

ADXL50

calibration are used and there is a desire to eliminate trim potentiometers, the design should leave room at either supply rail to account for signal swing and or variations in initial zero g bias. For example, in the circuit in Figure 19, the initial zero g bias of ±250 mV will be reflected to the output by the gain of the R3/R1 network, resulting in an output offset of ±526 mV worst case. The offset, combined with a full-scale signal of 50 g, (+2.0 V) will cause the output buffer amplifier to saturate at the supply rail.

The full ±2.25 V output swing of the buffer amplifier can be utilized if the user is able to trim the zero-g bias to exactly 2.5 V. In applications where the full-scale range will be ±25 g or less, a bias trim such as that shown in Figure 20 will almost always be required.

VARYING THE OUTPUT SENSITIVITY AND 0 g LEVEL USING THE INTERNAL BUFFER AMPLIFIER
The uncommitted buffer amplifier may be used to change the output sensitivity to provide useful full-scale ranges of ±50 g and below. Table II provides recommended resistor values for several standard ranges down to ±10 g. As the full-scale range is decreased, buffer amplifier gain is increased, and the noise contribution as a percentage of full scale will also increase. For all ranges, the signal-to-noise ratio can be improved by reducing the circuit bandwidth, either by increasing the demodulator capacitor, C1, or by adding a post filter using the buffer amplifier.

Table II. Recommended Resistor Values for Setting the Circuit of Figure 20 to Several Common Full-Scale Ranges

FS (g)	Buffer Gain	SF in mV/g	R1	R3	R2
±50.0	2.11	40	49.9 k	105 k	100 k
±40.0	2.63	50	39.2 k	103 k	100 k
±30.8	3.42	65	40.2 k	137 k	100 k
±26.7	3.95	75	28.7 k	113 k	100 k
±20.0	5.26	100	26.1 k	137 k	100 k
±10.0	10.53	200	23.7 k	249 k	100 k

Note that the value of resistor R1 should be selected to limit the output current flowing into VPR to less than 25 µA (to provide a safety margin). For a “J” grade device, this current is equal to:

$$I_{PR} = \frac{(2.05\text{ V} - \text{The peak full-scale output voltage at } V_{PR}) - 1.8\text{ V}}{R1\text{ in ohms}}$$

For a ±50 g full-scale range, R1 needs to be 49.9 kΩ or larger in value; but at the lower full-scale g ranges, if the VPR swing is much less, then it is possible to use much lower resistance values. For this table, the circuit of Figure 20 is used, as a 0 g offset trim will be required for most applications. In all cases, it is assumed that the zero-g bias level is 2.5 V with an output span of ±2 V.

Note that for full scales below ±20 g the self-test is unlikely to operate correctly because the VPR pull-down current is not guaranteed to be large enough to drive R1 to the required -1.0 V swing. In these cases, the self-test command will cause VOUT to saturate at the rail, and it will be necessary to monitor the self-test at VPR. Self-test can remain operational at VPR for all g

ranges listed by keeping R1 > 49.9 kΩ, with the subsequent tradeoff that the required values for R3 will become very large. The user always has the option of adding external gain and filtering stages after the ADXL50 to make lower full-scale ranges. **Measuring Full-Scale Accelerations Less than ±5 g**
Applications, such as motion detection, and tilt sensing, have signal amplitudes in the 1 g to 2 g range. Although designed for higher full-scale ranges, the ADXL50 may be adapted for use in

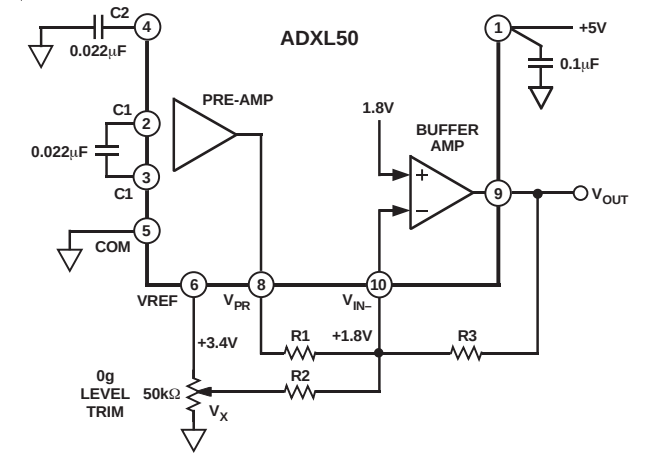


Figure 20. ADXL50 Circuit Using the Buffer Amplifier to Set the Output Scaling and 0 g Offset Level

low g applications; the two main design considerations are noise and 0 g offset drift (BH, KH grades recommended).

At its full 1 kHz bandwidth, the ADXL50 will typically exhibit 1 g p-p of noise. With ±50 g accelerations this is generally not a problem, but at a ±2 g full-scale level the signal-to-noise ratio will be very poor. However, reducing the bandwidth to 100 Hz or less considerably improves the S/N ratio. Figure 25 shows the relationship between ADXL50 bandwidth and noise.

The ADXL50 exhibits offset drifts that are typically 0.02 g per °C but which may be as large as 0.1 g per °C. With the buffer amplifier configured for a 2 g full scale, the ADXL50 will typically drift 1/2 of its full-scale range with a 50°C increase in temperature.

There are several cures for offset drift. If a dc response is not required, for example in motion sensing or vibration measurement applications, consider ac coupling the acceleration signal to remove the effects of offset drift. See the section on ac coupling.

Periodically recalibrating the accelerometer’s 0 g level is another option. Autozero or long term averaging can be used to remove long term drift using a microprocessor or the autozero circuit of Figure 29. Be sure to keep the buffer amplifier’s full-scale output range much larger than the measurement range to allow for the 0 g level drift.

CALCULATING COMPONENT VALUES FOR SCALE FACTOR AND 0 g SIGNAL LEVEL
The ADXL50 buffer’s scale factor is set by -R3/R1 (since the amplifier is in the inverter mode).

As an example, if the desired span is ± 2.0 V for a ± 50 g input, then $R3/R1$ should be chosen such that

$$R3/R1 = V_{OUT} Span / V_{PR} Span = 2.00/0.95 = 2.105 \quad (1)$$

where V_{PR} span is the output from the preamplifier and V_{OUT} span is the buffer amplifier's output, giving

$$R3 = 2.105 \times R1 \quad (2)$$

In noncritical applications, a resistor, $R2$, may simply be connected between V_{IN-} and common to provide an approximate 0 g offset level (see Figure 19). In this simplified configuration $R2$ is found using:

$$R2 = (1.8 \text{ V} \times R3) / (V_{OUT} @ 0 \text{ g} - 1.8 \text{ V})$$

When used with a trim potentiometer, as in Figure 20, resistor $R2$ sets the 0 g offset range and also sets the resolution of the offset trim. A value of 100 k Ω is typical. Increasing $R2$ above this value makes trimming the offset easier, but may not provide enough trim range to set V_{OUT} equal to ± 2.5 V for all devices.

To provide an output span of ± 2.00 V, with a 0 g output of ± 2.5 V, $R1$ could be set to the standard value of 49.9 k Ω and from Equation 2, $R3 = 105$ k Ω .

For Figure 20, the circuit transfer function is:

$$V_{OUT} = \left(\frac{R3}{R1} (1.8 \text{ V} - V_{PR}) \right) + \left(\frac{R3}{R2} (1.8 \text{ V} - V_X) \right) + 1.8 \text{ V}$$

The summing amplifier configuration allows noninteractive trimming of offset and span. Since V_{PR} is not always exactly 1.8 V at 0 g, it will contribute to output offset. Therefore, span must be trimmed first, followed by 0 g offset adjustment.

LOAD DRIVE CAPABILITIES OF THE V_{PR} AND BUFFER OUTPUTS

The V_{PR} and the buffer amplifier outputs are both capable of driving a load to voltage levels approaching that of the supply rail. However, both outputs are limited in how much current they can supply, affecting component selection.

V_{PR} Output

The V_{PR} pin has the ability to source current up to 500 μ A but only has a sinking capability of 30 μ A which limits its ability to drive loads. It is recommended that the buffer amplifier be used in most applications, to avoid loading down V_{PR} . In standard ± 50 g applications, the resistor $R1$ from V_{PR} to V_{IN-} is recommended to have a value greater than 50 k Ω to reduce loading effects.

Capacitive loading of the V_{PR} pin should be minimized. A load capacitance between the V_{PR} pin and common will introduce an offset of approximately 1 mV for every 10 pF of load. The V_{PR} pin may be used to directly drive an A/D input or other source as long as these sensitivities are taken into account. It is always preferable to drive A/D converters or other sources using the buffer amplifier (or an external op amp) instead of the V_{PR} pin.

Buffer Amplifier Output

The buffer output can drive a load to within 0.25 V of either power supply rail and is capable of driving 1000 pF capacitive

loads. Note that a capacitance connected across the buffer feedback resistor for low-pass filtering does not appear as a capacitive load to the buffer. The buffer amplifier is limited to sourcing or sinking a maximum of 100 μ A. Component values for the resistor network should be selected to ensure that the buffer amplifier can drive the filter under worst case transient conditions.

SELF-TEST FUNCTION

The digital self-test input is compatible with both CMOS and TTL signals. A Logic "1" applied to the self-test (ST) input will cause an electrostatic force to be applied to the sensor which will cause it to deflect to the approximate negative full-scale output of the device. Accordingly, a correctly functioning accelerometer will respond by initiating an approximate -1 volt output change at V_{PR} . If the ADXL50 is experiencing an acceleration when the self-test is initiated, the V_{PR} output will equal the algebraic sum of the two inputs. The output will stay at the self-test level as long as the ST input remains high and will return to the 0 g level when the ST voltage is removed.

A self-test output that varies more than $\pm 10\%$ from the nominal -1.0 V change indicates a defective beam or a circuit problem such as an open or shorted pin or component.

Operating the ADXL50's buffer amplifier at Gains > 2 , to provide full-scale outputs of less than ± 50 g, may cause the self-test output to overdrive the buffer into saturation. The self-test may still be used in the case, but the change in the output must then be monitored at the V_{PR} pin instead of the buffer output.

Note that the value of the self-test delta is not an exact indication of the sensitivity (mV/g) of the ADXL50 and, therefore, may not be used to calibrate the device for sensitivity error.

In critical applications, it may be desirable to monitor shifts in the zero-g bias voltage from its initial value. A shift in the 0 g bias level may indicate that the 0 g level has shifted which may warrant an alarm.

POWER SUPPLY DECOUPLING

The ADXL50 power supply should be decoupled with a 0.1 μ F ceramic capacitor from +5 V pin of the ADXL50 to common using very short component leads. For other decoupling considerations, see EMI/RFI section.

OSCILLATOR DECOUPLING CAPACITOR, C2

An oscillator decoupling capacitor, C2, is used to remove 1 MHz switching transients in the sensor excitation signal, and is required for proper operation of the ADXL50. A ceramic capacitor with a minimum value of 0.022 μ F is recommended from the oscillator decoupling capacitor pin to common. Small amounts of capacitor leakage due to a dc resistance greater than 1 M Ω will not affect operation (i.e., a high quality capacitor is not needed here). As with the power supply bypass capacitor, very short component leads are recommended. Although 0.022 μ F is a good typical value, it may be increased for reasons of convenience, but doing this will not improve the noise performance of the ADXL50.

ADXL50

DEMODULATOR CAPACITOR, C1

The demodulator capacitor is connected across Pins 2 and 3 to filter the demodulated signal from the sensor beam and to set the bandwidth of the force balance control loop. This capacitor may be used to approximately set the bandwidth of the accelerometer. A capacitor is always required for proper operation.

The frequency response of the ADXL50 exhibits a single pole roll-off response whose nominal 3 dB frequency is set by the following equation:

$$f_{3\text{ dB}} = (28.60/C1 \text{ in } \mu\text{F}) \pm 40\%$$

A nominal value of 0.022 μF is recommended for C1. In general, the design bandwidth should be set 40% higher than the minimum desired system bandwidth due to the $\pm 40\%$ tolerance.

A minimum value of 0.015 μF is required, (over temperature and system life), to prevent device instability or oscillation. The demodulation capacitor should be a low leakage, low drift ceramic type with an NPO (best) or X7R (good) dielectric.

In general, it's best to use the recommended 0.022 μF capacitor across the demodulator pins and perform any additional low-pass filtering using the buffer amplifier. Using a large denominator capacitor for low-pass filtering has the disadvantage that the capacitive sensor will be slow to respond to rapid changes in acceleration and, therefore, the full shock survivability of the device could be compromised. The use of the buffer for low-pass filtering generally results in smaller capacitance values and better overall performance. It is also a convenient and more precise way to set the system bandwidth. Post filtering allows bandwidth to be controlled accurately by component selection and avoids the $\pm 40\%$ demodulation tolerance. Note that signal noise is proportional to the square root of the bandwidth of the ADXL50 and may be a consideration in component selection—see section on noise.

Care should be taken to reduce or eliminate any leakage paths from the demodulator capacitor pins to common or to the +5 V pin. Even a small imbalance in the leakage paths from these pins will result in offset shifts in the zero-g bias level. As an example, an unbalanced parasitic resistance of 30 M Ω from either demodulator pin to ground will result in an offset shift at V_{PR} of approximately 50 mV. Conformal coating of PC boards with a high impedance material is recommended to avoid leakage problems due to aging or moisture.

REDUCING THE AVERAGE POWER CONSUMPTION OF THE ADXL50

The ADXL50 is a versatile accelerometer that can be used in a wide variety of applications. In some battery powered applications, such as shipping recorders, power consumption is a critical parameter. The ADXL50 typically draws 10 mA current from a 5 V power supply which may exceed the power budgeted for the accelerometer.

For such applications, the ADXL50 can be successfully power cycled, where the power is turned on only during the period when data is sampled. Figure 21 illustrates the power-on settling of the ADXL50 during cycling where the output amplifier has a gain of one with no filtering. The settling time-constant is approximately 0.12 ms, waiting 1 ms before sampling ensures maximally accurate readings.

For example, to reduce the average power to 5 mW from its typical 50 mW, the power should be on 10% of the time. With the power on for 1 ms and off for 9 ms, a maximum sample rate of 100 Hz is achievable. Further reduction in average power can be realized with lower sample rates.

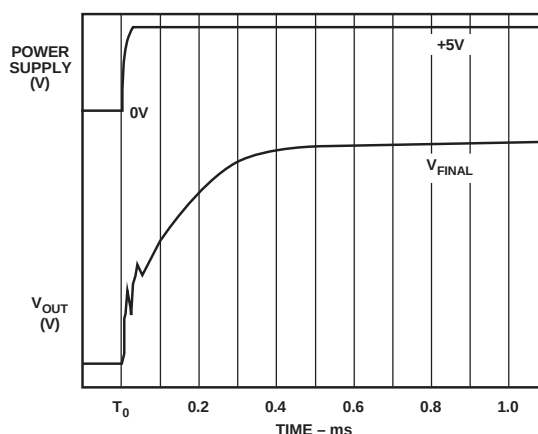


Figure 21. Power-On Settling Time when Power Cycling

SYSTEM BANDWIDTH CONTROL AND POST FILTERING

Unlike piezoresistive sensors, the resonant frequency of the ADXL50's capacitive sensor element is typically greater than 20 kHz and does not limit the useful bandwidth of the device. Usually, the resonant frequency of the beam appears as a peak in the bandwidth response at approximately 24 kHz with a Q of 3 to 4, as shown in Figure 22.

When using the recommended 0.022 μF demodulator capacitor, be advised that the nominal 1300 Hz pole it establishes within the device can vary $\pm 40\%$. Therefore, if additional low-pass filtering is used—at frequencies much above 600 Hz—the two poles may interact and result in a net circuit bandwidth that is lower than expected.

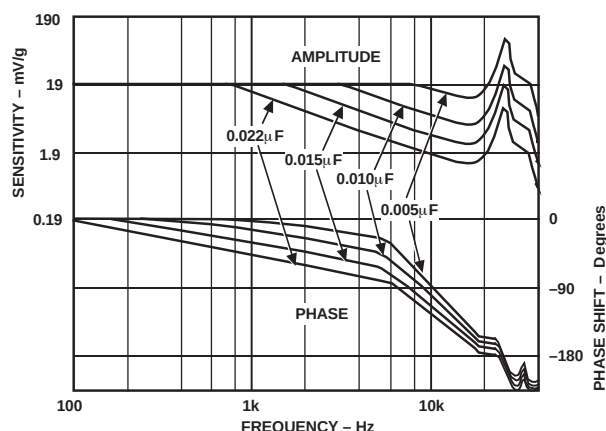


Figure 22. Frequency Response of the ADXL50 for Various Demodulator Capacitors

RECOMMENDED COMPONENT VALUES FOR VARIOUS FULL SCALE RANGES AND A 300Hz BANDWIDTH

FULL SCALE	mV per g	3dB BW (Hz)	R1a kΩ	R1b kΩ	R3 kΩ	R2 kΩ	C4 μF
±10 g	200	300	5	21.5	249	100	0.0022
±20 g	100	300	5	23.7	137	100	0.0039
±40 g	50	300	10	34	105	100	0.0056
±50 g	40	300	10	45.3	105	100	0.0056

$$3\text{dB BW} = \frac{1}{2\pi R_3 C_4}$$

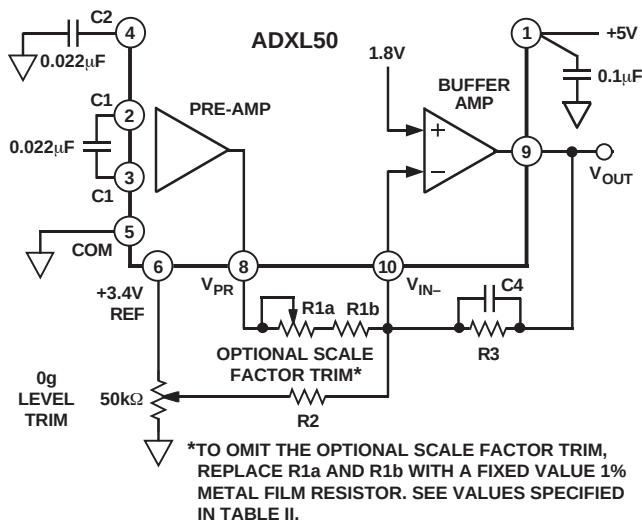


Figure 23. Using the Buffer Amplifier to Provide One Pole Post Filtering Plus Scale Factor and 0 g Level Trimming

ONE POLE POST FILTERING

Figure 23 shows the ADXL50 buffer amplifier connected to provide one pole post filtering, 0 g offset trimming, and output scaling. The table included with the figure lists practical component values for various full-scale g levels and approximate circuit bandwidths. For bandwidths other than those listed, use the formula:

$$\text{Capacitor } C_4 \text{ in Farads} = 1/(2\pi \times R_3 \text{ in Ohms} \times 3 \text{ dB BW in Hertz})$$

or simply scale the value of capacitor C4 accordingly; i.e., for a ±20 g application with a 50 Hz bandwidth, the value of C4 will

need to be twice as large as its 100 Hz value or $0.012 \mu\text{F} \times 2 = 0.024 \mu\text{F}$. The closest standard value of $0.022 \mu\text{F}$ should then be used.

TWO POLE POST FILTERING

Figure 24 shows a circuit which uses the ADXL50's buffer amplifier to provide two pole post filtering. An AD820 external op amp allows noninteractive adjustment of 0 g offset and scale factor. Component values for the two pole filter were selected to operate the buffer at unity gain with a Q of one.

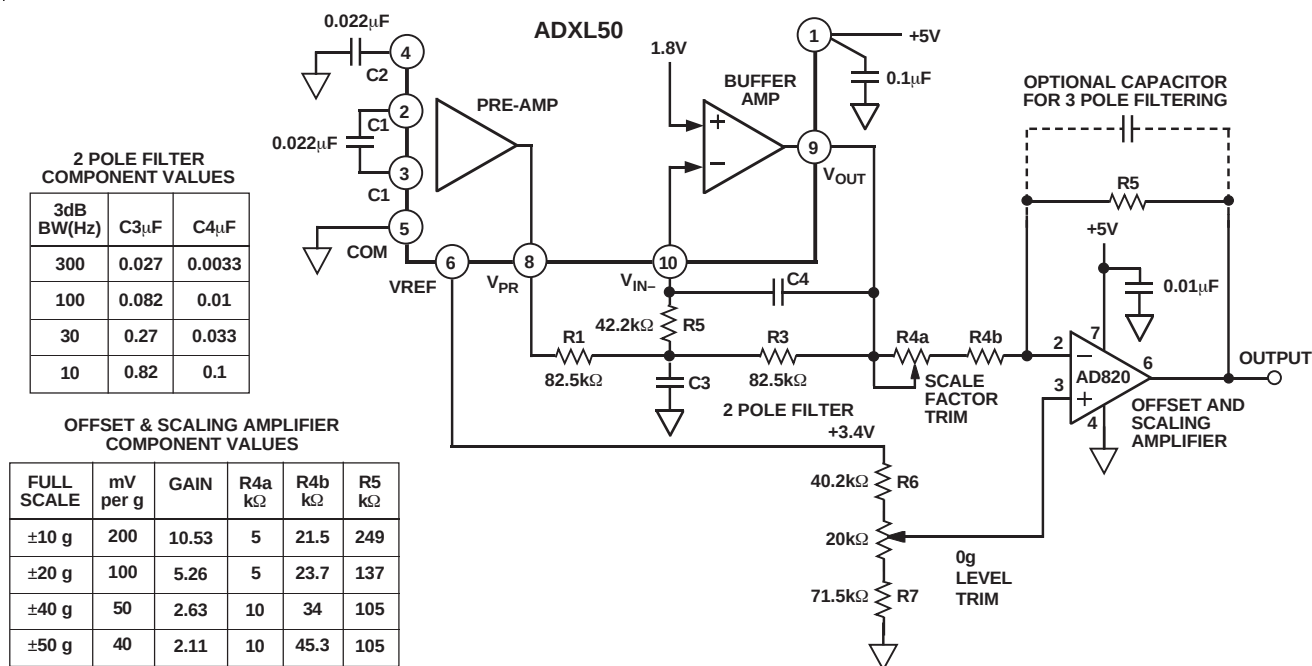


Figure 24. Circuit Providing Two Pole Post Filtering and 0 g Offset and Scale Factor Trimming

ADXL50

Capacitors C3 and C4 are chosen to provide the desired 3 dB bandwidth. Component values are specified for bandwidths of 10 Hz, 30 Hz, 100 Hz, and 300 Hz. For other 3 dB bandwidths simply scale the capacitor values; i.e., for a 3 dB bandwidth of 20 Hz, divide the 10 Hz bandwidth numbers by 2.0. The nominal buffer amplifier output will be +1.8 V ± 19 mV/g. Note that the ADXL50's self-test will be fully functional since the buffer amplifier is operated at unity gain and resistor R1 is large. The external op amp offsets and scales the output to provide a +2.5 V ± 2 V output over a wide range of full-scale g levels. The external op amp may be omitted in high g, low gain applications.

NOISE CONSIDERATIONS

The output noise of the ADXL50 scales with the square root of its bandwidth. The noise floor may be reduced by lowering the bandwidth of the ADXL50 either by increasing the value of the demodulator capacitor or by adding an external filter.

The typical rms noise of the ADXL50J with a bandwidth of 100 Hz and a noise density of 125 μV/√Hz is estimated as follows:

Noise (rms) = (125 μV/√Hz) √100 = 1.25 mV rms

Peak-to-peak noise may be estimated with the following equation:

Noise p-p = (6.6) Noise rms

Peak-to-peak noise is thus estimated at 8.25 mV or approximately 0.4 g p-p. The ADXL50 noise is characteristic of white noise. Typical rms and p-p noise for various 3 dB bandwidths is estimated in Figure 25.

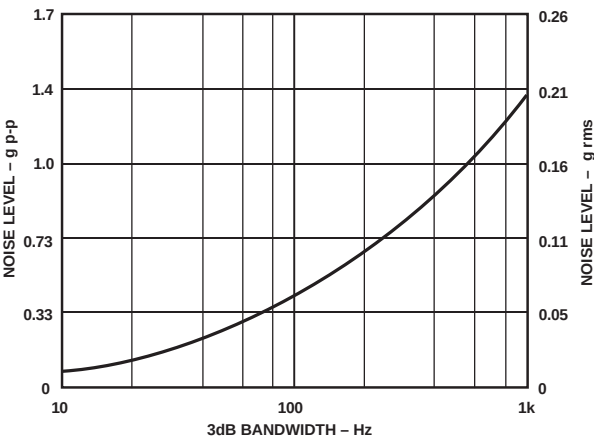


Figure 25. ADXL50 Noise Level and Resolution vs. -3 dB Bandwidth

Because the ADXL50's noise is for all practical purposes Gaussian in amplitude distribution, the highest noise amplitudes have the smallest (yet nonzero) probability. Peak-to-peak noise is, therefore, difficult to measure and can only be estimated due to its statistical nature. Table III is useful for estimating the probabilities of exceeding various peak values, given the rms value.

Table III.

Nominal Peak-to-Peak Value	% of Time that Noise will Exceed Nominal Peak-to-Peak Value
2.0 × rms	32%
3.0 × rms	13%
4.0 × rms	4.6%
5.0 × rms	1.2%
6.0 × rms	0.27%
6.6 × rms	0.1%
7.0 × rms	0.046%
8.0 × rms	0.006%

AC COUPLING VPR TO BUFFER INPUT

If a dc response is not required, as in applications such as motion detection or vibration measurement, then ac coupling should be considered. In low g applications, the output voltage change due to acceleration is small compared to the 0 g offset voltage drift. Because ac coupling removes the dc component of the output, the preamp output signal may be amplified considerably without increasing the 0 g level drift. The most effective way to ac couple the ADXL50 is between the preamp output at VPR and the buffer input, VIN-, as shown in Figure 26.

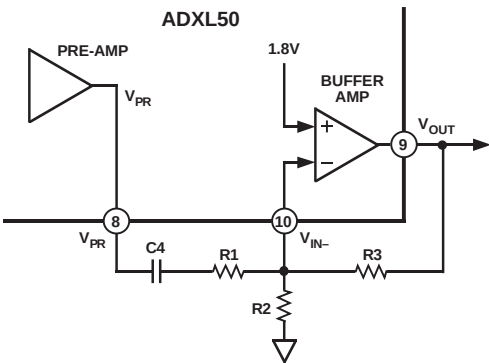


Figure 26. AC Coupling the VPR Output to the Buffer Input

Using this configuration, the system's ac response is now rolled off—at the low frequency end at FL, and at the high frequency end at FH. The normalized frequency response of the system can be seen in Figure 27.

The low frequency roll-off, FL, due to the ac coupling network is:

FL = 1/(2 π R1 C4)

The high frequency roll-off FH is determined by the dominant pole of the system which is controlled by either the demodulator capacitor and its associated time-constant or by a dominant post filter.

As a consequence of ac coupling, any constant acceleration component gravity will not be detected (because this too is a dc voltage present at the VPR output). The self-test feature, if used, must be monitored at VPR, rather than at the buffer output.

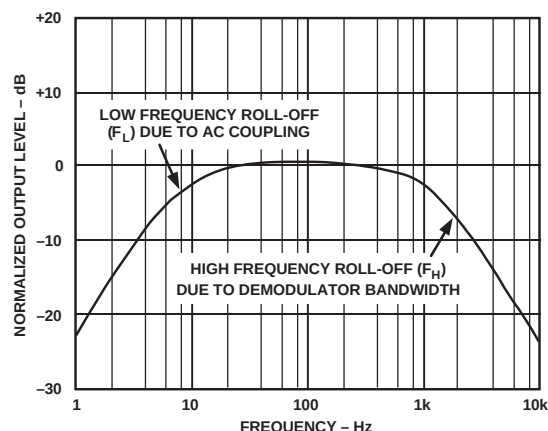


Figure 27. Normalized Output Level vs. Frequency for a Typical Application Using AC Coupling Between V_{PR} and Buffer Amplifier

MINIMIZING EMI/RFI

The architecture of the ADXL50 and its use of synchronous demodulation make the device immune to most electromagnetic (EMI) and radio frequency (RFI) interference. The use of synchronous demodulation allows the circuit to reject all signals except those at the frequency of the oscillator driving the sensor element. However, the ADXL50 does have a sensitivity to RFI that is within ± 5 kHz of the internal oscillator's nominal frequency of 1 MHz. The internal oscillator frequency will exhibit part to part variation in the range of 0.6 MHz to 1.4 MHz.

In general the effect is difficult to notice as the interference must match the internal oscillator within ± 5 kHz and must be large in amplitude. For example: a 1 MHz interference signal of 20 mV p-p applied to the +5 V power supply pin will produce a 200 mV p-p signal at the V_{PR} pin if the internal oscillator and interference signals are matched exactly. If the same 20 mV interference is applied but 5 kHz above or below the internal oscillator's frequency, the signal level at V_{PR} will only be 20 mV p-p in amplitude.

Power supply decoupling, short component leads (especially for capacitors C1 and C2), physically small (surface mount, etc.) components and attention to good grounding practices all help to prevent RFI and EMI problems. Please consult the factory for applications assistance in instances where this may be of concern.

SELF-CALIBRATING THE ADXL50

If a calibrated shaker is not available, both the 0 g level and scale factor of the ADXL50 may be easily set to fair accuracy by using a self-calibration technique based on the 1 g (average) acceleration of the earth's gravity. Figure 28 shows how gravity

and package orientation affect the ADXL50's output (TO-100 package shown). Note that the output polarity is that which appears at V_{PR} ; the output at V_{OUT} will have the opposite sign. With its axis of sensitivity in the vertical plane, the ADXL50 should register a 1 g acceleration, either positive or negative, depending on orientation. With the axis of sensitivity in the horizontal plane, no acceleration (0 g) should be indicated.

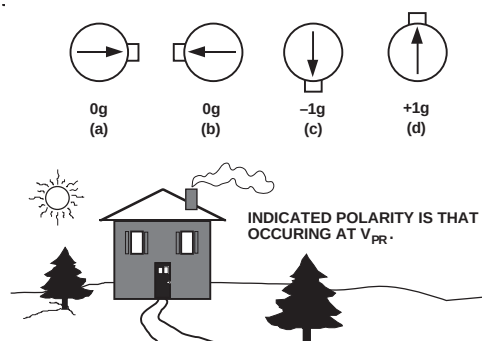


Figure 28. Using the Earth's Gravity to Self-Calibrate the ADXL50

To self-calibrate the ADXL50, place the accelerometer on its side with its axis of sensitivity oriented as shown in "a." The 0 g offset potentiometer, R_t , is then roughly adjusted for midscale: +2.5 V at the buffer output. If the optional scale factor trimming is to be used, it should be adjusted next.

Next, the package axis should be oriented as in "c" (pointing down) and the output reading noted. The package axis should then be rotated 180° to position "d" and the scale factor potentiometer, R_{1a} , adjusted so that the output voltage indicates a change of 2 g's in acceleration. For example, if the circuit scale factor at the buffer output is 100 mV per g, then the scale factor trim should be adjusted so that an output change of 200 mV is indicated.

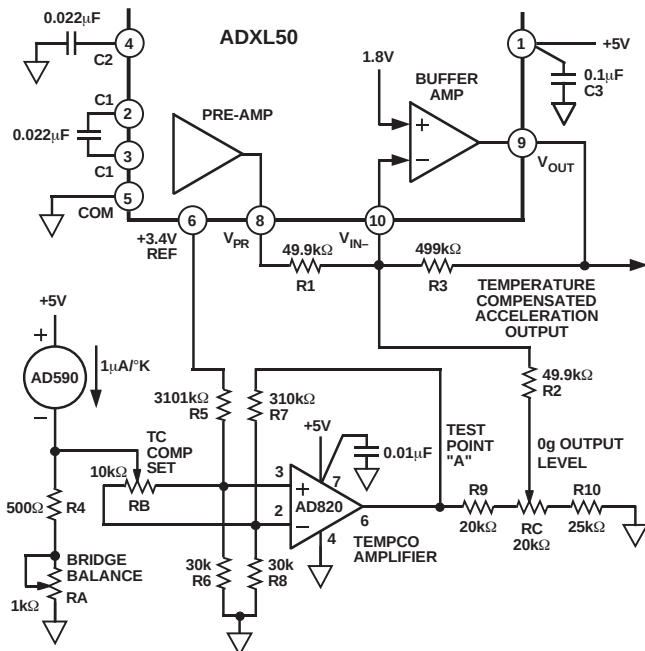
Adjusting the circuit's scale factor will have some effect on its 0 g level so this should be readjusted, as before, but this time checked in both positions "a" and "b." If there is a difference in the 0 g reading, a compromise setting should be selected so that the reading in each direction is equidistant from +2.5 V. Scale factor and 0 g offset adjustments should be repeated until both are correct. Temporarily placing a capacitor across the buffer amplifier's feedback resistor will reduce output noise and so aid in trimming the device. Note that, for high full-scale g ranges, ± 2 g may be a very small fraction of the full-scale range and device nonlinearity will, therefore, affect the circuit's high g level accuracy.

ADXL50

Compensating for the 0 g Drift of the ADXL50 Accelerometer
The circuit of Figure 29 provides a linear temperature compensation for the ADXL50. Figure 30 shows the 0 g drift over temperature for a typical ADXL50 with and without this circuit. As shown by Figure 30, the linear portion of the drift curve has been subtracted out. In effect, the curve has been rotated counterclockwise until it is horizontal, leaving just the bow of the curve: that portion which is not linear. As shown by Figure 30, over a +25°C to +70°C range, a 10× reduction in drift is achieved.

The circuit of Figure 29 is essentially a temperature sensor coupled to a Wheatstone bridge. The AD590 provides a 1 $\mu\text{A}/^\circ\text{K}$ current output whose voltage scale factor is set by resistor R_A . The bridge circuit subtracts out the nominal 298 mV output of the AD590 at +25°C and leaves only the change in temperature, which is what is needed. Without the bridge, the 298 mV room temperature “offset” would “swamp” the much smaller change in output with temperature.

Resistors R_5 and R_6 form a resistor divider (one half of the bridge) which divides down the +3.4 V reference output of the ADXL50 to 0.3 V which appears at the noninverting input of the AD820 op amp. Resistors R_7 and R_8 form the other half of the bridge, and because they have the same ratio as R_5 and R_6 , the op amp will have a +3.4 V output at room temperature.



CALIBRATION PROCEDURE:

AT T_{MIN} OR LOWER TEMP CAL POINT...

1. SET R_B ALL THE WAY TO ONE SIDE.
2. ADJUST R_A FOR +3.4V AT TEST POINT "A."
3. SET R_C FOR +2.5V V_{OUT} (AT PIN 9 OF ADXL50).
4. TEMPORARILY CONNECT A 1.5k Ω RESISTOR BETWEEN THE CENTER OF R_B AND GROUND.
5. ADJUST R_B FOR +2.5V AT V_{OUT} .
6. REMOVE THE 1.5k Ω RESISTOR. V_{OUT} SHOULD NOT CHANGE.
7. GO TO T_{MAX} OR HIGH TEMP CAL POINT.
8. READJUST R_B FOR +2.5V.
9. CALIBRATION COMPLETE.

Figure 29. ADXL50 0 g Drift Compensation Circuit

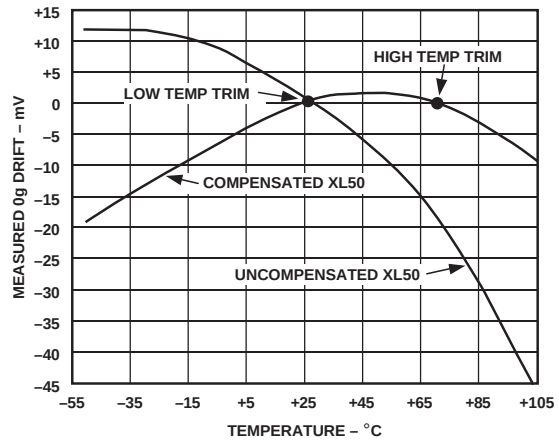


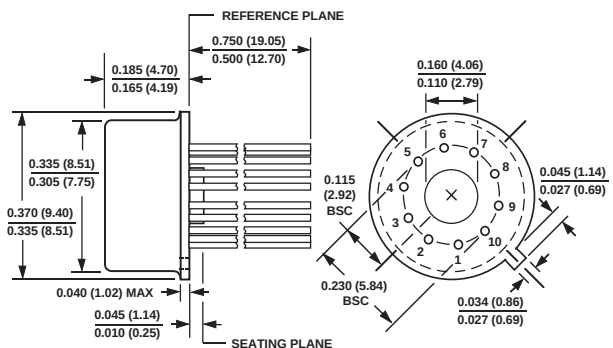
Figure 30. ADXL50 0 g Drift With and Without the Compensation Circuit of Figure 29

ADXL50 Application Literature Available

Contact the Analog Devices Literature Center for a copy of publication number G2112, the Accelerometer Application Guide. This includes all current application notes and data sheets for Analog Devices' accelerometers.

OUTLINE DIMENSIONS

Dimensions shown in inches and (mm).



C1808b-5-3/96

PRINTED IN U.S.A.

CERTIFICATE OF SERVICE (37 C.F.R. § 42.6(e))

I certify that the above-captioned **DECLARATION OF INGRID HSIEH-YEE, PH.D.** was served in its entirety on May 6, 2025 upon counsel of record in this proceeding via electronic mail:

Brenda Entzminger
Reg. No. 76,896
Corey Johanningmeier, Back-up Counsel
(*pro hac vice*)
Bunsow De Mory LLP
701 El Camino Real
Redwood City, CA 94063
Telephone: 650-351-7248
Facsimile: 415-426-4744
bentzminger@bdiplaw.com
cjohanningmeier@bdiplaw.com

Dated: May 6, 2025

/ Matthew J. Brigham /
Matthew J. Brigham
Reg. No. 44,047