NOISE & GROUNDING

OUTLINE:

- Noise Definition
- External Noise
  - Electromagnetic
  - Solutions
- Conducted Noise and Grounding
  - Issues
  - Solutions
- Environmental Noise
- Intrinsic Noise
  - Noise Types
    - Gaussian
    - Johnson
    - Shot
    - Flicker (1/F)
    - Quantization
  - Component Noise Calculations
  - Amplifier Noise Calculations

NOISE & GROUNDING LECTURE

I. Noise

A. Definition/Description

1. Noise = Anything other than desired signal
2. Described by:

$$SNR = \frac{Signal}{Noise} \cdot 100\%$$

or

$$SNR(dB) = 10 \cdot \log_{10} \frac{(Vsignal)^2}{(Vnoise)^2} = 20 \cdot \log_{10} \frac{(Vsignal)}{(Vnoise)}$$

B. Types of Noise

1. External, Conducted, Environmental, and Intrinsic
2. Different types of noise have different time and frequency domain characteristics and may be random or periodic.

EXTERNAL NOISE

II. Electromagnetic Noise – Near-field and Far-field characteristics

A. Electromagnetic Radiation described by Maxwell's Equations:

$$\nabla \times H = \varepsilon_0 \frac{\partial E}{\partial t} \qquad \nabla \times E = -\mu_0 \frac{\partial H}{\partial t}$$

$$\nabla \cdot E = 0 \qquad \nabla \cdot H = 0$$

1. Electromagnetic radiation has two time-varying components, the E-field (electric) and H-field (magnetic). Note: $B = \mu H$
2. The time-varying E-field and H-field cannot exist independently.

B. Far-field characteristics in free space

1. Far-field is defined as a distance greater than $\lambda/2\pi$ from the emission source, where λ is the wavelength of the electromagnetic wave in meters.

$$\lambda = \frac{c}{f}$$

Where:

c = speed of light = 3×10^8 (m/s)
f = frequency (Hz)

2. Examples of far-field noise sources include, among others, radio transmissions, lightning (non-conducted), ionospheric noise, distant high-voltage arcing, etc.
3. Far-field radiation is in the form of a uniform plane wave in free space and the ratio of E to H is constant and is determined by:

$$\frac{E}{H} = \sqrt{\frac{\mu_0}{\varepsilon_0}} = 377\Omega = Z_0$$

Where:

$\mu_o = 1.25 \times 10^8$ H/m, permeability

ε_o = 8.854 X 10^{-12} F/m, permittivity

This ratio is also known as the wave impedance (Z_o).

4. If the "receiving" circuitry has a low impedance, the H field is more efficiently transformed into an unwanted current. If the "receiving" circuitry has a high impedance, the E field is more efficiently coupled as an unwanted voltage. Note that in free space, however, that the E field is 377X "larger" than the H field.

C. Reduction of Far-field Noise

1. Conductive Shielding. Since E and H fields cannot exist without each other in the case of Far-field radiation, and since the H field is smaller than the E field, a simple conductive shield will completely attenuate the field.
2. Reduction of Input Impedance. Where complete shielding is impossible to implement, reduction of input impedance or other susceptible circuit impedances will reduce the E field pick-up.

3. Differential (Balanced) Inputs. The use of differential inputs reduces noise by subtracting the currents or voltages that have been equally induced into the lead wires (common mode rejection or CMR), leaving only the signal to be amplified (see Figure 1).

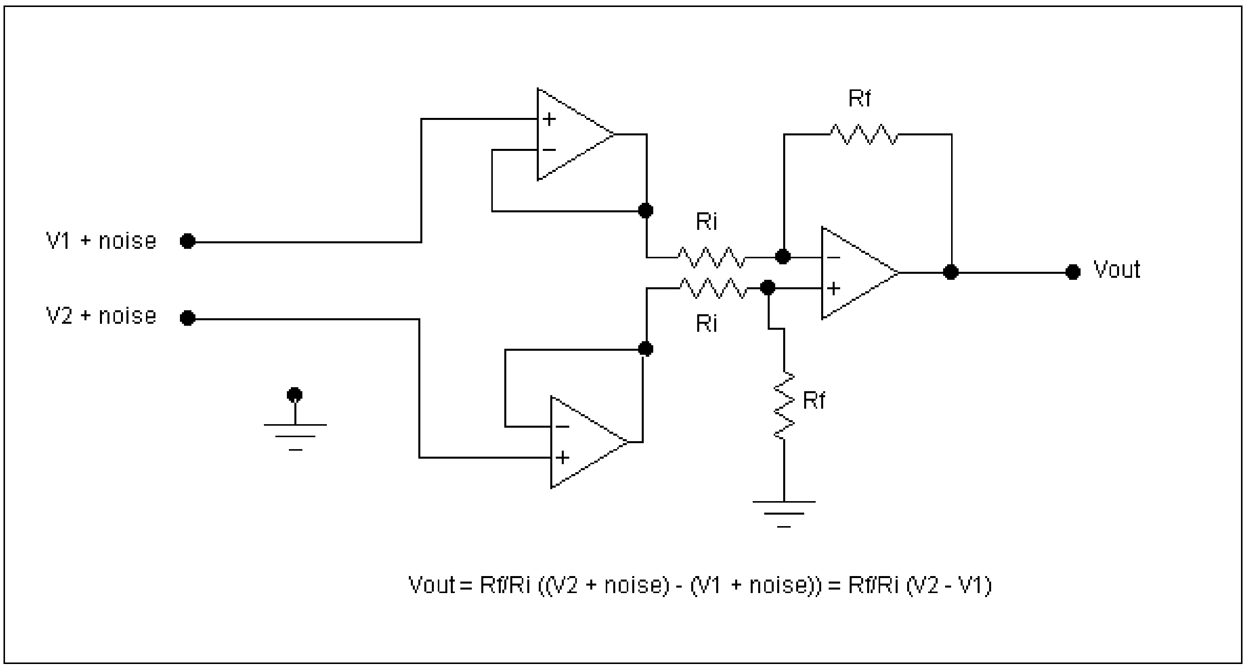


Figure 1. Differential input amplifier showing noise cancellation on the input.

D. Near-field characteristics
1. Near-field is defined as a distance less than $\lambda/2\pi$ from the emission source.
2. Near-field emissions are characterized by their "launching" or source impedance. The E/H ratio is not a constant in the near-field.
   a. Low source impedance. If the emission source has a low impedance and thus a high current, the H field is predominant over the E field. In this case the H field

falls off as $1/r^3$ while the E field falls off as $1/r^2$. Coupling of this signal into circuitry is accomplished mainly by magnetic (mutual) induction.

   b. High source impedance. If the emission source has a high impedance and thus a high voltage, the E field is predominant over the H field. In this case the E field falls off as $1/r^3$ while the H field falls off as $1/r^2$. At a distance near $\lambda/2\pi$, the E/H ratio becomes a constant 377 and the wave is now considered a far-field wave. Coupling of this type of signal into circuitry is accomplished mainly by means of capacitive coupling.

3. Examples of near-field noise sources:
   a. Low source impedance. Examples include line transformers, high current machinery including motors, line cords, lamps, switching power supplies, etc.
   b. High source impedance. Examples include adjacent circuitry (wires, PCB traces, components), large surface area conductors (that form capacitor plates), etc.

E. Reduction of Near-field Noise

1. Low source impedance, current-driven (magnetic induction):
   a. Reduce mutual inductance between emitter and receiver.
      i. Increase distance from emission source.
      ii. Avoid long wire lengths (thus reducing magnetic flux coupling).
      iii. Can sometimes be accomplished by placing the emission source at right angles to receiving apparatus.
      iv. Use toroidal transformers or other devices with low radiated fields.

b. Use high permeability material (i.e., mu-metal) enclosures.
c. Use twisted pair cable to carry large AC currents. Twisted pairs conduct AC currents 180 degrees out of phase in each wire, thus effectively canceling the radiated field at a distance from the wires.

2. High source impedance, voltage-driven (capacitive or antenna pick-up):
   a. Use conductive shielding.
   b. Increase distance from emission source.
   c. Reduce receiving circuit input impedance and component impedance.
   d. Use differential inputs to eliminate common mode voltages.
   e. Use ground planes on PC boards.
   f. Isolate analog and digital grounds when mixed technology is used.

III. Conducted Noise

A. Conducted noise is noise that for one reason or another cannot be completely eliminated at the front-end of a process and is being conducted along with the desired signal through circuitry.

B. Remediation Techniques

1. Noise Measurements
   a. Determine spectral content of desired signal by modeling or by measurement of noise-free signal using a spectrum analyzer or FFT.
   b. Determine the spectral content of the noise signal, typically by measurement.
   c. Determine the source of the noise. A probe consisting of a ferrite rod wound with wire connected to an oscilloscope makes a very useful noise probe.
2. Filtering

a. Low pass filters eliminate noise at frequencies above the signal of interest (equivalent of integration in the time domain, e.g., the same as averaging).
b. High pass filters eliminate noise at frequencies below the signal of interest (equivalent to taking the derivative in the time domain).
c. Band pass filters eliminate all but a band of desired frequencies.
d. Notch filters eliminate a very narrow band of frequencies (often used to eliminate 60 Hz line noise).
e. Comb filters have multiple notches spaced at harmonic intervals (often used to eliminate 60 Hz and its related harmonics at 120 Hz and above).

3. Filtering Issues
    a. Frequency dependent phase effects associated with filtering may impair the desired signal especially when temporal timing of the signal is important.
    b. Filter implementations are not perfect; beware of in- and out-of band effects such as ripple.
    c. Avoid circular logic in the design of a filter! Spectrally-rich noise can filtered through a narrow band pass filter and produce a sinewave at the center frequency that may look like the desired signal.
4. Other noise reduction techniques
    a. Phase cancellation. Periodic noise can be eliminated by shifting the phase of the noise signal by 180°, inverting it, and adding it back to the original signal containing the noise and the desired signal.
    b. Lock-in amplifiers. A lock-in essentially provides a very narrow band pass filter function at the input signal frequency (See Figure 2).

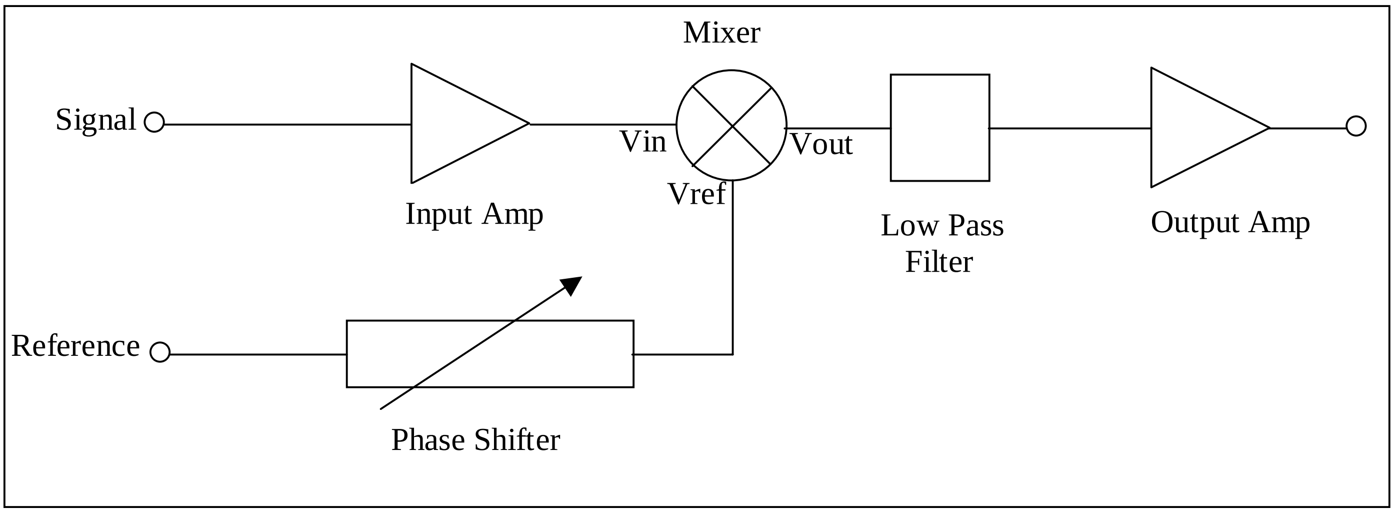


Figure 2. Lock-in Amplifier.

Time Domain Mixer Equations for the lock-in amplifier:

Input signal:

$$V_{in} = A\cos(\omega t) \qquad \omega = 2\pi F$$

Phase-shifted reference signal:

$$V_{ref} = B\cos(\omega t + \theta)$$

The mixer multiplies the two inputs:

$$
\begin{aligned}
V_{in} \times V_{ref} &= A\cos(\omega t) \times B\cos(\omega t + \theta) \\
&= AB\cos\omega t(\cos\omega t\cos\theta - \sin\omega t\sin\theta) \\
&= AB(\cos 2\omega t\cos\theta - \cos\omega t\sin\omega t\sin\theta) \\
&= AB(\tfrac{1}{2} + \tfrac{1}{2}\cos 2\omega t)\cos\theta - \tfrac{1}{2}\sin 2\omega t\sin\theta) \\
&= \tfrac{1}{2}AB((1 + \cos 2\omega t)\cos\theta - \sin 2\omega t\sin\theta) \\
&= \tfrac{1}{2}AB(\cos\theta + \cos 2\omega t\cos\theta - \sin 2\omega t\sin\theta) \\
&= \tfrac{1}{2}AB\cos\theta + \tfrac{1}{2}AB(\cos 2\omega t\cos\theta - 2\omega t\sin\theta) \\
&= \tfrac{1}{2}AB\cos\theta + \tfrac{1}{2}AB\cos(2\omega t + \theta)
\end{aligned}
$$

For most cases, θ is adjusted to 0° (or 180°) and both θ and B are held constant, therefore the last equation collapses to:

$$V_{in} \times V_{ref} = \tfrac{1}{2}AB + \tfrac{1}{2}AB + \tfrac{1}{2}AB\cos(2\omega t) = AB + \tfrac{1}{2}AB\cos(2\omega t)$$

This results in a DC level with magnitude AB and an AC signal at twice the frequency of the desired signal. A low pass filter removes the second harmonic AC signal leaving only the DC portion. If noise had been added to the input signal, Vin, but the noise was uncorrelated, it would not mix with the reference signal to form a part of the DC level. It would result in an AC signal that was the multiplication of the noise signal with the reference signal and would be remove with in the low pass filter.

IV. Conducted Noise – Grounding

A. Ground Loops

1. Ground loops occur when circuitry has multiple grounds. Each of the ground wires shown in Figure 3 are not perfect conductors and therefore have a resistance, R, associated with them. If the shields around each piece of equipment are doing their jobs and absorbing external noise, the ground wires will have a noise current, I, flowing in them towards earth ground. Therefore, instruments A and B will develop a voltage difference V = IR between them. Noise voltage V will then appear on signals processed by instruments A and B. In addition, the ground loop itself serves as a large pick-up loop to intercept further EM noise.

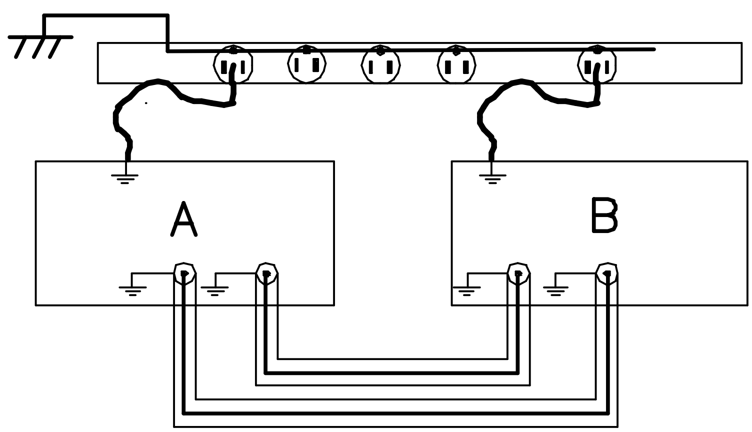


Figure 3. Ground loop example.

2. Identification of ground loops.
   a. An AC voltmeter may be used to detect a voltage difference exists, then either one piece of equipment is not grounded at all or a ground loop exists.
   b. An isolated 3-prong to 2-prong converter plug can be used on one of the instruments to isolate it from line

ground. This should be performed with great care because if an undetected wiring fault exists in one of the instruments, isolating it from ground could cause a shock hazard. If isolating the equipment reduces the noise problem, then a ground loop exists. DO NOT use an isolator to solve a ground loop problem permanently as a shock hazard could result.

c. Removal of connecting cables between instruments (or at least the ground shields) will cause the noise to decrease if a ground loop exists.

3. Remediation of Ground Loops. (Note: It may be necessary to use a trial-and-error approach to eliminate noise due to ground loops.)
   a. The ground loop can be allowed to stay as shown, but the resistance should be reduced between instruments A and B as well as between both A and B and earth ground. This can be accomplished by using heavy gauge wire or braid to connect A to B and to connect both A and B to earth ground. By reducing the resistance, the noise voltage developed between instruments is reduced.
   b. The shields from one end of each of the coaxial connections can be lifted, breaking the loop and allowing each instrument to be grounded in only one place through the line cord. Insulated BNC jacks can be used to accomplish this isolation.

B. Noisy ("dirty") Grounds

1. A noisy or "dirty ground typically refers to building electrical grounds present at the electrical receptacles. It can refer to a corrupted ground system or to a ground system designated as dirty because it carries the ground currents from heavy machinery. A dirty ground may have large noise currents flowing in it and can develop significant noise voltages from receptacle to receptacle.

2. "Clean" grounds are sometimes provided that have limited current-carrying capability and may have individual low resistance paths to earth ground.
3. Remediation methods for dirty grounds:
    a. Use only clean grounds for sensitive instruments.
    b. Do not plug high current devices (refrigerators, water baths, hot plates, etc.) into a clean ground system.
    c. Remove noise producing machinery from the circuit that shares a common ground system used for sensitive instruments. Unfortunately, this can be a major electrical task.
    d. Use of line conditioners or isolation transformers may help.

V. Environmental and Other Noise Types
  A. Temperature including air flow, conduction, convection, etc.
  B. Light
  C. Others: Pressure, galvanic (dissimilar metals) and electrolytic (dc current, dissimilar metals and electrolyte) action, triboelectric (electrostatic), acoustic (microphonic) ionospheric (sunspots), etc.

INTRINSIC (INTERNAL) NOISE

VI. Intrinsic Noise Types:

  A. Gaussian Noise
    1. Time and frequency relationships
        a. Random event occurrence in the time domain (Fig. 4)
        b. Constant spectral content (White) in the frequency domain (Fig. 5)
    2. Statistics

a. Voltage noise is commonly expressed in terms of root-mean-square voltage, Vrms:

$$Vrms = \sqrt{\left(\frac{1}{T}\int_0^T V^2(t)dt\right)}$$

Note: When the symbol Vn is used it will be assumed to refer to an rms value unless otherwise indicated.

b. Gaussian distribution:

$$P(Vn) = \left(\frac{1}{\sqrt{(2\pi)}\sigma}\right)e^{-\frac{Vn^2}{2\sigma^2}}$$

Where:

P(Vn) = Probability of the occurrence of a voltage level or value Vn
Vn = RMS voltage value
σ = Standard deviation of the noise

c. Correlation and addition of noise sources
   1. Random noise is uncorrelated (i.e., it has no phase or frequency correlation with any other process or series of events)
   2. Uncorrelated noise from two or more sources must be added as the square root of the sums of their squares:

$$Vn(total) = \sqrt{(V_{n1})^2 + (V_{n2})^2 + \dots\dots\dots}$$

d. Other characteristics

1. Time average value is zero (i.e., no DC component.

$$Vavg \quad = \ 0$$

2. RMS voltage value for random noise can be estimated as follows:

$$Vrms \cong \frac{Vpeak - to - peak}{6}$$

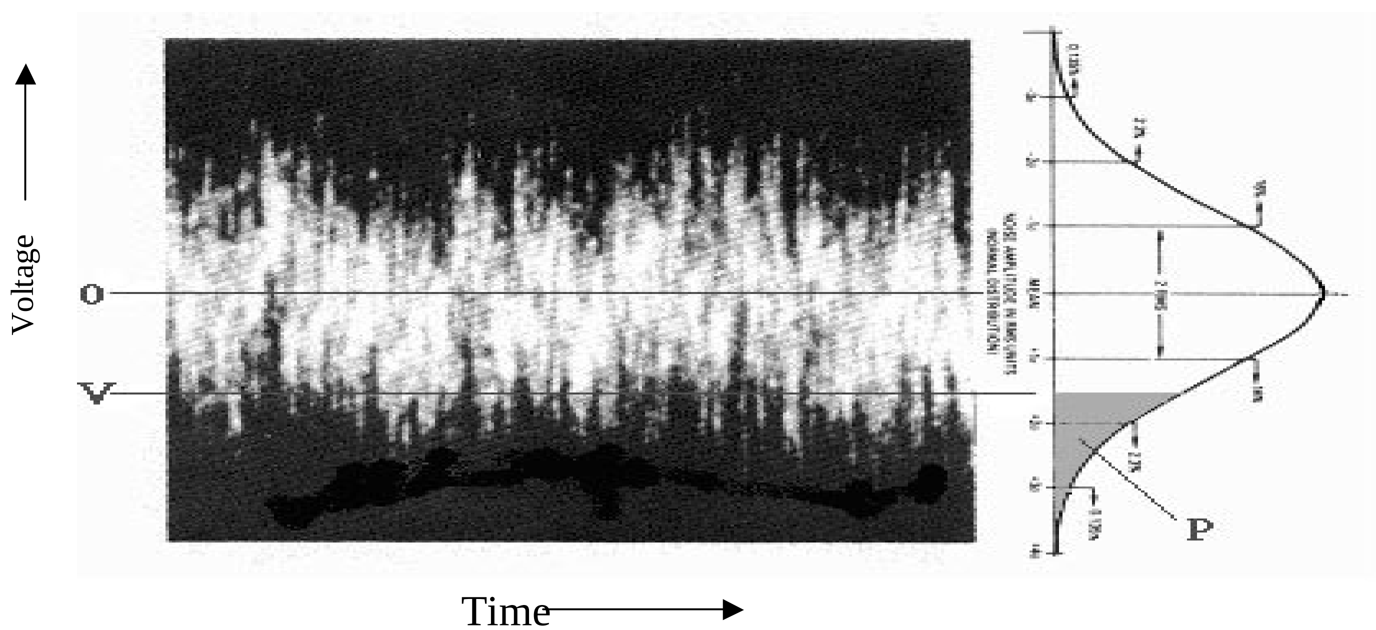


Figure 4. Voltage vs. Time plot (oscillographic) of Gaussian noise superimposed on the voltage distribution. Note that the area under the curve indicated at **P** is the probability of the instantaneous noise voltage exceeding voltage **V**.

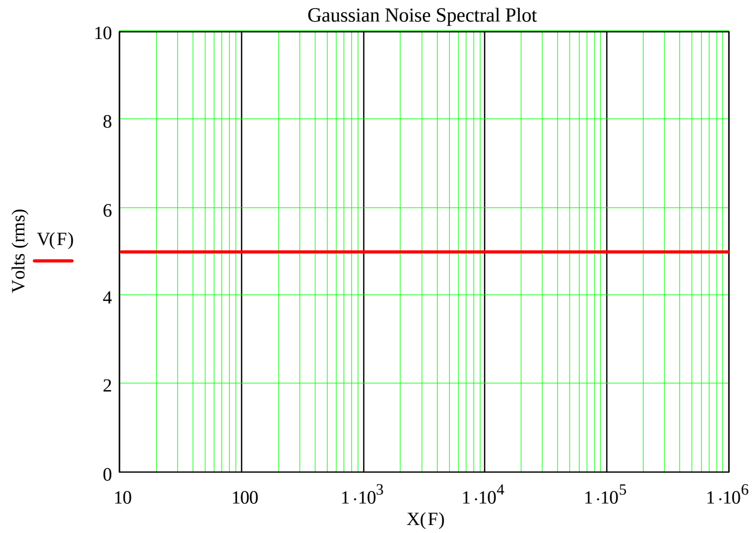


Figure 5. Gaussian rms voltage noise frequency spectrum plot.

VII. Johnson Noise (Nyquist noise, thermal noise) - due to random motion of carriers in any electrical conductor.

A. Gaussian time and frequency domain characteristics (white)

B. In a resistor, the Johnson noise can be described as:

$$V_{jn}(rms) = \sqrt{4kTRB}$$

$$I_{jn}(rms) = \sqrt{4kTB/R}$$

Where:

k = Boltzman's constant = 1.38 X 10^{-23} J/K
T = Temperature in K
R = Resistance in Ohms
B = Bandwidth in Hz

Notes:

1. Vjn (rms) = σ = standard deviation. This can be inserted into the Gaussian probability function to obtain a noise voltage distribution plot.
2. Spot noise V_{jn} or Ijn (rms) can be specified in terms of volts per square root-Hz by removing the bandwidth term, B as seen in Op Amp specification sheets. The equations above specified in terms of spot noise are:

$$V_{jn}(rms)\Big/\sqrt{Hz} = \sqrt{4kTR}$$

$$I_{jn}(rms)\Big/\sqrt{Hz} = \sqrt{4kT/R}$$

VIII. Shot Noise (diode noise) - due to the quantized flow of current (individual electrons) in presence of a potential barrier (e.g., semiconductor diode)

A. Shot noise has Gaussian time and frequency domain characteristics (white).

1. In a semiconductor device, the current noise is given by:

$$I_s rms = \sqrt{2qI_{DC}B}$$

Where:

q = electronic charge = 1.6 X 10^{-19} C
I_{DC} = DC bias current on junction in (A)
B = Bandwidth (Hz)

IX. Flicker Noise (1/F noise, excess noise, contact noise) – empirically determined noise in devices in excess of Johnson and other noises. Contact noise is a subset of flicker noise due to inhomogeneous contact areas.

A. The time and frequency domain characteristics are shown in Figure 6 and Figure 7, respectively. The frequency domain characteristic shows that it is bandwidth limited and typically called 1/F or pink noise and is only of concern in low frequency measurements. 1/F noise is found in resistors, diodes, switches, and transistors, among other components.

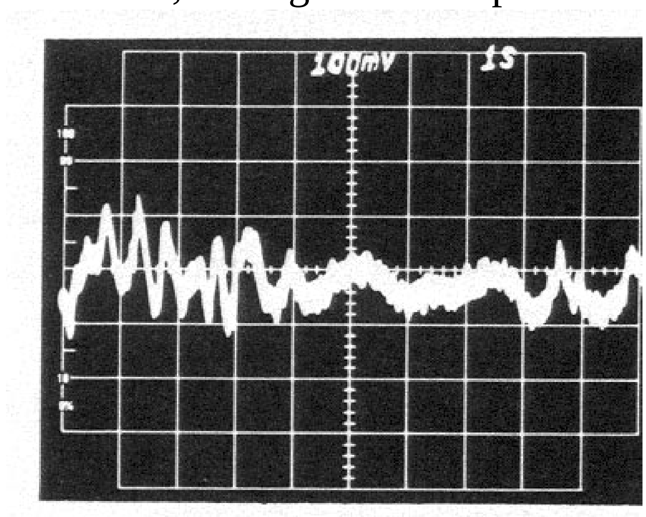


Figure 6. Time domain representation of 1/F noise.

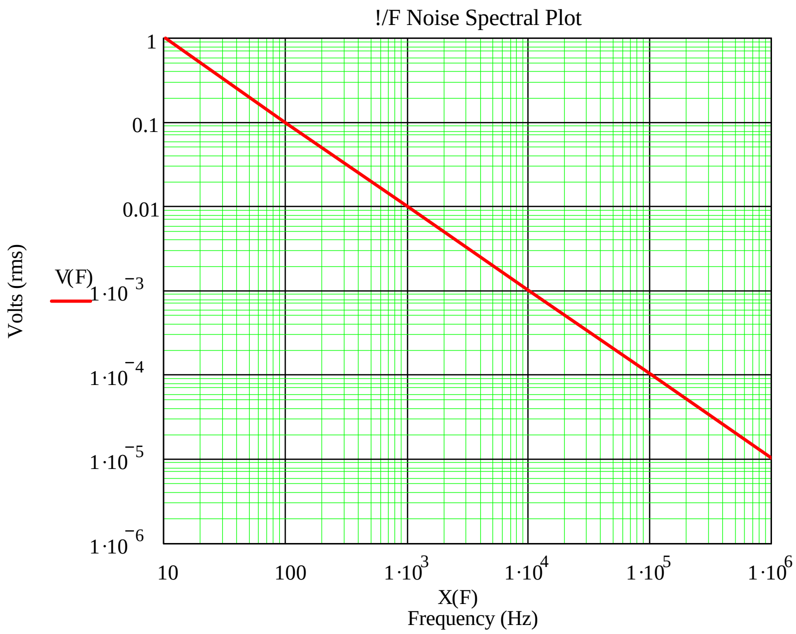


Figure 7. Frequency domain plot of 1/F rms voltage noise.

B. The 1/F rms noise current is governed by the following expression but typically must be measured empirically:

$$I_{fn} \approx \frac{K I_{DC} \sqrt{B}}{\sqrt{f}}$$

Where:

K = a constant dependent upon the material and construction of the device
I_{DC} = DC current flowing in device
B = bandwidth (Hz) centered around frequency f
F = center frequency

X. Quantization Noise

A. Due to discrete sampling of continuous waveforms.
B. Discrete steps with fast rise and fall times result in high order harmonic content (See Figure 8).
C. Low pass filtering can be used to effectively eliminate noise.

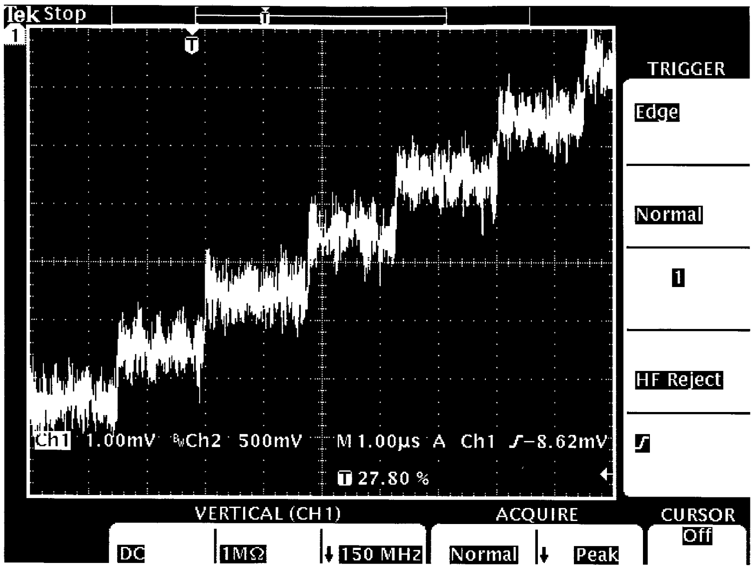


Figure 8. Example of stair-step quantization noise from a 40 MHz, 12 bit DAC. Each transition results in high-order harmonics.

XI. Noise Calculations

A. Noise in resistors

1. Resistor noise consists predominantly of Johnson noise which follows the equations given earlier:

$$V_{jn}(rms) = \sqrt{4kTRB}$$

$$I_{jn}(rms) = \sqrt{4kTB/R}$$

2. Resistor noise models that are used in circuit noise calculations are shown in Figure 9.

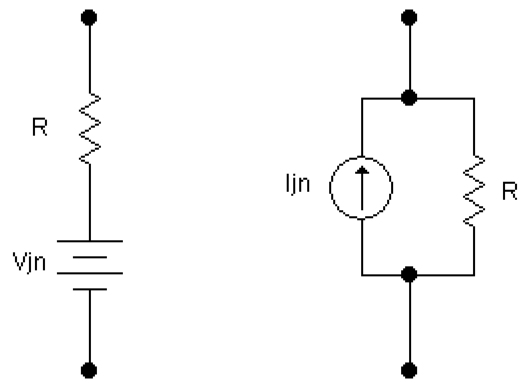


Figure 9. Resistor voltage and current noise models (Thevenin equivalent network on the left and Norton equivalent network on the right).

3. For Example: Noise Voltage for 1 kΩ, 27°C, 20 Hz – 20 kHz = 575 nVrms.
4. A plot of V_{jn} per square root-Hz for a range of resistance values at room temperature is shown in Figure 10.

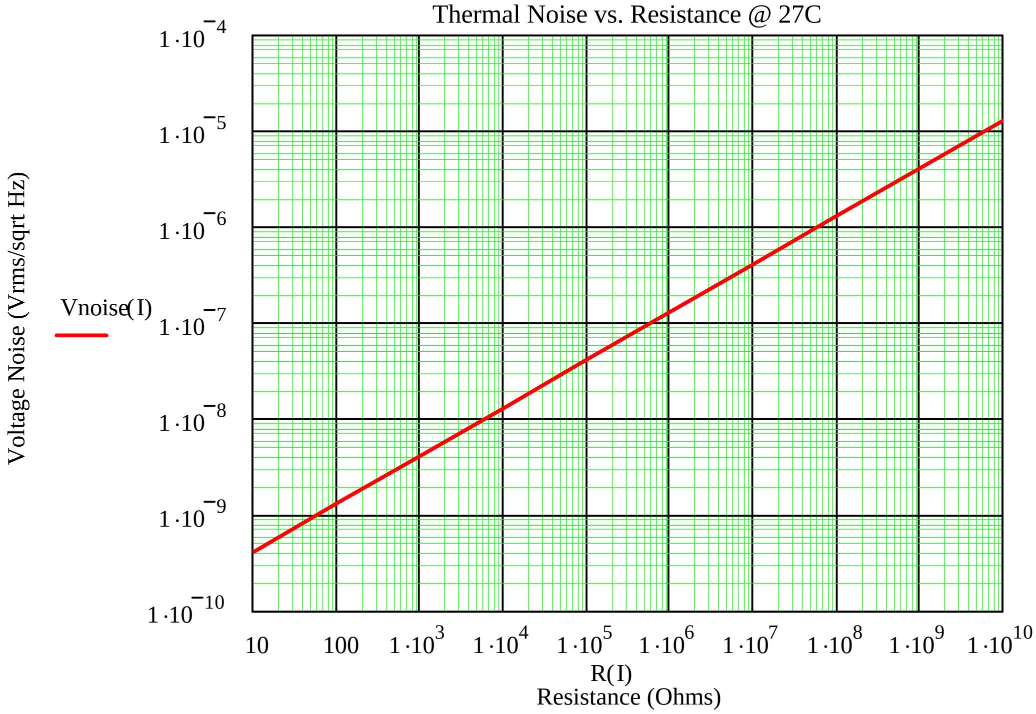


Figure 10. Johnson (Thermal) rms voltage noise vs. resistance at room temperature.

B. Noise in capacitors:

1. Noise in an ideal capacitor is zero.
2. Noise in a non-ideal capacitor is dependent upon the values of internal series and parallel leakage resistances. Typically, capacitors are considered noiseless.
3. Capacitors modify the frequency domain characteristics of the noise in circuits that contain noise sources.

C. Noise in inductors and transformers:

1. Noise in an ideal inductor is zero.
2. Noise in a non-ideal inductor is dependent upon the value of its series resistance (winding resistance) and any distributed capacitance. Typically, inductors are considered noiseless.

3. Noise in an ideal transformer is zero.

4. Noise in a non-ideal transformer is dependent upon the resistance of the windings, distributed capacitance, and coupling capacitance between windings. Typically, transformers are considered noiseless.
5. Inductors and transformers modify the frequency domain characteristics of the noise in circuits that contain noise sources.

D. Noise in discrete semiconductor devices.

1. The simplest semiconductor device is the diode and it can be used as a basis to understand noise issues in other semiconductor devices
2. Since diodes have a junction (barrier) and are typically biased with a current, Shot noise must be considered. In addition, Johnson noise resulting from the diode's equivalent forward or reverse-biased resistances must be considered.
3. A diode model using current noise sources (Norton equivalent for Johnson noise) is shown in Figure 11. The shot noise is a function of the applied bias current and is typically measured empirically. The shunt (channel) resistance, Rsh, must be known in order to calculate the Johnson noise (Rsh may be bias dependent).

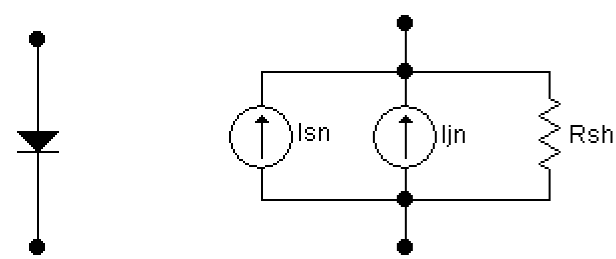


Figure 11. A diode noise model that shows Shot current noise with a Norton equivalent of Johnson noise.

E. Noise in OP-Amps

1. Op-Amp voltage and current noise is typically specified in the manufacturer's data sheets (See Appendix A). Unfortunately, specification data is not always complete or expressed in the same terms and conversions or assumptions must be made before they can be used in calculations.
2. These noise specifications must be taken together with the noise contributions of the external circuitry in order to determine the overall noise performance of the amplifier.

F. Noise in an Op Amp circuit

1. Generalities:
    a. Total noise is typically lowest for bipolar input Op-Amps when used with a low source impedance.
    b. Total noise is typically lowest for FET input Op-Amps when used with a high source impedance.

2. An example of noise calculations for a non-inverting Op-Amp circuit using a Burr-Brown OPA627 is shown in Figures 12 and 13.

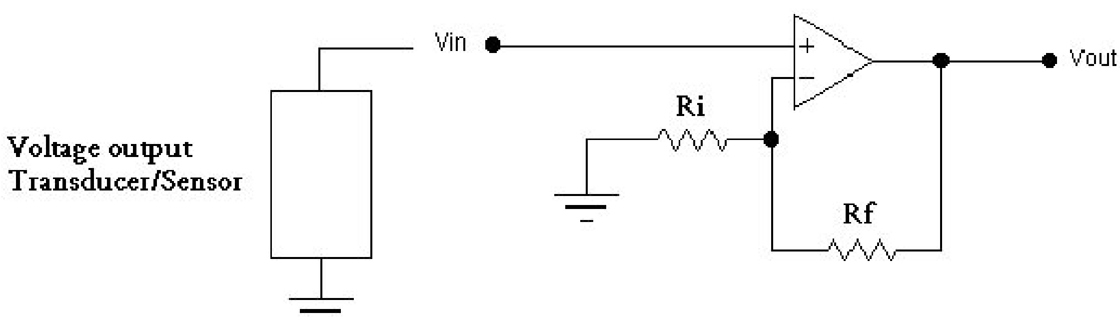


Figure 12. Typical non-inverting amplifier topology with the transfer function given by: Vout = Vin(Rf/Ri)

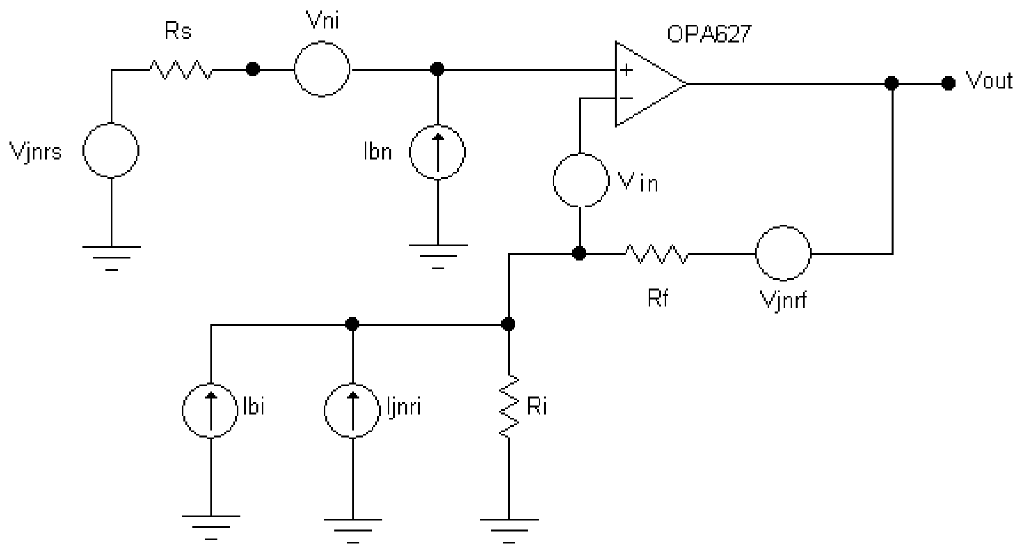


Figure 13. Model for amplifier noise calculations.

Definitions for the noise model in terms of spot noise:

Vni = Op-Amp non-inverting input voltage noise (from spec. sheet)
Vin = Op-Amp inverting input voltage noise (from spec. sheet, typically the same for both the non-inverting and inverting inputs)
Ibn = Op-Amp non-inverting input bias current noise (from spec. sheet)
Ibi = Op-Amp inverting input bias current noise (from spec. sheet)
Rs = Series resistance of transducer.
Vjnrs = Series resistance Rs Johnson rms voltage noise source = $\sqrt{4kTRs}$
Ijnri = Ri Johnson Norton equiv. rms voltage noise current source = $\sqrt{\frac{4kT}{Ri}}$
Vjnrf = Rf Johnson rms voltage noise source = $\sqrt{4kTRf}$
G = Circuit voltage gain = Rf/Ri

Op-Amp assumptions:

Non-inverting input impedance = ∞

Inverting input impedance = ∞

Bandwidth = DC to ∞
Output impedance = 0

Calculation Steps:

Reflect all noise sources to the input of the amplifier as seen in Figure 14:

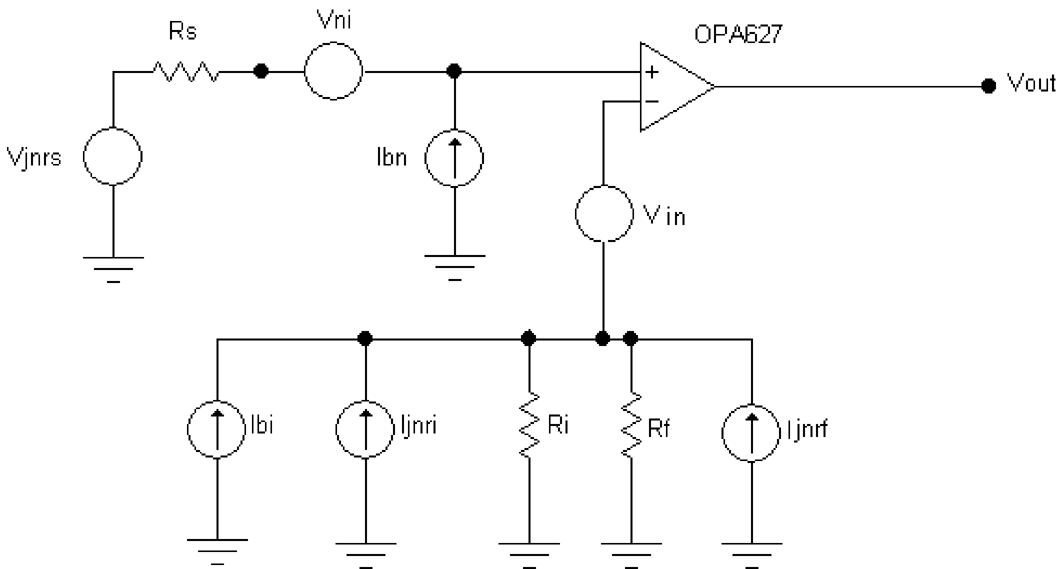


Figure 14. Non-inverting amplifier model with all noise sources reflected to the input of the amplifier.

Convert all current noise sources to voltage sources as seen in Figure 15:

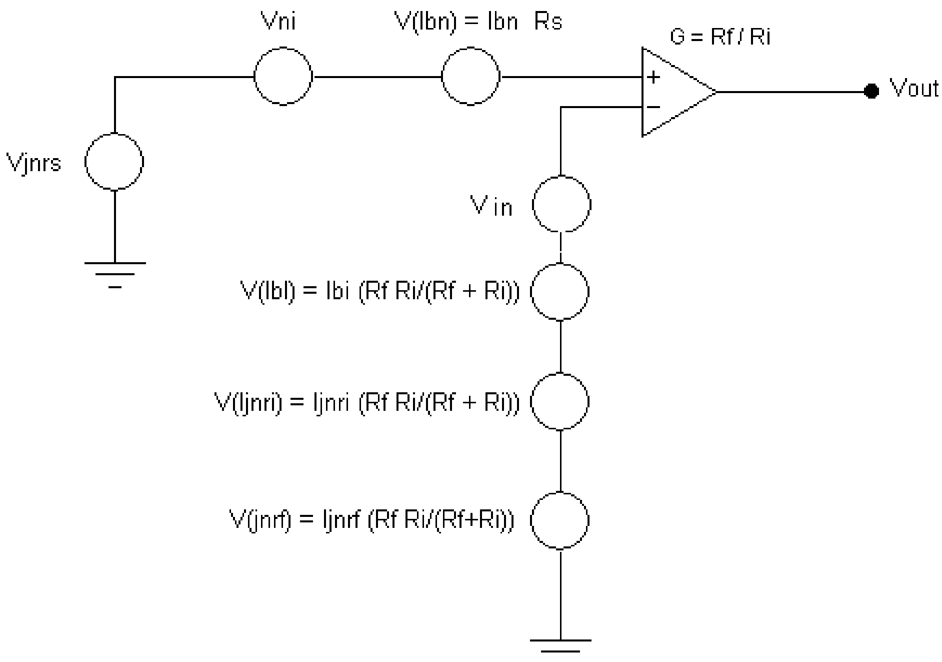


Figure 15. Non-inverting amplifier with all noise sources converted to voltage sources.

Calculate total noise at the output by multiplying each of the voltage sources by the voltage gain of the amplifier:

$$Vnout = \sqrt{((Vjnrs)(G))^2 + ((Vni)(G))^2 + ((Ibn)(Rs)(G))^2 + ((Vin)(G))^2 + \left(Ibi\left(\frac{RfRi}{Rf+Ri}\right)G\right)^2 + \left(Ijnri\left(\frac{RfRi}{Rf+Ri}\right)G\right)^2 + \left(Ijnrf\left(\frac{RfRi}{Rf+Ri}\right)G\right.}$$

Simplify parallel resistance combination using this expression for Rp:

$$Rp = \frac{RfRi}{Rf+Ri}$$

$$Vnout = \sqrt{((Vjnrs)(G))^2 + ((Vni)(G))^2 + ((Vni)(G))^2 + ((Ibn)(Rs)(G))^2 + (Ibi(Rp)G)^2 + (Ijnri(Rp)G)^2 + (Ijnrf\,(Rp)G)^2}$$

Substitute noise equations:

$$Vnout = \sqrt{\left(\left(\sqrt{4kTRs}\right)(G)\right)^2 + \left((Vni)(G)\right)^2 + \left((Vni)(G)\right)^2 + \left((Ibn)(Rs)(G)\right)^2 + \left((Ibi)(Rp)G\right)^2 + \left(\left(\sqrt{\frac{4kT}{Rp}}\right)(Rp)G\right)^2 + \left(\left(\sqrt{\frac{4kT}{Rp}}\right)(Rp)G \right.}$$

This final equation gives the spot voltage noise in Vrms/square root-Hz seen at the output of the non-inverting amplifier circuit. This equation, however, does not take into consideration 1/F noise that will most likely be the limiting noise below 100 Hz. It also does not take into consideration the effects of a real amplifier with a limited Gain-Bandwidth-Product and therefore will roll the noise off as frequency increases. In addition, if any reactive components (i.e., capacitors or inductors) were in the circuit, their effect on the frequency response of the circuit would have to be taken into effect as shown in Figure 16.

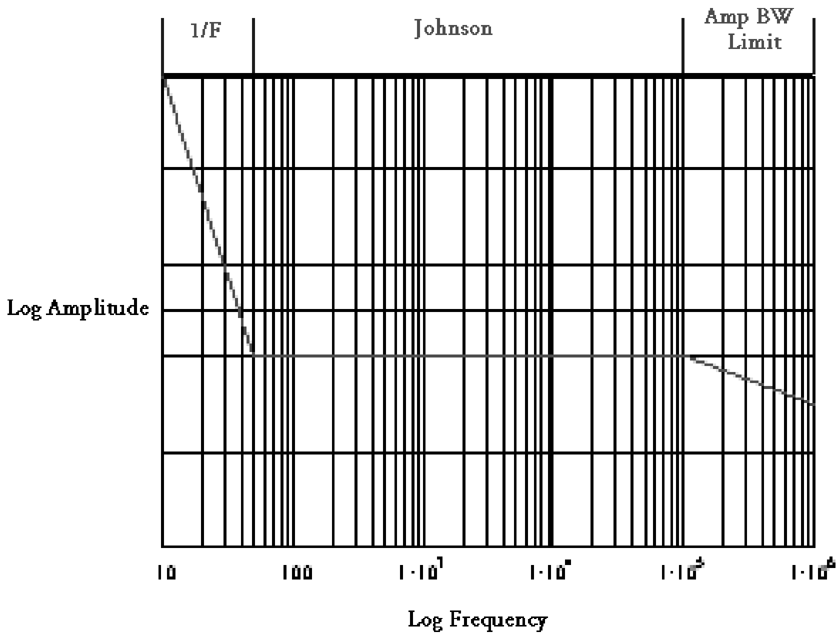
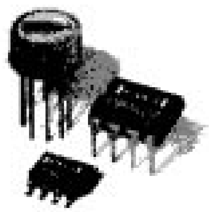


Figure 16. Noise vs. frequency plot for a realistic amplifier.

APPENDIX A

OPA627 DATA SHEETS

BURR - BROWN®



OPA627
OPA637

Precision High-Speed *Difet*® OPERATIONAL AMPLIFIERS

FEATURES

- **VERY LOW NOISE: 4.5nV/√Hz at 10kHz**
- **FAST SETTLING TIME:**
  OPA627—550ns to 0.01%
  OPA637—450ns to 0.01%
- **LOW V_{OS}: 100μV max**
- **LOW DRIFT: 0.8μV/°C max**
- **LOW I_B: 5pA max**
- **OPA627: Unity-Gain Stable**
- **OPA637: Stable in Gain ≥ 5**

APPLICATIONS

- **PRECISION INSTRUMENTATION**
- **FAST DATA ACQUISITION**
- **DAC OUTPUT AMPLIFIER**
- **OPTOELECTRONICS**
- **SONAR, ULTRASOUND**
- **HIGH-IMPEDANCE SENSOR AMPS**
- **HIGH-PERFORMANCE AUDIO CIRCUITRY**
- **ACTIVE FILTERS**

DESCRIPTION

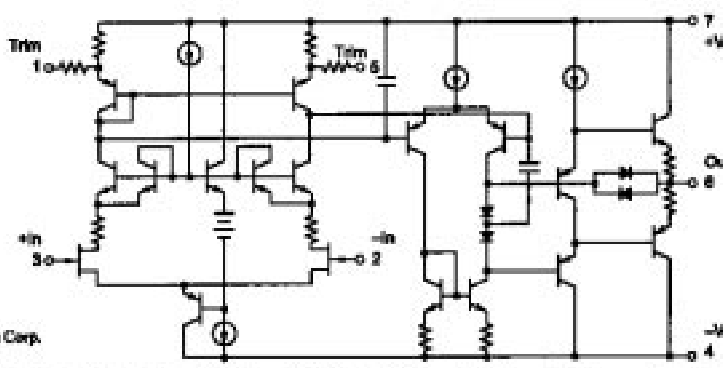
The OPA627 and OPA637 ***Difet*** operational amplifiers provide a new level of performance in a precision FET op amp. When compared to the popular OPA111 op amp, the OPA627/637 has lower noise, lower offset voltage, and much higher speed. It is useful in a broad range of precision and high speed analog circuitry.

The OPA627/637 is fabricated on a high-speed, dielectrically-isolated complementary NPN/PNP process. It operates over a wide range of power supply voltage—±4.5V to ±18V. Laser-trimmed ***Difet*** input circuitry provides high accuracy and low-noise performance comparable with the best bipolar-input op amps.

High frequency complementary transistors allow increased circuit bandwidth, attaining dynamic performance not possible with previous precision FET op amps. The OPA627 is unity-gain stable. The OPA637 is stable in gains equal to or greater than five.

Difet fabrication achieves extremely low input bias currents without compromising input voltage noise performance. Low input bias current is maintained over a wide input common-mode voltage range with unique cascode circuitry.

The OPA627/637 is available in plastic DIP, SOIC and metal TO-99 packages. Industrial and military temperature range models are available.



Difet®, Burr-Brown Corp.

International Airport Industrial Park • Mailing Address: PO Box 11400, Tucson, AZ 85734 • Street Address: 6730 S. Tucson Blvd., Tucson, AZ 85706 • Tel: (520) 746-1111 • Twx: 910-952-1111
Internet: http://www.burr-brown.com/ • FAXLine: (800) 548-6133 (US/Canada Only) • Cable: BBRCORP • Telex: 066-6491 • FAX: (520) 889-1510 • Immediate Product Info: (800) 548-6132

SPECIFICATIONS

ELECTRICAL

At T_A = +25°C, and V_S = ±15V, unless otherwise noted.

PARAMETER	CONDITIONS	OPA627BM, BP, SM OPA637BM, BP, SM MIN	TYP	MAX	OPA627AM, AP, AU OPA637AM, AP, AU MIN	TYP	MAX	UNITS
OFFSET VOLTAGE (1)								
Input Offset Voltage			40	100		130	250	µV
AP, BP, AU Grades			100	250		280	500	µV
Average Drift			0.4	0.8		1.2	2	µV/°C
AP, BP, AU Grades			0.8	2		2.5		µV/°C
Power Supply Rejection	V_S = ±4.5 to ±18V	106	120		100	116		dB
INPUT BIAS CURRENT (2)								
Input Bias Current	V_{CM} = 0V		1	5		2	10	pA
Over Specified Temperature	V_{CM} = 0V			1			2	nA
SM Grade	V_{CM} = 0V			50				nA
Over Common-Mode Voltage	V_{CM} = ±10V		1			2		pA
Input Offset Current	V_{CM} = 0V		0.5	5		1	10	pA
Over Specified Temperature	V_{CM} = 0V			1			2	nA
SM Grade				50				nA
NOISE								
Input Voltage Noise								
Noise Density: f = 10Hz	1/F Noise		15	40		20		nV/√Hz
f = 100Hz			8	20		10		nV/√Hz
f = 1kHz	Voltage Noise		5.2	8		5.6		nV/√Hz
f = 10kHz			4.5	6		4.8		nV/√Hz
Voltage Noise, BW = 0.1Hz to 10Hz			0.6	1.6		0.8		µVp-p
Input Bias Current Noise								
Noise Density, f = 100Hz	Current Noise		1.6	2.5		2.5		fA/√Hz
Current Noise, BW = 0.1Hz to 10Hz			30	60		48		fAp-p
INPUT IMPEDANCE								
Differential			10^{13} ‖ 8			*		Ω ‖ pF
Common-Mode			10^{13} ‖ 7			*		Ω ‖ pF
INPUT VOLTAGE RANGE								
Common-Mode Input Range		±11	±11.5		*	*		V
Over Specified Temperature		±10.5	±11		*	*		V
Common-Mode Rejection	V_{CM} = ±10.5V	106	116		100	110		dB
OPEN-LOOP GAIN								
Open-Loop Voltage Gain	V_O = ±10V, R_L = 1kΩ	112	120		106	116		dB
Over Specified Temperature	V_O = ±10V, R_L = 1kΩ	106	117		100	110		dB
SM Grade	V_O = ±10V, R_L = 1kΩ	100	114					dB
FREQUENCY RESPONSE								
Slew Rate: OPA627	G = –1, 10V Step	40	55		*	*		V/µs
OPA637	G = –4, 10V Step	100	135		*	*		V/µs
Settling Time: OPA627 0.01%	G = –1, 10V Step		550			*		ns
0.1%	G = –1, 10V Step		450			*		ns
OPA637 0.01%	G = –4, 10V Step		450			*		ns
0.1%	G = –4, 10V Step		300			*		ns
Gain-Bandwidth Product: OPA627	G = 1		16			*		MHz
OPA637	G = 10		80			*		MHz
Total Harmonic Distortion + Noise	G = +1, f = 1kHz		0.00003			*		%
POWER SUPPLY								
Specified Operating Voltage			±15			*		V
Operating Voltage Range		±4.5		±18	*		*	V
Current			±7	±7.5		*	*	mA
OUTPUT								
Voltage Output	R_L = 1kΩ	±11.5	±12.3		*	*		
Over Specified Temperature		±11	±11.5		*	*		V
Current Output	V_O = ±10V		±45			*		mA
Short-Circuit Current		±35	+70/–55	±100	*	*	*	mA
Output Impedance, Open-Loop	1MHz		55			*		Ω
TEMPERATURE RANGE								
Specification: AP, BP, AM, BM, AU		–25		+85	*		*	°C
SM		–55		+125				°C
Storage: AM, BM, SM		–60		+150	*		*	°C
AP, BP, AU		–40		+125	*		*	°C
θ_{J-A}: AM, BM, SM			200			*		°C/W
AP, BP			100			*		°C/W
AU			160					°C/W

* Specifications same as "B" grade.

NOTES: (1) Offset voltage measured fully warmed-up. (2) High-speed test at T_J = +25°C. See Typical Performance Curves for warmed-up performance.

References:

Henry Ott. Noise Reduction Techniques in Electronic Systems, Second Edition, John Wiley & Sons, New York, 1988.

Al Ryan and Tim Scranton. D-C Amplifier Noise Revisited, Analog Dialogue 18-1, 1984. Published by Analog Devices, Inc., Norwood, MA.

Paul Horowitz and Winfield Hill. *The Art of Electronics*, Cambridge University Press, Cambridge, pp. 286-315, 1983.

William Hayt. Engineering Electromagnetics, Third Edition, McGraw-Hill, New York, pp.359-406, 1974.