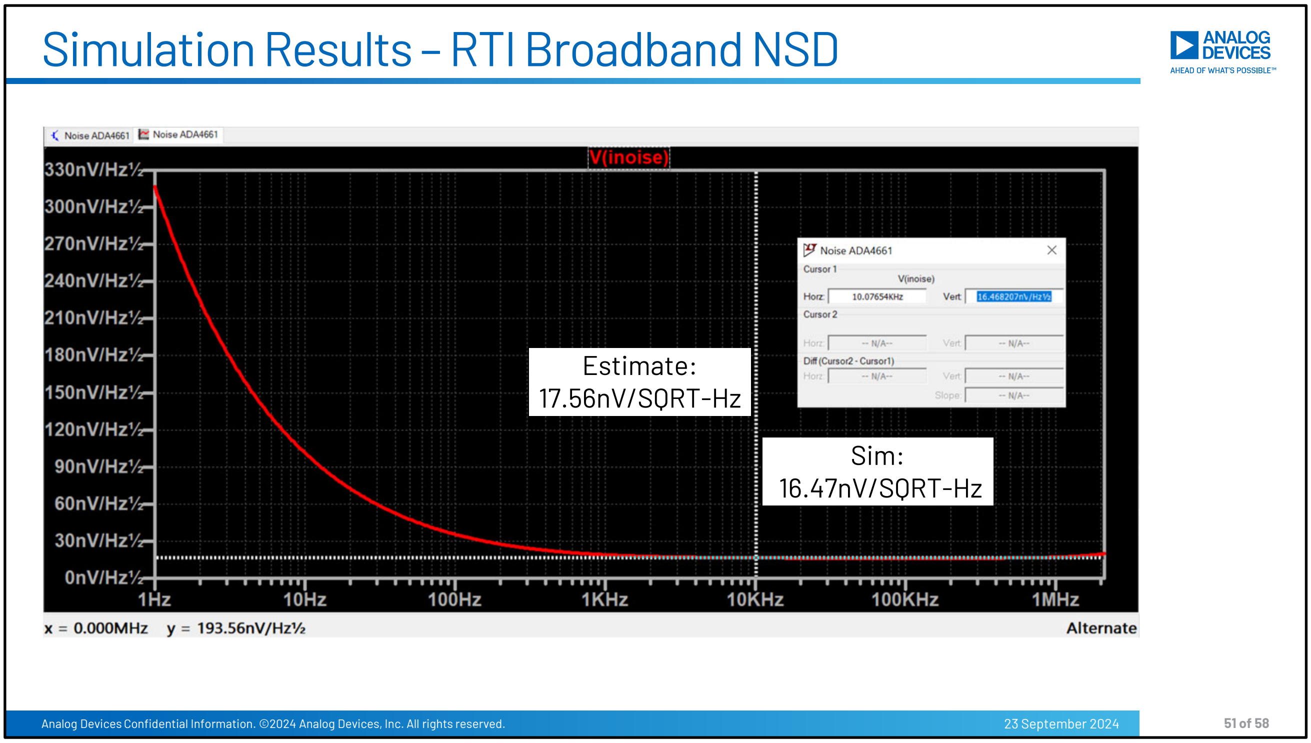
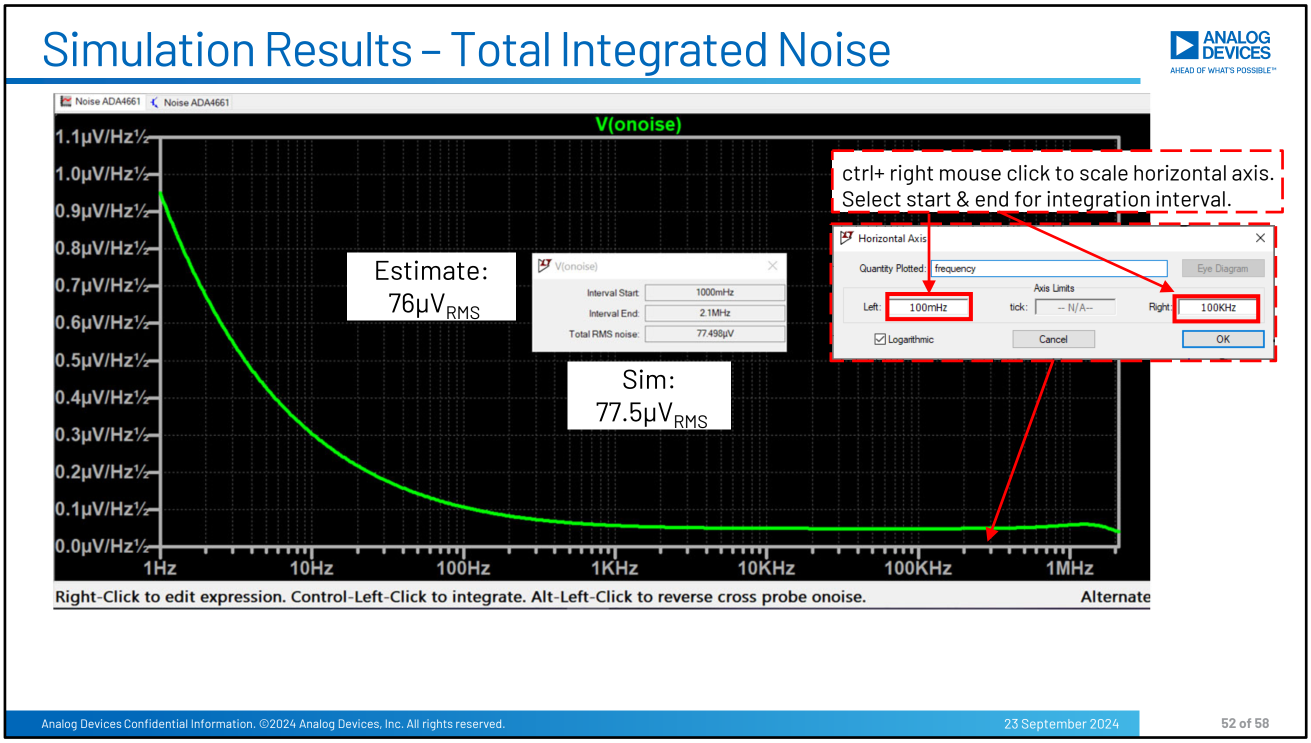


- Right click on V(onoise) and edit the text to say V(inoise) to see the RTI NSD.
- Left click on V(inoise) and a cursor should pop up.
- Compare to the hand analysis…16.47nV/SQRT-Hz simulated vs 17.56nV/SQRT-Hz estimated by hand.

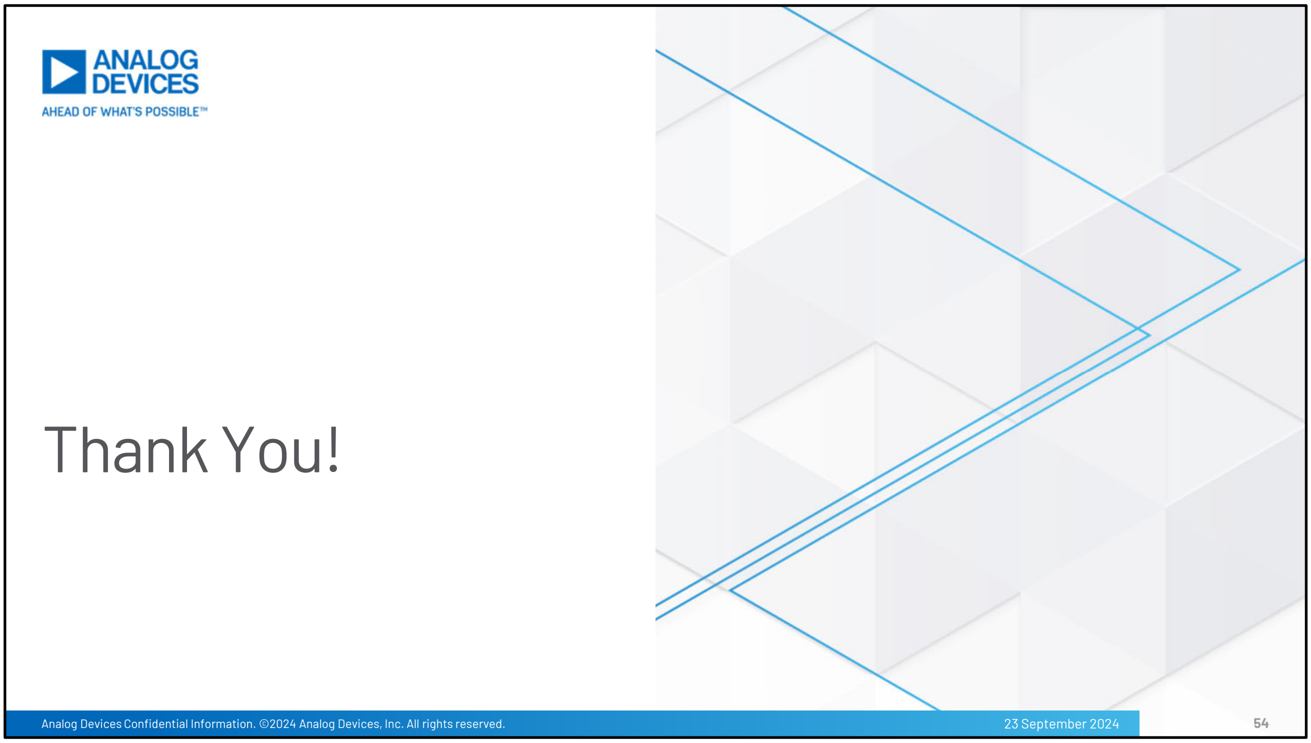


- Go back to V(onoise).
- Hold down the CTRL button and click on V(onoise) in the plot window to bring up the total integrated noise pop-up window. Bandwidth can be selected by first clicking on the X-axis settings and restricting the max frequency to the bandwidth of interest. Compare the simulated total noise to the estimated total noise over the noise bandwidth…Simulated 77.5uVRMS vs 76uVRMS estimated by hand.
- BW here is limited by the sim settings to 2.1MHz, since that's our noise BW.

Additional Reading and Resources

- 11 Myths About Analog Noise
- Noise Analysis of Precision Data Acquisition Signal Chain
- LTSpice Tutorial for AC & Noise Analysis (Video) | Analog Devices
- LTSpice: Noise Simulations
- LTSpice: Integrating Noise Over a Bandwidth | Analog Devices
- Step-by-Step Noise Analysis Guide for Your Signal Chain
- Low Frequency Noise Analysis for Sensor Signal Chains
- Noise Analysis in Precision Analog Designs
- Analysis of Input Current Noise with Even Harmonics Folding Effect in a Chopper Op Amp
- Practical Input-Referred Calculations in Precision Systems | Analog Devices
- Signal Chain Noise Calculator | Precision Studio | Analog Devices
- Low Noise Amplifier Selection Guide for Optimal Noise Performance | Analog Devices
- Understanding and Eliminating 1/f Noise | Analog Devices

- Art Kay, Operational Amplifier Noise Techniques and Tips for Analyzing and Reducing Noise, Boston, Newnes/Elsevier, 2012
- Paul R. Gray, Robert G. Meyer, Analysis and Design of Analog Integrated Circuits, Third Edition, New York, John Wiley & Sons, Inc., 1993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Appendix A: Terms and Definitions

Terms and Definitions

- **RTO** – Acronym meaning Referred to Output.
- **RTI** – Acronym meaning Referred to Input.
- **Noise Power Spectral Density** – Expressed in (W/Hz) is the square of either voltage noise spectral density or current noise spectral density.
- **Voltage Noise Spectral Density** - Voltage noise spectral density is a measurement of noise voltage per square-root hertz often expressed with units of nV_{RMS}/SQRT-Hz.
- **Current Noise Spectral Density** - Current noise spectral density is a measurement of noise current per square-root hertz often expressed with units of fA_{RMS}/SQRT-Hz.
- **Integrated Noise** – Also sometimes called total noise, can be thought of as the amount of "fuzz" you would see on a perfect, noiseless oscilloscope over a given measurement bandwidth. Integrated noise can be expressed in units RMS or peak-to-peak. It is called "integrated noise" as it is the area under the noise power spectral density curve when integrated over the frequency range of interest.
- **Noise Bandwidth** – Is different than the signal bandwidth and is a concept used to relate the area under the noise power spectral density curve as a function of the steepness of the roll-off at high frequency.

- Here are some common terms and phrases associated with the topic of noise in amplifiers. These terms will be used throughout the presentation today and an understanding of these terms will help you achieve the maximum performance from your designs.
- RMS is the power expressed as a signal.
- Peak to peak is literally the peak to peak.

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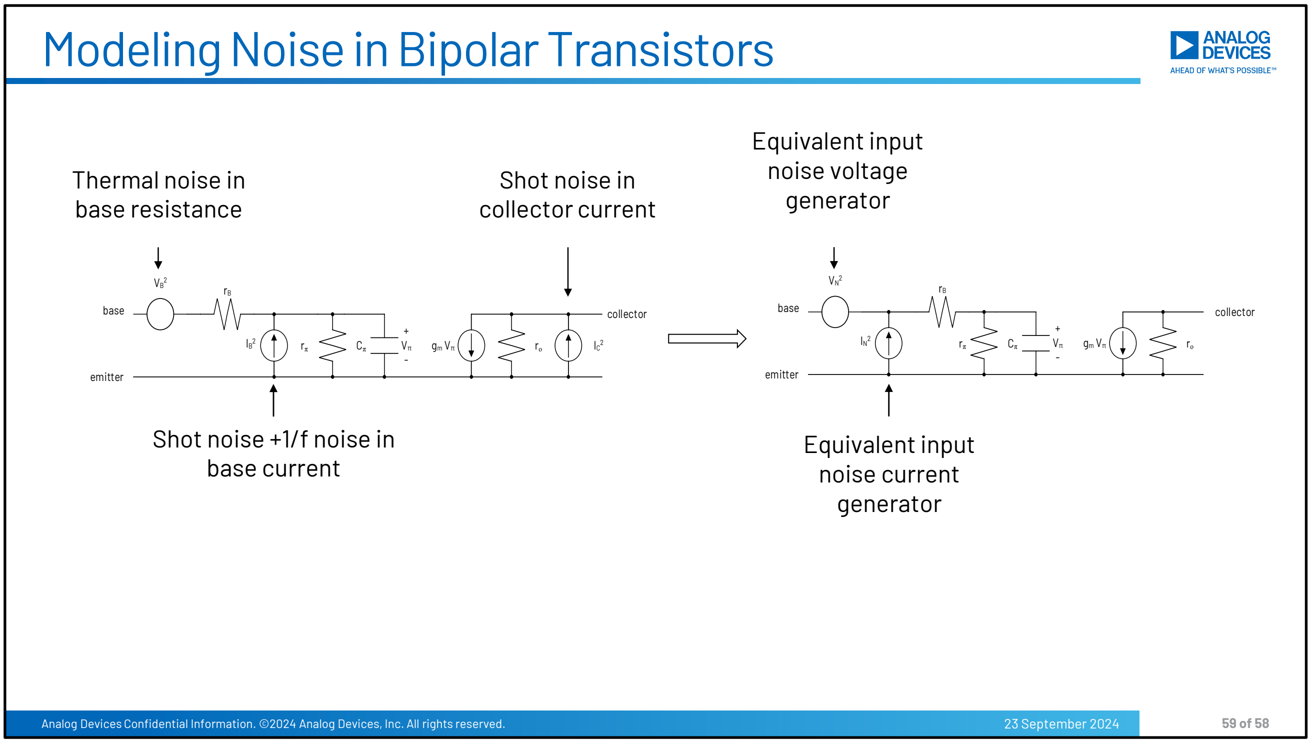
Terms and Definitions

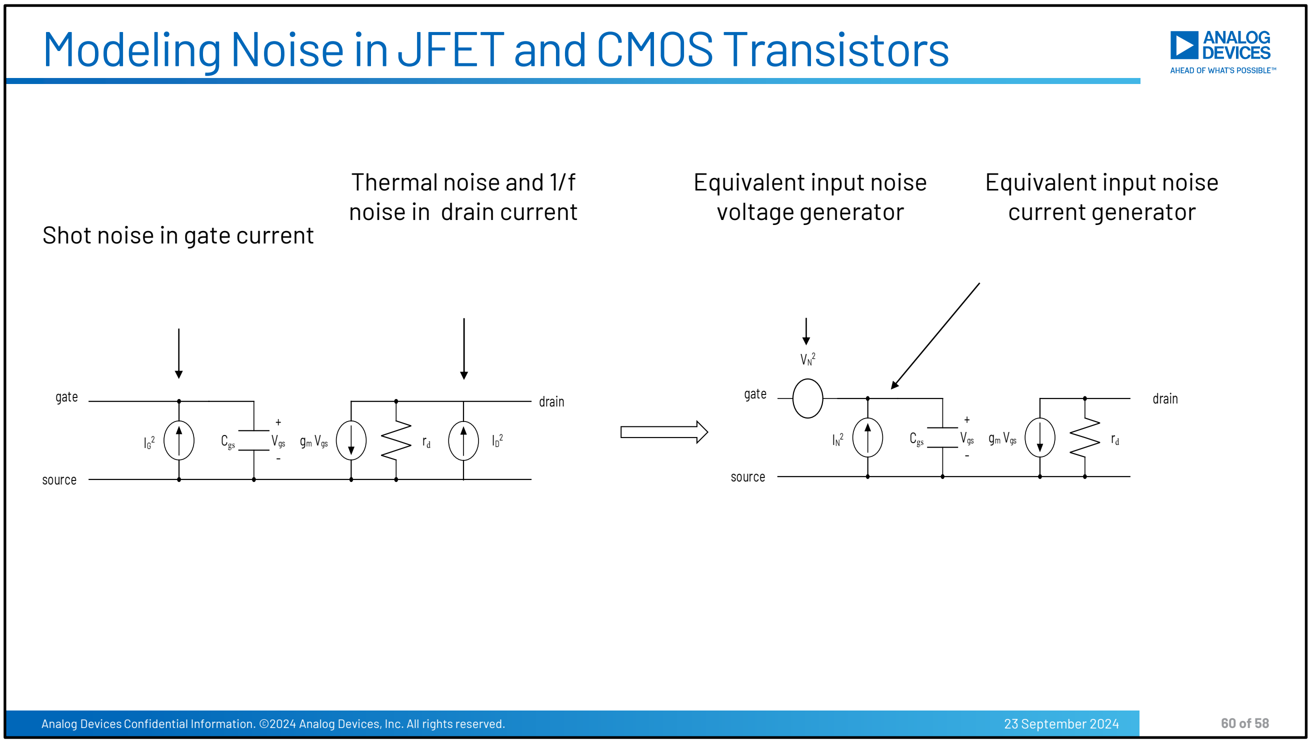
- **Noise Gain** – Can be different than the signal gain. Noise Gain throughout this presentation refers to the gain from the opamp's equivalent input noise voltage source. The gain will be the same gain as for the non-inverting opamp configuration, i.e. $1+Z_F/Z_G$
- **Equivalent Input Noise Voltage** – a voltage source with units of V/SQRT-Hz placed in series with IN+ terminal. This equivalent voltage source represents all of the noise sources internal to the opamp reflected to the input.
- **Equivalent Input Noise Current** – a current source with units of A/SQRT-Hz placed at each input terminal of an opamp and ground. This equivalent current source represents the noise associated with the input bias current and leakage current of an amplifiers input stage.
- **NSD –** Acronym meaning **N**oise **S**pectral **D**ensity. Describes the noise parameter as a function of frequency.
- **Corner Frequency** – The frequency at which the 1/f voltage noise will intercept the broadband voltage noise when using straight line approximations on the Voltage Noise Spectral Density vs Frequency Curve.
- **Flicker Corner** – Same as Corner Frequency.

- Here are some common terms and phrases associated with the topic of noise in amplifiers. These terms will be used throughout the presentation today and an understanding of these terms will help you achieve the maximum performance from your designs.

Appendix B: Modelling Noise in Transistors



- Bipolar transistors have shot noise associated with their DC collector current, shot noise and 1/f noise associated with their base current and thermal noise associated with the physical resistance in the base region.
- All these sources can be combined into simplified input referred equivalent noise voltage and noise current generators. It is worth noting that higher collector current results in lower input referred noise voltage.
- This is since the transconductance of the transistor is directly proportional to the DC collector current and the shot noise in the collector is proportional to the square root of the DC collector current. Higher quiescent current amplifiers tend to have lower noise voltage.
- Shot noise dominated for BJT.



- JFET and CMOS transistors have thermal noise and 1/f noise associated with their DC drain current, shot noise associated with their gate current. All these sources can be combined into simplified input referred equivalent noise voltage and noise current generators. It is worth noting that higher drain current results in lower input referred noise voltage.
- Thermal noise dominated for JFET, CMOS.

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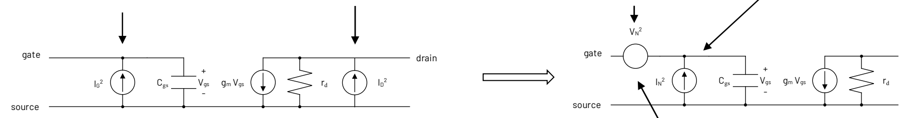
Modeling Noise in JFET and CMOS Transistors

Shot noise in gate current

Thermal noise and 1/f noise in drain current

Equivalent input noise voltage generator

Equivalent input noise current generator



$i_g^2 = 2qI_G\Delta f$

Where:

q = electronic charge, 1.602 x 10^{-19} Coulombs

I_G = gate leakage current

Δf = Noise bandwidth in Hertz

$$i_d^2 = 4kT\left(\frac{2}{3}g_m\right)\Delta f + K\frac{I_D^{\,a}}{f}\Delta f$$

Where:

k = Boltzmann's constant, 1.38 x 10^{-23} $\frac{\text{W-s}}{K}$

T = Temperature °K

g_m = the transcoductance of the device, $\frac{\partial I_D}{\partial V_{GS}}$

Δf = Noise bandwidth in Hertz

K = a constant for a given device and process technology

I_D = drain bias current

a = a constant between 0.5 and 2

f = frequency in Hz

$i_n^2 = 2qI_G\Delta f$

$$v_i^2 = 4kT\left(\frac{2}{3}\frac{1}{g_m}\right)\Delta f + K\frac{I_D^{\,a}}{g_m^{\,2}f}\Delta f$$

$$\mathrm{Re}q = \frac{2}{3}\frac{1}{g_m} + K'\frac{I_D^{\,a}}{g_m^{\,2}f}$$

Where;

$$K' = \frac{K}{4kT}$$

- Here is a closer look at the equations behind the scenes for a CMOS or JFET transistor