Signal conditioning for piezoelectric sensors

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Introduction

This article explains some of the principles of signal conditioning. Piezoelectric sensors have been chosen to illustrate these principles because their conditioning requires a mix of traditional tools and because they present some challenges that may not arise with other types of sensors.

Piezoelectric sensors

The application of piezoelectric transducers for sensing and actuation extends to many fields. This article focuses on the sensing of a group of physical magnitudes—acceleration, vibration, shock, and pressure—that from the perspective of the sensor and its requied signal conditioning can be considered similar.[1] In the case of acceleration, sensor sensitivity is usually expressed as a charge proportional to an external force or acceleration (many times described as gravitational acceleration, g). Nevertheless, in the strict physical sense, the sensor outputs a charge that is actually a function of its deformation/deflection.

For instance, Figure 1 shows a sensor fixed on the top while the bottom is being pulled by an external force, F_{ext}. In the case of an accelerometer, the fixed extreme (top) could be attached to the object whose acceleration is going to be measured, and the external force would be the inertia of a mass attached to the other extreme (bottom) that is trying to stay still. For the reference coordinate system fixed on the top extreme (assuming that the sensor is acting as a spring with a very high spring constant, K), the deflection x will create an opposing force of

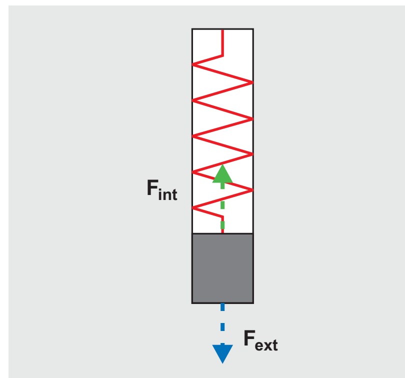
$$F_{int} = Kx. \qquad \textbf{(1)}$$

Eventually, the mass (the deflection of the sensor) will stop moving/changing at

$$F_{int} = F_{ext} = Kx. \qquad \textbf{(2)}$$

Since the charge, Q, is (to first order) proportional to deflection, and deflection is proportional to force, Q is proportional to force. Applying a sinusoidal force with a maximum F_{max}, will create a sinusoidal charge with a maximum Q_{max}. In other words, integrating the current coming from the sensor will yield Q_{max} when the sinusoidal force is at its maximum. Increasing the frequency of the sinusoid will increase the current; but the peak will be reached faster, i.e., keeping the integral (Q_{max}) constant. The manufacturer will provide the specification for sensitivity as the ratio of Q_{max} to F_{max} in the usable frequency range of the sensor. Nevertheless, due to the mechanical properties of the sensor, the sensor actually has a resonant frequency (above the usable frequency range) where even a small oscillatory force will produce relatively large displacements and therefore large output amplitudes.

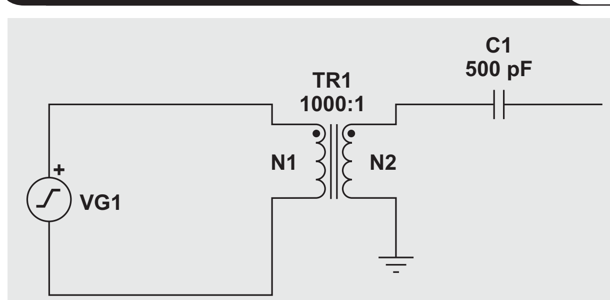
Figure 1. Sensor under acceleration



If the effects of the resonance are ignored, piezoelectric sensors can be modeled to first order as a current source in parallel with the sensor's parasitic capacitance, here referred to as C_d, or they can be modeled as a voltage source in series with C_d. This voltage is the equivalent voltage that would be seen on the plates of the sensor if the charge was just stored on them. Notice, nevertheless, that for the simulation of many applications, the second approach is more straightforward. As explained earlier, the current is proportional to the rate of change of deflection; so, for instance, for a sinusoidal AC sweep of accelerations with constant amplitude, the amplitude of the current generator would have to be changed depending on the frequency.

Finally, if the generator needs to represent the actual physical signal, a transformer can be used, as illustrated in Figure 2. In this example, a generator with a sensitivity of

Figure 2. Piezoelectric sensor model

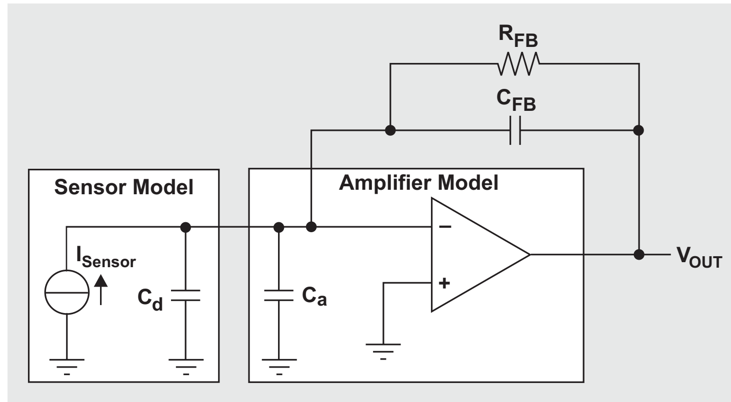


0.5 pC/g and a parasitic capacitance of 500 pF is modeled. The sinusoidal generator outputs 1 V for every unit of g to be simulated. The transformer scales that down to 1 mV on its secondary. A 1-mV swing applied to C1 (500 pF) would inject Q = VC = 0.5 pC on the next stage, as expected.

Analysis of the charge amplifier

Figure 3 shows the basic schematic of a classical charge amplifier that can be used as a signal-conditioning circuit. In this case, the current-source model was chosen to show that the sensor is mainly a device with high output impedance.

Figure 3. Charge amplifier used for signal conditioning



Input impedance

A signal-conditioning circuit must have very low input impedance to collect most of the charge output by the sensor. Thus, the charge amplifier is the ideal solution since its input presents a virtual ground to the sensor signal as long as the amplifier maintains high gain at those signal frequencies. In other words, if any charge coming from the sensor tries to build up on the plates of the sensor (C_d) or on the input parasitic capacitance of the amplifier (C_a), a voltage will be created across the input of the amplifier. This voltage will be immediately compensated for and nulled by pulling or sourcing the same amount of charge current through the negative feedback network, R_{FB} and C_{FB}.

Gain

Since the amplifier's signal input is a virtual ground, the input current creates an output-voltage swing; and the high-frequency gain is set by the value of C_{FB} (discounting the effect of R_{FB}, described next under "Bandwidth"). Notice that the smaller the capacitor is, the bigger the gain. An approximation for gain is

$$\text{Gain} = \frac{1}{C_{FB}}(\text{mV/C}). \tag{3}$$

Also notice that the gain of the circuit ultimately does not depend on the sensor's capacitance (C_d), although it is advisable to pay attention to this value's effect on noise.

Bandwidth

In order to bias the amplifier properly (offer a DC path for the input bias current of the amplifier), a feedback resistor (R_f) is necessary. At lower frequencies, the capacitive circuit in the feedback path becomes open and the feedback resistance become dominant, effectively reducing the gain. At higher frequencies the impedance of the capacitive circuit becomes smaller, effectively eliminating the effect of the resistive feedback path. The final circuit response, including the parasitic capacitor of the sensor, to an AC physical excitation, is that of a high-pass filter, with a pole at:

$$f_{HPF} = \frac{1}{2\pi R_{FB} C_{FB}}. \tag{4}$$

The signal bandwidth of interest is set by the application; so, as the capacitance is lowered to increase the gain, the resistance needs to be increased to keep the pole low. Increasing this resistance has consequences on other careabouts of the solution. Besides the effect on noise (described later under "Noise"), the higher the resistance is, the more difficult the practical implementation—from finding an off-the-shelf resistor to making sure that the trace-to-trace parasitic resistances on the PCB are much bigger than R_{FB} itself. For cases where the circuit specifications allow for the use of resistors on the order of a few hundred megohms, surface-mount resistors are readily available[2] and there are no requirements for advanced layout techniques (like using a guard band).

As mentioned before, another factor limiting the increase of the resistor value is the biasing of the circuit. The input bias current of the amplifier flows through this resistor and creates an output offset voltage. This can be minimized by choosing an amplifier with low input bias currents, such as a FET input amplifier. The input bias currents of this type of amplifier, usually below 100 pA, should be fine as long as the feedback resistor value is below 1 GΩ and the resulting offset can be filtered with AC coupling between the stages.

Note that, due to the difficulty of keeping the high-pass filter's pole low, it becomes increasingly difficult to use a piezoelectric sensor in near-DC applications (even if the leakage currents in the sensor itself are very small).

Although not a part of this amplification stage, a low-pass filter needs to be added at some point to reduce the circuit's response to unwanted signals at the sensor's resonant frequency and to reduce the overall digitized and aliased noise in the band of interest.

Noise

Finally, the signal-to-noise ratio (SNR) needs to be maximized. Performing a brief theoretical noise analysis before proceeding with the simulation will be helpful. Figure 4 shows the main noise sources in the charge amplifier. The output-noise spectral density can be expressed as

$$S_O(f) = I_{NA}^2 \times |Z_{FB}|^2 + E_A^2 \left|1 + \frac{Z_{FB}}{1/(C_d + C_a)s}\right|^2 + E_{R_{FB}}^2 \left|\frac{1}{1 + R_{FB}C_{FB}s}\right|^2, \quad \textbf{(5)}$$

where

$$Z_{FB} = \frac{R_{FB}}{R_{FB}C_{FB}s + 1} \quad \textbf{(6)}$$

and $s = 2\pi fj$. Equation 5 is the classical noise solution for a charge amplifier. C_a is typically very small compared to C_d. So, Equation 5 can be simplified to

$$S_O(f) = I_{NA}^2 \left|\frac{R_{FB}}{R_{FB}C_{FB}s + 1}\right|^2 + E_A^2 \left|1 + \frac{R_{FB}C_d s}{R_{FB}C_{FB}s + 1}\right|^2 + E_{R_{FB}}^2 \left|\frac{1}{R_{FB}C_{FB}s + 1}\right|^2. \quad \textbf{(7)}$$

In fact, the second term can be reduced even further if frequencies well above the high-pass filter's pole are considered:

$$S_O(f) = I_{NA}^2 \left|\frac{R_{FB}}{R_{FB}C_{FB}s + 1}\right|^2 + E_A^2 \left|1 + \frac{C_d}{C_{FB}}\right|^2 + E_{R_{FB}}^2 \left|\frac{1}{R_{FB}C_{FB}s + 1}\right|^2 \quad \textbf{(8)}$$

Trends can be analyzed in several ways. The pole (the term $R_{FB}C_{FB}s + 1$) can be considered constant since increasing R_{FB} would require a reduced C_{FB} or vice versa. From that perspective, increasing R_{FB} would increase the three terms in Equation 8. The voltage noise corresponding to the first term would increase linearly with R_{FB}; the voltage noise corresponding to the second term would also increase; and the voltage noise corresponding to the third term would increase as the square root of R_{FB}, since $E_{R_{FB}} = \sqrt{4kTR_{FB}}$, where k = Boltzmann's constant and T = temperature in degrees Kelvin. Nevertheless, the gain would increase with R_{FB} as C_{FB} became smaller (see Equation 3). This increase of signal with R_{FB} will be similar to any increase of the first two noise terms in Equation 8, but bigger than the increase of the last noise term, therefore improving the overall SNR. The bottom line is to increase R_{FB} as much as practically possible. Another trend to notice is that a sensor with more parasitic capacitance is less desirable from the noise perspective.

Figure 4. Charge amplifier's noise sources

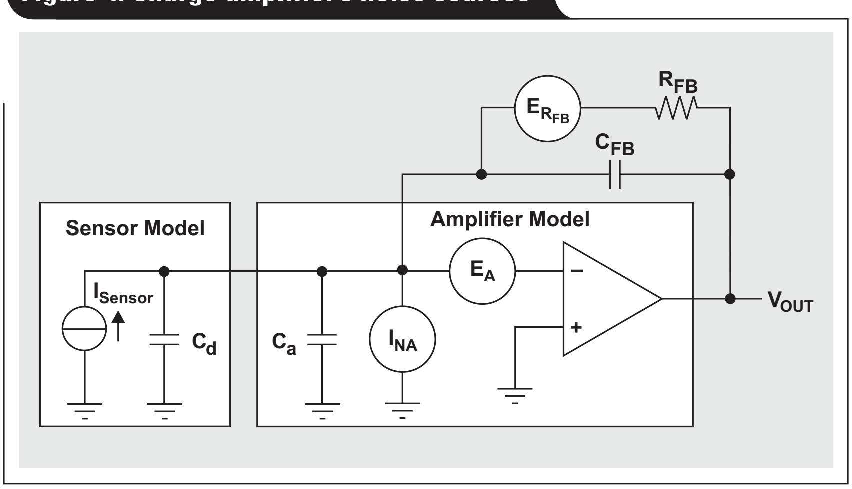
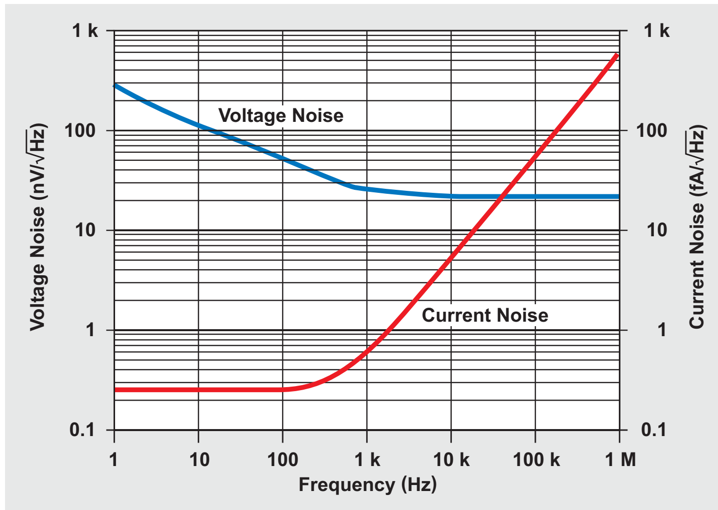


Figure 5. OPA337's input-voltage and input-current noise



Simulation results

For a more practical circuit implementation, the Texas Instruments (TI) OPA337 has been chosen. This amplifier offers low input-voltage and input-current noise (see Figure 5, taken from the data sheet[3]) while accepting a 3-V unipolar supply. Figure 6 shows a model of this circuit in TI's SPICE-based analog simulation program, TINA-TI™.

In this implementation, the pole is at 0.86 Hz. Equation 7 can be analyzed at 5 Hz just to double-check the accuracy of the formula:

- In the first term, if $I_{NA} \approx 0.01\ \mathrm{fA}/\sqrt{\mathrm{Hz}}$, and R_{FB} = 270 MΩ, this term's contribution to output noise is approximately $(2.7\ \mathrm{nV}/\sqrt{\mathrm{Hz}})/5.85 = 0.5\ \mathrm{nV}/\sqrt{\mathrm{Hz}}$.
- In the second term, if $E_A \approx 60\ \mathrm{nV}/\sqrt{\mathrm{Hz}}$, this term's contribution to output noise is approximately $120\ \mathrm{nV}/\sqrt{\mathrm{Hz}}$.
- In the third term, if R_{FB} = 270 MΩ, this term's contribution to output noise is approximately $(2\ \mu\mathrm{V}/\sqrt{\mathrm{Hz}})/5.85 = 340\ \mathrm{nV}/\sqrt{\mathrm{Hz}}$.

Adding all three terms together quadratically totals approximately $360\ \mathrm{nV}/\sqrt{\mathrm{Hz}}$, which is close to the simulation result in Figure 7. Notice, however, that the noise values used differ from the data-sheet values shown in Figure 5. The TINA-TI noise model

Figure 6. TINA-TI model of circuit using OPA337

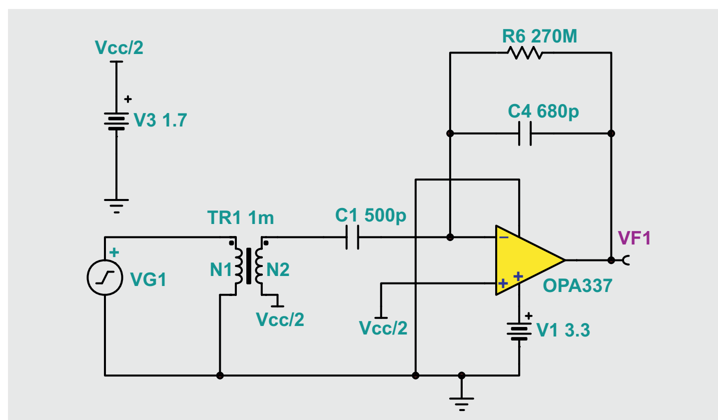


Figure 7. Simulation of output noise from model in Figure 6

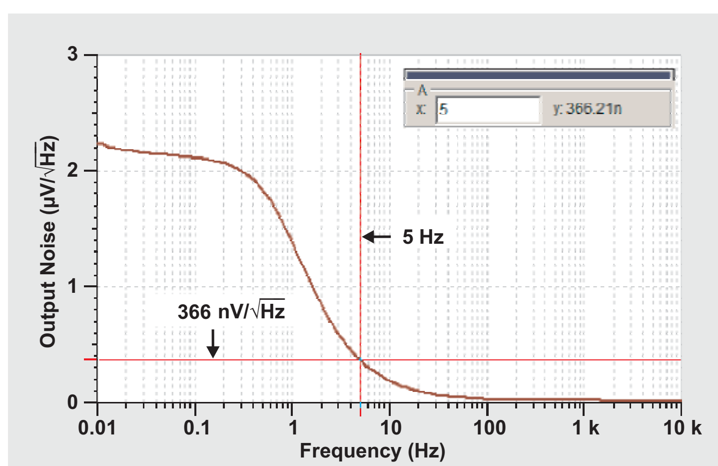


Figure 8. TINA-TI simulation circuit for amplifier noise analysis

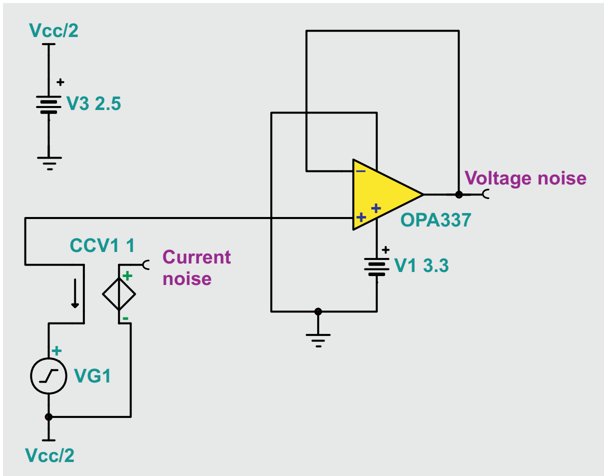
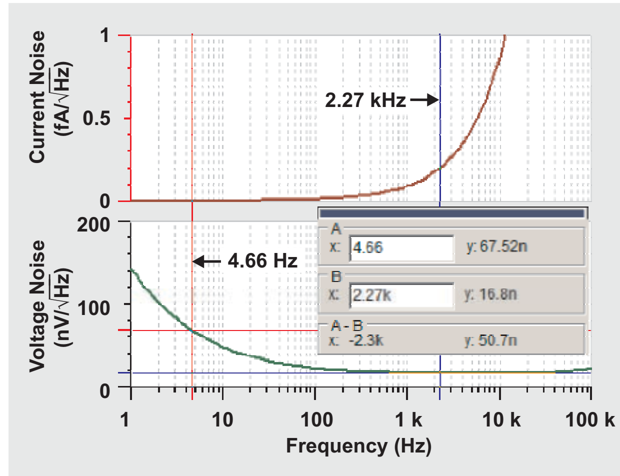


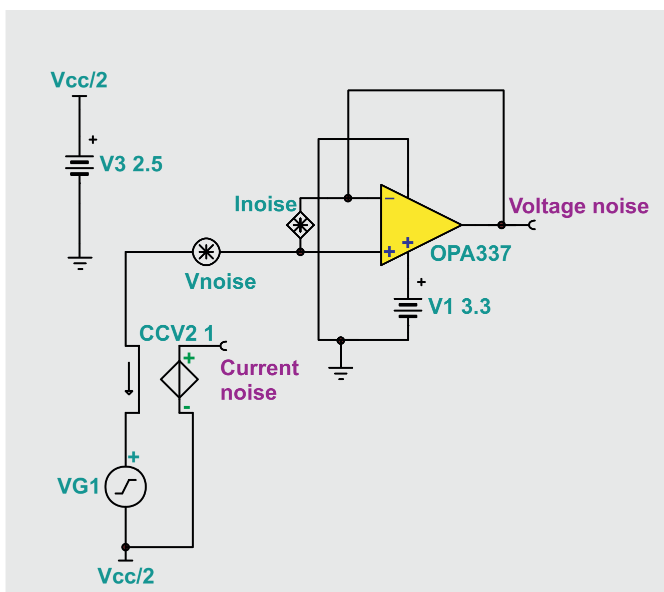
Figure 9. Simulation of output noise from circuit in Figure 8



for the OPA337 is not accurate, as can be proven by simulating the simplified circuit in Figure 8 and obtaining the results in Figure 9 (which should have been the same as in Figure 5).

These results emphasize the importance of a quick theoretical/hand analysis. The circuit of the amplifier is inaccurate and needs to be accounted for in TINA-TI to get realistic numbers. A way to do that can be found in Reference 4, which is Part IV of a series of very valuable articles by Art Kay on noise. A slightly simpler approach is to just add noise (V_{noise} and I_{noise} in Figure 10) to the circuit shown in Figure 8 to compensate for what is missing.

Figure 10. Noise added to circuit in Figure 8



Although not perfect, the results shown in Figure 11 look much more similar to the specification (Figure 5) than the results shown in Figure 9.

With the original circuit in Figure 6, the noise at 5 Hz can again be estimated with Equation 7 using the noise values originally specified.

- In the first term, if $I_{NA} \approx 0.3\ \text{fA}/\sqrt{\text{Hz}}$, and R_{FB} = 270 MΩ, this term's contribution to output noise is approximately $(80\ \text{nV}/\sqrt{\text{Hz}})/5.85 = 14\ \text{nV}/\sqrt{\text{Hz}}$.
- In the second term, if $E_A \approx 130\ \text{nV}/\sqrt{\text{Hz}}$, this term's contribution to output noise is approximately $260\ \text{nV}/\sqrt{\text{Hz}}$.
- In the third term, if R_{FB} = 270 MΩ, this term's contribution to output noise is approximately $(2\ \mu\text{V}/\sqrt{\text{Hz}})/5.85 = 340\ \text{nV}/\sqrt{\text{Hz}}$.

Adding all three terms together quadratically totals approximately $430\ \text{nV}/\sqrt{\text{Hz}}$, which as shown in Figure 13, is very close to the simulation result for the circuit in Figure 12 that includes corrected noise sources.

Now consider the variation of noise versus the feedback resistor. Changing R_{FB} in the first term of Equation 7 from 270 MΩ to 540 MΩ (and dividing C_{FB} by half, from 680 pF to 340 pF, to keep the pole constant) has the following effects on the output-referred noise:

- In the first term, if $I_{NA} \approx 0.3\ \text{fA}/\sqrt{\text{Hz}}$, and R_{FB} = 540 MΩ, this term's contribution to output noise is approximately $(160\ \text{nV}/\sqrt{\text{Hz}})/5.85 = 28\ \text{nV}/\sqrt{\text{Hz}}$.
- In the second term, if $E_A \approx 130\ \text{nV}/\sqrt{\text{Hz}}$, this term's contribution to output noise is approximately $320\ \text{nV}/\sqrt{\text{Hz}}$.
- In the third term, if R_{FB} = 540 MΩ, this term's contribution to output noise is approximately $(3\ \mu\text{V}/\sqrt{\text{Hz}})/5.85 = 510\ \text{nV}/\sqrt{\text{Hz}}$.

Adding all three terms together quadratically totals approximately $600\ \text{nV}/\sqrt{\text{Hz}}$, which is once again

Figure 11. Simulation of output noise from circuit in Figure 10

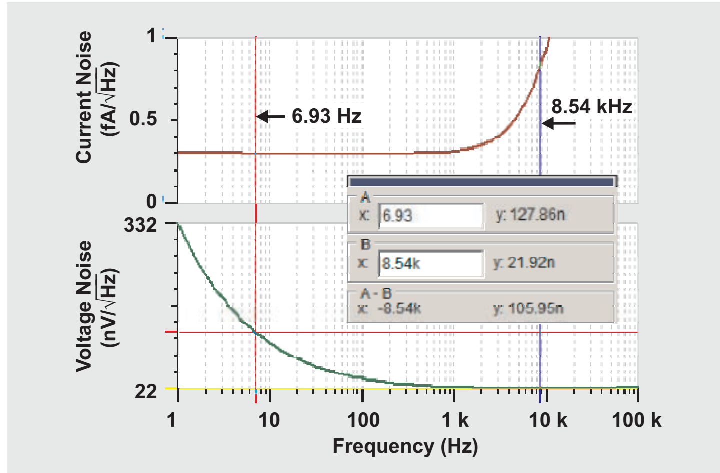


Figure 12. TINA-TI model of Figure 6 circuit with noise sources added

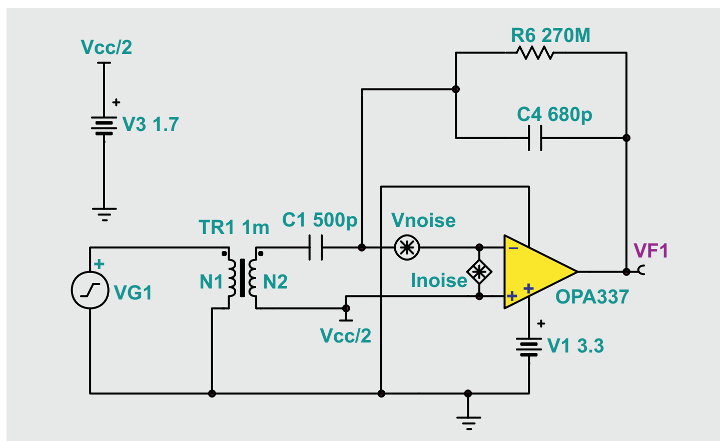


Figure 13. Simulation of output noise from circuit in Figure 12

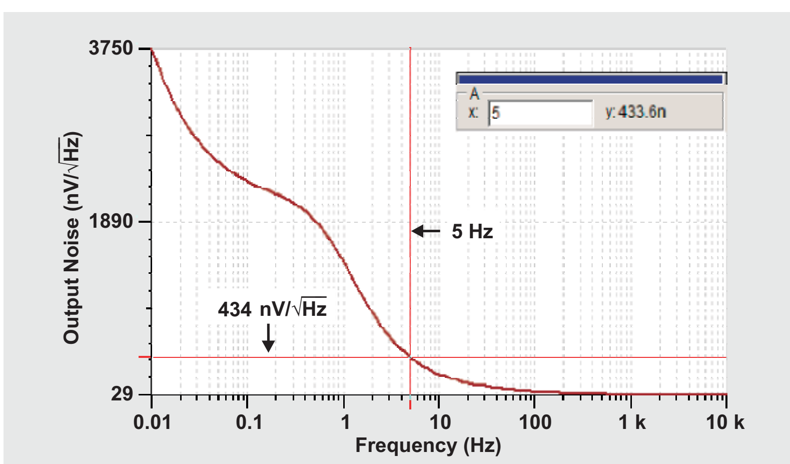
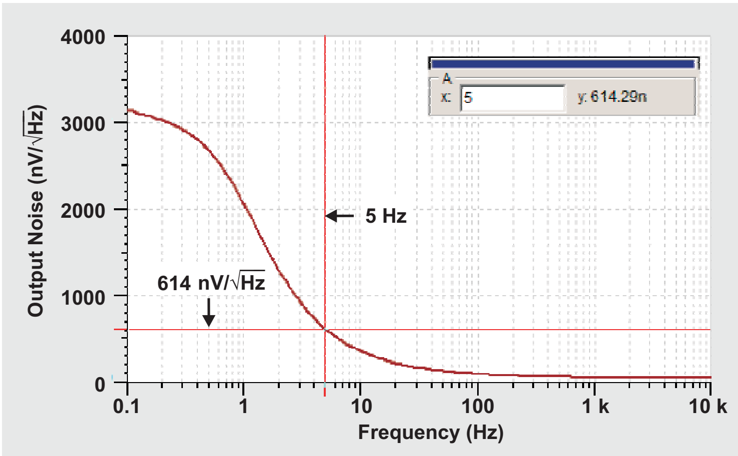


Figure 14. Simulation of output noise from circuit in Figure 12 with R_{FB} doubled and C_{FB} halved



close to the simulation result (see Figure 14). As expected, the output noise goes up. Nevertheless, doubling the resistance allows the capacitance to be divided by two, effectively doubling the gain (i.e., doubling the output signal). Even though R_{FB} is the dominant noise source, and increasing it increases its noise, an SNR improvement of 3 dB is realized because the doubled output signal far exceeds the added noise.

Other practical considerations

Creating equivalent larger resistors with a T network

When very large resistors in the feedback network are desired, it is tempting to create them by using a T network formed by smaller, more accessible components (see Figure 15). This practice is usually not recommended because the T network results in a large gain for the offset and noise, usually yielding a much worse SNR.

Using differential inputs

So far, the benefits of using differential inputs to reduce noise have been ignored. For simplicity, the amplifiers modeled have been analyzed as single-ended, but Figure 16 shows an improved configuration with differential inputs. This configuration has a double advantage:

1. It intrinsically has twice the gain of a circuit with a single-ended input (the charge gets integrated in C2 and C4), while noise increases only as a square-root function (i.e., the noise sources are uncorrelated).

Figure 15. T-network feedback circuit

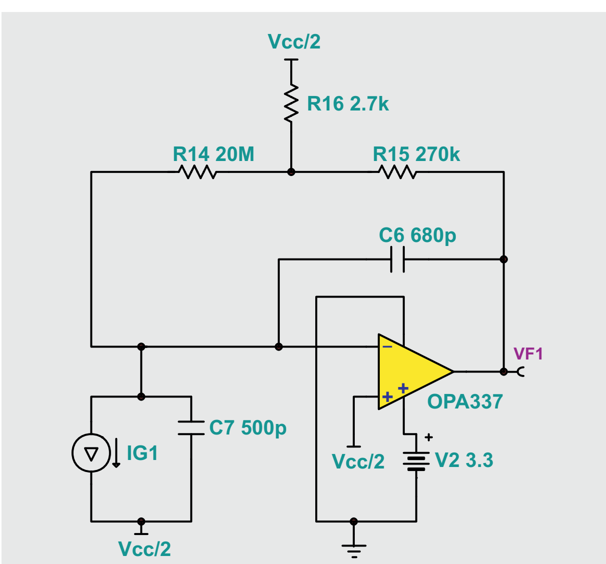
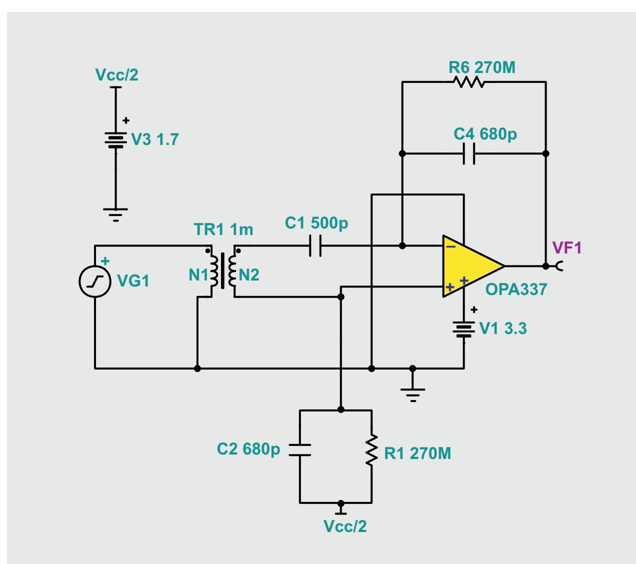


Figure 16. Improved circuit with differential inputs



2. The charge amplifier is a very sensitive (high-gain) circuit. Figure 17 shows that any capacitive coupling of an interferer (in this case the 60-Hz grid) with the input will effectively inject current. In the case of a single-ended amplifier, this means that one of the terminals injects current while the other goes to ground; i.e., the amplifier will just amplify the interferer. In the case of the differential input, common-mode signals applied to both terminals will cancel each other (assuming that the parasitics and feedback networks are the same). In Figure 18, notice the results of coupling to the 60-Hz grid with a single-ended input (blue trace) and how the 60-Hz common-mode noise is greatly reduced by differential inputs that cancel each other's interferer (yellow trace). For the purposes of this example, no effort was made to match the differential inputs beyond the 10% tolerances of the components.

Figure 17. Model of 60-Hz common-mode noise source to differential input amplifier

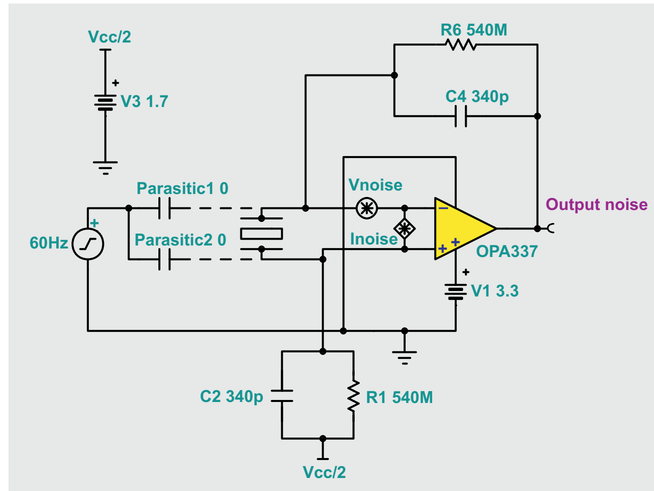
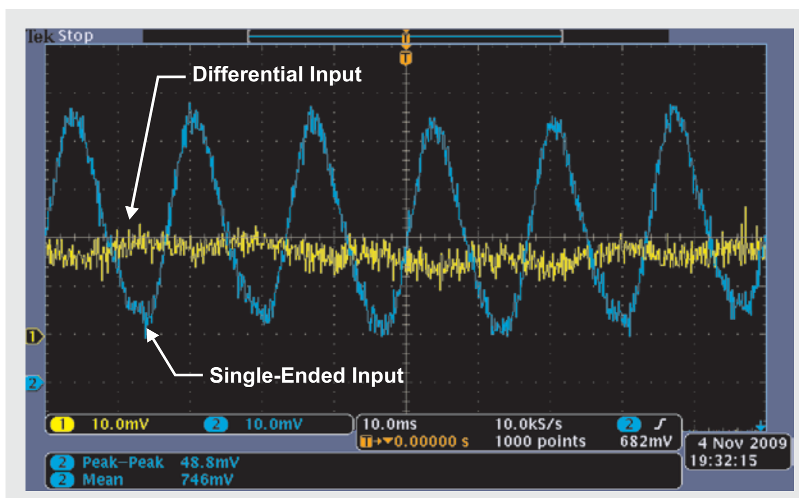


Figure 18. Differential amplifier nearly eliminates common-mode noise



Conclusion

Users can think of piezoelectric sensors as devices that output charge according to their deformation. As such, a charge amplifier is a good fit for this application. This article has presented some of the general rules of thumb to keep in mind when designing this circuit, such as increasing the feedback resistor as much as is practically possible, keeping an eye on the amplifier's input bias current, and using a differential structure. This article has also demonstrated the usefulness of conducting a theoretical analysis before attempting detailed simulations.

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