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IEEE SOLID-STATE SENSOR AND ACTUATOR

TECHNICAL DIGEST



IEEE

SOLID-STATE SENSOR AND ACTUATOR

Workshop



Sponsored by the IEEE Electron Devices Society

1992

Hilton Head Island, South Carolina
June 22-25

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1992 IEEE SOLID-STATE SENSOR AND ACTUATOR WORKSHOP
June 22-25, 1992
Marriott Hotel, Hilton Head Island, SC

PROGRAM

Sunday, June 21

7 - 9 PM WELCOME RECEPTION with no host bar

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David S. Eddy - Chairman

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SURFACE MICROMACHINED, DIGITALLY FORCE-BALANCED ACCELEROMETER WITH INTEGRATED CMOS DETECTION CIRCUITRY

WeiJie Yun, Roger T. Howe, and Paul R. Gray
University of California at Berkeley
Department of Electrical Engineering and Computer Sciences
Berkeley Sensor & Actuator Center
Berkeley, California 94720

ABSTRACT

A surface micromachined, force-balanced capacitive accelerometer is designed and fabricated with on-chip detection circuitry. It has demonstrated the modular integration of CMOS with polysilicon surface micromachining. The capacitive sensing structure consists of two polysilicon layers with the sensing and feedback electrodes underneath and the suspended plate as proof mass, and the sensing axis is perpendicular to the substrate. A unity gain buffer with low input capacitance is designed to actively drive (bootstrap) the ground plane to minimize the parasitic capacitance. Σ - Δ modulation technique is employed in the feedback control loop, in which the mechanical proof mass is used to replace the two integrators in the Σ - Δ loop. This approach provides large dynamic range and directly digital output. To reduce the temperature sensitivity, package-induced stress, and residual stress in the spring, two accelerometers with different proof mass are built side by side and their digital outputs are subtracted. The readout circuits have been characterized successfully. The accelerometer is tested in the self-testing mode and its DC and transient open loop performance have been measured.

INTRODUCTION

Silicon microaccelerometers are a rapidly expanding sensor market segment, with automotive applications driving the development of commercial products [1, 2]. The promise of silicon microaccelerometers is their potential low cost through the application of batch fabrication techniques. First-generation applications will require low-to-intermediate sensitivity: air-bag deployment (50-100g) and chassis control (1-5g) [3]. Microaccelerometers of higher sensitivity (sub milli-g) could be useful for automobile position localization in automated highways, as well as in aerospace navigation systems. As a research vehicle, a low-cost, high performance microaccelerometer is attractive for stimulating new approaches to the micromechanical sense element, position sensing structures and interface electronics, and the overall system architecture, as well as for identifying basic limits and tradeoffs. This paper describes research in the design, fabrication, simulation, and initial self-testing of a high-resolution (sub milli-g), high dynamic range microaccelerometer.

Surface micromachining has clear advantages for fabricating the suspension and proof mass for low-cost microaccelerometers. Thin-film microstructures have lateral dimensions on the order of 100-500 μm -- much smaller than a typical integrated-circuit die size of about 2.5 mm square. Multiple sensing elements to provide ratioed response, auto-ranging, or multi-axis sensing are feasible on a single chip. In contrast, proof masses fabricated by anisotropic etching through the silicon wafer are much larger, leading to a larger, more expensive chip. For surface microstructures, overrange protection is provided by the substrate in one direction, and can be supplied by other micromachined limit stops in other directions, if

necessary. There is no need for one or more capping wafers, as is the case for conventional bulk-micromachined sensing elements.

Capacitive measurement of the proof mass position is naturally suited to surface-micromachined sense elements, given that vertical or lateral gaps on the order of 1 μm are quite feasible. However, sense capacitances are on the order of 100 fF - 1 pF, with on-chip parasitic capacitances which can be substantially larger. The accurate measurement of the sense capacitance is much easier with an on-chip MOS buffer amplifier. Integration of surface microstructures with on-chip MOS electronics using a unique, merged fabrication sequence was demonstrated in 1984 [4, 5]. Recently, a commercial 50 g accelerometer has been reported which merges a single-level polysilicon surface micromachining process with a proprietary Bi-MOS precision analog IC process [6]. This monolithic accelerometer uses a force-balance approach, in which an analog feedback loop applies electrostatic forces to null deflections due to acceleration. Another monolithic accelerometer uses an open-loop design, implemented using an electroformed metal sense element fabricated after standard CMOS capacitive detection and signal-conditioning electronics [7]. Finally, a second polysilicon accelerometer is described for the first time at this workshop [8].

The microaccelerometer described here also uses a polysilicon suspension and proof mass, integrated with CMOS sense and control circuitry. However, we use a modular fabrication sequence, which has been developed in order to provide a generic solution to integrating polysilicon micromechanics with IC processes [9, 10]. In order to achieve linearity and high dynamic range, the proof mass position is sensed capacitively and then input to a one-bit quantizer, which provides a high-frequency pulse train. After digital processing, a 2-bit signal applies electrostatic force to the sense element, nulling its deflection. The suspended proof mass supplies a double integration in the feedback loop, by converting input acceleration into position. Digital force-balance schemes, based on Σ - Δ modulation technique [11], are attractive for achieving high dynamic range and good linearity [12]. Moreover, the asymmetrical feedback electrode layout in the research vehicle makes analog feedback approaches impractical, since they depend on symmetry for cancellation of non-linearities in the electrostatic force. The proof mass is levitated with a DC force generated by interdigitated comb fingers, whereas parallel-plate capacitors are used to provide the negative feedback force. Finally, we exploit the inherent matching of mechanical and electrical parameters between adjacent structures by using a dual-channel architecture, in order to suppress built-in offsets and to minimize long-term drifts.

After describing the modular fabrication process, we describe the sense element design, electronic interface, and digital feedback loop architecture. Experimental results are reported for the open-loop response of one channel to simulate accelerations generated by pulsing feedback and levitation electrodes.

FABRICATION TECHNOLOGY: THE MICS PROCESS

Surface micromachining of polysilicon structures utilizes standard integrated circuit materials and processes, enabling it to be integrated with an MOS fabrication sequence in several ways [5, 6, 13]. By interleaving the two processes -- using MOS steps in the microstructure process and vice versa -- it is possible to develop an optimized, custom process. However, this option is open only to those with control over an IC fabrication line.

In order to minimize the barrier to fabricating polysilicon structures together with CMOS integrated circuits, we have investigated a modular process in which the highly optimized CMOS process is carried out with no modification (making it possible to use a silicon foundry for this part of the process.) In order to raise the temperature ceiling for subsequent sensor processing, we replace aluminum with tungsten for the CMOS interconnect layer and add a TiN/TiSi₂ diffusion barrier at metal/silicon contacts [9, 10]. Low-pressure CVD oxide and nitride films are then deposited to isolate the CMOS from the polysilicon microstructure fabrication steps. The process is called the **Modular Integration of CMOS with microStructures**, or MICS.

The initial IC process is a conventional 3 μ m design rule, p-well, double-poly, single metal CMOS technology used at UC Berkeley [14]. Tungsten is selected as the material for metallization among other refractory metals for its close matching in thermal coefficient of expansion to that of silicon and its relatively low resistivity. Since tungsten will react with silicon at the contact at around 600°C, a titanium nitride layer is used as a barrier layer at the contact using a self-aligned process. After the contact opening, titanium is sputtered and rapid-thermal annealed at 600°C in nitrogen ambient for 30 seconds to form a titanium compound layer at the contact. The unreacted titanium layer is removed in a 3:1 NH₄:H₂O₂ solution selectively. The contact is then sintered by RTA at 1000°C in nitrogen for 10 seconds and titanium nitride and titanium silicide is formed at the contact. It has been shown that MOS transistors will not have any noticeable degradation for temperatures below 850°C

with this metallization technique [9]. Thus, polysilicon and nitride films can be deposited without degrading the circuit performance.

Figure 1 shows the cross section of CMOS integrated with surface micromachined structures fabricated in MICS process. After the completion of the CMOS process with tungsten metallization, a 3000Å LTO is deposited and the wafer is then encapsulated with a 2000Å LPCVD silicon nitride layer to protect the CMOS circuitry from subsequent etching steps in the sensor process. This nitride layer also serves as the starting layer of the microstructures. After etching the contacts in the nitride and oxide layer, the first polysilicon layer in the sensor process is deposited and forms the bottom electrodes for sensing and feedback capacitors, as well as the ground plane. This in-situ doped polysilicon layer (3500Å) is also used as the interconnection between sensor and the readout circuits. Note that instead of contacting with the tungsten layer directly, this polysilicon layer is contacted to capacitor poly which is used as a "jumper" between the first polysilicon layer and tungsten. This is to prevent any contamination in the polysilicon tube due to the out-diffusion of the mobile ions in the sputtered tungsten film. A 1 μ m phosphosilicate glass (PSG) layer is then deposited and used as the sacrificial layer. The second in-situ doped polysilicon layer is 2 μ m thick and forms the proof mass, the suspensions, and the levitation comb fingers. Before the etching of the second polysilicon, a 0.5 μ m PSG is deposited. The structural polysilicon is now sandwiched by PSG layers top and bottom so that stress gradient in the polysilicon is minimized after the stress anneal. Polysilicon stress is typically annealed by a 1050°C, an hour furnace anneal [15]. This is not permissible in MICS process since CMOS is completed. Stress anneal using rapid thermal processing is an alternative to the furnace annealing. Experiments have shown that residual stress in the polysilicon structures can be reduced after a 1 minute rapid-thermal annealing at 1000°C in nitrogen ambient. Higher temperature anneal can result better stress release, but blistering of the nitride layer is observed. After the stress anneal, the structural polysilicon is etched and the sacrificial oxide is removed in a buffered HF solution to release the structure from the substrate. The MICS is now completed and the wafers are ready to be diced and packaged.

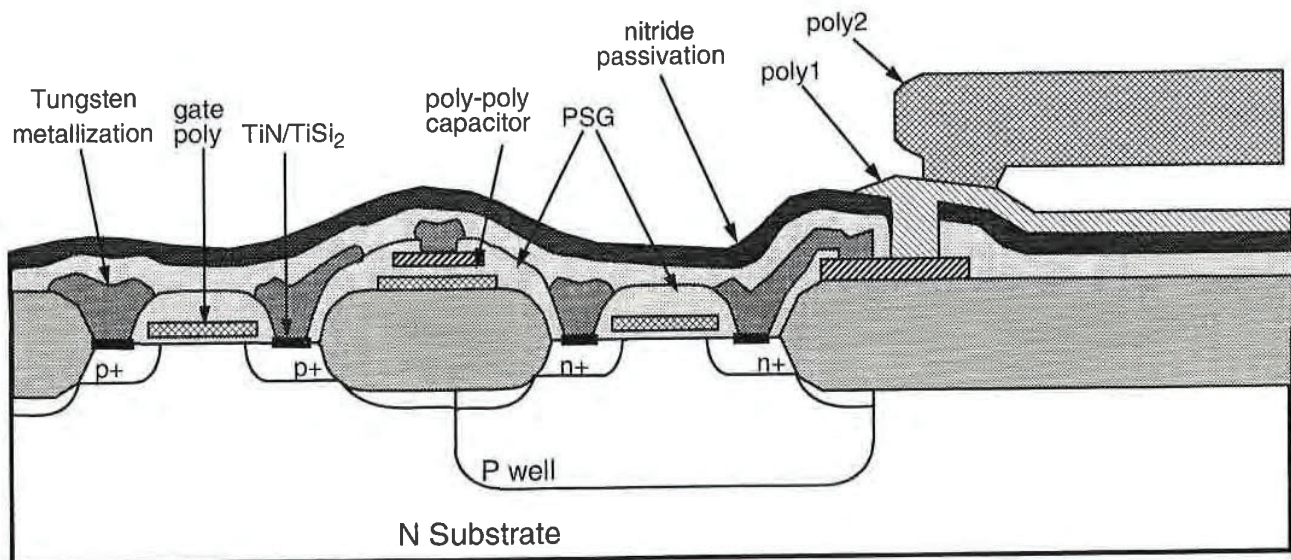


Fig. 1. Cross sectional view of CMOS and microstructure fabricated in MICS process.

SENSOR DESIGN

The capacitive sensing element is fabricated using surface micromachining technique and consists of two layers of in-situ doped polysilicon, shown in Fig. 2. The first polysilicon layer defines the sensing electrode, the feedback electrodes, and the ground plane. The bottom sensing electrode is $100\mu\text{m}$ on each side and is located at the center underneath the proof mass. The feedback electrodes are $25\mu\text{m} \times 25\mu\text{m}$ and are located symmetrically around the sensing electrode to avoid exciting higher-mode vibrations. The ground plane is bootstrapped to minimize the parasitic capacitance on the proof mass. The second polysilicon layer forms the proof mass, the interdigitated fingers for levitating, and the folded suspension. The proof mass is $2\mu\text{m}$ thick and $500\mu\text{m}$ on each side. The gap between the two polysilicon layers is $1\mu\text{m}$. Arrays of etch holes are used on the movable plate to help accelerate the final release etch. The suspension springs are anchored on the first polysilicon layer, which is connected to the readout electronics.

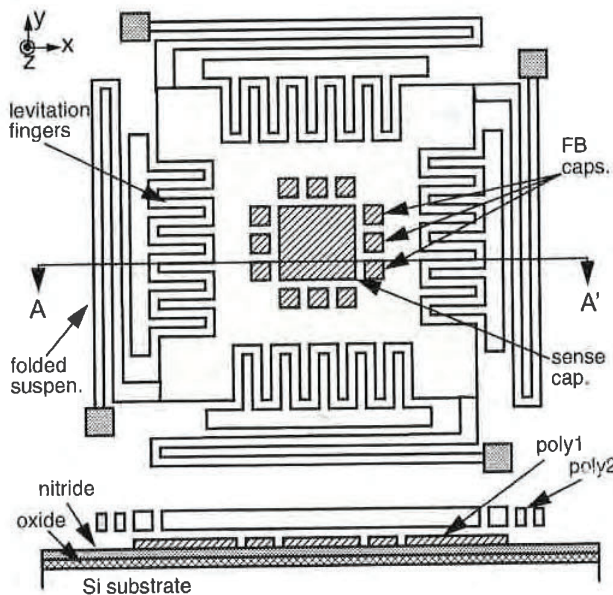


Fig. 2. Plane and cross sectional view (not to scale)

Folded suspensions are employed in the design and are attached to the four corners of the movable plate. Since the sensing axis is perpendicular to the substrate, the spring can be made very compliant by using another thin polysilicon layer for suspension, if sensitivity is the primary consideration. This design also provides low-cross axis sensitivity. The lateral movement of the plate in x-direction is constrained by the top and bottom springs, and the movement in y-direction is constrained by the left and right springs. Analysis shows the cross axis sensitivity is less than 0.1% [16]. The folded design relieves the residual stress in the suspension.

The feedback capacitors are driven by a high frequency, density-modulated pulses train. The pulse density is controlled by the feedback loop and is directly proportional to the acceleration. When there is no acceleration, the pulse train will have 50 percent duty cycle. Since these feedback capacitors generate a pull-down force and there is no feedback capacitors above the proof mass to have pull-up force, interdigitated fingers at the perimeter of the plate are used to generate a constant levitation force [17]. Because of the

presence of the ground plane underneath the fingers, there are more electrical field terminated on the top side of the fingers on the proof mass than that at the bottom side, which results a electrostatic levitation force on the proof mass fingers [17]. This levitation force is balanced by the pull down force from gravity, the sensing electrode, and the feedback electrodes so that the plate is at its neutral position. When acceleration causes the proof mass to move downward, the pulse density on the feedback capacitors reduces so that the pull-down force decreases and the plate moves back up to it is neutral position. Similarly, the pulse density increases when the plate moves upwards due to acceleration. Therefore, the movable plate can be maintained at a stable position.

Since the suspension is very compliant, fuses are used to hold the plate and the suspension in place during the process and are severed by sending a current through after the completion of all processing steps [19]. Figure 3 shows the SEM picture of a corner of the structure: the fuse-held folded suspension, the interdigitated levitation fingers, and the sensing plate with etching holes.

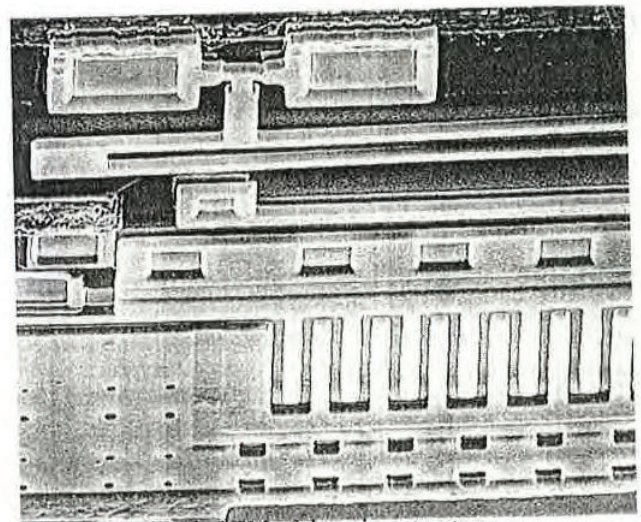


Fig. 3. SEM of the microstructure

SYSTEM ARCHITECTURE

Electrostatic force between the two electrodes of a capacitor is proportional to the square of the applied voltage and is always attractive regardless of the polarity of the voltage. In this design, the feedback capacitors are located underneath the movable plate. If there are feedback electrodes located on both sides of the proof mass symmetrically, the squared term between the feedback voltage and force can be canceled by applying a differential feedback signal with a DC bias to the two electrodes [18]. Since the feedback electrodes are all located underneath the plate, the feedback cannot be linearize by using this technique. If the feedback control is realized in the analog domain, a circuit that has a square-root function has to be designed in the feedback path in order to have a linear system. It is not trivial to design a circuit with that transfer function over a large dynamic range, which can degrade the linearity and the dynamic range of the accelerometer.

$\Sigma\Delta$ modulation technique is a popular technique to design analog-to-digital converters. By sampling the input signal at frequencies much higher than the Nyquist frequency, the modulator suppresses the quantization noise in the baseband where the input signal is, and pushes the noise to high frequency. Even though a 1-

bit quantizer is used in the modulator, high resolution can be achieved by filtering out the high frequency noise with a decimation filter [11, 12].

There are several advantages to use Σ - Δ modulator for feedback control in accelerometers. First, the feedback system has intrinsic linearity because the feedback is a 1-bit signal and the magnitude of the feedback force is proportional to the number of pulses in a given period. This property ensures the accelerometer to have large dynamic range. Second, the requirements for the readout circuits are less demanding because the 1-bit quantizer only concerns the polarity of the signal. This is especially important for integrated accelerometers where the transistor characteristics may be changed due to the sensor process. At last, its digital output makes it easy to interface with rest of the system.

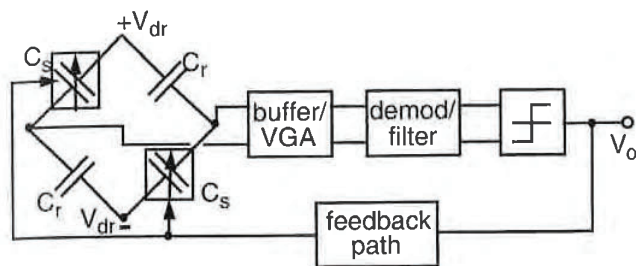


Fig. 4. System block diagram.

The system block diagram of this accelerometer is shown in Fig. 4, where two identical sensing capacitors (C_s 's) and two reference capacitors (C_r 's) of the same value forms the capacitive bridge. Note that two suspended plates are needed to have a full bridge, since the sense capacitor is one-sided. The bridge is driven by a 1MHz square wave. The output of the bridge is linearly proportional to the displacement of the proof mass. Since the displacement is the double integration of the input acceleration, the two integrators in a conventional second-order Σ - Δ modulator can be realized with the mechanical proof mass in this electro-mechanical system. The displacement signal is then quantized into a 1-bit signal. A compensation network is placed in the feedback path to insure loop stability and proper noise shaping. It consists of direct feedback and also feedback passing through a differentiator. The feedback signal through the differentiator is equivalently fed back as "electrical damping" on the plate. The output from the feedback path is a 2-bit signal because of the differentiation. The twelve feedback electrodes, shown in Fig. 2, are split into three sets symmetrically. The 2-bit feedback signal determines the number of feedback capacitor sets to be connected to the DC feedback voltage. The force-balanced accelerometer is therefore a second-order electro-mechanical Σ - Δ modulator.

Since the sensing capacitance is small, the output voltage of the capacitive bridge can be overwhelmed by the parasitic capacitance and the input capacitance to electronics. A unity gain buffer with low input capacitance is designed to bootstrap the ground plane, shown in Fig. 5. The input capacitance to the buffer consists of gate-to-source capacitance C_{gs} , gate-to-body capacitance C_{gb} , and gate-to-drain capacitance C_{gd} of the input transistor M_1 , where C_{gs} is the main contributor. The C_{gs} and C_{gb} can be minimized by using a source follower at the input. M_2 is a level shifter that insures the out-

put has the same DC level as the input. The C_{ds} capacitance is also bootstrapped using positive feedback. Transistors M_3 and M_4 take the output voltage and drive the drain the input transistor M_1 . The effective input capacitance is 16fF from simulation -- about 5% of the C_{gs} of the input transistor. Since the output of the buffer has the same DC level as the input, we can use it to bootstrap the ground plane without concern about introducing a DC electrostatic force on the plate. The buffer is simulated to have a bandwidth of 50MHz with 1pF capacitive load. A variable gain amplifier (VGA) follows the buffer to amplify the signal with its gain controlled externally.

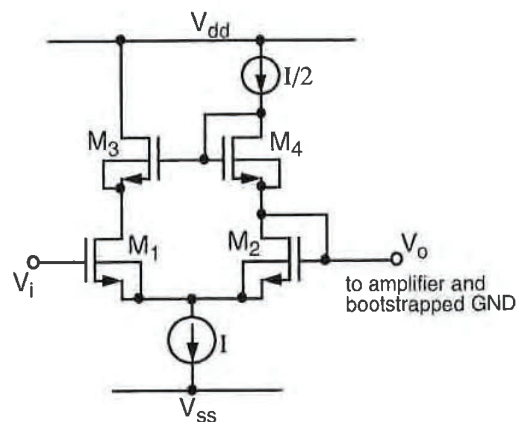


Fig. 5. Unity gain buffer with low input capacitance

Performance of accelerometers can be degraded due to signal feedthrough. In this design, sensing and feedback are operated in two non-overlapping clock phases. During the sensing clock phase, the feedback electrodes are grounded and detection circuits are enabled. During feedback clock phase, the readout circuit is disabled and the feedback electrodes are turned on. Thus, the signal feedthrough is minimized to great extent. Offset voltages from the buffer and the amplifier can also limit the resolution of the accelerometer. Since the offset voltages are at DC and the input signal is modulated to the carrier frequency, chopper stabilization technique can remove the offset with a cosine filter. To minimize the tempera-

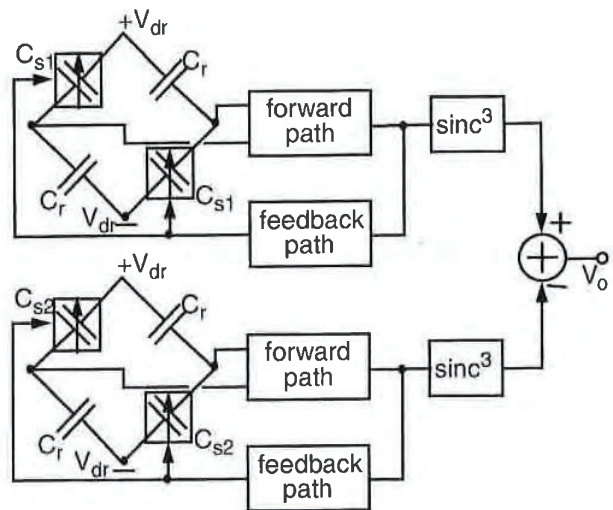


Fig. 6. System Block Diagram of Dual Loop Design

ture sensitivity, potential drift due to the mechanical suspension, and package induced stress effect, two accelerometers are built, where they are exactly identical (Fig. 6) except that the proof mass in one accelerometer (C_{s2} 's) is half of the other one (C_{s1} 's). Since they are built side by side, the temperature, the drift, and package induced stress are common mode signals for the two channels and can be eliminated by subtracting the output in digital domain.

EXPERIMENTAL RESULTS

The accelerometer was fabricated using MICS process in the Berkeley microfabrication facility. The die photo of the device is shown in Fig. 6. The two square structures on the right are the movable plates forming the sensing capacitor C_{s1} 's, and the two on the left are the sensing capacitors with half of the mass (C_{s2} 's). The two structures in the center form the four reference capacitor C_r 's, where two reference capacitors are shared with one top plate. The detection circuit is located in the middle, which includes the low input capacitance buffer, a variable gain amplifier, a demodulator and a cosine filter, a 1-bit quantizer, as well as the bias circuits. The layout is fully symmetrical to ensure that the environmental changes will affect the two accelerometers in the same fashion and are canceled at the output. The feedback control circuit is operated in digital domain and will be constructed off-chip potentially. The chip contains about 500 transistors and the die area is $2.5 \times 5 \text{ mm}^2$.

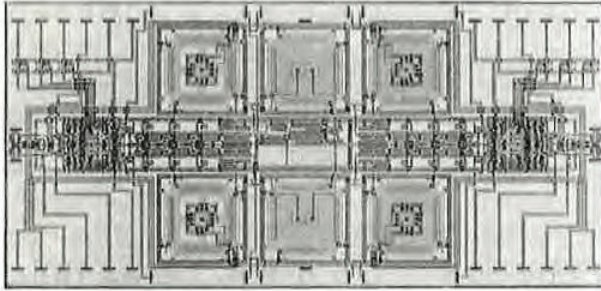


Fig. 7. Die photo of the accelerometer

The circuit performance was evaluated by wafer-level probing. The input buffer is measured to have a gain of 0.9 and a bandwidth around 500kHz which is limited by the parasitic capacitance from the probes and coaxial cables to the measurement instrument. The voltage gain of the variable gain amplifier is controlled externally and can be varied from unity to 40. Its differential output swing is 1.5V.

The accelerometer has been characterized in the open-loop self-testing mode by applying a voltage to the feedback electrodes and the levitation fingers. The suspended plate is moved and we measure the output. Figure 8 show the output voltage versus the levitation voltage at two different driving voltage V_{dr} , 250mV and 150mV. With feedback electrodes grounded, gravity and electrostatic force from sensing electrode pull the proof mass below the equilibrium position giving an offset voltage at the output. As the levitation voltage increases, the levitation force pulls the plate

towards to the equilibrium position and the output voltage decreases.

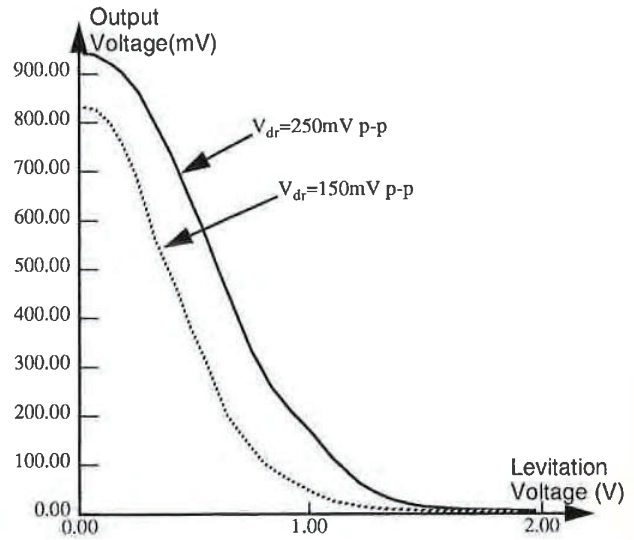


Fig. 8 Output voltage vs. levitation voltage

The open-loop dynamic behavior is also measured by applying a pulse to the feedback or levitation fingers and measuring the transient response at the output. Figure 9 shows a digital scope photo of the envelop of the high frequency sinusoidal output when a 500ms, 0.5V pulse is applied to the levitation fingers. The magnitude of the envelop is the output voltage. By measuring the settling time, the damping factor is derived to be around $1.2 \times 10^{-3} \text{ Nms}^{-1}$, while theoretical calculation predicts the damping factor to be 1 Nms^{-1} [20]. Similar output with opposite sign is also observed when the pulse is applied to the feedback electrodes.

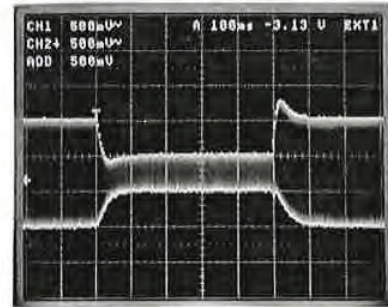


Fig. 9. Transient response of the accelerometer

CONCLUSION

In this paper, we have described a surface micromachined accelerometer with digital electrostatic feedback using Σ - Δ modulation technique, fabricated in a modular CMOS/microstructure process. Detection circuits have been tested successfully. The unity gain buffer with low input capacitance can also be used for other capacitive detection applications. The accelerometer has been characterized in the self-testing mode and has demonstrated the functionality of the device. The accelerometer will be packaged for 2g turnover test.

ACKNOWLEDGMENT

The authors would like to thank Clark Nguyen for his assistance in the fabrication and the staff at Berkeley Microfabrication laboratory. The Berkeley Sensor & Actuator Center (BSAC) is funded by National Science Foundation and a consortium of corporations and government agencies.

This work is performed as part of the PATH Program of the University of California, in cooperation with the State of California, Business and Transportation Agency, Department of Transportation, and the United States Department of Transportation, Federal Highway Administration.

The contents of this report reflect the views of the authors who are responsible for the facts and the accuracy of the data presented herein. The contents do not necessarily reflect the official views or policies of the State of California or the Federal Highway Administration. This report does not constitute a standard, specification, or regulation.

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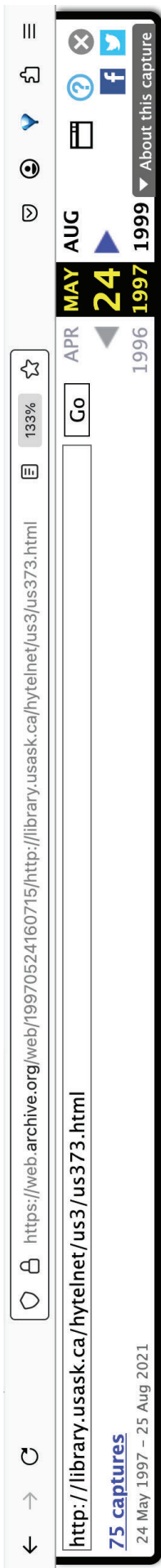
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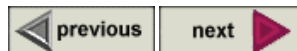
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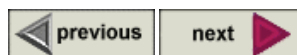
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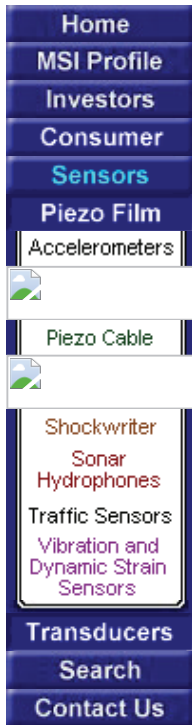
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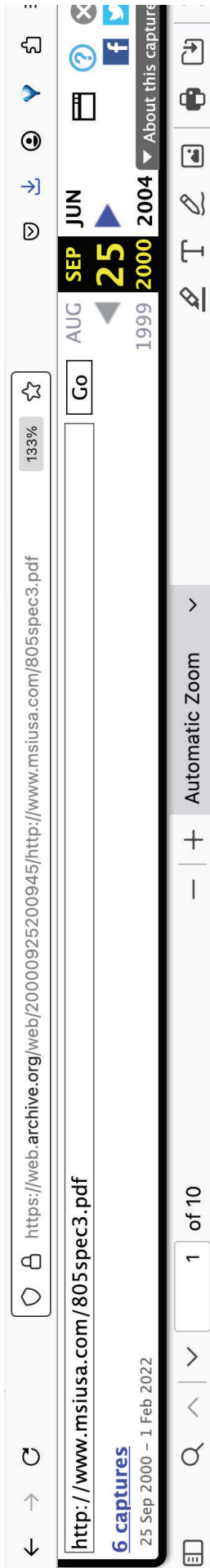
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S P E C I A L T I E S



Accelerometer ACH-04-08-05

Application Specification
114-27002
27 AUG 98 Rev A

1.0 INTRODUCTION

This specification covers the application requirements of Measurement Specialties Accelerometer ACH-04-08-05. The accelerometer contains three piezoelectric sensing elements oriented to simultaneously measure acceleration in three orthogonal, linear axes. Internal JFET's provide for a low impedance, buffered output with a wide variety of possible, application specific, signal conditioning circuitry. The sensor responds over a broad frequency range, from below 0.5Hz to above 5kHz, as a result of the integrated electronics and the damped sensing elements.

The ACH-04-08-05 accelerometer can be used in a wide range of applications including motion measurement and control systems, speaker feedback systems, vibration switches, shipment monitor and material handling systems, security systems, computer input/output devices, and data loggers. The device is not IR reflow solderable.

Features of the accelerometer include:

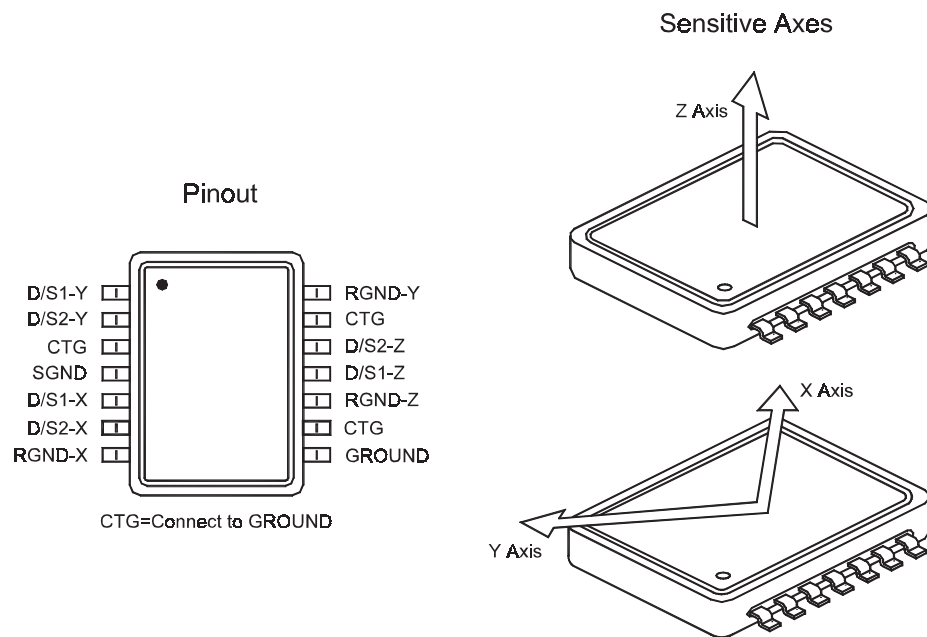
1.0 INTRODUCTION

This specification covers the application requirements of Measurement Specialties Accelerometer ACH-04-08-05. The accelerometer contains three piezoelectric sensing elements oriented to simultaneously measure acceleration in three orthogonal, linear axes. Internal JFET's provide for a low impedance, buffered output with a wide variety of possible, application specific, signal conditioning circuitry. The sensor responds over a broad frequency range, from below 0.5Hz to above 5kHz, as a result of the integrated electronics and the damped sensing elements.

The ACH-04-08-05 accelerometer can be used in a wide range of applications including motional measurement and control systems, speaker feedback systems, vibration switches, shipment monitor and material handling systems, security systems, computer input/output devices, and data loggers. The device is not IR reflow solderable.

Features of the accelerometer include:

- Fully Contained 3-Axis Accelerometer
- Low Frequency Response
- High Resonant Frequency & Low Q
- Excellent Linearity
- Three Simultaneous Analog Outputs
- Low-Impedance Output
- Ultra-Low Power
- Low Cost



2.0 REFERENCE MATERIALS

2.1 Revision Summary

This paragraph is reserved for a revision summary of changes and additions made to this specification. Revision A changes include: document format, improved transverse sensitivity, interface circuit corrections, Analog Test PCB Kit description, and additional performance curves.

2.2 Customer Assistance

Reference MSI Part Number 1003800-5 and call (610) 650-1500 in the USA or +49 6074 862 822 in Europe to obtain additional product information.

2.3 Drawings

MSI Customer Drawings are available for specific products. The information contained in Customer Drawings takes priority if there is a conflict with this specification or with any other technical document supplied by MSI.

2.4 Absolute Maximum Ratings

CHARACTERISTICS	VALUE	UNITS
Applied Voltage (V_{DD} with respect to GND)	+28	V
Package Power Dissipation	100	mW
ESD (Mil-Std-883, Method 3015.6)	2000	V

2.5 Pin Descriptions

PIN NUMBER	NAME	DESCRIPTION
1	D/S1-Y	Y-Axis JFET Drain or Source
2	D/S2-Y	Y-Axis JFET Source or Drain
3	CTG	Connect to GND
4	SGND	Sensor Common
5	D/S1-X	X-Axis JFET Drain or Source
6	D/S2-X	X-Axis JFET Source or Drain
7	RGND-X	Resistor GND for the X-Axis
8	GND	Device power - negative voltage with respect to V_{DD}
9	CTG	Connect to GND
10	RGND-Z	Resistor GND for the Z-Axis
11	D/S1-Z	Z-Axis JFET Drain or Source
12	D/S2-Z	Z-Axis JFET Source or Drain
13	CTG	Connect to GND
14	RGND-Y	Resistor GND for the Y-Axis

Note: JFET is symmetrical. The DRAIN and SOURCE are interchangeable.

2.6 Environmental Characteristics

CHARACTERISTICS ($T=25^{\circ}\text{C}$)	SYMBOL	MIN	TYP	MAX	UNITS
Operating Temperature †	T_{OP}	-40	-	+85	$^{\circ}\text{C}$
Storage Temperature	T_S	-40	-	+105	$^{\circ}\text{C}$
Relative Humidity ‡	RH	0	-	95	%R.H.
Maximum Shock (Any Axis)	S_{MAX}	1,000	-	-	g

† Nominal sensitivity will typically change less than $\pm 15\%$ over temperature range.

‡ Humidity must be non-condensing.

2.7 Specification

CHARACTERISTIC (T=25°C) ●	SYMBOL	MIN	TYP	MAX	UNITS
X Axis Sensitivity ■ ▲	M_x	1.07	1.80	-	mV/g
Y Axis Sensitivity ■ ▲	M_y	1.07	1.80	-	mV/g
Z Axis Sensitivity ■ ▲	M_z	1.07	1.35	-	mV/g
Lower Frequency Limit (3dB Point)	f_{L3dB}	-	0.35	0.7	Hz
Upper Frequency Limit (3dB Point)	f_{U3dB}	4,000	5,000	-	Hz
Resonant Frequency	f_o	-	9,200	-	Hz
Resonant Q	Q_R	-	10	-	Hz/Hz
Transverse Sensitivity	M_T	-	10	-	%
Base Strain Sensitivity	-	-	0.05	-	$g/\mu\epsilon$
Temperature Transient Sensitivity	-	-	0.2	-	$g/^\circ\text{C}/\text{min}$
Dynamic Range	-	-	± 250	-	$g's$
Linearity	-	-	0.1	1.0	%
Equivalent Noise (100Hz)	e_{100}	-	0.2	-	$mg/\sqrt{\text{Hz}}$
Weight	-	-	0.35	-	grams

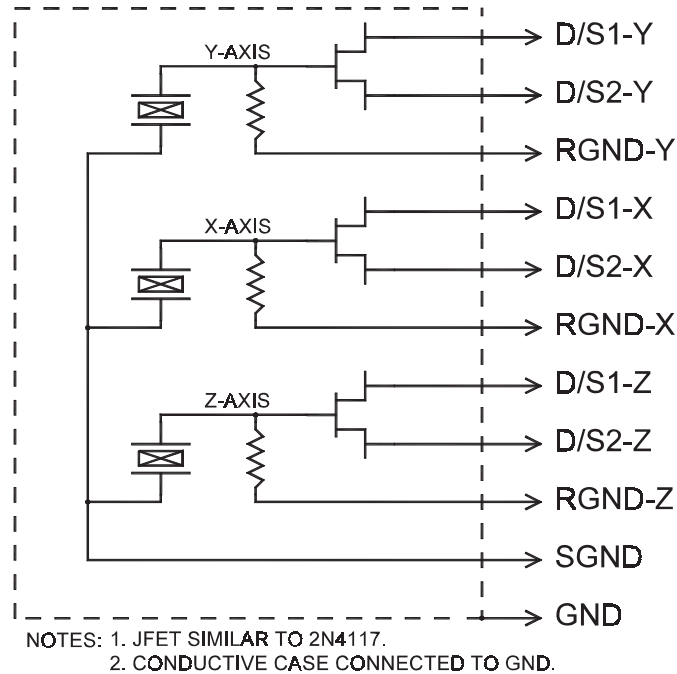
- Reference Frequency is 10 Hz.
- Assumes constant current bias, guarded Drain & Source
- ▲ Tighter sensitivity tolerances available. Consult factory for options.

2.8 Electrical Specifications

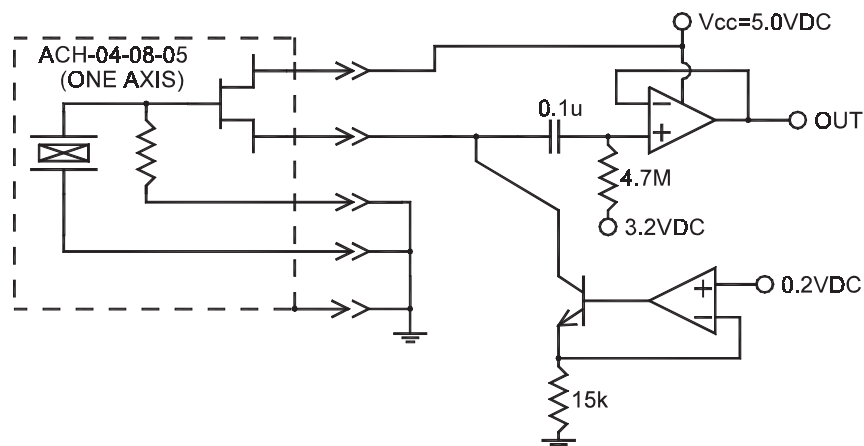
CHARACTERISTICS (T=25°C) ◆	SYMBOL	MIN	TYP	MAX	UNITS
Gate-Drain Voltage	-	-30	-	-	V
Gate-Source Voltage	-	-30	-	-	V
Gate Current	-	-	-	10	mA
Power Dissipation	-	-	-	360	mW
Power Derating	-	-	3.27	-	$\text{mW}/^\circ\text{C}$
Gate-Source Cut-Off Voltage	$V_{GS(OFF)}$	-0.5	-	-1.8	V
Saturation Drain Current ◆	I_{DSS}	10	-	150	μA
CS Forward Transconductance ◆	g_{FS}	50	-	210	μS

- ◆ JFET's similar to 2N4117. Three per unit.
- ◆ Specifications modified from standard 2N4117 JFET's.

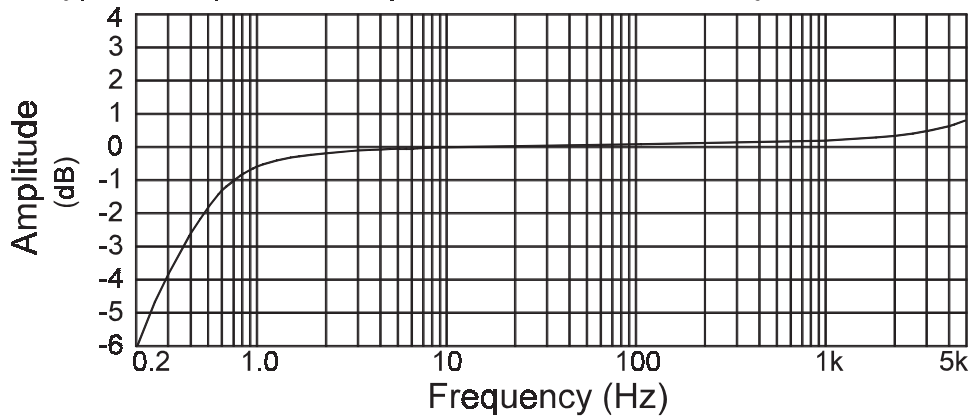
2.9.1 Equivalent Electrical Schematic



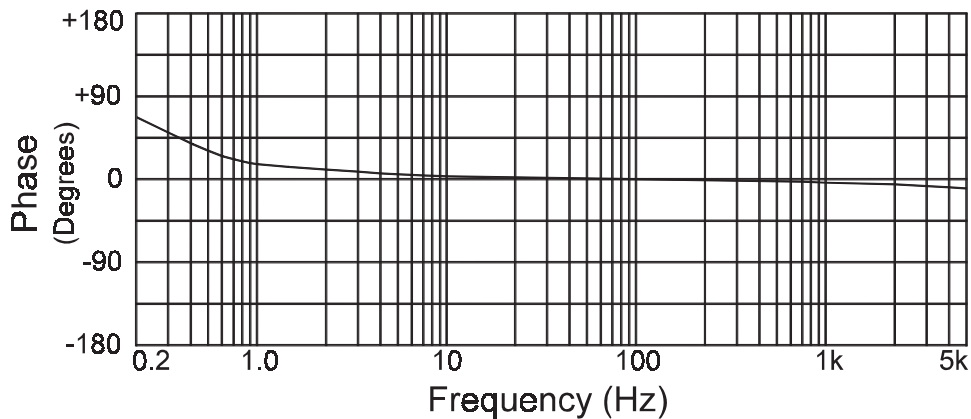
2.9.2 Interface Test Circuit (One Channel Shown)



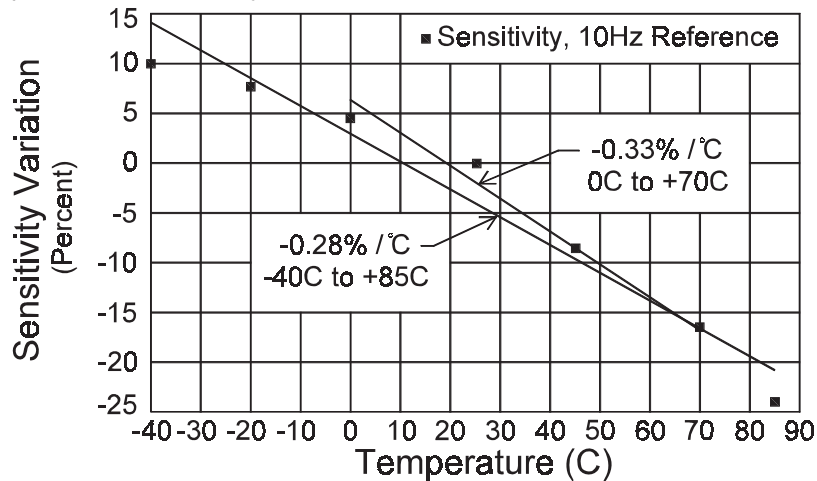
2.9.3 Typical Amplitude Response versus Frequency



2.9.4 Typical Phase Response versus Frequency



2.9.5 Typical Sensitivity Variation versus Temperature

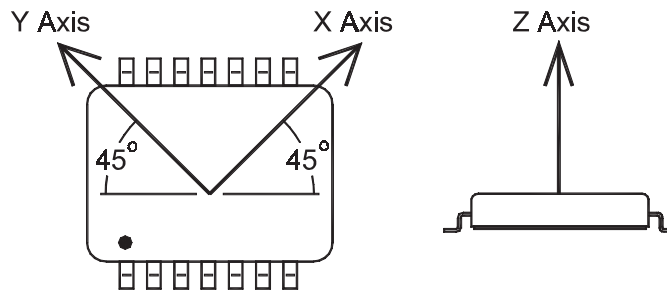


3.0 REQUIREMENTS

3.1 Detailed Description

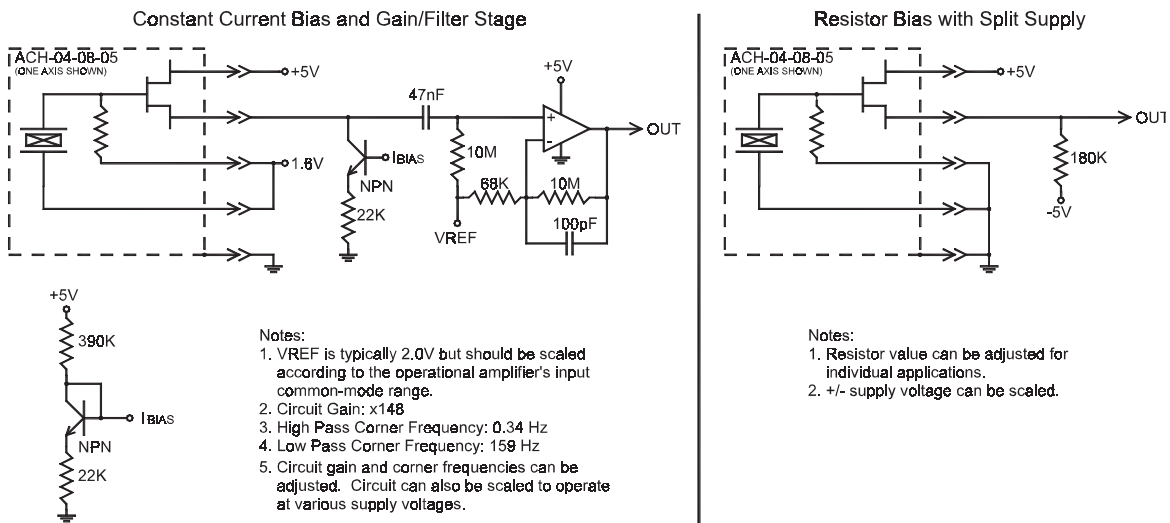
Each of the three sensing elements is a small cantilever beam consisting of a metal substrate with a piezoelectric polymer element affixed to one side. Each beam is supported at one end while the opposite end is allowed to flex in response to acceleration. The beam flex strains the piezoelectric material which in turn generates a charge proportional to the applied acceleration.

The beams are oriented to sense acceleration in the defined X, Y, and Z axes. Note that the X and Y axes are at a 45° angle relative to the accelerometer's package.



3.2 Electrical Interface

The ACH-04-08-05 contains three JFET's, one for each channel, which function as impedance converters easing electrical interface requirements. The JFET's are similar to the industry standard part number 2N4117. The JFET's **MUST** be biased for the accelerometer to function properly. There are many possible interface circuits. The JFET biasing circuit, used in MSI's final production test, is shown on page 4. Two other example circuits are shown below.

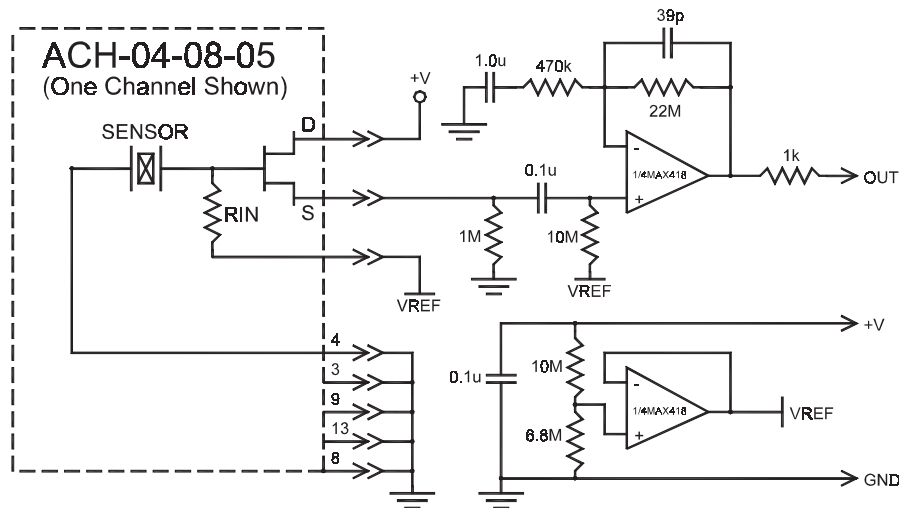


The constant-current bias circuit minimizes device-to-device and temperature related JFET transconductance variations keeping the JFET gain near one (1). With the additional operational amplifier, the circuit has a gain of x148, a one-pole low-pass filter at 159Hz, and a one-pole high-pass filter at 0.34Hz. The circuit's component values and DC reference/supply voltages can be scaled to fit many applications.

The resistor bias circuit positions the operating point in the middle of the transconductance curve again minimizing transconductance variations. This circuit can also be powered from a single-supply voltage by creating an artificial ground, VREF.

The following figure shows the circuit used in MSI's Analog Test PCB (only one channel shown) which is available to quickly evaluate the basic characteristics of the ACH-04-08-05 accelerometer. Contact the factory to receive more information about the Analog Test PCB. Here, a resistor biasing scheme is used. The circuit runs from a single supply voltage by generating a reference voltage, VREF. The circuit has a gain of x47.8, a high-pass corner frequency of 0.34Hz, and a low-pass corner frequency of 185Hz. These values, as well as the operational amplifier, are easily changed to suit specific applications. As shown, the circuit typically draws only 13 μ A for all three axes.

Analog Test PCB: Low Power, Single Supply Circuit



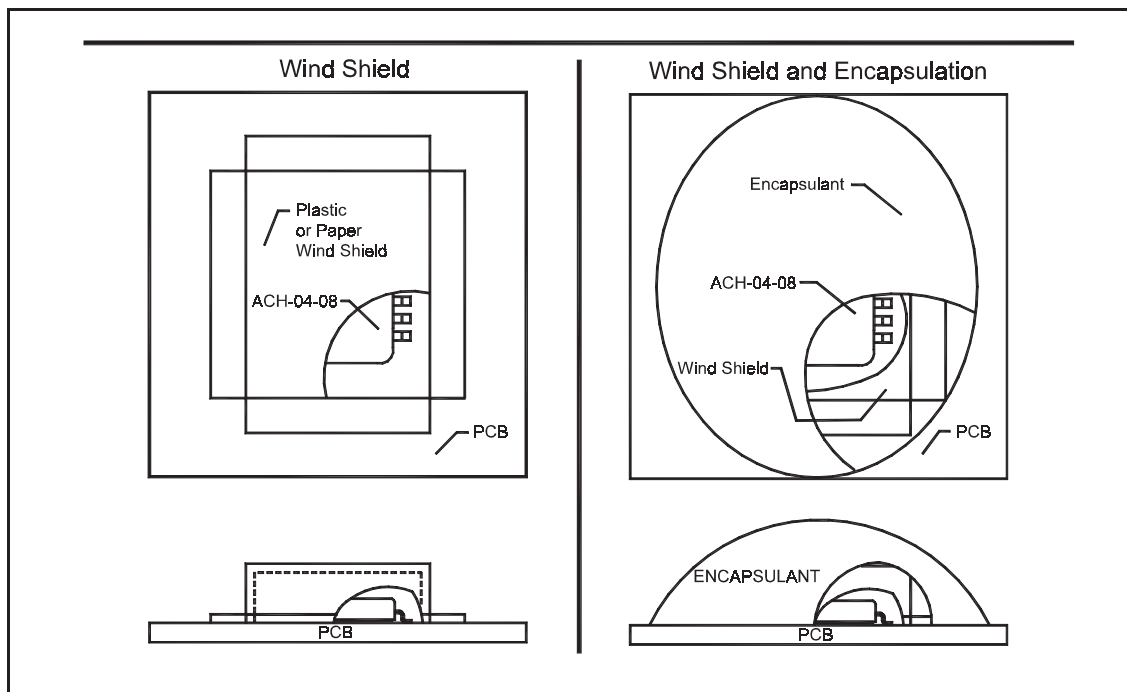
3.3 Sensitivity Variation versus Temperature

The ACH-04-08-05's typical sensitivity variation versus temperature is shown on Page 5. As shown, the sensor output varies +10% / -24% from -40°C to +85°C, respectively, relative to 25°C. From 0°C to +70°C, the sensitivity variation is typically +5% / -16%, respectively. For many applications, this variation is acceptable (less than ± 2 dB from -40°C to +85°C). If a tighter tolerance is required, the ACH-04-08-05 is easily temperature compensated. A

simple method is to use linear temperature coefficient resistors in the feedback network of the operational amplifier used to condition the accelerometer's signal. KOA Speer Electronics, Inc.® (Bradford, PA) manufactures a variety of these resistors (LT73 and MLT product family). Using a -2800ppm/°C temperature compensating resistor yields a sensitivity variation of $\pm 4\%$ from -40°C to +85°C. From 0°C to +70°C, a variation of only $\pm 2\%$ is obtained using a -3300ppm/°C resistor. Note that since the resistors have negative temperature coefficients, they need to be placed in the circuit such that the gain increases with increasing temperature.

3.4 Temperature Transient Sensitivity

Because of its very low operating frequency, the ACH-04-08-05 exhibits high Temperature Transient Sensitivity (TTS). TTS is defined as the accelerometer's output response to RAPID temperature changes and is expressed in equivalent g's per degree C. This effect should not be confused with the ACH-04-08-05's sensitivity shift versus



temperature which is defined at static (unchanging) temperatures and is unrelated to TTS (see above). The ACH-04-08-05's TTS will manifest itself as a low frequency ("DC") drift when the device is subjected to temperature gradients. Even small temperature changes can produce significant DC drift.

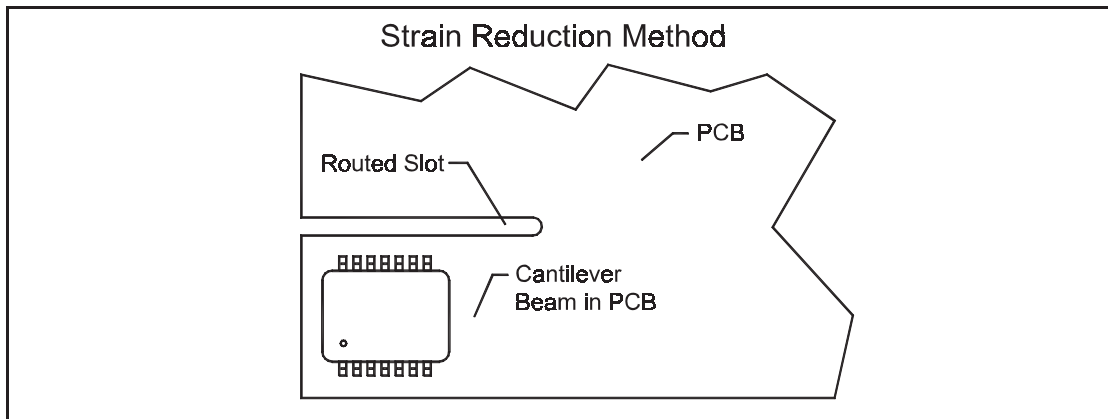
TTS results primarily from differences in the thermal coefficients of expansion (TCE) of the various materials in the accelerometer. During product development, every effort is made to match TCE's, however, many mismatches are unavoidable. Even small TCE mismatches can lead to high TTS when combined with low frequency measurement capability since TTS is primarily a low frequency phenomena.

Fortunately, it is easy to minimize TTS effects. The key is to prevent air currents, which are at various temperatures, from circulating around the ACH-04-08-05. These "wind currents" can be minimized by encasing the accelerometer in a shell. The shell can be made of almost anything. Successful TTS minimization shells have been made from

plastic, paper, metal, etc. In fact, wind shields have been successfully integrated into the housing of the final circuit board assembly negating the need for additional parts. While the shell does not have to be air-tight, tighter seals provide better performance. Encapsulating the ACH-04-08-05 can also provide an effective wind shield. To ensure that the ACH-04-08-05's housing is not compressed under the weight of the encapsulant, MSI recommends that a spacer be placed above the unit. Note that the best performance is obtained by combining a wind shell with encapsulation.

3.5 Base Strain Sensitivity

The ACH-04-08-05 displays Base Strain Sensitivity (BSS) for many of the same reasons that it displays TTS; namely, low frequency operation and low modulus/high TCE materials. BSS is defined as the amount of output signal generated as a result of bending in the accelerometer's mounting surface and is expressed in terms of equivalent g's per input micro-strain, g/ $\mu\epsilon$.



The best way to minimize BSS effects is to stiffen the mounting structure. Stiffer mounting structures are typically less susceptible to bending. However, since the ACH-04-08-05 is designed to be attached to a circuit board, it may be impractical to stiffen the mounting surface. In this situation, the accelerometer can be mounted at the end of a "cantilever beam" routed into the circuit board, where minimal strain is coupled into the accelerometer. Mounting arrangements such as this are very effective for low frequency applications. However, they are unacceptable for high frequency measurements where the signal to be measured is near or above the resonant frequency of the circuit board, cantilever beam. Implementation of a cantilever beam, strain minimization design requires careful analysis to accurately position the cantilever resonance well above the frequency range of interest.

3.6 IC Package

To insure proper protection against RF and EM interference, the outer surface of the package is conductive and electrically connected to the GND pin. Exercise caution when laying out PC boards to ensure that no uninsulated traces are located under the package. They could contact the package surface and be shorted to GND. It is recommended that all 14 pins of the package be soldered to the PC board to insure good mechanical coupling of acceleration, motion, and shock. Complete soldering will also eliminate undesirable mechanical resonances of the IC package.

The package is sealed to eliminate the effects of normal exposure to the environment and typical cleaning processes found in an electronic manufacturing facility. However, the package seal is not hermetic and exposure to acidic, basic, corrosive, or aqueous environments is not recommended. Because of process variations, the ACH-04-08-05 should be qualified through the intended production assembly processes (soldering, wash, etc.) before releasing to production. **The device is NOT IR reflow solderable.** Recommended alternative soldering processes include laser, hand and hot bar soldering. The ACH-04-08-05 can be manufactured to suit wave (not IR) soldering processes. Contact MSI for more detailed information.

3.7 ACH-04-08-05 Mechanical Dimensions

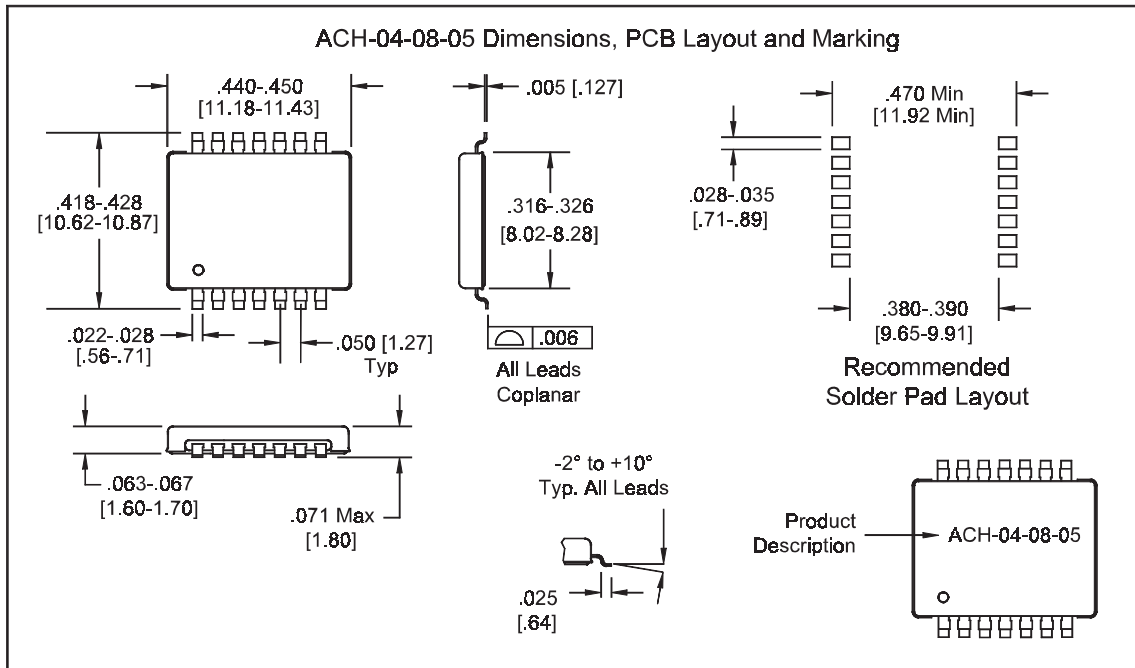


EXHIBIT 1063B

3-V Accelerometer Featuring TLV2772

Application Report

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Application Report

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Abstract

This paper describes a complete solution for digital measurement of acceleration. An AMP accelerometer sensor is used for the conversion between mechanical acceleration and electrical analog. This electrical signal is then conditioned using Texas Instruments' TLV2772 op amp (on the Universal Op Amp EVM), digitized using the TLV1544 ADC EVM, and processed with the TMS320C5X EVM. This provides the user with a quick and easy way to evaluate a complete 3-axis accelerometer solution.

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3-V Accelerometer Featuring TLV2772

1 Introduction

Accelerometers are used in aerospace, defense, automotive, household appliances, instrumentation, audio, transport, material handling, etc. This application report develops a data collection system for acceleration in 3 axis.

Figure 1 shows a block diagram of a typical analog data collection system. This application presents information about the sensor, signal conditioning, ADC, processor, display, and memory.

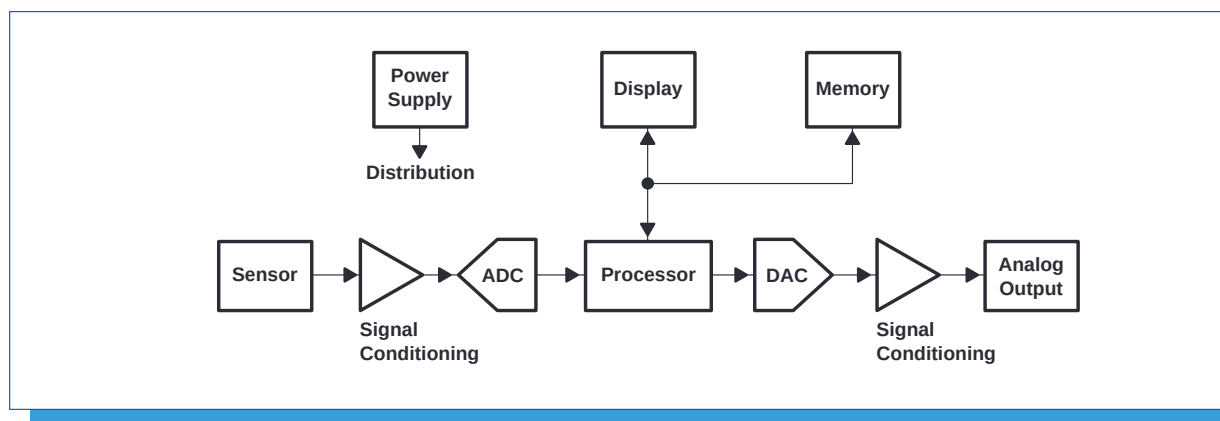


Figure 1–1. Typical Analog Data Collection System

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2 System Description

Figure 2.1 shows the accelerometer system diagram

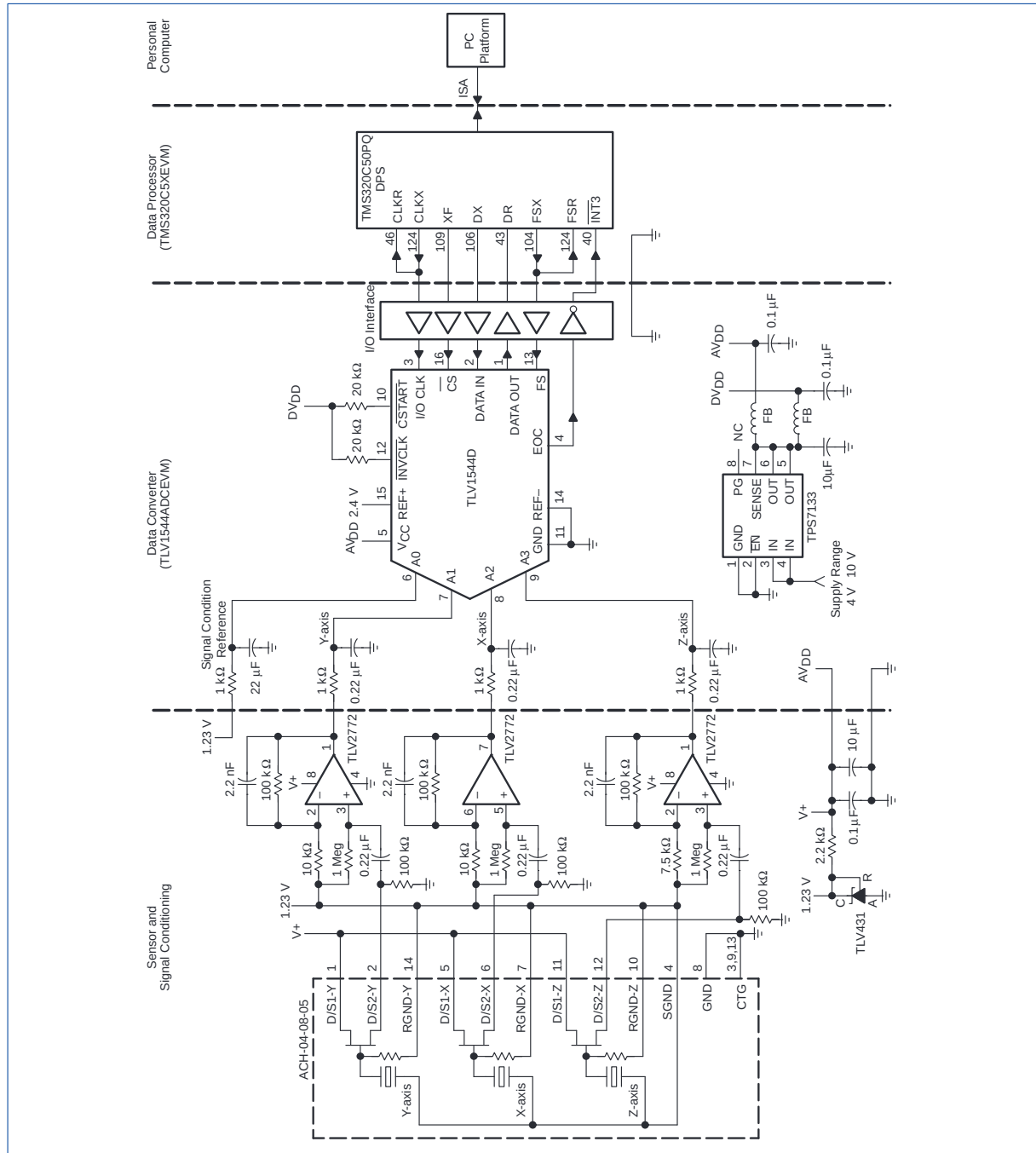


Figure 2-1. Accelerometer System Diagram

3-V Accelerometer Featuring TLV2772

2.1 Sensor

An AMP ACH04-08-05 shock sensor converts mechanical acceleration into electrical signals. The shock sensor contains three piezoelectric sensing elements oriented to simultaneously measure acceleration in three orthogonal linear axes. The sensor responds from 0.5 Hz to above 5 kHz. An internal JFET buffers the output. Typical output voltage for x and y axis is 1.80 mV/g. Typical output voltage for the z axis is 1.35 mV/g. Refer to AMP's web site at <http://www.amp.com/sensors> for in-depth information about this sensor, piezo materials in general, and other related products.

2.2 Signal Conditioning

Circuitry using the Texas Instruments TLV2772 operational amplifier provides amplification and frequency shaping of the shock sensor output. Due to its high slew rate and bandwidth, rail-to-rail output swing, high input impedance, high output drive and excellent dc precision the TLV2772 is ideal for this application. The device provides 10.5 V/ μ s slew rate and 5.1 MHz gain bandwidth product while consuming only 1 mA of supply current per amplifier. The rail-to-rail output swing and high output drive make this device ideal for driving the analog input to the TLV1544 analog-to-digital converter. The amplifier typically has 360 μ V input offset voltage, 17 nV/ $\sqrt{\text{Hz}}$ input noise voltage, and 2 pA input bias current. Refer to Texas Instruments' web site at <http://www.ti.com> and search on TLV2772 to download a TLV2772 data sheet, literature #SLOS209.

The Universal Operational Amplifier EVM is used to construct the ACH04-08-05 shock sensor and TLV2772 operational amplifier circuitry. The Universal Operational Amplifier EVM facilitates construction of surface mount operational amplifier circuits for engineering evaluation. Refer to Texas Instruments' web site at <http://www.ti.com> to download the Universal Operational Amplifier EVM User's Manual, literature #SLVU006.

2.3 ADC

The TLV1544 EVM provides analog-to-digital conversion. The TLV1544 is a low-voltage (2.7 V to 5.5 V dc single supply), 10-bit analog-to-digital converter (ADC) with serial control, 4 analog inputs, conversion time = 10 μ s, and programmable 1 μ A power down mode. Refer to Texas Instruments' web site at <http://www.ti.com> to download a TLV1544 data sheet, literature #SLAS139B, TLV1544 EVM User's Manual, literature #SLAU014, and related information.

2.4 Processor, Memory, and Display

The TMS320C5x EVM controls and collects data samples from the TLV1544. Refer to Texas Instruments' web site at <http://www.ti.com> to download the TMS320C5x EVM Technical Reference, literature #SPRU087, and related information.

The TMS320C5x EVM installs into a PC platform. The PC provides programming and control of the TMS320C5x EVM, and provides resources for file storage or other processing of the collected data.

3 System Specification Requirements

The following system specification requirements were derived to guide the design.

3.1 G-Force Measurement Requirements

Range: $\pm 50g$
Noise: 0.05g pk-pk (equivalent input noise measured in g)
Resolution: 0.1g
Frequency Response: 1 Hz to 500 Hz (min 3 dB BW)

3.2 Power Requirements

Input Voltage: 3 V $\pm 10\%$, Noise = 30 mV pk-pk (20 MHz BW)
Input Current: 5 mA (max) (power requirements for sensor and signal conditioning circuitry)

3-V Accelerometer Featuring TLV2772

4 Sensor and Signal Conditioning Design

Circuit design is a 3-step process:

1. Hand analysis
2. SPICE simulation
3. Circuit breadboard and lab testing

The signal from the sensor must be amplified and frequency shaped to provide a signal that the ADC can properly convert into a digital number.

The schematic in Figure 4–1 shows the topology used in this application for 1 axis of the sensor and signal conditioning circuit.

Input power is 3 V and ground. The TLV431 precision voltage regulator, when configured as shown, produces a nominal 1.23-V reference voltage. This voltage provides the signal reference for the signal conditioning circuitry and the bias voltage for the internal JFETs in the shock sensor.

The transfer function of the signal conditioning circuit is derived by several means, the easiest of which may be by using super position. Perform a dc analysis, perform an ac analysis, and superimpose the results.

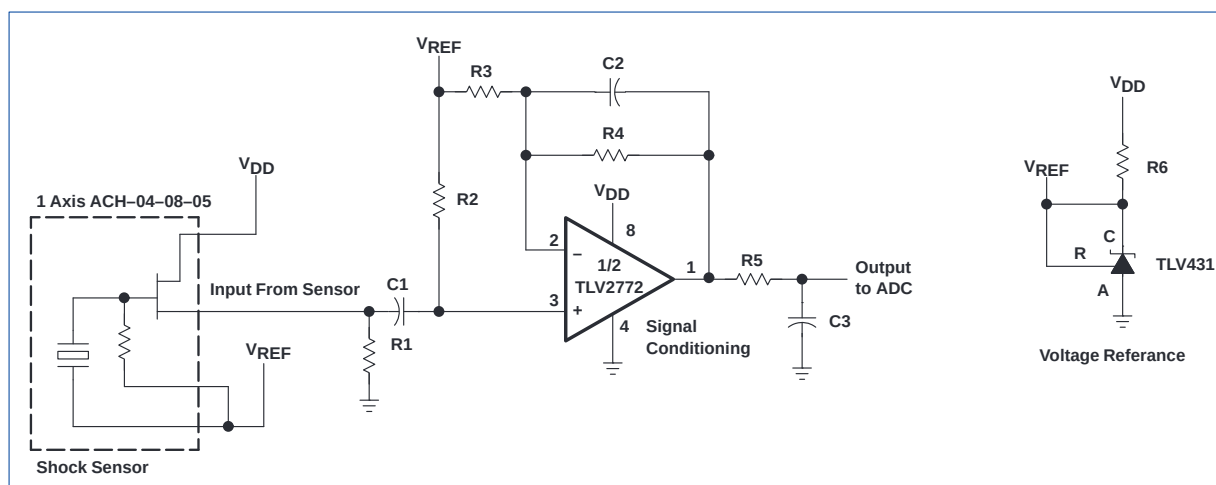


Figure 4–1. 1-Axis Accelerometer Sensor and Signal Conditioning Circuit

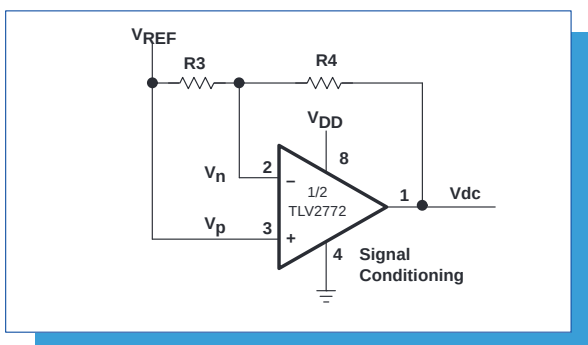


Figure 4–2. DC Circuit Model

4.1 Hand Analysis

In hand analysis, simplifying assumptions make solutions easier to derive. If the circuit does not function as anticipated, these assumption must be reevaluated.

4.1.1 DC Analysis

To perform a dc analysis, assume all inductors are short circuits and all capacitors are open circuits.

Assume that the resistance of R2 is insignificant compared to the input impedance of the op amp. Therefore, Vref will appear at the positive input to the op amp.

Assuming the ADC input does not impose a significant load on the circuit, the voltage divider formed between the ADC input and R5 can be disregarded.

3-V Accelerometer Featuring TLV2772

The dc model shown in Figure 4–2 is based on the assumptions made above.

The gain of the amplifier with reference to the negative input, V_n , is:

$$\frac{V_{dc}}{V_{ref}} = -\frac{R_4}{R_3} \text{ or } V_{dc} = -V_{ref} \frac{R_4}{R_3}$$

The gain of the amplifier with reference to the positive input, V_p is:

$$\frac{V_{dc}}{V_{ref}} = \left(1 + \frac{R_4}{R_3}\right) \text{ or } V_{dc} = V_{ref} \left(1 + \frac{R_4}{R_3}\right)$$

Superimposing the positive and negative dc gains of the amplifier results in:

$$V_{dc} = V_{ref}$$

The output of the amplifier is referenced to V_{ref} . The ac response is superimposed upon this dc level.

4.1.2 AC Analysis

For ac analysis break the circuit into 3 parts and determine the transfer functions;

- $H1(s) = V_p / V_i$
- $H2(s) = V_o / V_p$
- $H3(s) = V_{adc} / V_o$

Combine the results to obtain the overall transfer function:

- $H(s) = V_{adc} / V_i$

Where V_i is the input signal from the shock sensor and V_{adc} is the output signal to the analog-to-digital converter. Figure 4–3 shows the ac model with the dc sources shorted.

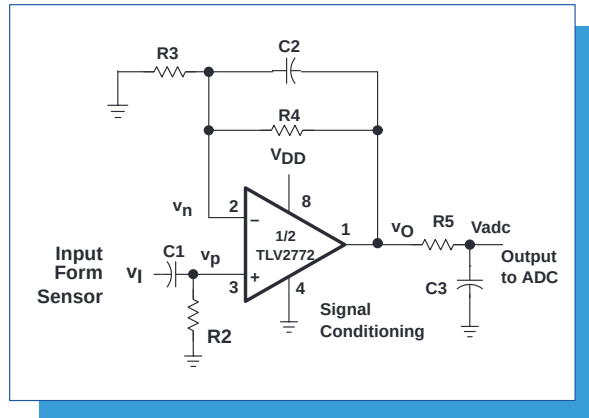


Figure 4–3. AC Circuit Model

4.1.2.1 $H1(s) = V_p / V_i$

Capacitor C1 and resistor R2 form a passive high-pass filter from the sensor input to the high impedance positive input to the TLV2772 where:

$$V_p = V_i \frac{R_2}{R_2 + 1/sC_1} \text{ or}$$

$$H1(s) = \frac{V_p}{V_i} = \frac{R_2}{R_2 + 1/sC_1}$$

4.1.2.2 $H2(s) = V_o / V_p$

The amplifier gain is found by solving for $H2(s) = V_o / V_p$. The solution is a non-inverting amplifier with:

$$H2(s) = \frac{V_o}{V_p} = \left(1 + \frac{Z}{R_3}\right)$$

Where:

$$Z = \frac{R_4}{1 + sC_2R_4}$$

substituting

$$H2(s) = \frac{V_o}{V_p} = 1 + \frac{R_4}{R_3} \left(\frac{1}{1 + sC_2R_4} \right)$$

4.1.2.3 $H3(s) = V_{adc} / V_o$

Assuming that the input impedance to the TLV1544 ADC is very high in comparison to the impedance of C3 and R5, C3 and R5 form a passive low-pass filter where:

$$V_{adc} = V_o \frac{1}{1 + sC_3R_5} \text{ or}$$

$$H3(s) = \frac{V_{adc}}{V_o} = \frac{1}{1 + sC_3R_5}$$

4.1.2.4 $H(s) = V_{adc} / V_i$

Superimposing the results from above gives the overall transfer function:

$$H(s) = \frac{V_{adc}}{V_i} = \left(\frac{R_2}{R_2 + 1/sC_1} \right) \times \left(1 + \frac{R_4}{R_3} \left(\frac{1}{1 + sC_2R_4} \right) \right) \times \left(\frac{1}{1 + sC_3R_5} \right)$$

To find the complete response, add the ac and dc components so that:

$$V_{adc} = V_i H(s) + V_{ref}$$

4.1.3 Gain Calculation

Since the TLV2772 is capable of rail-to-rail output, with a 3 V supply, $V_{out\ min} = 0\text{ V}$ and $V_{out\ max} = 3\text{ V}$. With no signal from the sensor, $V_{out\ nom} = \text{reference voltage} = 1.23\text{ V}$. Therefore, the maximum negative swing from nominal is $0\text{ V} - 1.23\text{ V} = -1.23\text{ V}$, and the maximum positive swing is $3\text{ V} - 1.23\text{ V} = 1.77\text{ V}$.

Model the shock sensor as a low impedance voltage source with output of 2.25 mV/g max in the x and y axis and 1.70 mV/g max in the z axis, and calculate the required amplification of the signal conditioning circuit as follows:

$$\text{Gain} = \text{Output Swing} \div (\text{Sensor Sensitivity} \cdot \text{Acceleration})$$

3-V Accelerometer Featuring TLV2772

To avoid saturating the op amp, base the gain calculations on the maximum negative swing of -1.23 V and the maximum sensor output of 2.25 mV/g for the x and y axis, and 1.70 mV/g for the z axis. Therefore:

$$\text{Gain}(x,y) = -1.23\text{ V} \div (2.25\text{ mV/g} \times (-50\text{g})) = 10.9 \text{ and}$$

$$\text{Gain}(z) = -1.23\text{ V} \div (1.70\text{ mV/g} \times (-50\text{g})) = 14.5$$

Choosing $R3 = 10\text{ k}\Omega$ and $R4 = 100\text{ k}\Omega$, gives a gain of 11 in the x and y channels. Choosing $R3 = 7.5\text{ k}\Omega$ and $R4 = 100\text{ k}\Omega$, gives a gain of 14.3 in the z channel.

4.1.4 Bandwidth Calculations

To calculate the component values for the frequency shaping characteristics of the signal conditioning circuit, use 1 Hz and 500 Hz as the minimum required 3-dB bandwidth from the specifications requirements.

4.1.4.1 Component Values for $H1(s) = V_p / V_i$

To minimize the value of the input capacitor required to set the lower cutoff frequency, a large value resistor is required for $R2$. A $1\text{ M}\Omega$ resistor is used here. To set the lower cutoff frequency, the value for capacitor $C1$ must be:

$$C1 = 1 \div (2\pi \times \text{lower cutoff frequency (Hz)} \times R2 (\Omega))$$

$$C1 = 1 \div (6.28 \times 1\text{ Hz} \times 1\text{ M}\Omega) = 0.159\text{ }\mu\text{F}$$

A more common $0.22\text{-}\mu\text{F}$ capacitor is used for $C1$. This moves the lower cutoff frequency to 0.724 Hz .

Figure 4–4 shows the bode plot approximation to the input transfer function, $H1(s) = V_p/V_i$.

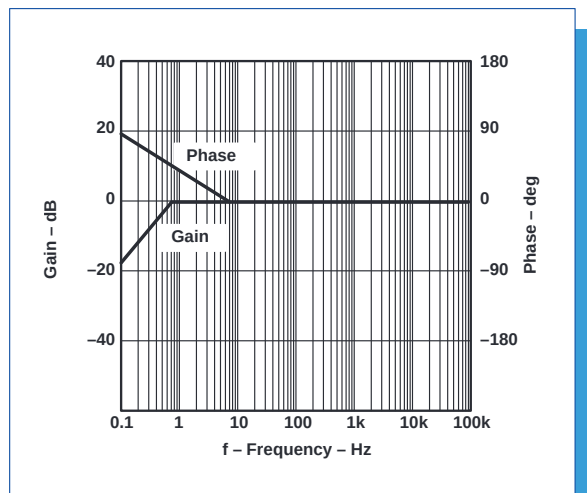


Figure 4–4. Bode Plot $H1(s) = V_p/V_i$

4.1.4.2 Component Values for $H2(s) = V_o/V_p$

To minimize phase shift in the feedback loop caused by the input capacitance of the TLV2772, it is best to minimize the value of feedback resistor $R4$. Also, to reduce the required capacitance in the feedback loop, a large-value resistor is required for $R4$. A compromise value of $100\text{ k}\Omega$ is used for $R4$. To set the upper cutoff frequency, the required capacitor value for $C2$ is:

$$C2 = 1 \div (2\pi \times \text{upper cutoff frequency (Hz)} \times R4 (\Omega))$$

$$C2 = 1 \div (6.28 \times 500\text{ Hz} \times 100\text{ k}\Omega) = 3.18\text{ nF}$$

A more common 2.2 nF capacitor is used for $C2$. This changes the upper cutoff frequency to 724 Hz .

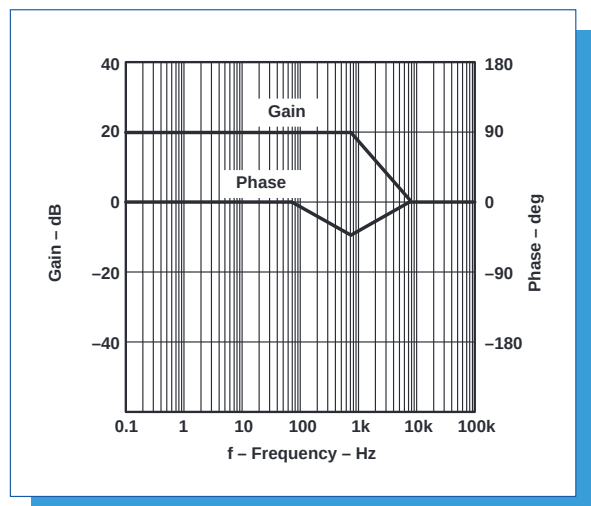


Figure 4–5. Bode Plot of $H2(s) = V_o/V_p$

Using the amplifier gain calculated above, Figure 4–5 shows the bode plot approximation to the transfer function $H2(s) = V_o/V_p$ for the x and y channels. The z channel is the same except the gain is slightly higher in the pass band.

4.1.4.3 Component Values for $H3(s) = V_{adc}/V_o$

Resistor $R5$ and capacitor $C3$ cause the signal response to roll-off further. To set the frequency for this roll off to begin at the upper cutoff frequency, select $1 \div (2\pi \cdot C3(F) \times R4(\Omega)) = \text{upper cutoff frequency(Hz)}$.

With $R5 = 1\text{ k}\Omega$ and $C3 = 0.22\text{ }\mu\text{F}$, the roll-off frequency $= 1 \div (6.28 \times 0.22\text{ }\mu\text{F} \times 1\text{ k}\Omega) = 724\text{ Hz}$.

Figure 4–6 shows the bode plot approximation to the transfer function $H3(s) = V_{adc}/V_o$.

3-V Accelerometer Featuring TLV2772

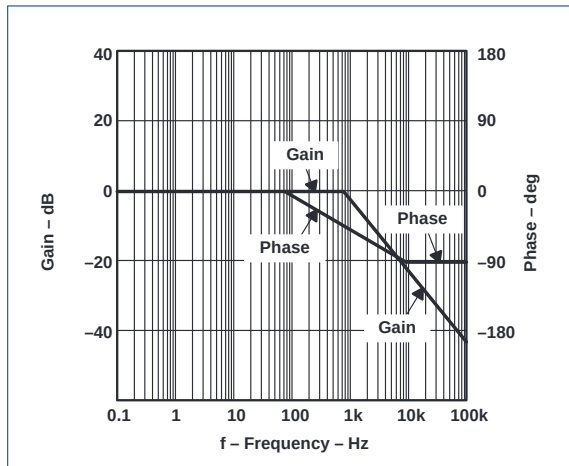


Figure 4-6. Bode Plot of $H_3(s) = V_{adc}/V_o$

4.1.4.4 Transfer Function $H(s) = V_{adc}/V_i$

Superimposing the previous bode plot approximations results in the bode plot approximation for the overall transfer function to be expected from the signal conditioning circuit as shown in Figure 4-7.

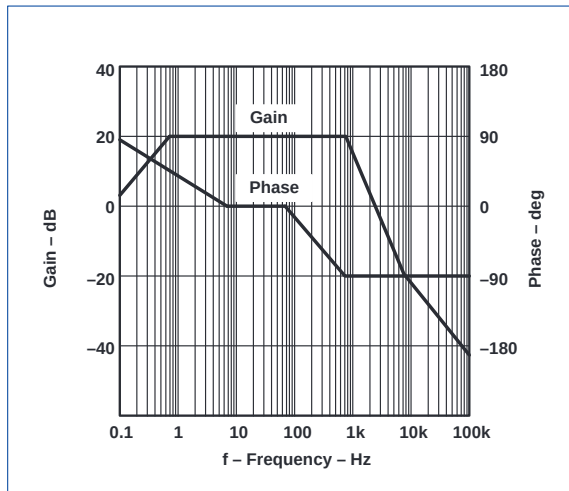


Figure 4-7. Bode Plot of $H(s) = V_{adc}/V_i$

4.2 Spice Simulation

Spice simulation verifies the results of hand analysis and provides a more accurate result than what is practical by hand.

The proper models must be used to perform SPICE simulation of the shock sensor and signal conditioning circuit.

The data sheet for the ACH-04-05-08 shock sensor states that the internal JFET used to drive the output is similar to the industry standard 2N4117. To model the sensor, a signal source is used to drive the gate of a 2N4117 JFET and proper bias is applied.

Modeling the signal conditioning circuit is straight forward except that most available SPICE versions do not have a library model for the TLV2772. This is easily remedied. Place a similar part on the schematic and modify its model with the model editor to match that of the TLV2772. Figure 4-8 below shows the TLV2772 sub-circuit model (as printed in the TLV2772 data sheet).

3-V Accelerometer Featuring TLV2772

```
.SUBCKT TLV2772-X 1 2 3 4 5
C1 11 12 2.3094E-12
C2 6 7 8.0000E-12
CSS 10 99 2.1042E-12
DC 5 53 DY
DE 54 5 DY
DLP 90 91 DX
DLN 92 90 DX
DP 4 3 DX
EGND 99 0 POLY(2) (3,0) (4,0) 0 .5 .5
FB 7 99 POLY(5) VB VC VE VLP VLN 0 19.391E6 - 1E3 1E3 19E6 - 19E6
GA 6 0 11 12 150.80E-6
GCM 06 10 99 7.5576E-9
ISS 3 10 DC 116.40E-6
HLIM 90 0 VLIM IK
J1 11 2 10 JX1
J2 12 1 10 JX2
R2 6 9 100.0E3
RD1 4 11 6.6315E3
RD2 4 12 6.6315E3
RO1 8 5 17.140
RO2 7 99 17.140
RP 3 4 4.5455E3
RSS 10 99 1.7182E6
VB 9 0 DC 0
VC 3 53 DC .1
VE 54 4 DC .1
VLIM 7 8 DC 0
VLP 91 0 DC 47
VLN 0 92 DC 47
.MODEL DX D(IS=800.0E-18)
.MODEL DY D(IS=800.0E-18 Rs=1m Cjo=10p)
.MODEL JX1 PJF(IS=2.250E-12 BETA=195.36E-6 VTO=-1)
.MODEL JX2 PJF(IS=1.750E-12 BETA=195.36E-6 VTO=-1)
.ENDS
```

Figure 4–8. TLV2772 Sub-Circuit Model

Figure 4–9 shows the schematic used for SPICE simulation. Figures 4–10 show the simulation results.

3-V Accelerometer Featuring TLV2772

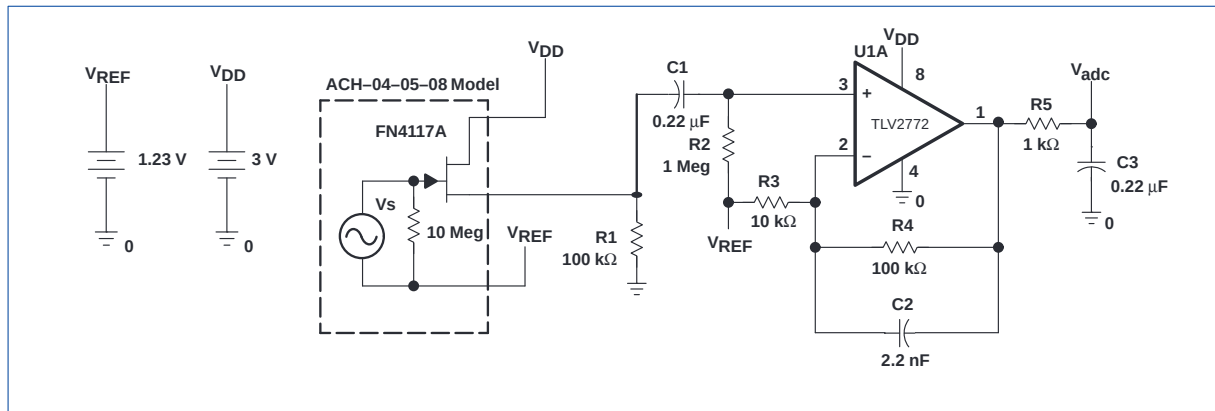


Figure 4–9. SPICE Simulation Schematic

The SPICE simulations show agreement with the hand calculations and the bode plot approximations.

The analyses above do not consider the frequency response of the shock sensor, which falls off below 0.5 Hz and above 5 kHz. The data sheet for the

ACH-04-08-05 does not specify the roll-off characteristics of the shock sensor's frequency response, but in general it is expected that there will be more attenuation in the signal below 0.5 Hz and above 5 kHz.

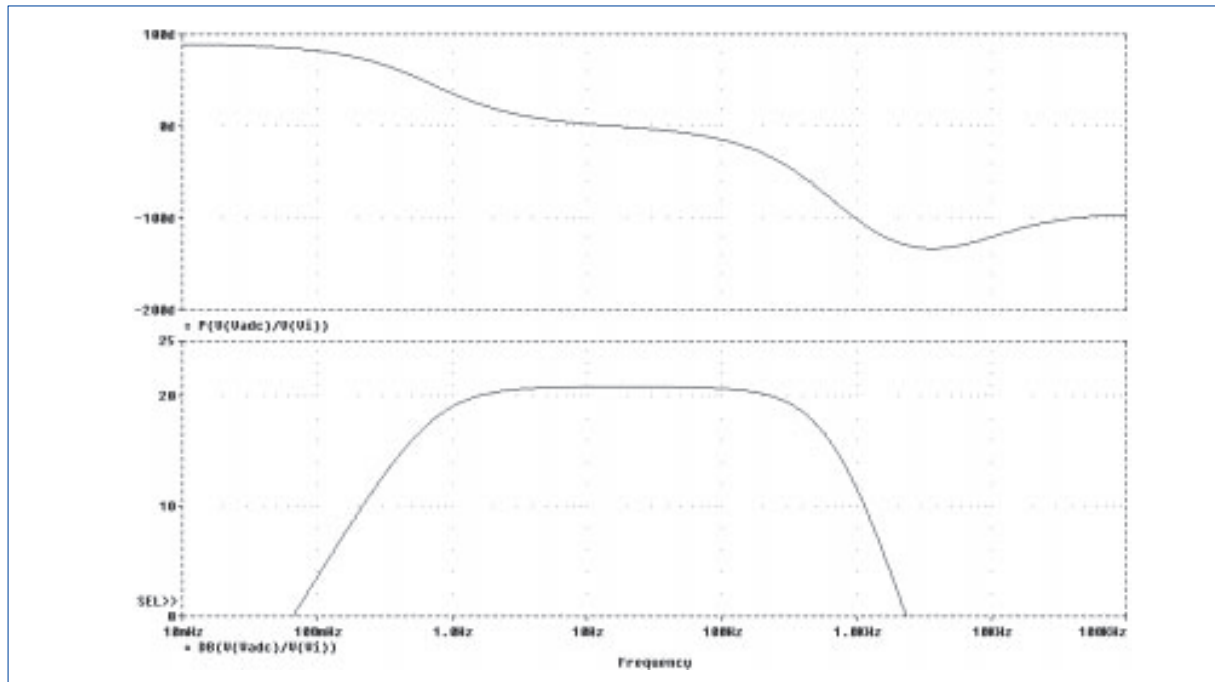


Figure 4–10. SPICE Simulation Results

3-V Accelerometer Featuring TLV2772

5 Circuit Realization

The shock sensor and signal conditioning circuits are built on area 100 of two Universal Operational Amplifier EVM boards (SLOP 120–1). One EVM board holds a TLV2772, an ACH-04-05-08, a TLV431, and required ancillary devices. The other board has only a TLV2772 and required ancillary devices. The signal conditioning circuit is constructed by installing required components and wiring. The two boards share signals and sources

through board-to-board connectors. Standoffs and screws secure the two boards together. Figure 5–1 shows the schematic diagram with reference designators for using area 100 on two Universal Operational Amplifier EVM boards.

Tables 5–1 and 5–2 summarize reference designator part descriptions to construct the shock sensor and signal conditioning circuits.

Table 5–1. Board 1 – Universal Operational Amplifier EVM Area 100

REFERENCE DESIGNATOR	DESCRIPTION	REFERENCE DESIGNATOR	DESCRIPTION
R101	1 M Ω 1% SMT	C101	Not used
R102	10 k Ω 1% SMT	C102	2200 pF 5% NPO SMT
R103	Not used	C103	2.2 μ F 20% Y5V SMT
R104	100 k Ω 1% SMT	C104	0.1 μ F 10% X7R SMT
R105	100 k Ω 1% SMT	C105	2200 pF 5% NPO SMT
R106	Not used	C106	Not used
R107	10 k Ω 1% SMT	C107	Not used
R108	1 M Ω 1% SMT	C108	Not used
R109	Use 0.22 μ F 10% X7R SMT Capacitor	C109	0 Ω or Jumper
R110	Use 0.22 μ F 10% X7R SMT Capacitor	C110	Not used
R111	0 Ω or Jumper	C111	Not used
R112	Not used	C112	Not used
R113	Not used	U101	TLV2772CD
R114	2.2 k Ω 5% SMT	U102	TLV431ACDBV5
R115	0 Ω or Jumper	V1+	3 V power input
R116	Not used	VREF1	Signal conditioning reference
R117	Not used	GND1	Signal and power ground
R118	Not used	R119	0 Ω or Jumper
A1OUT	X Axis output	B104+	Y Axis input form ACH-04-05-08 pin 2
A101–	Not used	B103+	Jump to VREF1
A102–	Jump to VREF1	B102–	Jump VREF1
A103+	Jump to VREF1	B101–	Not used
A104+	X Axis input from ACH-04-05-08 pin 6	B1OUT	Y Axis output

Table 5–2. Board 2 – Universal Operational Amplifier EVM Area 100

REFERENCE DESIGNATOR	DESCRIPTION	REFERENCE DESIGNATOR	DESCRIPTION
R101	Not used	C101	Not used
R102	Not used	C102	Not used
R103	Not used	C103	Not used
R104	0 Ω or Jumper	C104	0.1 μ F 10% X7R SMT
R105	100 k Ω 1% SMT	C105	2200 pF 5% NPO SMT

3-V Accelerometer Featuring TLV2772

Table 5–2. Board 2 – Universal Operational Amplifier EVM Area 100

REFERENCE DESIGNATOR	DESCRIPTION	REFERENCE DESIGNATOR	DESCRIPTION
R106	Not used	C106	Not used
R107	7.5 kΩ 1% SMT	C107	Not used
R108	1 MΩ 1% SMT	C108	Not used
R109	Use 0.22 μF 10% X7R SMT Capacitor	C109	0 Ω or Jumper
R110	Not used	C110	Not used
R111	Not used	C111	Not used
R112	Not used	C112	Not used
R113	0 Ω or Jumper	U101	TLV2772CD
R114	Not used	U102	Not used
R115	Not used	V1+	3 V from Board 1
R116	Not used	VREF1	Signal conditioning reference from Board 1
R117	Not used	GND1	Circuit common connect to board 1 GND1
R118	Not used	R119	0 Ω or Jumper
A1OUT	Z Axis output	B104+	Not used
A101–	Not used	B103+	Not used
A102–	Jump to VREF1	B102–	Not used
A103+	Jump to VREF1	B101–	Not used
A104+	Z Axis input from ACH-04-05-08 pin 12	B1OUT	Not used

The ACH-04-08-05 shock sensor mounts in the breadboard area on board 1. Table 5–3 summarizes the connections.

Table 5–3. Board 1 – Universal Operational Amplifier EVM Area 100 ACH-04-08-05 Connections

ACH-04-08-05		
PIN	NAME	BOARD 1 CONNECTIONS
1	D/S1–Y	V1+
2	D/S2–Y	B104+
3	CTG	GND1
4	SGND	VREF1
5	D/S1–X	V1+
6	D/S2–X	A104+
7	RGND–X	VREF1
8	GND	GND1
9	CTG	GND1
10	RGND–Z	VREF1
11	D/S1/Z	V1+
12	D/S2–Z	To board 2
13	CTG	GND1
14	RGND–Y	VREF1

Board-to-board connectors provide required connections between board 1 and board 2. The required board-to-board connections are GND1, V1+, and VREF. Also, pin 12 of the ACH-04-08–05 must be connected to A104+ on board 2. This is accomplished by using board-to-board connectors in the breadboard area and connecting the appropriate nodes thereto.

Figure 5–1 is the schematic of the 3-axis realization of the 1-axis circuit shown in Figure 4–1, except for output filter components R5 and C3. To ease component placement, the output filters for each channel are placed on the TLV1544 EVM. For standalone testing of the signal conditioning circuit without the TLV1544 EVM, temporarily place the filter components on board 1 and board 2. Remove them for final integration.

3-V Accelerometer Featuring TLV2772

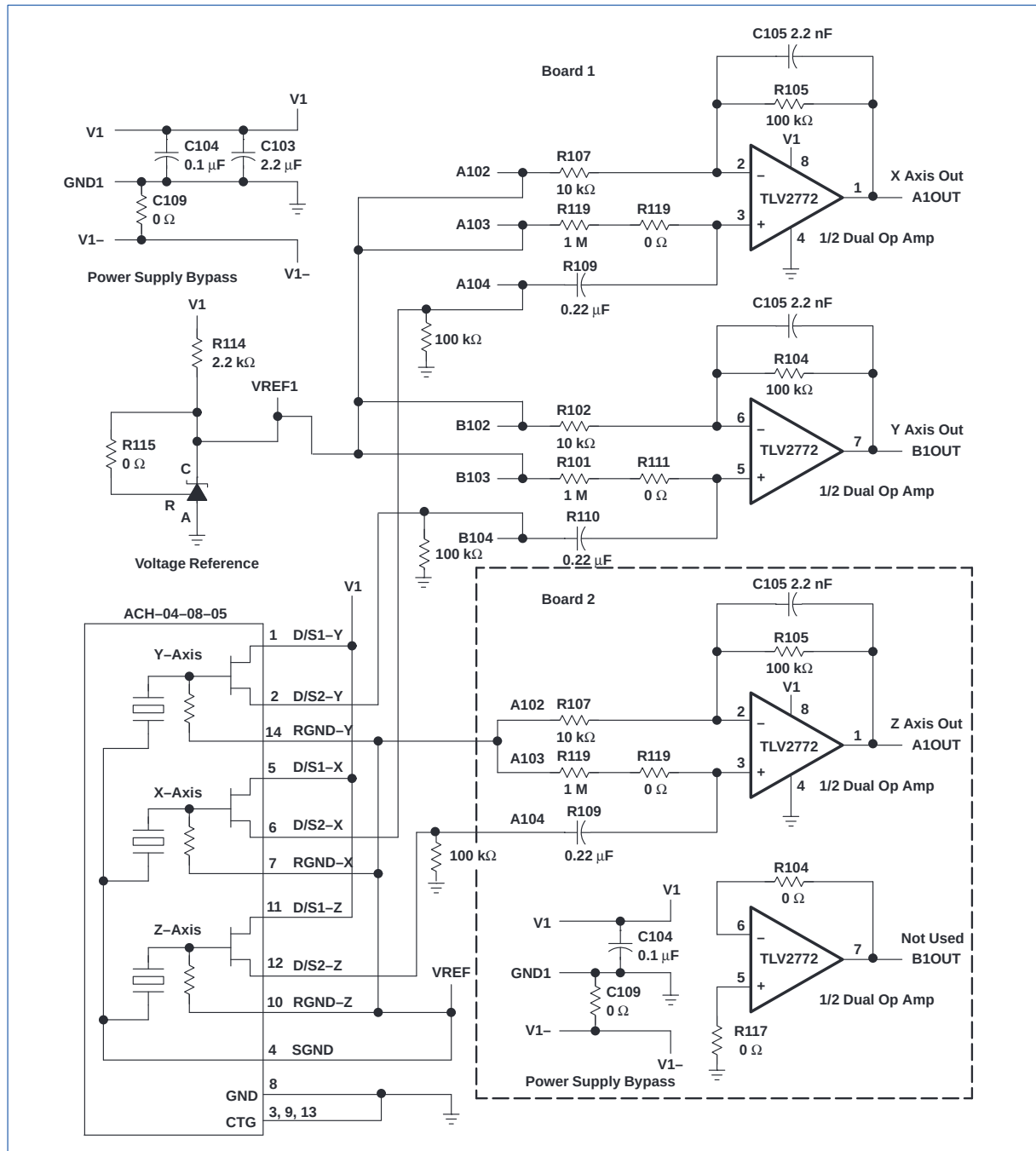


Figure 5-1. Signal Conditioning Schematic using Two Universal Operational Amplifier EVM Boards

3-V Accelerometer Featuring TLV2772

5.1 Test of Signal Conditioning Circuit

After constructing the signal conditioning circuit, and before connecting the shock sensor, tests were run to verify the signal conditioning circuit performed as

expected. An AP Instruments network analyzer was used to measure the gain and phase response. Figure 5–2 shows the results are very much the same as the SPICE simulation predicted.

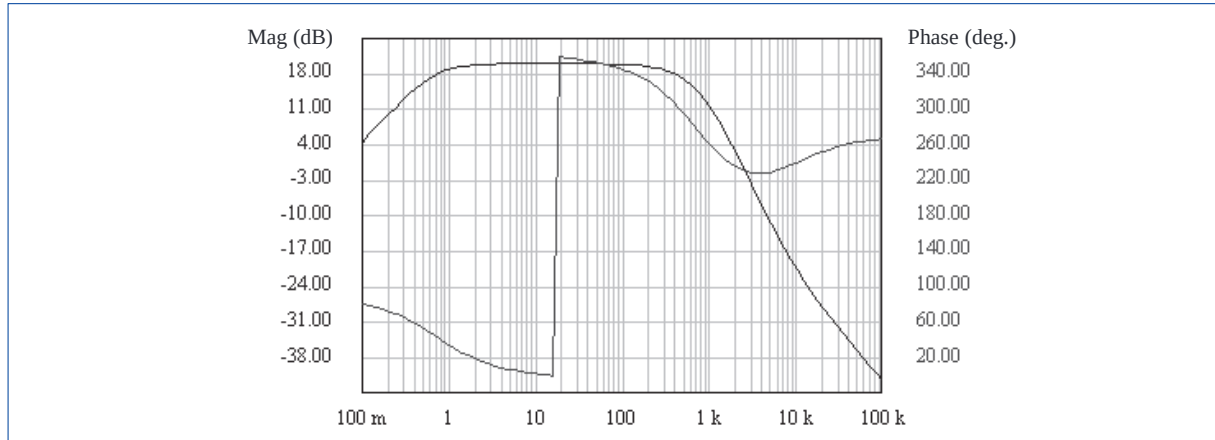


Figure 5–2. Network Analyzer Display

5.2 Test of Shock Sensor and Signal Conditioning Circuit

After connecting the shock sensor, the measurement of acceleration was verified. A spring loaded test fixture was constructed to induce simple harmonic motion into the sensor. Figure 5–3 shows a diagram of the test fixture. By measuring the deflection from center and the frequency of the oscillation, the acceleration is calculated as:

$$a = -\omega^2 \times \cos(\omega t + \phi)$$

Where

a is acceleration

ω is frequency of oscillation in radians

x is the maximum deflection from center

ϕ is the phase

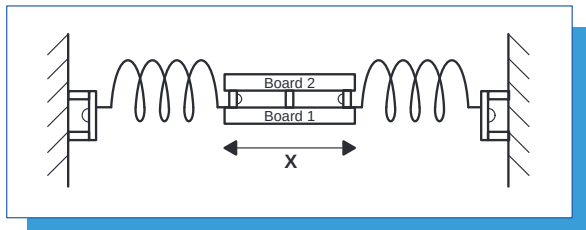


Figure 5–3. Spring Test Fixture

While monitoring the output of all 3 axes with an oscilloscope, the system was set into motion along the x-axis. Figure 5–4 shows the result displayed on an oscilloscope. Channel 1 is the x-axis, channel 2 is the y-axis, and channel 3 is the z-axis.

About 0.25" mechanical deflection was measured visually. An oscilloscope was used to measure the frequency of oscillation and the peak voltage amplitude. Using the above formula to find the peak acceleration expected from the mechanical movement (ignoring the sign):

$$a \text{ (peak)} = (2\pi \times 16)^2 \cdot (0.25 \text{ in/s}^2) = 2524 \text{ in/s}^2 = 6.58g$$

Where $g = 32 \text{ ft/s}^2$

To check the output of the signal conditioning circuit, use the formula:

$$a \text{ (peak)} = \frac{\text{Peak voltage output}}{\text{Sensor output} \cdot \text{Gain}} = 128 \text{ mV} \cdot \frac{g}{1.80 \text{ mV}} \cdot \frac{1}{11} = 6.46 g$$

Using the typical x-axis output = 1.80mV/g, there is good agreement between the two calculations.

3-V Accelerometer Featuring TLV2772

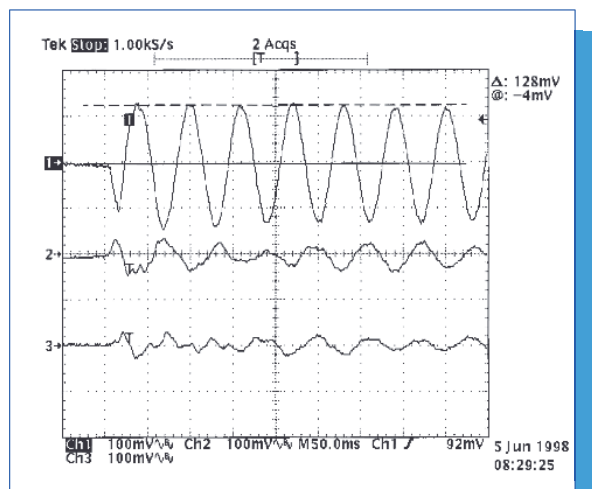


Figure 5-4. Output Displayed on Oscilloscope

Note that output is seen on the y-axis and z-axis. The ACH-04-08-05 data sheet specifies a transverse sensitivity of 15%. Fifteen percent of the x-axis peak signal output is 19.2 mV. The peak y axis signal is 40 mV and the peak z-axis signal is 26 mV. Transverse sensitivity does not account for the observation. While the primary motion was in the x-axis, motion was noticed in the other 2 axes as well.

5.3 TLV1544 EVM

For this application the fast conversion mode of the TLV1544 ADC is used and three of the four analog inputs are converted one after the other. This results in a sampling frequency of 28 ksp/s. By Nyquist's sampling theorem the frequency content of the input analog signal must be insignificant for frequencies $\frac{1}{2}$ the sampling frequency and above. In this system, the signal is attenuated by over 40 dB at the Nyquist frequency.

5.3.1 Interfacing the Signal Conditioning Circuit to the TLV1544 EVM

The TLV1544 EVM provides a power supply, analog signal conditioning circuitry, ADC (TLV1544), and circuitry to support the interface to a DSP or micro controller.

Refer to the TLV1544 EVM User's Guide, literature #SLAU014, for detailed descriptions of the board and its functions.

To interface the shock sensor and signal conditioning circuit to the TLV1544 EVM, minor modifications are required on the TLV1544 EVM. To ease circuit realization, place the output filter for each axis signal, R5 and C3 in Figure 4-1, on the TLV1544 EVM.

The TLV1544 has 4 analog inputs. In this application, channel 0 samples the reference voltage from the signal conditioning circuit, channel 1 samples the x-axis output, channel 2 samples the y-axis output, and channel 3 samples the z-axis output. Make the following modifications to the TLV1544 EVM, REV-:

Channel 0 Modifications:

- Remove U4, R26, R27, and R28.
- Jumper U4-2 to U4-3.
- Place a 1-k Ω resistor for R28.
- Place a 22- μ F capacitor between A0 and ground.

Channel 1 Modifications:

- Remove U5, R29, and C19.
- Jumper U5-3 to U5-6.
- Place a 1-k Ω resistor for R29.
- Place a 0.22- μ F capacitor for C19.

Channel 2 Modifications:

- Place a 1-k Ω resistor for JP3.
- Place a 0.22- μ F capacitor across D6.

Channel 3 Modifications:

- Place a 1-k Ω resistor for JP4.
- Place a 0.22- μ F capacitor across D8.

Switch SW1 to operate the TLV1544 EVM at 3 V. Set the TLV1544 reference voltage input to 2.4 V by installing jumpers JP1-1, 2 and JP2-1, 2. Use the 3 V power supply on the TLV1544 EVM to supply power to the shock sensor and the signal conditioning circuit.

Table 5-4 shows the interface connections between the signal conditioning circuit and the TLV1544 EVM, REV-.

Table 5-4. Interface Between Signal Conditioning Circuit and TLV1544 EVM

SIGNAL CONDITIONING CIRCUIT	TLV1544 EVM
Board 1 V1+	AV _{DD}
Board 1 GND1	GND
Board 1 VREF1 (signal reference)	JP19 (U2-2)
Board 1 A1OUT (X axis output)	Analog Input 1
Board 1 B1OUT (Y axis output)	Analog Input 2
Board 2 A1OUT (Z axis output)	Analog Input 3

Figure 5-5 is a schematic drawing for the interface between the signal conditioning circuit and the TLV1544.

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5.4 Interfacing the TLV1544 EVM to the TMS320C5X EVM

Figure 5–6 shows the connections required to interface the TLV1544 EVM to the TMS320C5X EVM. For

further details refer to *Interfacing the TLV1544 Analog-to-Digital Converter to the TMS320C50 DSP Applications Report* literature # SLAA025.

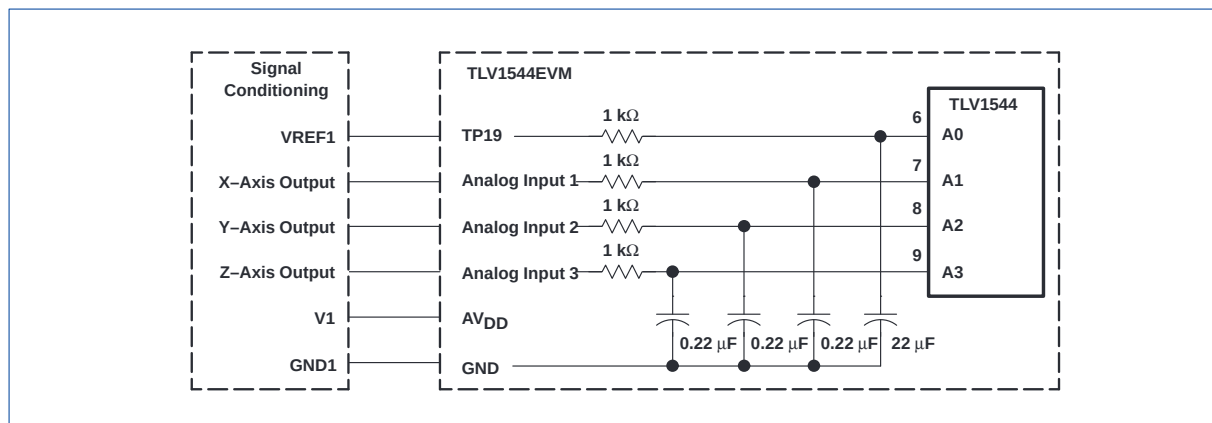


Figure 5–5. Schematic – Signal Conditioning to TLV1544 ADC

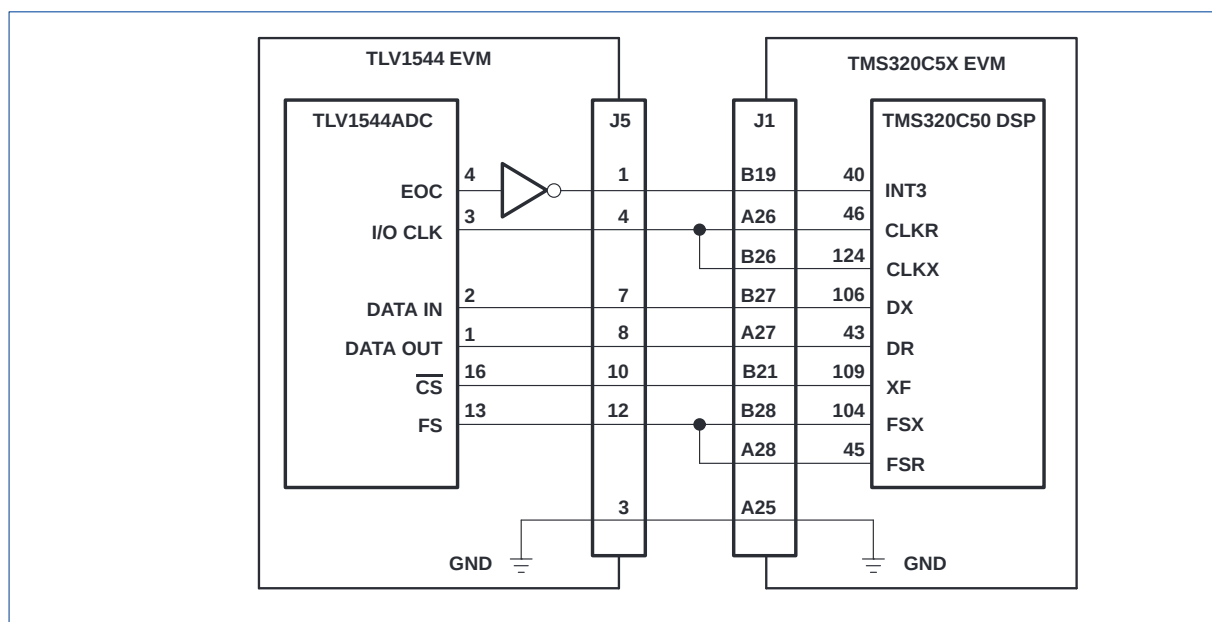


Figure 5–6. Interface Between TLV1544 EVM and TMS320C5X EVM

5.5 The TMS320C5X EVM

The TMS320C5X EVM plugs into an ISA slot in a PC. Refer to *TMS320C5X Evaluation Module Technical Reference* literature # SPRU087 for detailed information on its use. The TMS320C5X C Source Debugger is also required. Refer to *TMS320C5X C*

Source Debugger User's Guide literature # SPRU055B for user information.

6 Error and Noise Analysis

Error analysis can predict the errors to be expected under normal operating conditions. Two analyses are presented: gain error and noise.

Gain errors contribute to uncertainty in the system output given a known input.

Noise is a random electrical event that occurs in all conductors. Noise imposes a statistical uncertainty on the system's electrical signals.

The system can be calibrated to compensate for the effect of gain errors. The uncertainty associated with noise, on the other hand, prohibits any type of compensation.

Two of the most widely used types of error analysis are worst case analysis and sum of squares analysis. The major task in performing an error analysis is to identify and quantify the error sources.

Worst case analysis assumes that all errors compound, one on top of the other. Therefore, the maximum error in all circuit elements is added to arrive at the total system error.

Sum of squares analysis assumes that some errors will tend to cancel each other. By summing the square of all the errors and taking the square root, the result takes this cancellation effect into account. A sum of squares analysis gives a realistic idea of how well the typical system will perform. This is the type of analysis that will be presented. A sum of squares analysis uses the following formula;

$$\text{Total System Error} = \sqrt{\sum_{n=1}^k \text{Error}_n^2}$$

Where the subscript n indicates an individual error component and k is the total number of error components.

System gain error analysis is presented first, followed by system noise analysis.

6.1 System Gain Error Analysis

Gain errors are errors that contribute to uncertainty in the system output, given a known input.

The system is essentially an ac system in which dc errors can be compensated. Therefore, dc errors caused by reference voltage errors, offset voltages, and the like are not included in this error analysis.

6.1.1 Sensor Gain Error

The shock sensor's output varies, and this contributes to uncertainty in the system output given a known input. The data sheet for the ACH-04-08-05 states the output sensitivity:

CHARACTERISTIC	MIN	TYP	MAX	UNIT
X-axis sensitivity	1.35	1.80	2.25	mV/g
Y-axis sensitivity	1.35	1.80	2.25	mV/g
Z-axis sensitivity	1	1.35	1.70	mV/g

This is a $\pm 25\%$ tolerance in the output voltage for the same acceleration input. Thus the error assigned to this component is 25%.

6.1.2 Signal Conditioning Gain Error

Errors in the signal conditioning circuit also contribute to uncertainty in the system output, given a known input. The analysis is limited to the pass band of the amplifier circuit. In its pass band the gain of the amplifier is given by:

$$A = \left(\frac{1}{b}\right) \left[\frac{1}{1 + \frac{1}{ab}} \right]$$

Where:

a = open loop amplification of the operational amplifier

And

$$b = \frac{R4}{R4 + R3}$$

The term $1/ab$ is seen as an error term in this equation because it is desired to set the gain of the circuit with the $1/b$ term (ratio of $R3$ to $R4$). To minimize this error, the open loop amplification of the operational amplifier should be as large as possible so that $1/ab$ is very close to 0. In this application the TLV2772's open loop amplification is the lowest at the upper cutoff frequency. At 1 kHz the open loop amplification is about 75 dB or 5600. Using this value: $1/ab = 0.000016$, and the deviation of the amplifier gain from ideal is 0.002%.

The error caused by the op amp gain setting resistors is the tolerance of the resistors. In this case, 1% resistors add an error of 1%.

6.1.3 ADC Gain Errors

Errors caused by sampling of the signal by the ADC also need to be analyzed. Figure 6–1 shows a simplified schematic to use for analysis.

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Set initial conditions where $V_{adc} = V_o$ and $V_{sample} = 0$. To simplify the analysis, notice that the time constant $R5C3 = 0.22 \text{ ms}$ is much longer than the time constant $R_{dac}C_{dac} = 55 \text{ ns}$, because $C3$ is 4000 times larger than C_{dac} . In this system the TLV1544's I/O CLK is being driven at 5 MHz. The sampling period = 6 I/O CLK periods = $1.2 \text{ } \mu\text{s}$. Because $C3 \gg C_{dac}$, when the switch closes during sampling, $C3$ acts like a voltage reservoir for charging C_{dac} via R_{dac} . During the sampling period very little charge is transferred from V_o . The sampling period, $1.2 \text{ } \mu\text{s} = 29 \times 55 \text{ ns}$ or 29 $R_{dac}C_{dac}$ time constants. Therefore, at the end of the sampling the period, the voltage across C_{dac} and $C3$ is essentially equal, the sample voltage is set by the ratio of $C3$ to C_{dac} ;

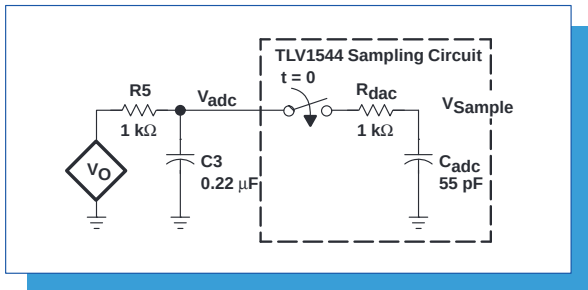


Figure 6-1. Sampling Input Model

$$V_{sample} = V_{adc} \left(\frac{C3}{C_{dac} + C3} \right) = 0.99975 V_{adc}$$

The error attributed is 0.025%.

The other sources of error in the ADC can be read from the data sheet. As stated earlier, the system is essentially an ac system, and dc errors are not included in the error analysis. Therefore, only the maximum linearity error from the data sheet, which is $\pm 1 \text{ LSB}$, is included. With 10 bits of resolution or 1024 codes the error = 0.0976%.

6.1.4 Total Gain Error

From this point on in the system, the signal is digital, and it is assumed there are no errors which cannot be corrected for.

Calculating the total error as the square root of the sum of the error squared results in:

$$\sqrt{25\%^2 + 0.002\%^2 + 1\%^2 + 0.025\%^2 + 0.0976\%^2} = 25.02\%$$

The uncertainty associated with the shock sensor output dominates the error analysis. Calibration is required to insure accurate measurement of acceleration.

6.2 System Noise Analysis

The first step to analyzing system noise is to qualify and quantify noise sources within the system. Next, the components must be added to see the overall effect on the system.

In this system the noise sources are the shock sensor, the resistors, and the amplifier. Since all three channels are essentially the same, the noise analysis is performed on one channel. Reference designators from Figure 4-1 are used with the component values as calculated above for channel z.

Because the z-axis of the shock sensor has about 30% less sensitivity than either x-axis or y-axis, 30% more gain is required in channel z. The results of the analysis show the dominant noise source is the sensor. The noise specification for the sensor is given in $\text{mg}/\sqrt{\text{Hz}}$ and converted to equivalent input noise voltage. Therefore, the noise voltage, as calculated in this analysis, will be higher in channel z, but not significantly so.

It is typical in amplifier circuits to identify equivalent input noise voltage per root Hertz for system components. This is then multiplied by the equivalent noise bandwidth and amplifier noise gain. The result is the total system noise. It is convenient to work in RMS values to compute the total system noise, then to convert to peak-to-peak (pk-pk) and compare to 1 LSB to see the effect on the ADC's output. The following formula is used:

$$\text{Total System Noise (in RMS Volts)} = \sqrt{\sum_{m=1}^k \text{ENB} \cdot e_n^2 \cdot A_n^2}$$

Where ENB is the equivalent noise bandwidth in hertz, the subscript m indicated an individual equivalent input noise source, e_n , in RMS volts per root Hertz, A_n is the noise gain of the amplifier, and k is the total number of noise sources referred to the input.

Normally, noise in RMS voltage is multiplied by numbers ranging from 5 to 7 to convert to peak-to-peak (pk-pk) voltage. This accounts for the large crest factor seen in noise signals. The pk-pk value of noise voltage is of concern when driving the input to an ADC converter. A multiplication factor of 6 is used in this analysis to convert RMS into pk-pk.

The equivalent noise bandwidth (ENB) accounts for the noise transmitted above the upper cutoff frequency as a result of the finite roll-off, or tail of the frequency response curve. For a 2nd order system, ENB is 1.11 times the normal bandwidth. For this system, $\text{ENB} = 1.11(724 - 0.7) = 802.9 \text{ Hz}$. It is convenient and acceptable to use $\text{ENB} = 803 \text{ Hz}$.

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The amplifier noise gain is the noninverting gain of the amplifier, which is equal to 14.3.

6.2.1 Sensor Noise Analysis

From the ACH-04-08-05 data sheet the shock sensor has a specification for equivalent noise at 100 Hz equal to 0.2 mg/ $\sqrt{\text{Hz}}$. Using the typical z-axis value for sensitivity, 1.35 mV/g, the equivalent noise in volts/ $\sqrt{\text{Hz}}$ is 0.27 $\mu\text{V}/\sqrt{\text{Hz}}$. The data sheet gives no further details as to the spectral shape of the noise from the shock sensor. For this analysis, assume the noise voltage from the ACH-04-08-05 is flat over the bandwidth of the system.

6.2.2 Signal Conditioning Noise Analysis

For Vdd = 2.7 V, the TLV2772 data sheet specifies equivalent input noise voltage of 147 nV/ $\sqrt{\text{Hz}}$ at 10 Hz and 21 nV/ $\sqrt{\text{Hz}}$ at 1 kHz. Because the TLV2772 is a CMOS input op amp, the input noise current is insignificant with the circuit values being used.

The resistors in the signal conditioning amplifier's input circuit contribute thermal noise voltage:

$$= \sqrt{\frac{4kTR}{\text{Hz}}}$$

Where k is Boltzmann's Constant = $1.38 \times 10^{-23} \text{J/K}$

T is temperature in Kelvin

R is resistance in ohms

Hz is frequency in Hertz

R2 contributes 128 nV/ $\sqrt{\text{Hz}}$. R3 and R4 appear in parallel at the negative input for noise calculations. Together they contribute 10.7 nV/ $\sqrt{\text{Hz}}$. Resistor R1 adds 40 nV/ $\sqrt{\text{Hz}}$.

Resistor R5 adds thermal noise voltage = $\sqrt{4kTR/\text{Hz}}$ to the system. The value of the resistor is small and the noise is not amplified by the amplifier. Therefore the effect of R5 on the overall system noise is negligible.

Adding the foregoing is straightforward for all components except the amplifier's input noise. Referring to the equivalent input noise voltage vs frequency graph in the data sheet for the TLV2772 (Figure 39) it can be seen that the equivalent input noise voltage is characterized by 1/f noise over the bandwidth of the system's operation. The input noise is a combination of the 1/f noise and the white noise of the op amp: $e_n^2 = (K^2/f) + e_w^2$ where e_n is the equivalent input noise voltage per root Hertz, K is a proportionality constant and f is frequency in Hertz, and e_w is the white noise specification for the op amp. K can be calculated from the values on the equivalent input noise vs frequency graph to be about 450 nV. By integrating the noise density over the frequency range of operation, the input noise can be computed by:

$$e_{oa} = \sqrt{ENB e_w^2 + K^2 \ln \frac{f_H}{f_L}}$$

Where e_{oa} is the equivalent input noise voltage over the bandwidth in RMS volts

K is a proportionality constant, equal to 450 nV RMS

f_L is the lower cutoff frequency, 0.724 Hz and

f_H is the upper cutoff frequency, 724 Hz.

The result is $e_{oa} = 1.32 \mu\text{V RMS}$.

6.2.3 Total Noise Analysis

Adding the system noise sources results in:

Total System Noise =
(in RMS Volts)

$$14.3 \times \sqrt{803[(0.27E-6)^2 + (128E-9)^2 + (10.7E-9)^2 + (40E-9)^2] + (1.32E-6)^2}$$

$$= 0.124 \text{ mV RMS}$$

To convert to pk-pk voltage, multiply by 6. Then Total System Noise pk-pk = 0.743 mV pk-pk.

The TLV1544 ADC has 10 bits of resolution. With a voltage reference = 2.4 V, 0.743 mV pk-pk of noise is equivalent to about 1/3 LSB.

7 System Test and Evaluation

With the shock sensor and signal conditioning circuit interfaced to the TLV1544 EVM and the TMS320C5X EVM, proceed with testing and evaluation of the system performance.

The program listings as published in the *Interfacing the TLV1544 Analog-to-Digital Converter to the TMS320C50 DSP Applications Report* literature #SLAA025 can be used as a starting point. Appendix A contains program listings based on these, except they modify program flow so that the TLV1544 performs consecutive conversions of three of its four channels. When motion in any axis is detected, the program stores 3000h samples in memory from each channel.

Once the data is stored, it can be saved as a COFF file using the "store data" utility provided with the TMS320C5X C-Source Debugger. To read the data into an Excel spread sheet, a utility program is required to

convert the hexadecimal numbers into decimal ASCII format. Source code listings are included in Appendix A.

With the data converted and loaded into Excel, it can be converted into equivalent acceleration in g using the formula:

$$g = \left(\frac{ADC \text{ Sample} \times ADC \text{ Ref Volts}}{2^{10}} - Sig \text{ Ref Volts} \right) \times \left(\frac{1}{Sensor \text{ output} \times Gain} \right)$$

Figure 7–1 shows the x-axis acceleration graphed in Excel from data collected when testing the signal conditioning and shock sensor circuits. It can be seen that the results match the results as presented previously.

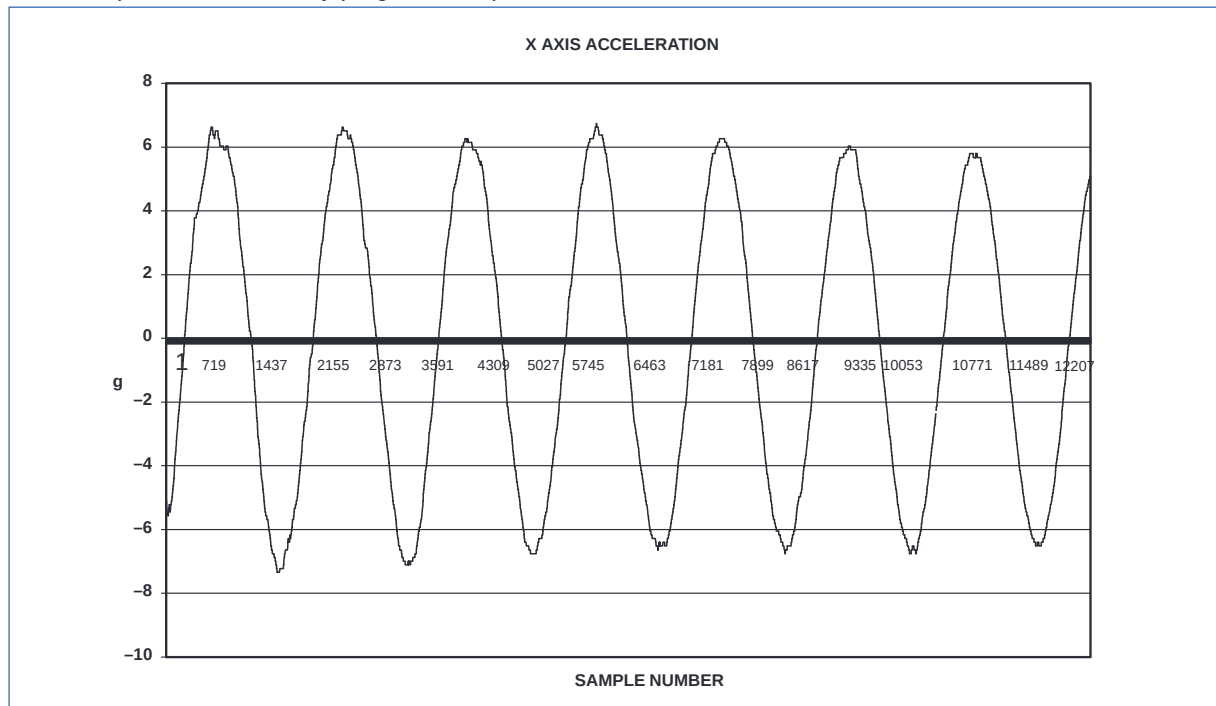


Figure 7–1. X-Axis Acceleration Graphed in Excel

8 Calibration Data/Analysis

8.1 Calibration

The acceleration sensor and the signal conditioning circuitry were housed in a Bud CU-234 box, which was machined to mount on a fixture plate for calibrated testing. The calibrated acceleration source used was an Unholtz Dickie T-1000 vibration tester. The specifications for this machine are:

- 1 inch maximum displacement
- 70 in/s maximum velocity
- 75g maximum sine acceleration $\pm 10\%$ operational tolerance
- 5 Hz to 5 kHz frequency range $\pm 1/4$ Hz frequency operational tolerance
- Cross axis output 25% maximum

Signals from the signal conditioning circuitry were routed to the TLV1544 ADC via coaxial cables. Sine wave acceleration was used and data was collected at various acceleration levels and frequencies on all 3 axis. Data for each axis was plotted over frequency and acceleration using the mean sensitivity. The data was analyzed and the errors calculated. Tables 8–1 through 8–6 and Figures 8–1 through 8–6 show the result. Figure 8–7 shows the output vs. input for each axis when averaged over frequency.

A maximum cross axis sensitivity of 17.6% from y-axis to x-axis was observed. Resonance was observed in the z-axis at 2 kHz.

Table 8–1. X-Axis Output Vs. Input Using X-Axis
Mean Sensitivity = 1.16 mV/g

g/f	10 Hz	20 Hz	50 Hz	100 Hz	500 Hz
1	1.20	1.20	1.20	1.05	0.83
10		11.50	11.00	10.56	8.70
20			21.86	20.94	17.45
50			53.80	51.15	42.60

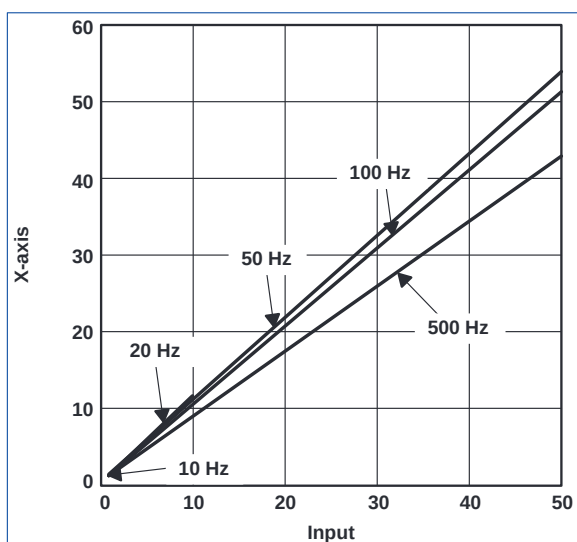


Figure 8–1. X-Axis Output vs. Input

3-V Accelerometer Featuring TLV2772

Table 8-2. X-Axis % Error

g/f	10 Hz	20 Hz	50 Hz	100 Hz	500 Hz
1	20.00%	20.00%	20.00%	5.00%	-17.50%
10		15.00%	10.00%	5.60%	-13.00%
20			9.30%	4.70%	-12.75%
50			7.60%	2.30%	-14.80%

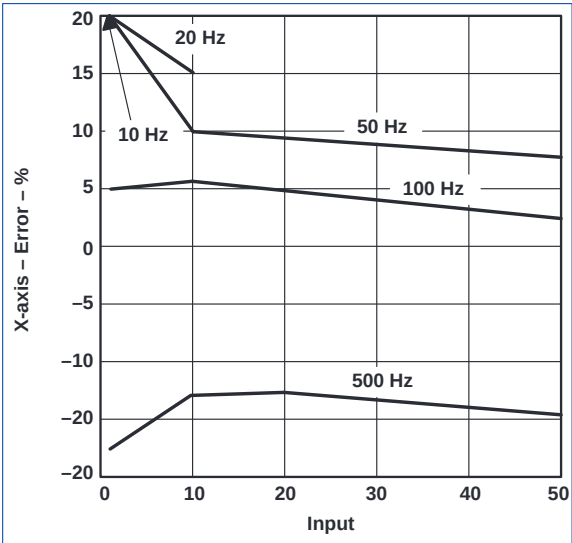


Figure 8-2. X-Axis % Error

Table 8-3. Y-Axis Output Vs. Input Using Y-Axis Mean Sensitivity = 1.35 mV/g

g/f	10 Hz	20 Hz	50 Hz	100 Hz	500 Hz
1	1.03	1.03	1.11	1.03	0.88
10		10.50	9.80	9.70	9.85
20			19.65	19.18	20.10
50			49.65	47.00	52.55

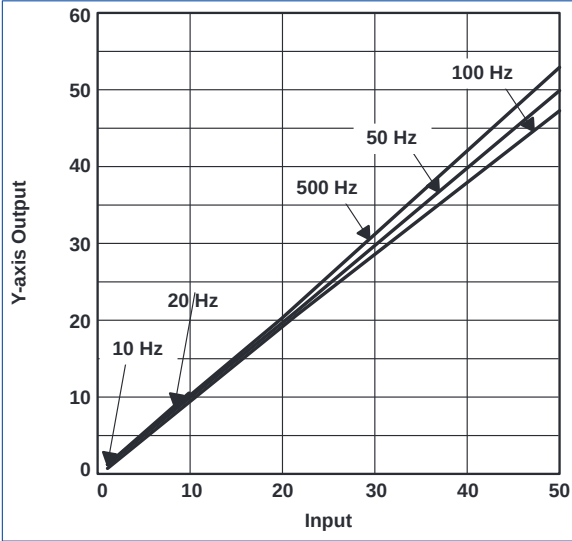


Figure 8-3. Y-Axis Output vs. Input

3-V Accelerometer Featuring TLV2772

Table 8-4. Y-Axis % Error

g/f	10 Hz	20 Hz	50 Hz	100 Hz	500 Hz
1	2.50%	2.50%	10.50%	-2.50%	-12.50%
10		5.00%	-2.00%	-3.00%	-1.50%
20			-1.75%	-4.13%	0.50%
50			-0.70%	-6.00%	5.10%

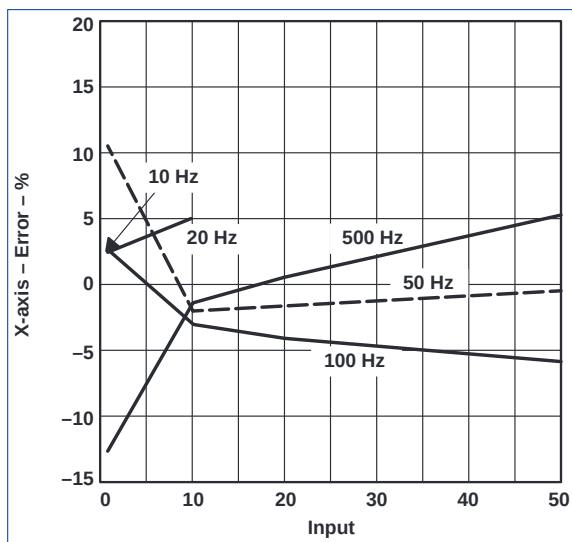


Figure 8-4. Y-Axis % Error

Table 8-5. Z-Axis Output Vs. Input Using Z-Axis Mean Sensitivity = 1.01 mV/g

g/f	10 Hz	20 Hz	50 Hz	100 Hz	500 Hz
1	0.89	0.97	0.97	0.97	1.13
10		8.20	8.35	8.20	11.70
20			16.80	16.55	23.40
50			41.30	41.55	59.10

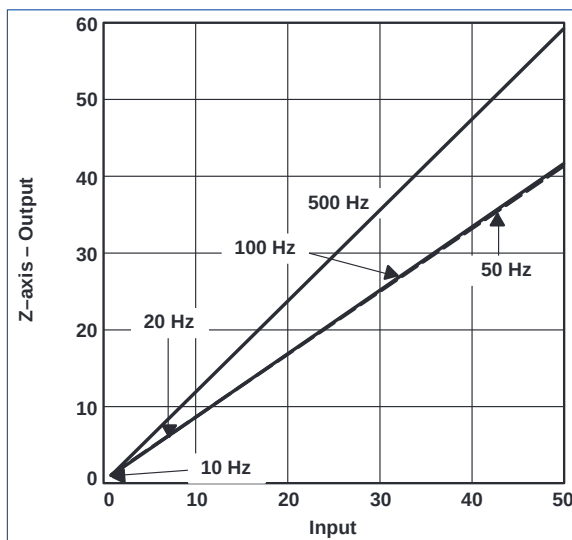


Figure 8-5. Z-Axis Output vs. Input

3-V Accelerometer Featuring TLV2772

Table 8–6. Z-Axis % Error

g/f	10 Hz	20 Hz	50 Hz	100 Hz	500 Hz
1	–11.00%	–3.00%	–3.00%	–3.00%	13.00%
10		–18.00%	–16.50%	–18.00%	17.00%
20			–16.00%	–17.25%	17.00%
50			–17.40%	–16.90%	18.20%

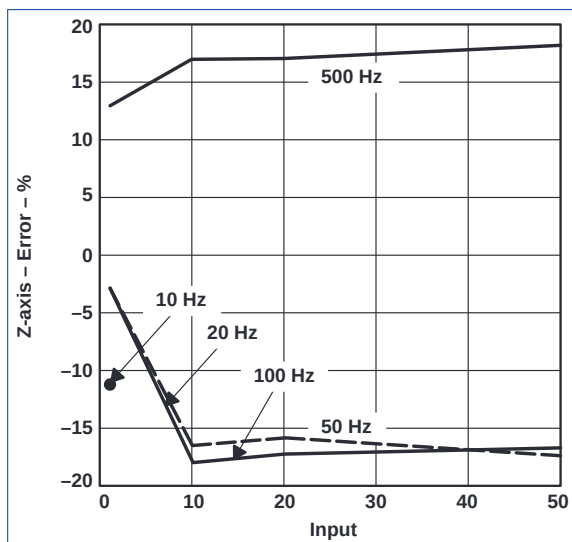


Figure 8–6. Z-Axis % Error

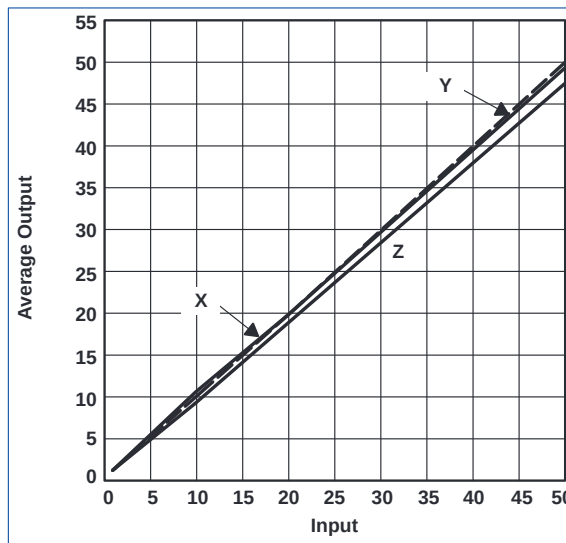


Figure 8–7. Average Output vs. Input Over Frequency for Each Axis

9 References

1. ACH04-08-05 Data Sheet
<http://www.amp.com/sensors>
2. TLV2772 Data Sheet, literature #SLOS209
3. Universal Operational Amplifier EVM User's Manual, literature #SLVU006
4. TLV1544 Data Sheet, literature #SLAS139B
5. TLV1544 EVM User's Manual, literature #SLAU014
6. Interfacing the TLV1544 Analog-to-Digital Converter to the TMS320C50 DSP Applications Report, literature #SLAA025
7. TMS320C5X Evaluation Module Technical Reference, literature #SPRU087
8. TMS320C5X C Source Debugger User's Guide, literature #SPRU055B
9. TMS320C5X EVM Technical Reference, literature #SPRU087

Appendix A. Source Code Listings

```

*****
* TITLE           : ADC Interface routine                                     *
* FILE            : ACCELER1.ASM                                           *
* FUNCTION        : MAIN                                                  *
* PROTOTYPE       : void MAIN ()                                           *
* CALLS           : Wait ()                                               *
*                 Channels ()                                              *
* PRECONDITION    : N/A                                                  *
* POSTCONDITION   : N/A                                                  *
* DESCRIPTION     :                                                         *
* AUTHOR          : AAP Application Group, Dallas                         *
*                 CREATED 1998(C) BY TEXAS INSTRUMENTS INCORPORATED.      *
* REFERENCE       : TMS320C5x User's Guide, TI 1997                      *
*                 : Data Aquisition Circuits, TI 1998                     *
*****

    .mmregs
    .sect ".vectors"
    .copy "vectors.asm"

    .sect ".data"
    .copy "values.asm"
    .def    _MAIN
    .def    ACT_CHANNEL
    .def    ADWORD
    .def    ADCOUNT
    .def    ADMEM
    .def    isr save
    .def    FINITO
    .def    data_loc_point0
    .def    data_loc_point1
    .def    data_loc_point2
    .def    data_loc_point3
    .def    MEMCOUNT
    .def    TST_CH1;
    .def    TST_CH2;
    .def    TST_CH3;
AD_DP      .usect ".variabl", 0      ;
ACT_CHANNEL .usect ".variabl", 1      ; jump address to init. new channel
ADWORD     .usect ".variabl", 1      ; send-bytes to the ADC
ADCOUNT   .usect ".variabl", 1      ; counter for one channel
ADMEM      .usect ".variabl", 1      ; points to act. memory save location

```

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```

isr_save      .usect ".variabl", 1                ; memory location to save AR7 during interrupts
FINITO        .usect ".variabl", 1                ; shows, that the sampling is completed
MEMCOUNT     .usect ".variabl", 1                ; counter for samples per channel
TST_CH1       .usect ".variabl", 1                ;
TST_CH2       .usect ".variabl", 1                ;
TST_CH3       .usect ".variabl", 1                ;
data_loc_point0 .usect ".variabl", 1              ;
data_loc_point1 .usect ".variabl", 1              ;
data_loc_point2 .usect ".variabl", 1              ;
data_loc_point3 .usect ".variabl", 1              ;
               .sect ".text"

_MAIN:
*****

* INITIALIZATION BODY
*****

* DSP INITIALIZATION
    SETC      INTM                ; DISABLE GLOBAL INTERRUPTS
    LDP        #0;
    OPL        #0038h, PMST        ; Configure PMST (allocate IRQ)
    APL        #0000h, PDWSR       ; clear PDWSR (zero wait states)
    APL        #00F0h, CWSR        ; clear CWSR (zero wait states)

* SERIAL PORT INITIALIZATION
SPI_INI:
    APL        #07F38h, SPC        ; clear Res, DLB, FO, XRST, RRST and FREE in SPC
    OPL        #00038h, SPC        ; Set Burst Mode,
                                   ; CLKX=1/4CLKOUT1,
                                   ; FSX generated by DSP
    OPL        #080C0h, SPC        ; activate transmitter and receiver

* INITIALIZE USER INFORMATIONS FOR DATA AQUISITION
    CALL       CHANNELS           ; load informations about ADC channel and
                                   ; memory location for data saving

* Enable Interrupts
    LDP        #0
    OPL        #00014h, IMR        ; Unmask RINT and INT3
    CLRC      INTM                ; enable global interrupts

* FIRST SEND OPERATION MODE TO TLV1544
    LDP        #ADWORD
    LACL      ADWORD
    CLRC      XF                  ; Enable Chip Select
    SAMM      DXR                 ; move ADWORD into DXR (DXR = fast conversation mode)
*****

* initialize MEMCOUNT and data_loc_pointx
*****

```

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```
SPLK    #data_loc_0,data_loc_point0  ;
SPLK    #data_loc_1,data_loc_point1  ;
SPLK    #data_loc_2,data_loc_point2  ;
SPLK    #data_loc_3,data_loc_point3  ;
SPLK    #TRIG_REF,TST_CH1   ;
SPLK    #TRIG_REF,TST_CH2   ;
SPLK    #TRIG_REF,TST_CH3   ;
SPLK    #numb_data_all,MEMCOUNT  ;

*****
* MAIN BODY FOR A PROGRAM, UNRELATED TO AD-CONVERTING                      *
* PROCESSOR SLEEP AT NONACTIVITY TIMES                                    *
* INSTEAD OF THE IDLE INSTRUCTION, USER CAN RUN THEIR OWN PROGRAM        *
*****

USERINTERFACE:
    SPLK    #000000h,FINITO      ; sampling not finished yet                *

*****
* testing the input samples against trigger values to start                *
* the final sampling of the accelerator                                    *
*****

TRIP_LOOP:
    LDP     #TST_CH1             ;
    LACC    TST_CH1              ;
    SUB     #TRIP_HIGH           ;
    BCND    TRIG,GEQ             ;
    LACC    TST_CH1              ;
    SUB     #TRIP_LOW            ;
    BCND    TRIG,LEQ             ;
    LDP     #TST_CH2             ;
    LACC    TST_CH2              ;
    SUB     #TRIP_HIGH           ;
    BCND    TRIG,GEQ             ;
    LACC    TST_CH2              ;
    SUB     #TRIP_LOW            ;
    BCND    TRIG,LEQ             ;
    LDP     #TST_CH3             ;
    LACC    TST_CH3              ;
    SUB     #TRIP_HIGH           ;
    BCND    TRIG,GEQ             ;
    LACC    TST_CH3              ;
    SUB     #TRIP_LOW            ;
    BCND    TRIG,LEQ             ;
    B       TRIP_LOOP           ;

*****
```

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```
* start the final sampling of the accelerator
*****
TRIG:
    SPLK    MAIN_CHANNEL0,ACT_CHANNEL    ; set the next INTERRUPT-BRANCH-ADDRESS to the
                                           ; MAIN sampling

M1:
    LDP     #FINITO;
    LACC    FINITO ;
    SUB     #00001h          ;
    BCND    M1, LT
*****
* sampling is done, wait until user interrupt
*****
M2:
    NOP                                ; no operation
    B       M2;
*****
* END_CONV_IRQ:
*   Interrupthandler for the external INT3-IRQ, which is the end of conversation signal
*   of the ADC. The Routine will enable the Chip Signal again and send the next sample
*   instruction, contained in ADWORD, to the ADC.
*****
END_CONV_IRQ:
* MAIN IRQ ROUTINE
    LDP     #ADWORD
    LACL    ADWORD
    CLRC    XF                      ; Enable Chip Select
    SAMP    DXR                     ; move ADWORD into DXR
    RETE                                ; Return to Wait loop
*****
* RECEIVE_IRQ:
*   Interrupthandler for serial receive of the SPI interface. The Routine will disable the
*   Chip Signal and store the received sample into the by ADMEM specified memory location.
*   It also checks, how many samples were made of the actual channel and if all done, it
*   will call the procedure CHANNELS to load the next channel specifications.
*****
RECEIVE_IRQ:
* save AR7 before handling the IRQ routine
    ldp     #isr_save
    sar     ar7,isr_save            ; save AR7 in isr_save

* MAIN IRQ ROUTINE (Data collection from ADC)
    CLRC    SXM                      ; Clear sign bit
    LDP     #0;
```

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```

LACC   DRR,10                ; load content of DRR in ACCh; shift 10
LDP    #ADMEM;
MAR    *,AR7;
LAR    AR7, ADMEM            ; AR7 = ADMEM
SACH   *+                    ; save sample in the memory
SAR    AR7, ADMEM            ; ADMEM++
LAR    AR7, ADCOUNT;         ; AR7 = ADCOUNT
LACL   ADCOUNT;
SUB    #1                    ; increment ACC
SACL   ADCOUNT               ; ADCOUNT=ADCOUNT-1
BCND   KEEP_ON,NEQ           ; if ADCOUNT not 0 jump to KEEP_ON
LACC   ACT_CHANNEL           ; load addr. to set new sampling channel
CALA                   ; jump into the subroutine
KEEP_ON:
SETC   XF                    ; Disable Chip Select
* restore AR7 before jump back into the user software
ldp    #isr_save
lar    ar7,isr_save          ; load old AR0
RETE                   ; Return from Interrupt

.copy "channels.asm"
.end

*****
* TITLE       : TLV1544C ADC Interface routine                      *
* FILE        : Channels.ASM                                         *
* FUNCTION     : CHANNELS                                             *
* PROTOTYPE    : void CHANNELS ()                                    *
* CALLS        : N/A                                                 *
* PRECONDITION : N/A                                                 *
* POSTCONDITION : N/A                                               *
* DESCRIPTION   :                                                    *
* AUTHOR       : AAP Application Group, Dallas                      *
*               : CREATED 1998(C) BY TEXAS INSTRUMENTS INCORPORATED. *
* REFERENCE    : TMS320C5x User's Guide, TI 1997                    *
*               : Data Aquisition Circuits, TI 1998                  *
*****

.text
CHANNELS:
WAKE_UP:
LDP    #AD_DP                ; point to AD Data Page
SPLK   #fast_conv,ADWORD     ; The first Word to send is fast_conv to wake up the ADC
                                ; from power down
SPLK   #tresh,ADMEM          ; DATA LOCATION = THRES, because first received byte
                                ; is meaningless

```

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```

    SPLK    #1,ADCOUNT                ; NUMBER OF DATA 0
    SPLK    #one_more,ACT_CHANNEL      ; set new jump level
    RET

one_more:
    LDP     #AD_DP                      ; point to AD Data Page
    SPLK    #channel_1,ADWORD          ; The first Word to send is fast_conv to
                                        ; wake up the ADC from power down
    SPLK    #tresh,ADMEM               ; DATA LOCATION = THRES, because first
                                        ; received byte is meaningless

    SPLK    #4,ADCOUNT                ; NUMBER OF DATA 0
    SPLK    #CHANNEL1,ACT_CHANNEL      ; set new jump level
    RET

CHANNEL1:
    LDP     #AD_DP                      ; point to AD Data Page
    SPLK    #channel_1,ADWORD          ; SELECT CHANNEL 0
    SPLK    #TST_CH3,ADMEM             ; DATA LOCATION 0
    SPLK    #00001h,ADCOUNT          ; NUMBER OF DATA 0
    SPLK    #CHANNEL2,ACT_CHANNEL      ; set new jump level\
    RET

CHANNEL2:
    LDP     #AD_DP                      ; point to AD Data Page
    SPLK    #channel_2,ADWORD          ; SELECT CHANNEL 0
    SPLK    #TST_CH1,ADMEM             ; DATA LOCATION 0
    SPLK    #00001h,ADCOUNT          ; NUMBER OF DATA 0
    SPLK    #CHANNEL3,ACT_CHANNEL      ; set new jump level
    RET

CHANNEL3:
    LDP     #AD_DP                      ; point to AD Data Page
    SPLK    #channel_3,ADWORD          ; SELECT CHANNEL 0
    SPLK    #TST_CH2,ADMEM             ; DATA LOCATION 0
    SPLK    #00001h,ADCOUNT          ; NUMBER OF DATA 0
    SPLK    #CHANNEL1,ACT_CHANNEL      ; set new jump level
    RET

MAIN_CHANNEL0:
    LDP     #AD_DP                      ; point to AD Data Page
    SPLK    #channel_0,ADWORD          ; SELECT CHANNEL
    LACL    data_loc_point3            ;
    SACL    ADMEM                      ; load actual memory pointer
    ADD     #1                          ; increment ACC
    SACL    data_loc_point3            ; save new memory pointer
    SPLK    #00001h,ADCOUNT          ; NUMBER OF DATA
    SPLK    #MAIN_CHANNEL1,ACT_CHANNEL ; set new jump level
    RET

```

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MAIN_CHANNEL1:

```
LDP    #AD_DP                ; point to AD Data Page
SPLK   #channel_1,ADWORD     ; SELECT CHANNEL
LACL   data_loc_point0      ;
SACL   ADMEM                 ; load actual memory pointer
ADD    #1                    ; increment ACC
SACL   data_loc_point0      ; save new memory pointer
SPLK   #00001h,ADCOUNT       ; NUMBER OF DATA
SPLK   #MAIN_CHANNEL2,ACT_CHANNEL ; set new jump level\
RET
```

MAIN_CHANNEL2:

```
LDP    #AD_DP                ; point to AD Data Page
SPLK   #channel_2,ADWORD     ; SELECT CHANNEL
LACL   data_loc_point1      ;
SACL   ADMEM                 ; load actual memory pointer
ADD    #1                    ; increment ACC
SACL   data_loc_point1      ; save new memory pointer
SPLK   #00001h,ADCOUNT       ; NUMBER OF DATA
SPLK   #MAIN_CHANNEL3,ACT_CHANNEL ; set new jump level
RET
```

MAIN_CHANNEL3:

```
LDP    #AD_DP                ; point to AD Data Page
SPLK   #channel_3,ADWORD     ; SELECT CHANNEL
LACL   data_loc_point2      ;
SACL   ADMEM                 ; load actual memory pointer
ADD    #1                    ; increment ACC
SACL   data_loc_point2      ; save new memory pointer
SPLK   #00001h,ADCOUNT       ; NUMBER OF DATA
LACL   MEMCOUNT            ;
SUB    #1                    ; increment ACC
SACL   MEMCOUNT            ; ADCOUNT=ADCOUNT-1
BCND   FINITISIMO,EQ         ; if ADCOUNT not 0 jump to KEEP_ON
SPLK   #MAIN_CHANNEL0,ACT_CHANNEL ; set new jump level
RET
```

FINITISIMO:

```
SPLK   #00001h,FINITO       ; set finish sign
SPLK   #CHANNEL_END,ACT_CHANNEL ; set new jump level
RET
```

* TLV1544 POWER DOWN AFTER TRANSITIONS

CHANNEL_END:

```
LDP    #AD_DP                ; point to AD Data Page
SPLK   #power_down,ADWORD    ; SELECT CHANNEL 0
SPLK   #0,ADCOUNT            ; NUMBER OF DATA 0
```

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```

        SPLK      #CHANNEL_END,ACT_CHANNEL ; set new jump level
        RET
*****
* TITLE          : TLV1544C ADC Interface routine                                     *
* FILE           : values.ASM                                                         *
* FUNCTION       : N/A                                                                *
* PROTOTYPE      : N/A                                                                *
* CALLS          : N/A                                                                *
* PRECONDITION   : N/A                                                                *
* POSTCONDITION  : N/A                                                                *
* DESCRIPTION    : contains the control data for the TLV 1544, the starting address to  *
*                  save the samples to the DSP memory and the number of samples       *
*                  according each channel                                             *
* AUTHOR         : AAP Application Group, Dallas                                     *
*                  CREATED 1998(C) BY TEXAS INSTRUMENTS INCORPORATED.                *
* REFERENCE      : TMS320C5x User's Guide, TI 1997                                  *
*                  : Data Aquisition Circuits, TI 1998                              *
*****
channel_0        .set      0000h                ; Select TLV1544 channel 0
channel_1        .set      2000h                ; Select TLV1544 channel 1
channel_2        .set      4000h                ; Select TLV1544 channel 2
channel_3        .set      6000h                ; Select TLV1544 channel 3
power_down       .set      8000h                ; Software Power Down
fast_conv        .set      9000h                ; Fast Conversion Rate
slow_conv        .set      0A00h                ; Slow Conversion Rate
test_200         .set      0B00h                ; (Vreg+ - Vreg-)/2
test_000         .set      0C00h                ; Vreg-
tets_3FF         .set      0D000h               ; Vreg+
num_data_0       .set      1FFh                 ; Number of Data from channel 0
num_data_1       .set      200h                 ; Number of Data from channel 1
num_data_2       .set      200h                 ; Number of Data from channel 2
num_data_3       .set      200h                 ; Number of Data from channel 3
tresh            .set      1000h                ; address to waste the first input sample
                ; after initialization
data_loc_0       .set      01000h               ; Start data location for channel 0
data_loc_1       .set      04000h               ; Start data location for channel 1
data_loc_2       .set      07000h               ; Start data location for channel 2
data_loc_3       .set      0A000h               ; Start data location for channel 3
numb_data_all    .set      03000h               ; how many samples per channel to do
TRIP_HIGH        .set      023bh                ; high trigger value = 1.34 V
TRIP_LOW         .set      01e7h                ; low trigger value = 1.14 V
TRIG_REF         .set      0211h                ; mean value of no triggering = 1.24 V
*****
* TITLE          : TLV1544C ADC Interface routine                                     *

```

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```

* FILE           : vectors.ASM
* FUNCTION        : N/A
* PROTOTYPE       : N/A
* CALLS           : N/A
* PRECONDITION    : N/A
* POSTCONDITION   : N/A
* DESCRIPTION     : defines the interrupt vector table.
*                  if RINT occurs: vector points to RECEIVE_IRQ subroutine
*                  if INT3 occurs: vector points to END_CONV_IRQ subrout.
* AUTHOR          : AAP Application Group, Dallas
*                  CREATED 1998(C) BY TEXAS INSTRUMENTS INCORPORATED.
* REFERENCE       : TMS320C5x User's Guide, TI 1997
*****
        .mmregs

RS      b    _MAIN          ;0x00; RESET
INT1    b    INT1           ;0x02; external user interrupt #1
INT2    b    INT2           ;0x04; external user interrupt #2
INT3    b    END_CONV_IRQ   ;0x06; external user interrupt #3
TINT    b    TINT           ;0x08; internal timer interrupt
RINT    b    RECEIVE_IRQ    ;0x0A; Serial Port receive interrupt
XINT    b    XINT           ;0x0C; Serial Port transmit interrupt
TRNT    b    TRNT           ;0x0E; TDM receive interrupt
TXNT    b    TXNT           ;0x10; TDM transmit interrupt
INT4    b    INT4           ;0x12; external user interrupt #4
        .space 14*16        ;0x14-0x21; reserved area
TRAP    b    TRAP           ;0x22; trap instruction vector
NMI     b    NMI            ;0x24; non-maskable interrupt
/*****/
/* TITLE          : Hexadecimal to ASCII - Decimal Converter
/* FILE           : B_H_C502.PAS
/* FUNCTION        : N/A
/* PROTOTYPE       : N/A
/* CALLS           : N/A
/* PRECONDITION    : N/A
/* POSTCONDITION   : N/A
/* DESCRIPTION     : strip header from COFF file and convert
/*                  hexadecimal numbers into ASCII-decimal
/* AUTHOR          : AAP Application Group, Dallas
/*                  CREATED 1998(C) BY TEXAS INSTRUMENTS INCORPORATED.
/* REFERENCE       :
/* NOTE            : Written in PASCAL
/*****/
uses crt,dos,printer;

```

3-V Accelerometer Featuring TLV2772

```
type
  positionstyp=record
    AAAA :word;
  END;
var
  pos          :positionstyp;
  satz1  :file of positionstyp;
  infile,ffile :string[20];
  na,c        :char;
  s2          :string[2];
  i           :integer;
  A,n         :char;
  AA          :byte;
  tin         :file of byte;
  tout        :text;
  offset      :byte;
  nb          : byte;
  number      :longint;
  number_low  :longint;
  number_high :longint;
begin
repeat
  clrscr;
  writeln;
  writeln(' TEXAS INSTRUMENTS (C) SOFTWARE');
  writeln(' 1998 ');
  writeln;
  write(' hard drive for input binary hexadecimal file(16bit) : '); na:=readkey;
  writeln(na,':');
  write(' file name: ');
  readln(ffile);ffile:=na+':'+ffile;
  assign(tin,ffile);{$i-} reset(tin);{$i+}
  i:=IoResult;
  if (i<>0) then
    begin
      repeat
        writeln(' file not found, new Input Name? [Y/N]');
        n:=uppercase(readkey);
        until ((n='Y') or (n='N'));
      end;
      if i=0 then n:='w';
    until ((n='N') or (n='w'));
    if n<>'N' then
      begin
```

3-V Accelerometer Featuring TLV2772

```
inpfile:=ffile;
repeat
  writeln;
  write(' hard drive for decimal output file : '); na:=readkey; writeln(na,':');
  write(' file name: '); readln(ffile);ffile:=na+':'+ffile;
  assign(tout,ffile);{$i-} reset(tout);{$i+}
  i:=IoResult;
  if (i=0) then
    begin
      repeat
        writeln(' file already found, should I overwrite it? [Y/N]');
        n:=upcase(readkey);
        until ((n='Y') or (n='N'));
      end
    else n:='Y';
  until n='Y';
  rewrite(tout);writeln;
  writeln(tout,'binary Hex_dec-conversion, made by TI: ');
  writeln(tout,'Input_file: ',inpfile);
  read(tin,nb);
  reset(tin);
  for i:=1 to 70 do read(tin,nb);
  repeat
    read(tin,nb);
    number_low:=nb;
    read(tin,nb);
    number_high:=nb;
    number:=number_high*256+number_low;
    writeln(number);
    writeln(tout,number,',');
  until eof(tin);
  close(tin);
  close(tout);
  readln;
end;
end.
{A=97 0=48}
```

EXHIBIT 1063C



Piezo Film Sensors

Technical Manual

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INTRODUCTION

Transducer materials convert one form of energy into another, and are widely used in sensing applications. The tremendous growth in the use of microprocessors has propelled the demand for sensors in diverse applications. Today, **PIEZOELECTRIC POLYMER SENSORS** are among the fastest growing of the technologies within the \$18 billion worldwide sensor market. Like any new technology, there have been an extraordinary number of applications where **"PIEZO FILM"** has been considered for the sensor solution. In the 20 years since the discovery of piezoelectric polymer, the technology has matured, practical applications have emerged from a long list of possibilities, and the rate of commercialization of the technology is accelerating.

These documents provide an overview of piezoelectric polymer technology and nomenclature, its properties, and sensor design considerations. It also explores a range of sensor applications that have been successfully developed in recent years.

Solving unique sensor problems is a particular strength of our group of applications engineers. We welcome the opportunity to provide assistance to you during your evaluation of piezo film sensors for your design.

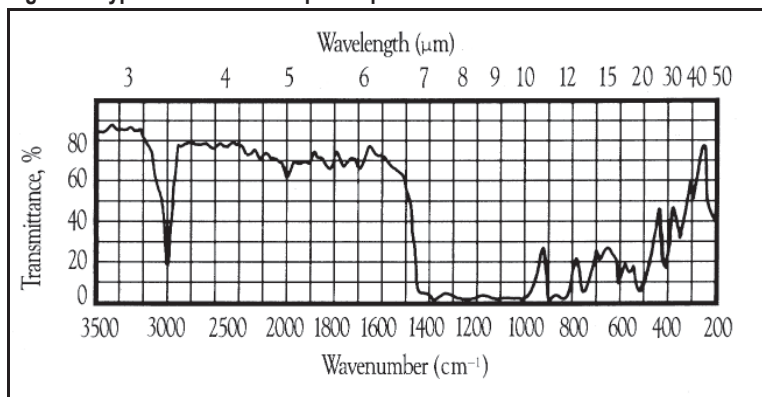
BACKGROUND

Piezoelectricity, Greek for "pressure" electricity, was discovered by the Curie brothers more than 100 years ago. They found that quartz changed its dimensions when subjected to an electrical field, and conversely, generated electrical charge when mechanically deformed. One of the first practical applications of the technology was made in the 1920's by another Frenchman, Langevin, who developed a quartz transmitter and receiver for underwater sound - the first SONAR. Before World War II, researchers discovered that certain ceramic materials could be made piezoelectric when subjected to a high polarizing voltage, a process analogous to magnetizing a ferrous material.

By the 1960's, researchers had discovered a weak piezoelectric effect in whale bone and tendon. This began an intense search for other organic materials that might exhibit piezoelectricity. In 1969, Kawai found very high piezo-activity in the polarized fluoropolymer, polyvinylidene fluoride (PVDF). While other materials, like nylon and PVC exhibit the effect, none are as highly piezoelectric as PVDF and its copolymers.

Like some other ferroelectric materials, PVDF is also pyroelectric, producing electrical charge in response to a change in temperature. PVDF strongly absorbs infrared energy in the 7-20 μ m wavelengths (see Figure 1), covering the same wavelength spectrum as heat from the human body. Accordingly, PVDF makes a useful human motion sensor as well as pyroelectric sensor for more sophisticated applications like vidicon cameras for night vision and laser beam profiling sensors. A dense infrared array has been recently introduced that identifies one's fingerprint pattern using the pyro effect of piezo polymer.

Figure 1. Typical infrared absorption spectrum of PVDF film.



New copolymers of PVDF, developed over the last few years, have expanded the applications of piezoelectric polymer sensors. These copolymers permit use at higher temperatures (135°C) and offer desirable new sensor shapes, like cylinders and hemispheres. Thickness extremes are possible with copolymer that cannot be readily attained with PVDF. These include ultrathin (200 Å) spin-cast coatings that enable new sensor-on-silicon applications, and cylinders with wall thicknesses in excess of 1200µm for sonar. Piezo cable is also produced using copolymer.

PIEZOELECTRIC FILM PROPERTIES

Piezo film is a flexible, lightweight, tough engineering plastic available in a wide variety of thicknesses and large areas. Its properties as a transducer include:

- Wide frequency range—0.001 Hz to 10^9 Hz.
- Vast dynamic range (10^{-8} to 10^6 psi or µ torr to Mbar).
- Low acoustic impedance—close match to water, human tissue and adhesive systems.
- High elastic compliance
- High voltage output—10 times higher than piezo ceramics for the same force input.
- High dielectric strength—withstanding strong fields (75V/µm) where most piezo ceramics depolarize.
- High mechanical strength and impact resistance (10^9 — 10^{10} Pascal modulus).
- High stability—resisting moisture (<0.02% moisture absorption), most chemicals, oxidants, and intense ultraviolet and nuclear radiation.
- Can be fabricated into unusual designs.
- Can be glued with commercial adhesives.

One major advantage of piezo film over piezo ceramic is its low acoustic impedance which is closer to that of water, human tissue and other organic materials. For example, the acoustic impedance ($Z_o = \rho v$) of piezo film is only 2.6 times that of water, whereas piezo ceramics are typically 11 times greater. A close impedance match permits more efficient transduction of acoustic signals in water and tissue.

Piezo film does have some limitations for certain applications. It makes a relatively weak electromechanical transmitter when compared to ceramics, particularly at resonance and in low frequency applications. The copolymer film has maximum operating/storage temperatures as high as 135°C, while PVDF is not recommended for use or storage above 100 °C. Also, if the electrodes on the film are exposed, the sensor can be sensitive to electromagnetic radiation. Good shielding techniques are available for high EMI/RFI environments.

Table 1 lists typical properties of piezo film. Table 2 provides a comparison of the piezoelectric properties of PVDF polymer and two popular piezoelectric ceramic materials.

Piezo film has low density and excellent sensitivity, and is mechanically tough. The compliance of piezo film is 10 times greater than the compliance of ceramics. When extruded into thin film, piezoelectric polymers can be directly attached to a structure without disturbing its mechanical motion. Piezo film is well suited to strain sensing applications requiring very wide bandwidth and high sensitivity. As an actuator, the polymer's low acoustic impedance permits the efficient transfer of a broadband of energy into air and other gases.

Table 1. Typical properties of piezo film

Symbol	Parameter		PVDF	Copolymer	Units
t	Thickness		9, 28, 52, 110	<1 to 1200	μm (micron, 10 ⁻⁶)
d ₃₁	Piezo Strain Constant		23	11	10 ⁻¹² $\frac{\text{m/m}}{\text{V/m}}$ <i>or</i> $\frac{\text{C/m}^2}{\text{N/m}^2}$
d ₃₃			-33	-38	
g ₃₁	Piezo Stress constant		216	162	10 ⁻³ $\frac{\text{V/m}}{\text{N/m}^2}$ <i>or</i> $\frac{\text{m/m}}{\text{C/m}^2}$
g ₃₃			-330	-542	
k ₃₁	Electromechanical Coupling Factor		12%	20%	
k _t			14%	25-29%	
C	Capacitance		380 for 28μm	68 for 100μm	pF/cm ² @ 1KHz
Y	Young’s Modulus		2-4	3-5	10 ⁹ N/m ²
V ₀	Speed of Sound	stretch: thickness:	1.5	2.3	10 ³ m/s
			2.2	2.4	
p	Pyroelectric Coefficient		30	40	10 ⁻⁶ C/m ² °K
ε	Permittivity		106-113	65-75	10 ⁻¹² F/m
ε/ε ₀	Relative Permittivity		12-13	7-8	
ρ _m	Mass Density		1.78	1.82	10 ³ kg/m
ρ _e	Volume Resistivity		>10 ¹³	>10 ¹⁴	Ohm meters
R _□	Surface Metallization Resistivity		<3.0	<3.0	Ohms/square for NiAl
R _□			0.1	0.1	Ohms/square for Ag Ink
tan δ _e	Loss Tangent		0.02	0.015	@ 1KHz
	Yield Strength		45-55	20-30	10 ⁶ N/m ² (stretch axis)
	Temperature Range		-40 to 80...100	-40 to 115...145	°C
	Water Absorption		<0.02	<0.02	% H ₂ O
	Maximum Operating Voltage		750 (30)	750 (30)	V/mil(V/μm), DC, @ 25°C
	Breakdown Voltage		2000 (80)	2000 (80)	V/mil(V/μm), DC, @ 25°C

Table 2. Comparison of piezoelectric materials

Property	Units	PVDF Film	PZT	BaTiO ₃
Density	10 ³ kg/m ³	1.78	7.5	5.7
Relative Permittivity	ϵ/ϵ_0	12	1,200	1,700
d_{31} Constant	(10 ⁻¹²)C/N	23	110	78
g_{31} Constant	(10 ⁻³)Vm/N	216	10	5
k_{31} Constant	% at 1 KHz	12	30	21
Acoustic Impedance	(10 ⁶)kg/m ² -sec.	2.7	30	30

OPERATING PROPERTIES FOR A TYPICAL PIEZO FILM ELEMENT

The DT1 element is a standard piezo film configuration consisting of a 12x30 mm active area printed with silver ink electrodes on both surfaces of a 15x40 mm die-cut piezo polymer substrate.

1. Electro-Mechanical Conversion

(1 direction) 23 x 10⁻¹²m/V, 700 x 10⁻⁶N/V
(3 direction) -33 x 10⁻¹²m/V

2. Mechano-Electrical Conversion

(1 direction) 12 x 10⁻³V per microstrain, 400 x 10⁻³V/μm, 14.4V/N
(3 direction) 13 x 10⁻³V/N

3. Pyro-Electrical Conversion

8V/°K (@ 25°C)

4. Capacitance

1.36 x 10⁻⁹F; Dissipation Factor of 0.018 @ 10 KHz; Impedance of 12 KΩ @ 10 KHz

5. Maximum Operating Voltage

DC: 280 V (yields 7 μm displacement in 1 direction)
AC: 840 V (yields 21 μm displacement in 1 direction)

6. Maximum Applied Force (at break, 1 direction)

6-9 kgF (yields voltage output of 830 to 1275 V)

Electrical to Mechanical Conversion

Large displacements of forces are not generally available from piezo film. This becomes apparent when designing loudspeaker elements for instance, as low frequency performance (below 500Hz) tends to be limited. Even a large sheet of film is unable to create high amplitude pressure pulses as low audio frequencies. This does not apply, however, to low to high frequency ultrasonic frequencies, as seen in current designs for ultrasound air ranging transducers (40-50 KHz) and in medical ultrasonic imaging applications. In enclosed air cavities (headset speakers, hearing aids), the low frequency response of piezo film is excellent. For air ranging ultrasound, the piezo film element height controls vertical beam angle and the curvature and width of the transducer controls horizontal beam pattern. Piezo film air ranging transducers can provide up to 360° field of view, ranging object from a few centimeters to several meters with high resolution.

Bimorph configurations (like a bimetal strip) allow the small differential displacement of two reverse-connected elements to be translated into substantial flexural motion. Small fans or optical deflectors can thus be created. Such devices consume very little real power (being capacitive in nature). Large devices may be difficult to drive due to high capacitance, especially when transformers are used to step up the drive voltage. Good amplifier design is important. Nevertheless, conventional fan and blower technologies generate higher flow rates and back pressures than piezo bimorphs.

Although the forces involved are small, the film can be used to excite other mechanical structures over a very wide frequency range. If a second element of film is used to receive the induced vibration, the system can possess a very high dynamic range, even though the overall "insertion loss" due to the film is about -66 dB typically for a structure at resonance. If sufficient gain is applied between these elements, the structure will self-oscillate at its natural frequency. For these resonant mechanical systems, high voltage drive is not required. The amplifier circuit may function adequately from a normal dual rail op-amp supply, or even from a single 9 volt battery. For analysis purposes, even lower applied voltages, e.g., the noise source of a spectrum analyzer at 70 mVrms, are sufficient to insert the mechanical energy into a structure when piezo film is also used to monitor the result.

Mechanical to Electrical Conversion

The sensitivity of piezo film as a receiver of mechanical work input is awesome. In its simplest mode the film behaves like a dynamic strain gage except that it requires no external power source and generates signals greater than those from conventional foil strain gages *after* amplification. Frequency response is thus free from any limitations imposed by the need for high gains and will extend up to the wavelength limit of the given transducer.

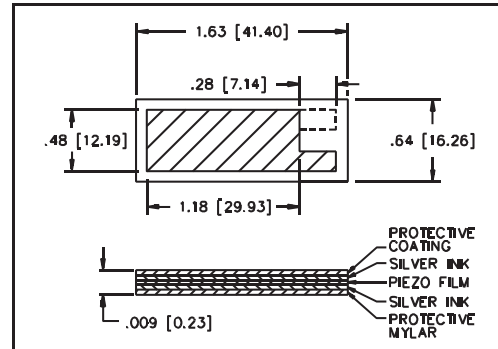
The extreme sensitivity is largely due to the format of the piezo film material. The low thickness of the film makes, in turn, a very small cross-sectional area and thus relatively small longitudinal forces create very large stresses within the material. It is easy to exploit this aspect to enhance the sensitivity parallel to the machine axis. If a laminated element of film (for example an LDT1-028K) is placed between two layers of compliant material then any compressive forces are converted into much larger longitudinal extensive forces. In fact, this effect tends to predominate in most circumstances since most substances are compliant to some extent and the ratio of effective sensitivity in the 1 (length) vs 3 (thickness) directions is typically 1000:1.

Piezo film transducers may often cover a much larger area than normal strain gages so any direct comparisons should be performed in a *uniform* strain field for meaningful results. Obviously "point"-type transducers could be used where required although the capacitance of a very small area will require consideration. The low frequency limit of operation will be defined by the greatest resistive load achievable, or by the largest capacitance load that still allows the signal to be easily detected. Operation down to fractions of Hz can be achieved using either conventional charge amplifiers or, since signal levels are relatively high, simple high impedance FET buffer circuits.

Pyro to Electrical Conversion

Piezo film absorbs strongly in the region of 7 to 20 μm which corresponds to well beyond both operating temperature limits of the film. It thus makes a sensitive pyroelectric detectors for, say, human body radiation. Since the pyro sensitivity is strong, care must be taken when designing low (<0.01 to 1Hz) frequency mechanical sensors to avoid ambient temperature changes swamping

Figure 3. DT1 Element in [mm]



the output with pyro-generated signal. If a very long time constant is in use, then the film will generate a voltage corresponding to the change in temperature since switch-on. Since the output will be several volts per degree C, substantial offsets may be noticed.

In general, however, most piezo applications will have a cut-off frequency of several Hertz or more. Connecting a device of 1nF capacitance to an oscilloscope input, even at 10 MΩ impedance, will produce a roll off below 16 Hz. Only a more rapid change in the film temperature will generate a detectable signal.

Common-mode rejection can be used to isolate either very low frequency mechanical strain from simultaneous pyro-effects or vice-versa. These straight-forward techniques are quite familiar to MSI applications engineers who are available for design assistance.

Electrical Design Considerations

A useful model for piezo film which applies for most cases except ultrasonic applications is a strain-dependent voltage source in series with a capacitance. Thus any resistive load will form a divider network with a simple RC high-pass filter characteristic. The cut-off frequency is given by

$f_c = \frac{1}{2\pi RC}$ and the time constant $\tau = RC$. Operation below the cut-off frequency will give an output signal proportional to the rate of change of the input parameter (differentiator). Application of a constant stress will generate an initial level followed by an exponential decay of rate $\exp(RC)^{-1}$.

A capacitive load will extend the time constant but reduce the magnitude of the response. Energy is always lost when transferring charge from one capacitor to another. Large capacitive loads are useful for attenuating the very large signals arising from powerful impacts—often hundreds of volts.

When driving the film at high voltage and high frequency, the dissipation factor of the film may result in substantial energy loss in the form of heat. Also, the surface resistivity of the electrodes may become significant, especially with vacuum metallized film. Very high localized currents may be encountered. Operation within the field limits given in the Technical Manual is strongly recommended since any arcing will normally destroy the device.

Silver ink, screenprinted onto both film surfaces, has been developed to withstand high voltage and high localized currents. The silver ink metallization has been successfully used in tweeters and active vibration damping applications. The DT1 sample is electroded with the silver ink. The unmetallized border mitigates potential for arcing across the film's thickness. The offset lead attach tabs also preclude high voltage breakdown, as the conductor at each lead attach site is on one side only.

Mechanical Design Considerations

The output energy is proportional to the volume of film stressed. Film thickness may be chosen to optimize the electrical signal or in view of mechanical strength considerations. Thicker films generate higher voltages but form smaller capacitors, so a laminate of thinner film with a compatible, passive material such as polyester (i.e. the LDT1-028K) may be preferable to a single thicker film. Any area of film that is not undergoing stress will act as a capacitive load on the "active" area and should be minimized if required.

Most metallizations are subject to corrosion, especially when handled. Thin conformal coatings or laminates are frequently applied to maintain surface quality. Acrylics adhesives, synthetic rubber resins, epoxies and cyano-acrylates are all frequently employed in lamination and assembly.

Some designs may use external metallic or conductive substrates as the electrodes, in which case unmetallized film may be used to good advantage. The external metal surface can be in direct

contact with the unmetallized film to collect the charge, or, capacitive coupling through thin adhesive tapes or epoxy layers can be employed for ac applications. Patterning of the electrodes is especially useful for defining specific active areas on a continuous sheet and also to allow die-cutting of elements with a clear border around the cut area. Displacement (offset) of upper and lower electrode tabs at the connection point is good practice to prevent unpredictable piezo behavior in this area caused by the influence of the wire terminations. This also allows low cost penetrative lead-attach methods to be used (crimps or eyelets).

Joint Electrical and Mechanical Design Considerations

The capacitive nature of piezo film devices implies that they are susceptible to Electro Magnetic Interference (EMI). This becomes increasingly more important as the output signal level drops. EMI can be ignored where the output is high or when the film is being driven in a non-critical environment. A.C. mains interference may become a problem with unshielded devices. Another potential problem exists when one electrode element is being driven and another is receiving the vibration signal. Care must be taken to avoid "crosstalk".

Use of ready-made shielded elements (SDT1-028K) supplied with coaxial cable eliminates these problems, but simple measures may be taken with any device to avoid interference.

Unwanted frequencies may be filtered out electronically. If the sensor is to be mounted on a conductive substrate, then this may form one half of a grounded envelope, with the outer electrode forming the other half. Lightweight shielded cable is readily available and is an alternative to twisted pair wires. Attention should be paid to the point of connection itself as this is also an area of EMI vulnerability.

Durable lead attachment techniques have been fully developed by MSI, and most products are supplied with leads preattached. As indicated, some form of coaxial cable is often employed and must be interfaced to a very thin flexible material. Reinforcement at the lead attach site may be required, which can introduce some acoustic effects into the transducer if the interconnection site is free to vibrate.

Thin copper foil backed with a conductive adhesive can provide excellent but non-permanent connections to the film. An area of 1 cm² will give a contact resistance of a few mΩ s. Crimp-through connectors as used for flexible circuits are routinely used with offset electrode patterns, but thin films require some physical reinforcement for good results. Polyester reinforcement at the lead attach site is a common method to ruggedize the interconnection. The stiffener may lie between the crimp and the electrode with only minor degradation of contact resistance. Typical values are 150-500 mΩ. Miniature rivets, eyelets and even nuts and bolts, with washers, all combine great strength with good contact resistance at typically less than 100 mΩ. These techniques may be used to connect to cables using solder tags, or direct onto printed circuit boards.

Clamping methods, either direct to the conductive traces on the PCB or using conductive rubber, ZEBRA® connectors, lugs and washers have all been used with success. Direct connection using silver-loaded (conductive) epoxy also works well, but requires curing time, often at elevated temperature, for best results.

As indicated earlier, other materials may form the electrodes themselves, such as PCB traces or conductive rubber. Capacitive coupling through thin adhesive layers is practical under some a.c. circumstances, allowing some unusual transducer designs with apparently no lead attachment at all!

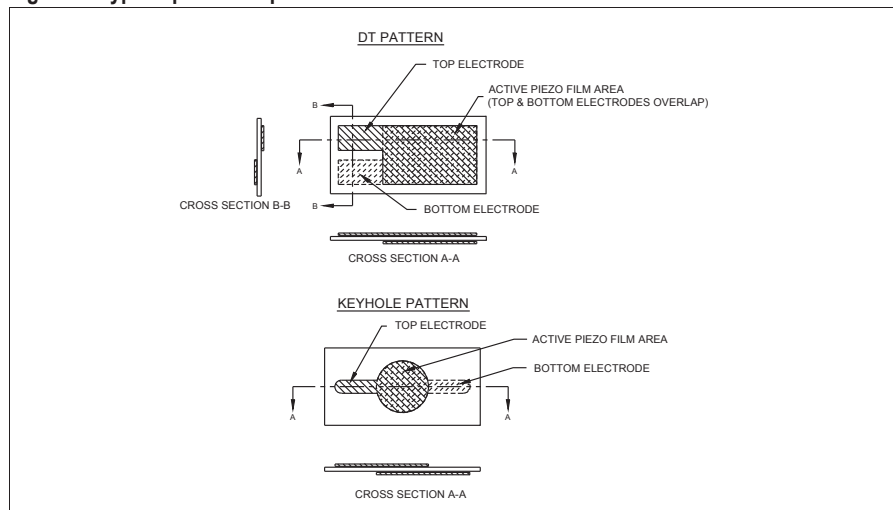
ZEBRA is a registered trademark of Fujipoly.

LEAD ATTACHMENT TECHNIQUES FOR PIEZO FILM SENSORS

Introduction

How to make reliable interconnection to piezo film is one of the most frequently asked customer questions. With this in mind, MSI has paid great attention to the development of techniques to simplify interconnection to piezo film elements. Today, most of the sensor elements supplied to customers from our Division have leads already attached. The aim of this article is to examine and discuss available interconnection options.

Figure 4. Typical piezo film patterns



Some of the most convenient interconnection techniques require that MSI apply patterned electrodes on one or both surfaces of the piezo film—this can always be done to customers' requirements during manufacture— alternatively, a simple method achieving the same goal is presented at the end of the text. In general, patterned electrodes are achieved during piezo film manufacturing by screen printing conductive inks, metal masking during sputtered electrode deposition, or chemically etching patterns by photolithographic techniques.

The Targets

Considered here are the design objectives desired for the lead-attach method. Not all objectives can be achieved with any one technique. Designers should identify the most important objectives and select among the interconnection options accordingly.

- ! High conductivity/low resistance — surprisingly, high conductivity interconnection is not a particularly important parameter for most piezoelectric applications. Piezo transducers are frequently used in high-impedance circuits where inclusion of a few ohms does not usually affect performance. More important, however, is consistence—the resistance should not fluctuate during use since this will introduce a source of electrical noise.
- ! Low mass — this is especially important when the piezo film is not to be clamped to a mechanical support structure. The acoustic effect created by the mechanical vibration of the mass of the interconnection on an otherwise flexible structure can be dramatic.
- ! Low profile — many piezo film applications arise by virtue of the low thicknesses of piezo film. Interrupting this with bulk terminations is often prohibited. Contact vibration sensors can show distinct resonances if film is not bonded flush to the contact surface to include the interconnection.
- ! Flexibility — here again is a property that must often match that of the film itself. Some degree of flexibility is a distinct advantage in many applications.
- ! Low area — useful piezo devices can be quite literally be employed as "point" receivers. Small piezo-active areas (where the top and bottom conductors fully overlap) can be configured with

displaced or off-set lead-attach tabs. The top and bottom tabs are off-set with respect to each other (when viewed through the film thickness). This allows a precisely defined active area (overlapped electrodes) with non-piezo conductors (off-set tabs) leading to remote bonding sites, a technique most frequently employed for "small" devices.

- ! Mechanical Strength — very often the greatest strain experienced by a polymer transducer is around the connection, whether by accident (tripping over the cable) or by design. In general, those methods which involve the interconnection penetrating through the film at the off-set tab locations with crimps, eyelets or rivets yield the best ultimate strain resistance. Often the lead attach area is reinforced with polyester to improve the strength of the penetrative interconnection.
- ! Long-term Stability — including all the usual environmental parameters. Most interconnections have unlimited life (crimps, eyelets, conductive rubber connectors). Others have a more limited shelf life (conductive tapes)
- ! Speed and Ease of application — of particular importance when high volume production is planned. Many interconnection techniques are supported by semi-automatic equipment for volume production (crimps, eyelets) while others are labor intensive (conductive adhesives).
- ! Electrical strength — an issue associated mainly with electrically driven (high voltage) elements such as loudspeakers and actuators.

The Design Considerations

Two major issues control the selection of lead-attach methods:

- ! Is anchorage of the film allowed at the site of lead-attach? This can be a major advantage, for example, direct connection or capacitive coupling to the conductive traces of a printed-circuit board.
- ! Is special patterning of the film available, which would allow penetrative techniques? (with MSI Sensors custom patterning service, the answer is almost always "yes.") Simple experimental methods allow the same result.

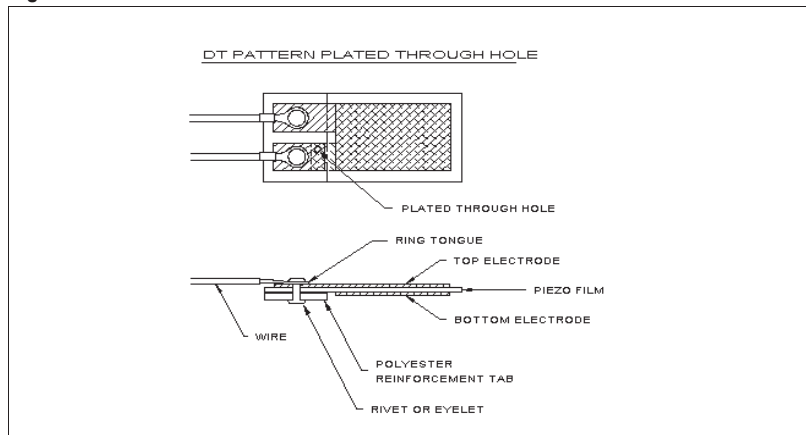
This concludes the "questions" section—now, hopefully, are the "answers."

The Methods

Penetrative - Here the techniques involve piercing the film (and possibly additional reinforcing laminates to give sufficient thickness and strength), and thus the film should be patterned with a displaced or off-set lead-out arrangement to prevent shorting of upper and lower electrodes by the inserted connector.

- ! Rivets or eyelets can be affixed to the off-set conductive traces on the piezo film. Included between the eyelet or rivet can be a ring tongue lug terminal with wire attached. The eyelet or rivet mechanically presses the conductive ring against the off-set patterned electrode to make reliable interconnection.

Figure 4.



! To affix the piezo film directly to a PCB, small "POP" or "blind" rivets or eyelets can be used in conjunction with patterned film electrodes and the conductive tracks on the PCB to allow a single operation to form the interconnection. During screen printing of conductive ink electrodes, a small "plated through hole" can be formed in one of the off-set tabs, thereby bringing both conductors to the same side of the piezo film. This greatly facilitates riveting the film electrode tabs to the corresponding PCB traces. If the "plated through hole" technique is not used, then the top film electrode can be electrically connected by the rivet to a conductive trace on the underside of the PCB. The bottom film electrode is electrically connected to a corresponding trace on the top of the PCB and held in intimate contact by the pressure exerted by the rivet.

! Nuts and bolts - Wires terminated with washers, ring-tongue lugs, solder-tags, etc. can easily be incorporated with small nuts and bolts.

! Crimp Connectors — generally, crimps designed for flexible circuit technology work well with piezo film elements. Crimps can have solder tabs for affixing wires, or the crimp ends can be inserted into corresponding holes in a PCB and soldered to the underside of the PCB (maximum of a few second soldering time so as not to overheat the film).

Like the eyelets mentioned above, crimps are normally designed to work with a specified thickness of "substrate," so film may require "padding" on one side (i.e., polyester reinforcement) to accommodate the crimp connectors. Additionally, a complete multi-way connector may be crimped to a more complex device, giving straight plug-in compatibility with other connectors.

Figure 5.

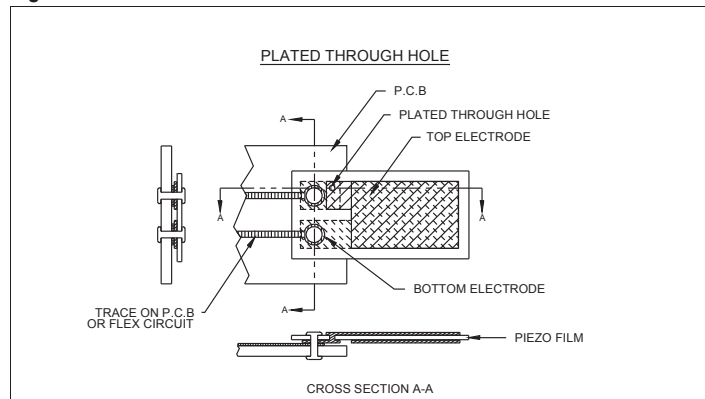


Figure 6.

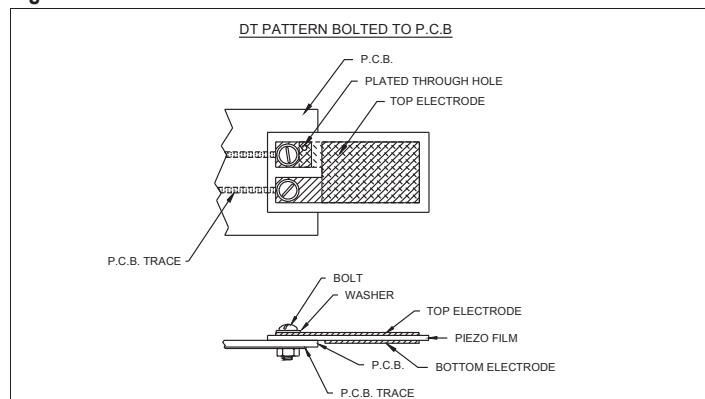
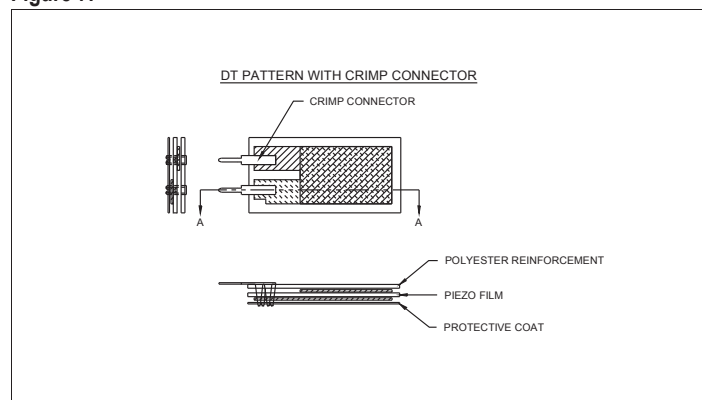


Figure 7.



Non-penetrative (and temporary) - Conductive-adhesive coated Copper Foil Tape (e.g., 3M #1181)—available in widths from 3mm up to 25mm. Best results are obtained by...

! Using a "reasonable" area of tape (perhaps about 1cm or more). Small pieces do tend to lift off easily.

! Soldering wires to the tape FIRST, then removing the liner and adhering with gentle pressure to film. If small areas are to be used, solder before cutting the contact pad down to size, thus leaving the excess area to act as a heat sink. Soldering does appear to degrade the adhesive properties in the vicinity of the joint. NOTE: 3M does not recommend relying on the conductive adhesive in this way and suggest an embossed

version of the same tape. The tape is really designed for large area contacts to metal, but results have shown this method to be an effective, if not guaranteed, technique. An aluminum version of this product is available (Part No. 1170). Beware of similar tapes that do not have conductive adhesives (although these can be used for shielding, etc.)

! Conductive Transfer Tape—e.g., 3M #9702 (Preliminary product). An acrylic adhesive layer loaded with conductive particles giving excellent "Z-axis" conductivity (i.e., through the thickness of the tape) with very high resistivity in the X and Y axes. Thus single or multiple-way connections may be made with a single strip. This material is relatively new. Initial results seem very promising. Obviously this can be used to make direct connection with PC board or strip, or to sections of foil with soldered leads.

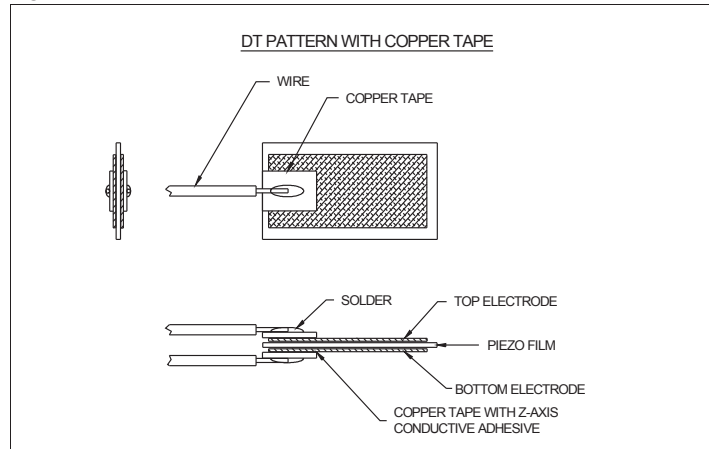
! Negative aspects are a) high cost, and b) like all transfer adhesives, there is a tendency for the material to adhere to its own liners around the edge so that "stringing" occurs on liner removal. NOTE: Since time of writing, this product has been superseded by an improved version (#9703) with an easy-release liner. This may not yet be generally available.

! Conductive Epoxy. This is usually available in two-part form (adhesive and hardener). Precise metering and mixing of the small quantities usually required is rather difficult and messy. One-part, pre-mixed material is available as a product which is stored at very low temperature and should be used and cured at room temperature. Curing of any epoxy mix can usually be accelerated by use of higher temperature, but since the piezo film has a modest high-temp capability, curing is often a long term process (many hours, a day). Some mechanical clamping is usually required on the parts to be bonded. Final reinforcement with "ordinary" epoxy can be reassuring. Negative aspects: difficulty of use, cure time, higher cost, short "shelf life."

! Low melting-point Alloys—some alloys (e.g., Indium/Tin/Bismuth) which are known as "fusible alloys" rather than "solders," melt at temperatures which allow them to be used on piezo film with suitable metallization (e.g., gold, copper, silver or silver ink). Rather aggressive fluxes are often required, and the joint may be brittle. Mechanical strength is limited by the adhesion of the metallization onto the film surface, so once again, reinforcement with epoxy may help. For joints that must be very small and do not need undue mechanical strength this may prove a valuable technique. Negative aspects: only certain metallizations are appropriate, sample quantities hard to come by. Mechanical strength limited [Indium Corp.]

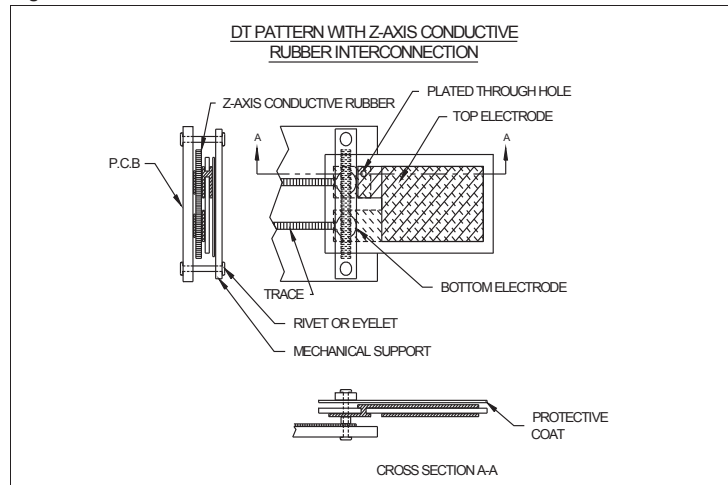
! Zebra® Connectors — Conductive rubber spliced with insulating rubber as used to form contacts to LCD displays. High density multiple-way contacts may be made. External clamping of contacts is required.

Figure 9.



- ! Mechanical clamping—simply sandwiching the film between two conductive surfaces (possibly using a thin layer of conductive-loaded rubber) can provide excellent results. Two rings can provide useful support for diaphragms, speakers, etc.
- ! Capacitive Coupling - In certain applications, no metal electrode is required on the piezo film itself. Thin, non-conductive adhesives can affix the unmetallized film to a conductive surface. The conductive surface in effect provides the film's electrodes in ac applications. A PCB, having conductive pads on one surface corresponding to the desired active sensor area, is an embodiment of this concept. The opposite piezo film surface can be metallized with a ground electrode. The film can be sandwiched between two conductive surfaces with or without adhesive to form electrodes.

Figure 10.



User Etching of Piezo Film Electrodes

Patterned electrodes are available from MSI in either silver screen printed ink or as sputtered electrodes. In some instances, customers purchase fully metallized sheets for experimentation, and want to produce their own patterns. This is very difficult with screen printed inks as they cannot be easily etched or mechanically braided. For sputtered electrodes, standard photolithographic techniques work quite well.

In order to pattern piezo film in such a way as to allow penetration of film without shorting top and bottom electrodes, a very simple technique may be employed which works on any vacuum deposited electrodes (NOTE: not recommended for Ag Ink.)

One terminal of a power source (bench p.s.u. or 9 volt battery) is connected via a conductor pad or block by mechanical pressure to the piece of film in question. The other terminal is brought to a conductive point (needle, wire-end, blunt scalpel, etc.) and the area required to be isolated simply drawn around. Sufficient current normally passes to cause arcing at the point contact and the metallization is vaporized. Concentric "guard rings" may be drawn for extra confidence.

For more complex patterning of thin sputtered metallization, it is possible to coat the piezo film with photoresist aerosol (both sides if necessary). The cured spray can then be exposed through a mask using UV light, as with conventional PCB techniques, and then dipped in an etchant. Complete etching of the very thin metal layer occurs in seconds.

Copper/Nickel metallizations etch very well with standard PCB etchant (ferric chloride). Other metals require special etchants for good results (Aquaregia for gold). Remember that the metallization layer may only be a few hundred atoms thick (300-700 Å), and therefore fine traces are very vulnerable to scratching or cracking.

High Voltage Techniques

The use of piezo film as a vibration exciter requires separate consideration. Since the impedance of a capacitive transducer decreases with frequency and approaches infinity for low frequencies, very high voltages (a few hundred volts typically) may be required to drive, for example, full audio-range loudspeakers. Frequently, transformers are used to step up moderate voltages to supply the required drive signal. Under these circumstances, extreme stresses may be placed upon the connections. Consider first applying a voltage step of 30V to a capacitor of 100nF with an overall circuit resistance of 2 ohms. The initial current pulse peaks at 15 amps (assuming the supply is capable of supplying this). Such a current "spike" may well show up defects in connectors.

Consider next a transformer which steps 12V signals up to 240V. A DC current in the primary of 200 μ A (corresponding to an applied voltage of 0.5 volts), when broken, may cause a voltage surge of 830 volts across the secondary circuit, well in the excess of the expected X 20 magnification factor. Even with heavy capacitive loading, high voltages may be seen. Worse still, if the secondary circuit is broken, current pulses exceeding 60A with durations of only tens of nanoseconds may arise. Such phenomena should not trouble well-formed connections. But if a lead-attach method has been used which has any trapped air, the effect of the reduced dielectric constant may be to promote breakdown. Such events may be catastrophic, as the familiar crackling sound and lively blue sparks will testify.

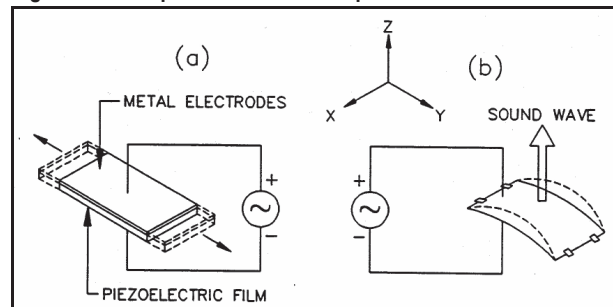
Solutions are:

1. Silver ink electrodes are a must - the thin sputtered electrodes cannot withstand the high voltages
2. Large area contacts to reduce stress. We paint silver ink around eyelets/rivets to provide extra conduction paths to the film electrode.
3. (Possibly) a semi-resistive contact pad to reduce current surges—equivalent to including a series resistance in the circuit. Practical values up to about 1 k will produce only a fractional loss in output and will reduce the magnitude of current spikes.

FREQUENCY RESPONSE

Unlike piezo ceramic transducers, piezo film transducers offer wide dynamic range and are also broadband. These wide band characteristics (near dc to 2GHz) and low Q are partly attributable to the polymers' softness. As audio transmitters, a curved piezo film element, clamped at each end, vibrates in the length (d_{31}) mode, as shown in Figure 10. Piezo film is a very high fidelity tweeter, also used in novelty speakers for toys, inflatables and apparel. The d_{31} configuration (Figure 10) is also used for air ultrasound ranging applications up to frequencies of about 50 KHz.

Figure 10. Clamped film in d_{31} mode produces sound



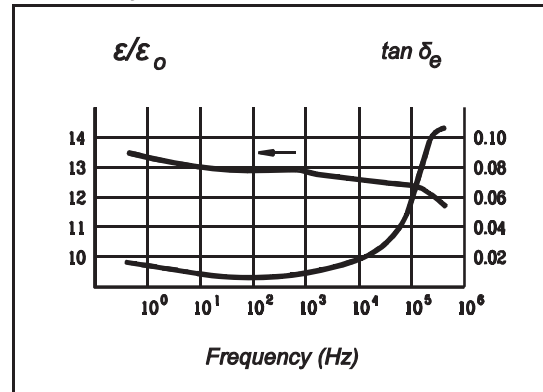
When used as a high ultrasonic transmitter (generally >500KHz), piezo film is normally operated in the thickness (d_{33}) mode. Maximum transmission occurs at thickness resonance. The basic half-wavelength resonance of 28 μ m piezo film is about 40 MHz:

$$f_r = \frac{v}{2t} = \frac{2.2 \times 10^3 \text{ m/sec}}{2 \times 28 \times 10^{-6} \text{ m}}$$

Resonance values thus depend on film thickness. They range from low MHz for thick films ($1,000\mu m$) to $>100\text{MHz}$ for very thin films.

Figure 11 shows the effect that frequency has on permittivity and dissipation factor at room temperature. As a result of its very low permittivity ϵ (1 percent that of piezo ceramics), the film exhibits g -constants (voltage output coefficients) that are significantly greater than piezo ceramics ($g = d/\epsilon$).

Figure 11. Dielectric permittivity and dissipation factor vs. frequency



PIEZO FILM AT LOW FREQUENCIES

Introduction

The behavior of a piezo film component at low frequencies is fairly straightforward to describe in electrical terms, yet is quite frequently misunderstood. Since any practical application of the technology will most likely involve some consideration of this topic, it is the intent of this article to examine the subject at some length. The treatment is made as non-mathematical as possible, with verbal descriptions and real-world examples being used to illustrate the concepts. Some familiarity with the use of FFT techniques to transform between time-domain and frequency-domain descriptions is assumed, but not essential.

Connecting Up

In most instances, the first evaluation of piezo film begins with connecting a piezo component to an oscilloscope via a probe ("scope probe"). Under normal electronics circumstances, a scope probe can be considered to be an "infinite impedance" - so high, that its effect on the circuit under test can be neglected. **Not so with piezo film** - in many cases, a scope probe can act almost like a short-circuit. Typical probes, when plugged in to an oscilloscope, have an effective *resistance* of $1\text{M}\Omega$ (one million ohms). Others may be fixed at $10\text{M}\Omega$, while many are conveniently switchable between "x1" ($1\text{M}\Omega$) and "x10" ($10\text{M}\Omega$). Note that the physical element comprising the $1\text{M}\Omega$ resistance is usually built into the oscilloscope input stage, rather than being a discrete component within the probe itself. A "x1" probe is thus basically a length of shielded cable with suitable contacts attached to each end.

Source Capacitance

To analyze what will happen when the probe is connected, we now need to consider the properties of the piezo film element. Perhaps the most important characteristic (after the piezoelectric property, of course) is the material's *capacitance*. Capacitance is a measure of any component's ability to store electrical charge, and is always present when two conductive plates are brought close together. In our case, the conductive plates are the conductive electrodes printed or metallized onto each surface of the film. The capacitance of the device is strongly affected by the properties of the insulator serving to space the plates apart, and the measure of the insulator's capacity to store charge is given by its *dielectric constant* or *permittivity*.

PVDF has a high dielectric constant compared with most polymers, with its value being about 12 (relative to the permittivity of free space).

Obviously, the capacitance of an element will increase as its plate area increases, so a large sheet of film will have a larger capacitance than a small element. Capacitance also increases as the film thickness *decreases*, so for the same surface geometry, a thin film will have a higher capacitance than a thick film.

These factors are formally related in the equation:

$$C = \epsilon \frac{A}{t}$$

where C is the capacitance of the film,
 ϵ is the permittivity (which can also be expressed in the form

$\epsilon = \epsilon_r \epsilon_0$ where ϵ_r is the relative permittivity (about 12 for PVDF), and ϵ_0 is the permittivity of free space (a constant, 8.854×10^{-12} F/m)

A is the active (overlap) area of the film's electrodes
and t is the film thickness

The units of capacitance are Farads (F), but usually much smaller sub-multiples are encountered: microfarads (μF or 10^{-6} F), nanofarads (nF or 10^{-9} F) and picofarads (pF or 10^{-12} F).

The capacitance of any piezo film element can be calculated using the formula, or measured directly using a hand-held capacitance meter, or bench-top instrument such as an "LCR bridge".

Capacitance values should be quoted at a given measurement frequency - where this is not given, a frequency of 1 KHz is often assumed. Capacitance values of piezo film components usually decrease as the measurement frequency increases.

Table 3. Capacitance values of common piezo film components

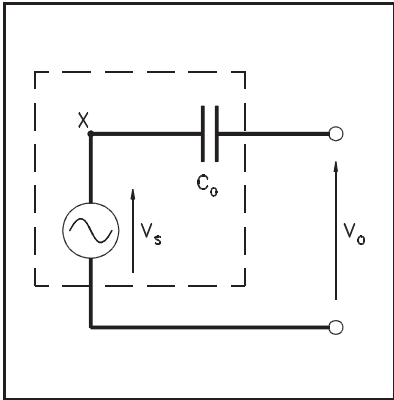
Description	Part No.	Capacitance
LDT0-028K/L	0-1002794-1	500 pF
DT1-028K/L	1-1002908-0	1.3 nF
DT1-052K/L	2-1002908-0	650 pF
DT2-028K/L	1-1003744-0	2.6 nF
DT4-028K/L	1-1002150-0	9 nF
8" x 11" 28 μm	1-1003702-4	30 nF
HYD-CYL-100	0-1001911-1	43 pF

Equivalent Circuit of Piezo Film

We are now ready to draw out an electrical equivalent of the piezo film element. There are two equally valid "models" - one is a *voltage* source in series with a capacitance, the other a *charge* generator in parallel with a capacitance - but the latter is uncommon in electrical circuit analysis and we will concentrate on the voltage source (see Figure 12).

The dashed line represents the "contents" of the piezo film component. The voltage source V_s is the piezoelectric generator itself, and this source is directly proportional to the applied stimulus (pressure, strain, etc). It is not the purpose of

Figure 12. Piezo film element as a simple voltage generator



this article to elaborate further on the calculations involved, but it is important to realize that this voltage will absolutely follow the applied stimulus - it is a "perfect" source.

Note, however, that the node marked "X" can never be accessed! The film's capacitance C_0 will always be present and connected when we monitor the "output" of the film at the electrodes.

Adding in a resistive load

Now we can add in the effect of connecting up to the oscilloscope. The oscilloscope and its probe are modeled simply as a pure resistance, although in reality there will be a very small capacitance associated with the probe and the cable (usually in the region of 30 to 50 pF). This can be neglected if the film capacitance is significantly higher in value.

The voltage measured across the load resistor R_L will **not** necessarily be the same voltage developed by the "perfect" source (V_s).

To see why, it is helpful to redraw this circuit in another way.

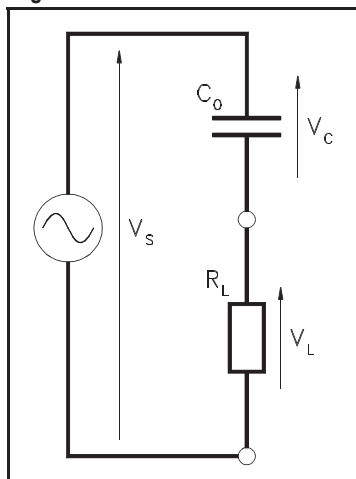
Potential Divider

With the circuit shown in Figure 13 redrawn as in Figure 14, it is easier to see why the full source voltage does not always appear across the resistive load.

A *potential divider* is formed by the *series* connection of the capacitance and the resistance. Since the capacitance has an impedance which varies with frequency, the share of the full source voltage which appears across R_L also varies with frequency.

The proportion (V_L) of V_s which appears across R_L is given by:

Figure 14. Potential divider

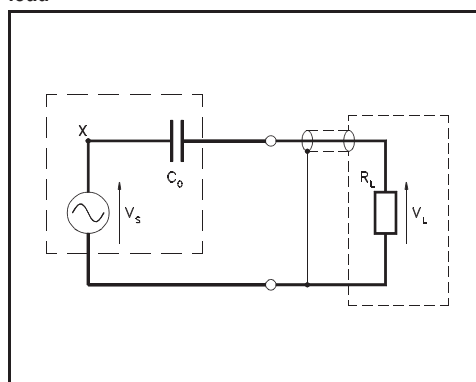


where

(j denoting $\sqrt{-1}$, and X_C being the reactance of the capacitive element. For simplicity, we ignore any resistive component of the film's impedance).

The above equations may be used in simple ways to calculate the voltage level expected to be observed in simple cases where the frequency of excitation is constant, and so a value of f can simply be substituted. In many real-world cases, however, there may be a distribution of signal energy over a band of frequencies. Then it becomes useful to consider the "frequency response" of the network.

Figure 13. Adding the oscilloscope as resistive load



$$\frac{Z_C}{V_L} = \frac{-jX_C}{R_L + Z_C} = \frac{-j}{2\pi fC}$$

Frequency Response

This is illustrated in the following example graphs. First, a lin/lin plot is shown (Figure 15, linear y-scale or amplitude, plotted against linear x-scale or frequency) with the corresponding phase plot (Figure 16) also shown in lin/lin form. Following these is a log/log plot (Figure 17), which will be dealt with in a little greater detail.

Note that the phase curve indicates that at very low frequencies, the observed voltage will show significant phase deviation from the source (limiting at -90° or $-\pi/2$ radians at "dc" or zero Hz). The significance of this effect is great if the piezo film element is to be used as part of a control loop.

Figure 15. Magnitude response of R-C filter

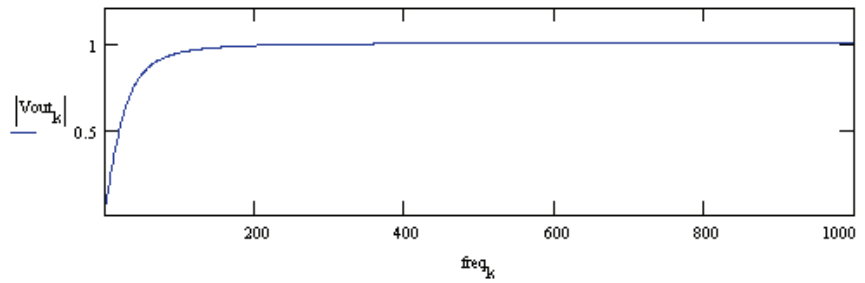


Figure 16. Phase response of R-C filter

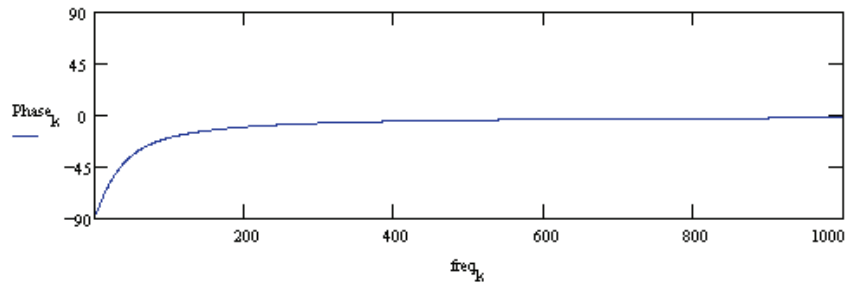
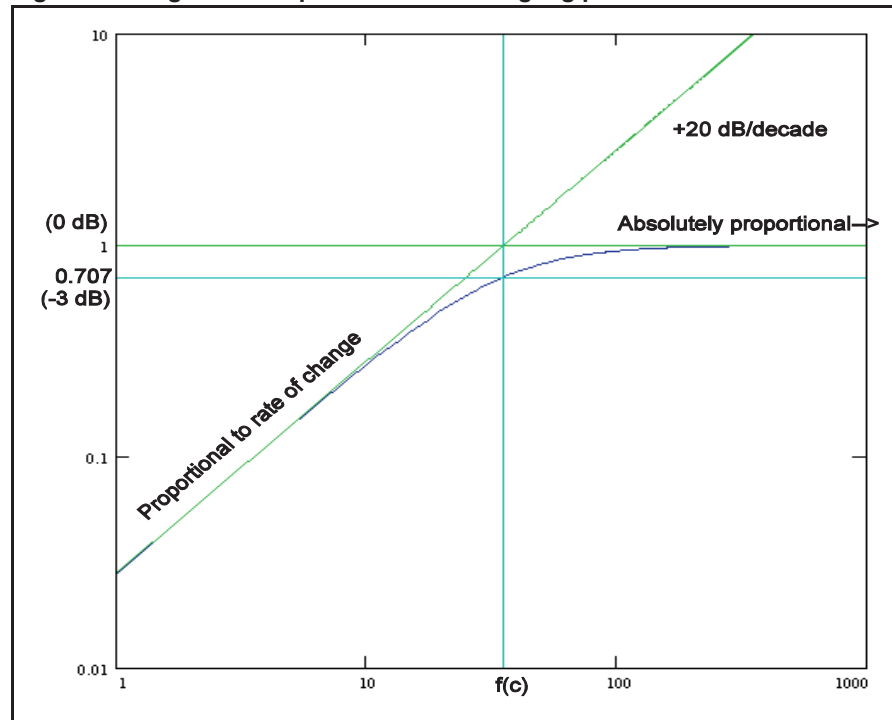


Figure 17. Magnitude response shown as log/log plot



Analysis of the log/log R-C frequency response curve

Some key features:

- # the overall characteristic of this network is known as a high-pass filter
- # the frequency at which the magnitude falls to 0.707 or -3 dB is known as the "cut-off" or "corner" frequency of the high-pass filter
- # this frequency can be calculated as $f(c) = 1/(2\pi RC)$, when both the resistance R and capacitance C are known
- # at frequencies well *below* the cut-off frequency, the plot has the form of a straight line with gradient +20 dB/decade (in other words, doubling the frequency will double the signal amplitude) - this characteristic is identical with that of a *differentiator* network, and gives an output which is proportional to the *rate of change* of the input quantity
- # at frequencies well *above* the cut-off frequency, the plot is level at "unity gain" and the output is directly proportional to the input quantity
- # the filter characteristic can be approximated by these two intersecting straight lines, but the magnitude actually follows an asymptotic curve, with magnitude -3 dB at the cut-off frequency where the straight lines cross
- # the filter characteristic can then be applied to the frequency-domain description of any practical signal by multiplying the filter transfer characteristic with the spectrum of the input signal, and deriving a response curve (output) which can in turn be transformed back into a time-domain signal.

Some practical examples of the effect of this filter characteristic will be shown next. For each signal, the time-domain description of the "perfect source" (e.g. the waveform which would be seen on an oscilloscope if the filter characteristic was absent) is given first, followed by its spectrum (obtained by use of the FFT [Fast Fourier Transform] algorithm supplied in the analysis software), then the filter characteristic (identical for all examples, but shown to emphasize the effect), then the resulting output signal spectrum obtained by multiplying the complex input spectrum by the complex filter characteristic, and finally the corresponding time-domain description obtained by inverse FFT, which shows the waveform an engineer would expect to observe in reality.

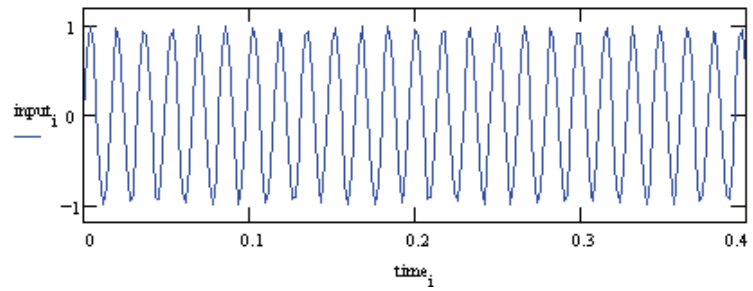
Note: in Figures 15, 16 and 17 the R-C values used to generate the curve were $R = 1\text{M}\Omega$ and $C = 4.5\text{ nF}$. In the following plots, the value of C was reduced to 1.5 nF. These values were chosen somewhat arbitrarily to demonstrate the principle, and so the scaling on the curves has not been annotated. But the time waveforms can be read in x units of seconds, and the frequency curves with x units of Hz. The cut-off frequency for $R = 1\text{M}\Omega$ and $C = 1.5\text{ nF}$ is approximately 106 Hz.

Key to following figures

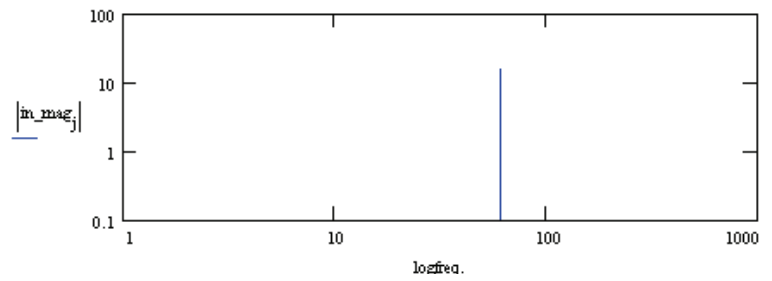
- Figure 18** shows a relatively high-frequency sine wave passing through the network. In the input spectrum, the signal is represented by a single spectral line at the appropriate frequency. This frequency is just below the filter "cut-off", and so is only slightly attenuated by the network. The resulting output wave is diminished in amplitude, and slightly shifted in phase.
- Figure 19** shows the same process applied to a slower sine wave. In this case, the attenuation is much greater, and the phase shift more significant. This situation occurs when trying to monitor steady vibration at "too low" a frequency using a piezo sensor. The phase behavior may be significant if a control loop is to be implemented.
- Figure 20** shows a harmonic series, with a number of discrete spectral lines all lying below the cut-off frequency. Each is attenuated to a different extent, and so the "balance" of harmonics in the output signal is altered.
- Figure 21** shows a slow half-sine input pulse (typical of many mechanical impact signals). Although the high-frequency content is largely unaltered, the output waveform appears heavily "distorted" and clearly shows both positive and negative excursions, whereas the input waveform is unipolar.
- Figure 22** shows a sawtooth waveform with slowly rising "leading edge" followed by a "snap" descent back to zero. Many piezo switches detect this form of mechanical event. In the output waveform, the "leading edge" has almost disappeared, but the "snap" gives almost full amplitude. Note the polarity of the output pulse relative to the input waveform.

Figure 18. Effect of R-C filter on High Frequency Sine Wave input waveform

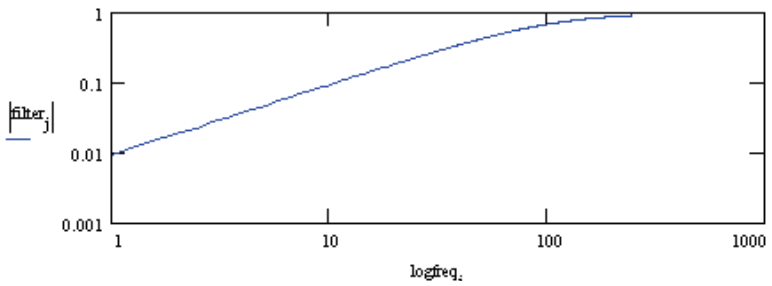
a) Input waveform



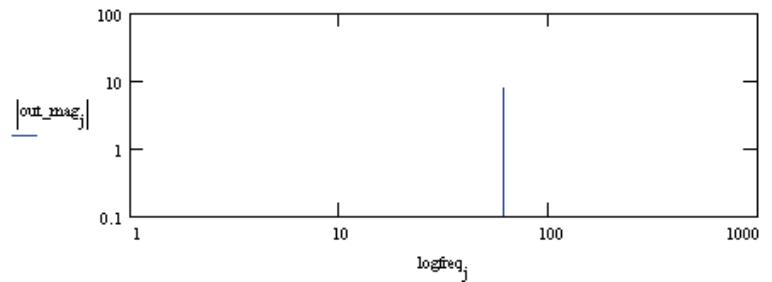
b) Input spectrum



c) Filter characteristic



d) Output spectrum



e) Output waveform

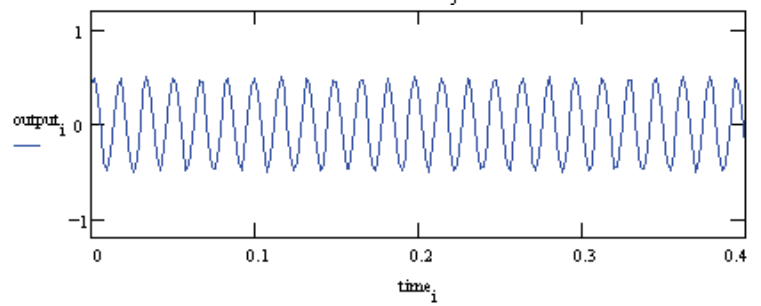
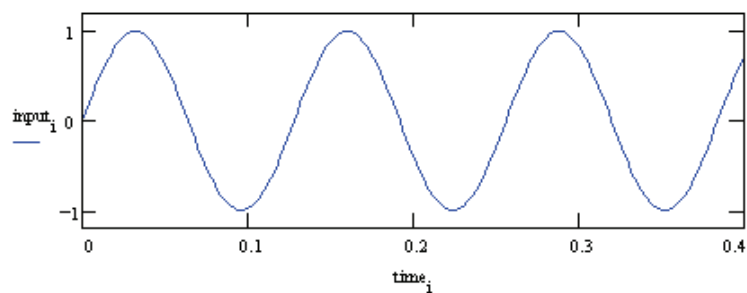
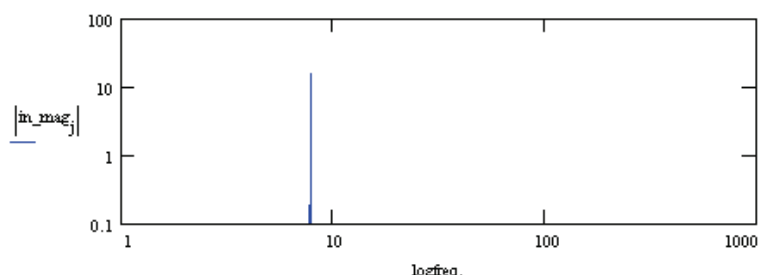


Figure 19. Effect of R-C filter on Low Frequency Sine Wave input waveform

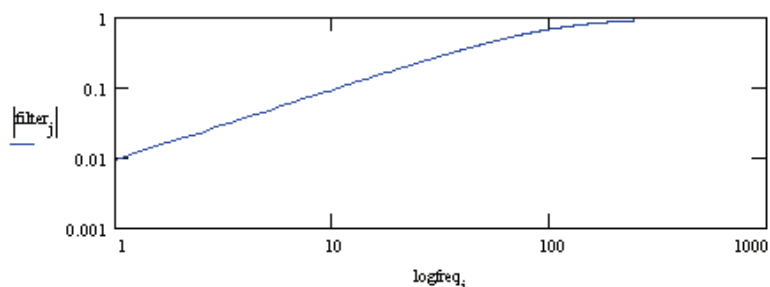
a) Input waveform



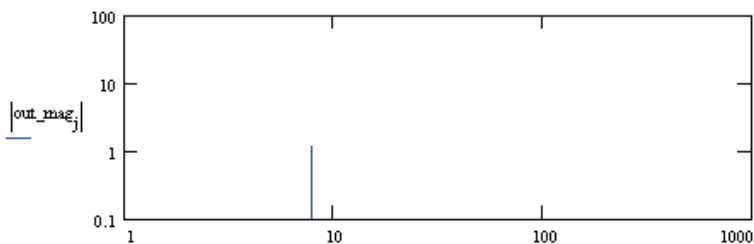
b) Input spectrum



c) Filter characteristic



d) Output spectrum



e) Output waveform

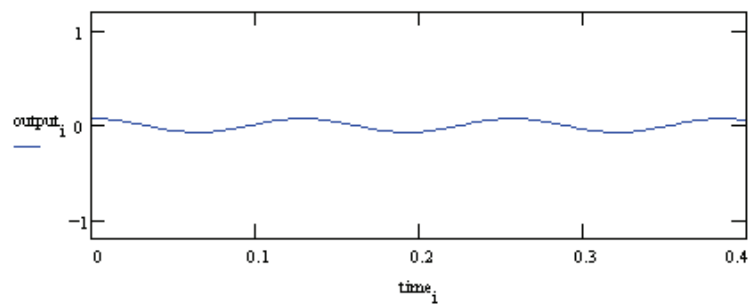
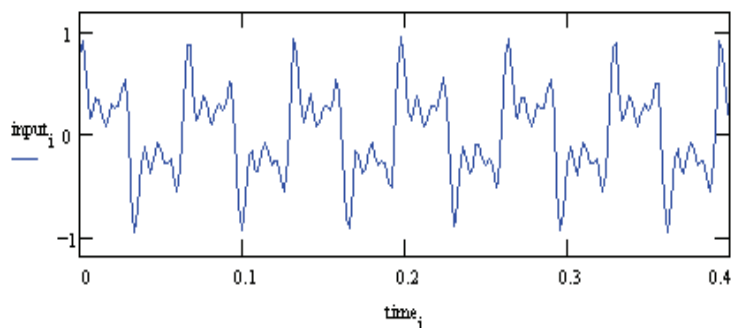
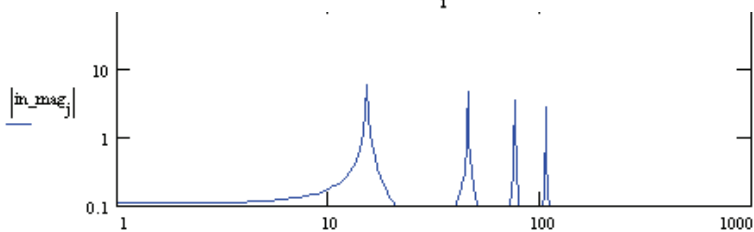


Figure 20. Effect of R-C filter on Harmonic Series input waveform

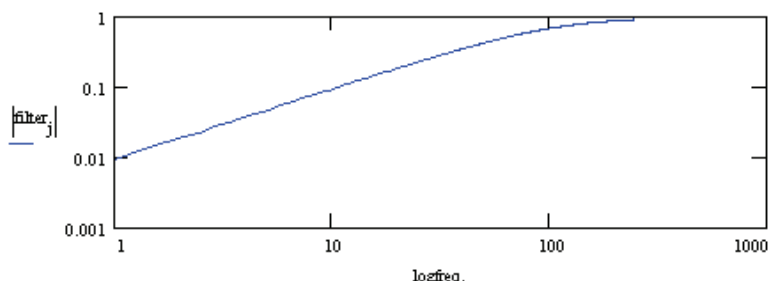
a) Input waveform



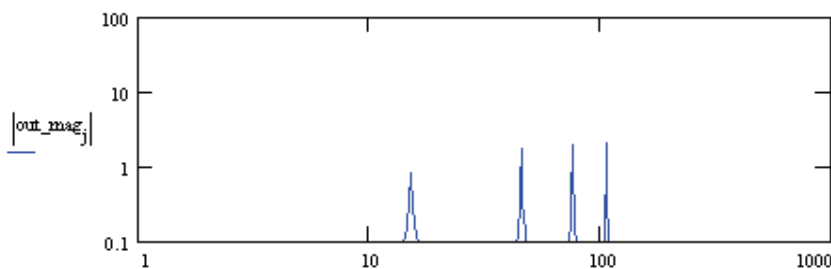
b) Input spectrum



c) Filter characteristic



d) Output spectrum



e) Output waveform

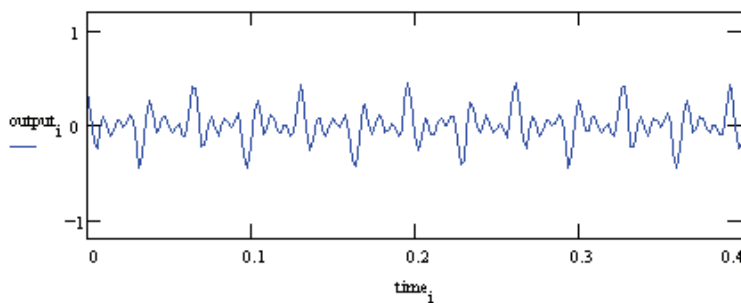
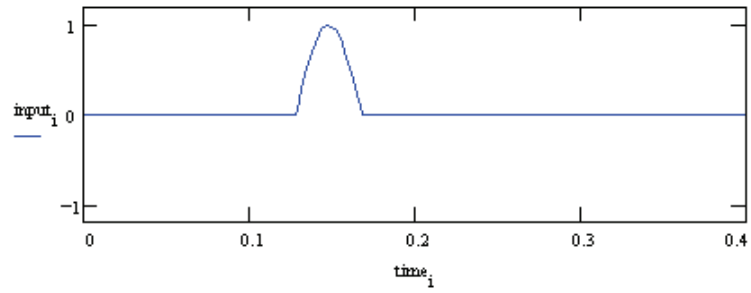
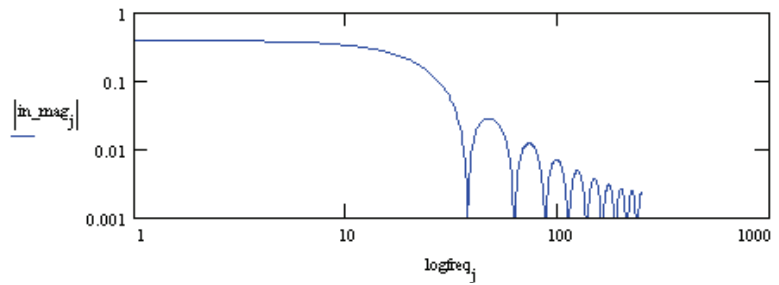


Figure 21. Effect of R-C filter on Slow Half-Sine Transient input waveform

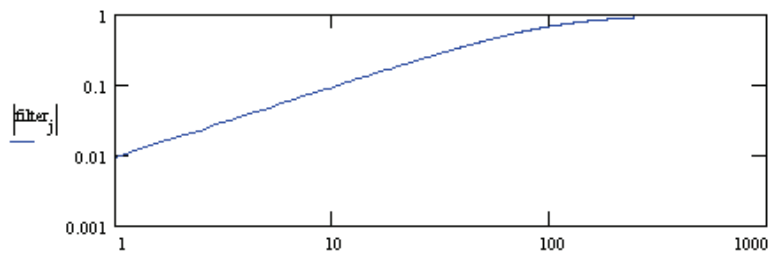
a) Input waveform



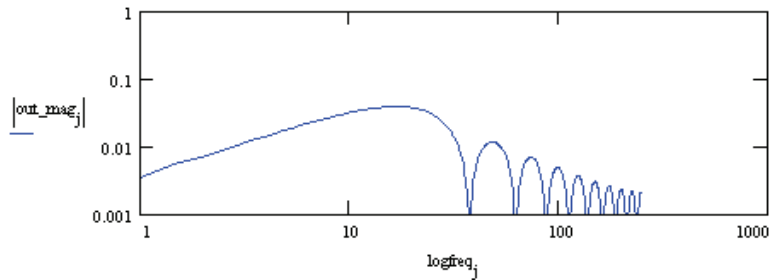
b) Input spectrum



c) Filter characteristic



d) Output spectrum



e) Output waveform

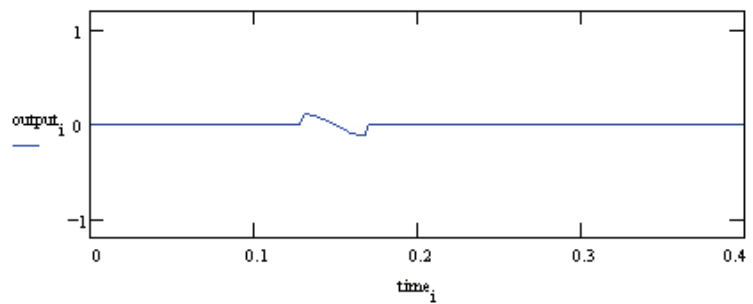
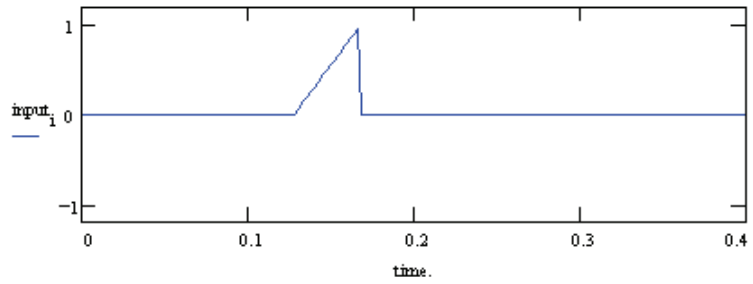
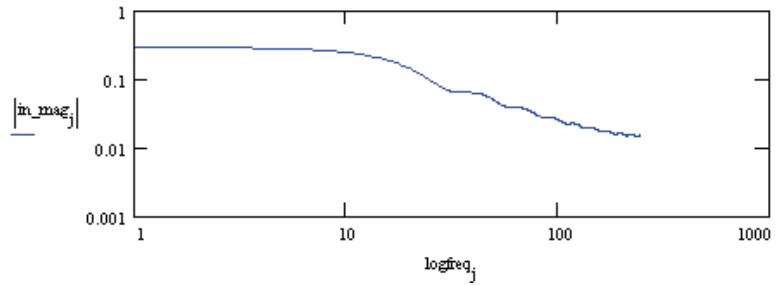


Figure 22. Effect of R-C filter on Slow Sawtooth Transient input waveform

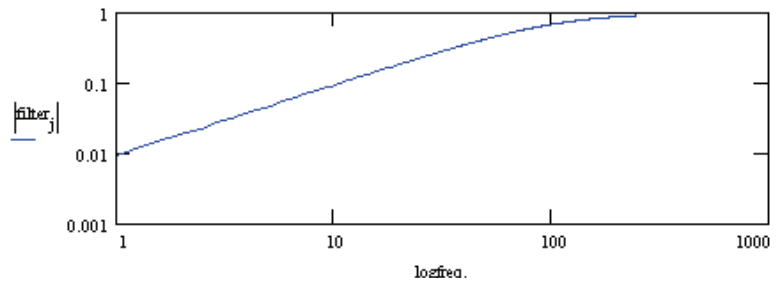
a) Input waveform



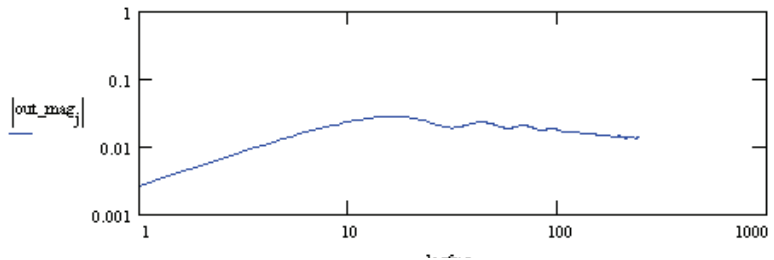
b) Input spectrum



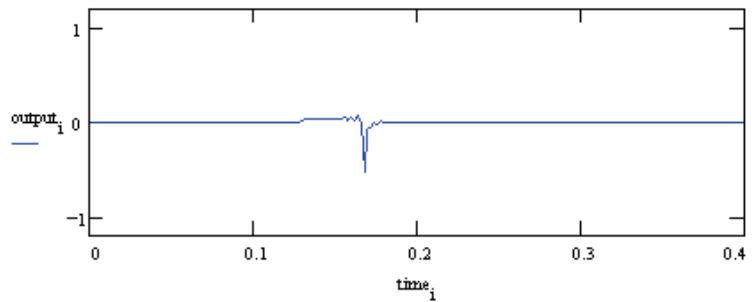
c) Filter characteristic



d) Output spectrum



e) Output waveform

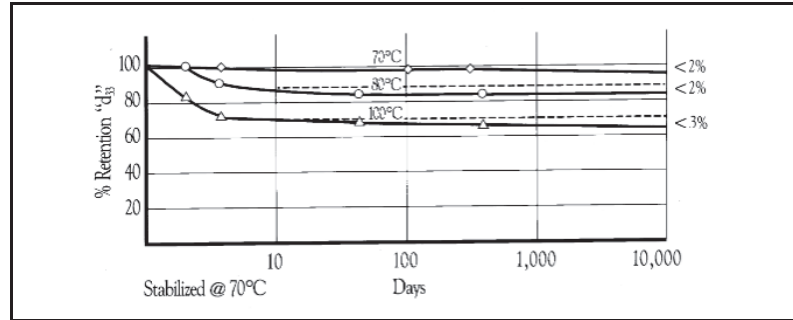


TEMPERATURE EFFECTS

Many of the properties of piezo film change with excitation frequency and temperature. These properties are reversible and repeatable with either frequency or temperature cycling.

In addition, Figure 23 shows the permanent decay of the piezoelectric strain constant d_{33} for PVDF, annealed at 70°C, after long term exposure to elevated temperatures.

Figure 23. Thermal stability of d_{33} constant - 70 °C annealed PVDF



Having reached a stabilizing temperature, the material properties then remain constant with time. Piezo film can be annealed to specific operating (or maximum storage) temperatures to achieve long-term stability for high temperature applications. Figure 24 shows the reversible temperature effects on d_{33} and g_{31} coefficients for PVDF.

In Figures 25a and 25b, the effect of temperature on the dielectric constant (ϵ/ϵ_0) and dissipation factor ($\tan \delta_e$) are shown for copolymer films.

Piezo films have been shown to offer excellent transducer properties at very low (cryogenic) temperatures.

Figure 24. Temperature coefficient for d_{33} and g_{31} constants - PVDF

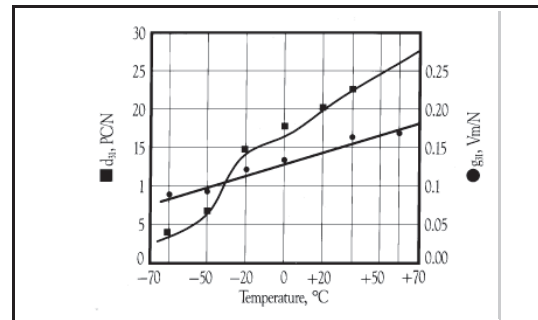


Figure 25a. Dielectric loss tangent vs. temperature COPOLYMER

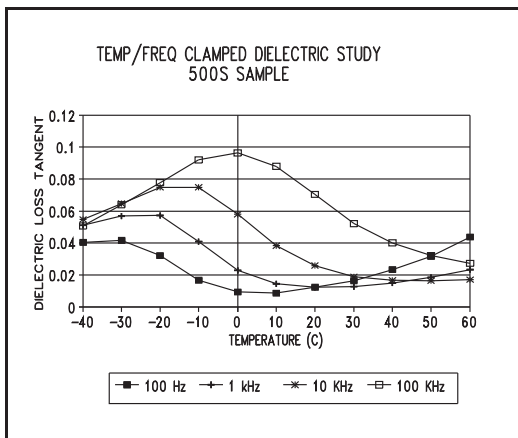
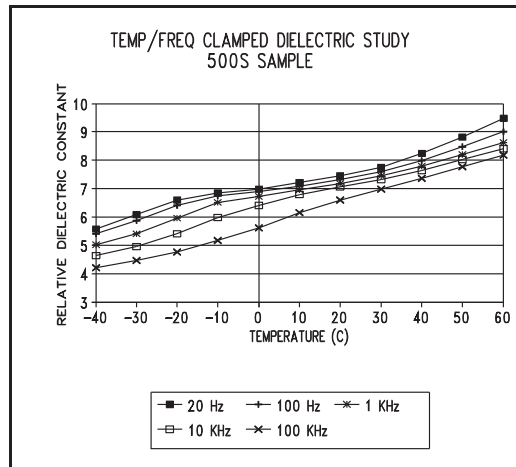


Figure 25b. Dielectric constant vs. temperature COPOLYMER



PIEZOELECTRIC CABLE AND PROPERTIES

One of the most recent developments in piezo polymer technology is piezo cable. The cable has the appearance of standard coaxial cable, but is constructed with a piezoelectric polymer insulator between the copper braid outer shield and the inner conductor (Figure 26).

Protected by a rugged polyethylene jacket, the cable is used in buried or fence security systems, traffic sensors including vehicle classification and weight-in-motion systems, and taxiway sensors for aircraft identification, safety and security applications. Other applications include sensors for anti-tampering, door edge safety monitoring, floor mats, touch pads and panels, and patient mattress monitors. The new cables feature the same piezoelectric properties that are characteristic of piezo film sensors. The electrical output is proportional to the stress imparted to the cable. The long, thin piezoelectric insulating layer provides a relatively low output impedance (600 pF/m), unusual for a piezoelectric device. The dynamic range of the cable is substantial (>200 dB), sensing distant, small amplitude vibrations caused by rain or hail, yet responding linearly to the impacts of heavy trucks. The cables have withstood pressures of 100 MPa. The typical operating temperature range is -40 to +125°C. Table 4 lists typical properties for piezo cable.

Figure 26. Piezo cable construction

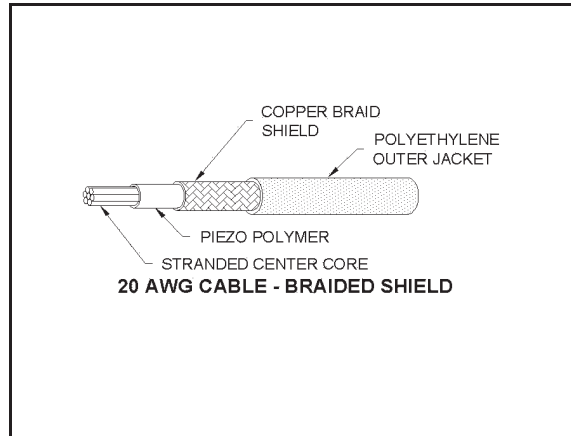


Table 4. Piezo Cable Typical Properties

Parameter	Units	Value
Capacitance @ 1KHz	pF/m	600
Tensile Strength	MPa	60
Young's Modulus	GPa	2.3
Density	kg/m ³	1890
Acoustic Impedance	MRayl	4.0
Relative Permittivity	@1KHz	9
tan δ_e	@1KHz	0.017
Hydrostatic Piezo Coefficient	pC/N	15
Longitudinal Piezo Coefficient	Vm/N	250 x 10 ⁻³
Hydrostatic Piezo Coefficient	Vm/N	150 x 10 ⁻³
Electromechanical Coupling	%	20
Energy Output	mJ/Strain (%)	10
Voltage Output	kV/Strain (%)	5

Cable Typical Properties

The output sensitivity of piezo cable in response to increasing impact load is shown in Figure 27a. The linearity in output for increasing force as shown in Figure 27b is typical of all piezo cable gages.

PIEZOELECTRIC BASICS

Figure 27a. Sensitivity vs. load

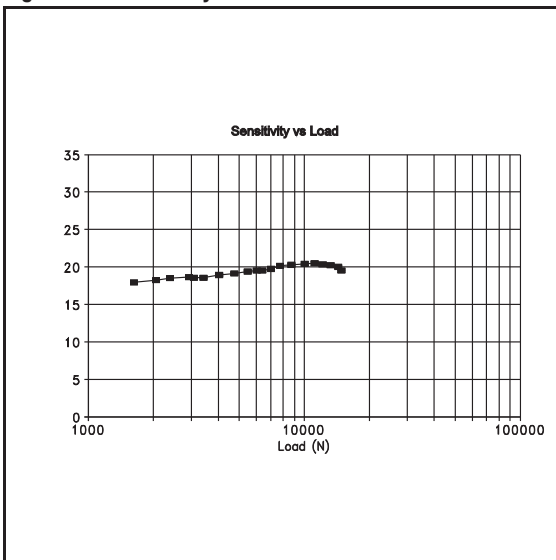
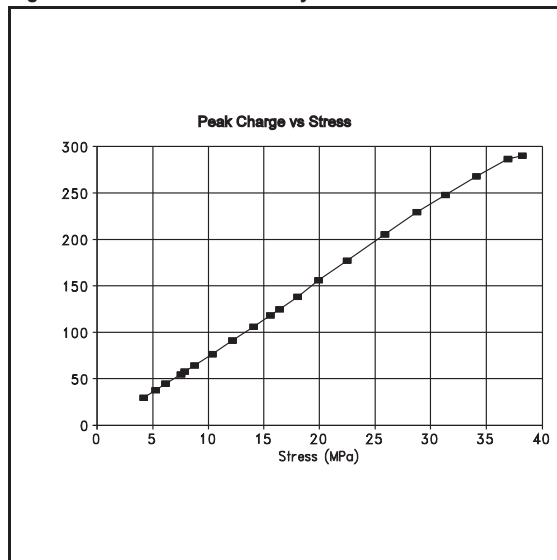


Figure 27b. Piezo cable linearity



Mechanical to Electrical

Like water from a sponge, piezoelectric materials generate charge when squeezed. The amplitude and frequency of the signal is directly proportional to the mechanical deformation of the piezoelectric material. The resulting deformation causes a change in the surface charge density of the material so that a voltage appears between the electroded surfaces. When the force is reversed, the output voltage is of opposite polarity. A reciprocating force thus results in an alternating output voltage.

Piezo film, like all piezoelectric materials, is a dynamic material that develops an electrical charge proportional to a change in mechanical stress. Piezoelectric materials are not suitable static measurements (true dc) due to their internal resistance. The electrical charges developed by piezo film decay with a time constant that is determined by the dielectric constant and the internal resistance of the film, as well as the input impedance of the interface electronics to which the film is connected. Practically speaking, the lowest frequency measurable with piezo film is in the order of 0.001Hz. There are methods to achieve true dc response, but these require using the piezo film as both an actuator and sensor, monitoring change in the actuation resulting from the dc event.

The fundamental piezoelectric coefficients for charge or voltage predict, for small stress (or strain) levels, the charge density (charge per unit area) or voltage field (voltage per unit thickness) developed by the piezo polymer.

Charge Mode:

Under conditions approaching a short circuit, the generated charge density is given by:

$$D = Q/A = d_{3n}X_n \quad (n = 1, 2, \text{ or } 3)$$

The mechanical axis (n) of the applied stress (or strain), by convention, is:

- 1 = length (or stretch) direction
- 2 = width (or transverse) direction
- 3 = thickness direction

where

D = charge density developed

Q = charge developed

A = conductive electrode area

d_{3n} = appropriate piezoelectric coefficient for the axis of applied stress or strain

n = axis of applied stress or strain

X_n = stress applied in the relevant direction

It is important to note that the d_{3n} coefficient is commonly expressed in pico-Coulombs per Newton (pC/N), but the more correct form would be (pC/m²)/(N/m²) since the areas (m²) upon which the stresses or strains applied are very often different and cannot be "canceled".

Voltage Mode:

The open-circuit output voltage is given by:

$$V_o = g_{3n}X_n t \quad (n = 1, 2, \text{ or } 3, \text{ as above})$$

where

g = appropriate piezoelectric coefficient for the axis of applied stress or strain

X_n = applied stress in the relevant direction

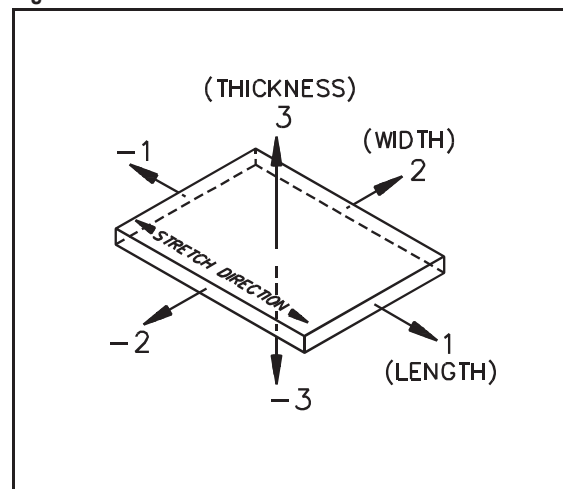
t = the film thickness

Piezo Coefficients:

The most widely used piezo coefficients, d_{3n} and g_{3n} , charge and voltage respectively, possess two subscripts. The first refers to the electrical axis, while the second subscript refers to the mechanical axis. Because piezo film is thin, the electrodes are only applied to the top and bottom film surfaces. Accordingly, the electrical axis is always "3", as the charge or voltage is always transferred through the thickness ($n = 3$) of the film. The mechanical axis can be either 1, 2, or 3, since the stress can be applied to any of these axes, as shown in Figure 28.

Typically, piezo film is used in the mechanical 1 direction for low frequency sensing and actuation (< 100KHz) and in the mechanical 3 direction for high ultrasound sensing and actuation (> 100KHz).

Figure 28. Numerical classification of axes



Directionality:

Piezoelectric materials are anisotropic. This means that their electrical and mechanical responses differ depending upon the axis of applied electrical field or axis of mechanical stress or strain. Calculations involving piezo activity must account for this directionality.

EXAMPLE 1:

A 1.45 psi load (10,000 N/m²) is applied to a piezo film switch of 2.54 cm length, 2.54 cm width and 110µm in film thickness. The switch element is rigidly backed, so the force acts to compress the film's thickness (therefore g₃₃ mode). In this example the load acts on the length by width area of the piezo film. The open circuit voltage developed across the thickness of the piezo film is:

$$V_o = -g_{33} X t$$

$$g_{33} = -339 \times 10^{-3} \frac{V/m}{N/m^2}$$

where:

V/m is Volts out per meter of piezo film thickness

N/m² is stress applied to the relevant film area. The conversion from psi to

N/m² is approximately 7,000.

$$V_o = - \left(-339 \times 10^{-3} \frac{V/m}{N/m^2} \right) (-10,000 N/m^2) (110 \times 10^{-6} m)$$

$$V_o = -0.373 \text{ volts}$$

EXAMPLE 2:

The same piezo film element as in EXAMPLE 1 is subjected to a force (10,000 N/m² x 0.0254m² = 6.45 Newtons), but in this example, the film switch is configured as a membrane having a compliant backing. Now, the force acts on the thickness cross-sectional area (wt). The piezo film is being stretched by the load, so it is acting in the g₃₁ mode.

$$V_o = -(3g_{31} \left(\frac{F}{wt} \right) (t) = -(g_{31}) \left(\frac{F}{w} \right)$$

$$g_{31} = 216 \times 10^{-3} \frac{V/m}{N/m^2}$$

$$V_o = - \left(216 \times 10^{-3} \frac{V/m}{N/m^2} \right) \left(\frac{6.45 N}{2.54 \times 10^{-2} m} \right)$$

$$V_o = -54.9 \text{ volts}$$

The sharp increase in output voltage results because the force is applied to the much smaller cross-sectional area of the film. The small area results in a correspondingly higher stress.

Dynamic Range

Piezo film has a vast dynamic range. The sensor has been used to detect the impact of high speed particles in space having a mass of 10^{-12} grams, and at the other extreme, measures shock waves at 300,000 atmospheres produced during weapons testing. A recent study was conducted to determine the maximum output energy of a 52 μ m thick film, having an area of 155.5 mm x 18.5 mm. The film was subjected to approximately 350 MPa (in the stretch or "n = 1" direction) without failure. The charge generated was found to be very linear, with the following measurements made at maximum applied stress:

Maximum Charge Observed: 20 μ C, giving 6.95 mC/m²
Maximum Voltage Observed: 1600 V, giving 30.8×10^6 V/m
Maximum Energy Converted: 30.9 mJ, giving 207 kJ/m³

Later experiments showed that about 10% of the above energy levels can be sustained for long periods of time without measurable damage to the piezo film element.

Electrical to Mechanical

When a voltage is applied to a sheet of piezo film, it causes the film to change dimensions due to the attraction or repulsion of internal dipoles to the applied field.. With one voltage polarity is applied, the piezo film becomes thinner, longer and wider. The opposite polarity causes the film to contract in length and width and become thicker. An ac voltage causes the film to "vibrate".

The amount of deformation is given by the piezoelectric " d_{3n} " constant:

$$\text{for length change} \quad \Delta l = l d_{31} V/t$$

where

Δl = change in film length in meters
 l = original film length in meters
 d_{31} = piezoelectric coefficient for length ("n=1" direction) change in meters per volt
 V = applied voltage across the thickness (t)

$$\text{for width change} \quad \Delta w = w d_{32} V/t$$

where

d_{32} = piezoelectric coefficient for width ("n=2" direction) change

$$\text{for thickness change} \quad \Delta t = t d_{33} V/t = d_{33} V$$

where

d_{33} = piezoelectric coefficient for thickness ("n=3" direction) change

EXAMPLE 3:

A piezo film of 3 cm length (l), 2 cm width (w) and 9 μ m thickness (t) is subjected to an applied voltage of V=200 volts in the 3 (thickness) direction. The amount of strain S resulting from this electrical input is d times the applied field.

In the l direction:

$$S_1 = \frac{\Delta l}{l} = d_{31}(V/t) \quad \text{where } d_{31} = 23 \times 10^{-12} \frac{\text{m/m}}{\text{V/m}}$$

$$\Delta l = d_{31}(V/t) l = \left(23 \times 10^{-12} \frac{\text{m/m}}{\text{V/m}} \right) \frac{(200 \text{ V})(3 \times 10^{-2} \text{ m})}{(9 \times 10^{-6} \text{ m})}$$

$$\Delta l = 1.53 \times 10^{-5} \text{ m} \quad \text{or} \quad 15.3 \mu\text{m}$$

In the t direction:

$$\Delta t = t d_{33}(V/t) l = d_{33} V = \left(-33 \times 10^{-12} \frac{\text{m/m}}{\text{V/m}} \right) (200 \text{ V}) = 6.6 \times 10^{-9} \text{ m} \quad \text{or} \quad 66 \text{ \AA}$$

Actuators

Generally, piezo film actuator designs depend on the application requirements such as operating speed, displacement, generated force, and available electrical power. Piezo film technology offers various design options to meet such application requirements. Those design options include:

- ! Customized electrode patterns on one or both sides of the piezo film sheet.
- ! Multilaminate structures or bimorphs.
- ! Fold-over or scrolled multilayer structures.
- ! Extruded piezo tubes and piezo cables.
- ! Cast piezo polymer on various substrates
- ! Molded 3-D structures.

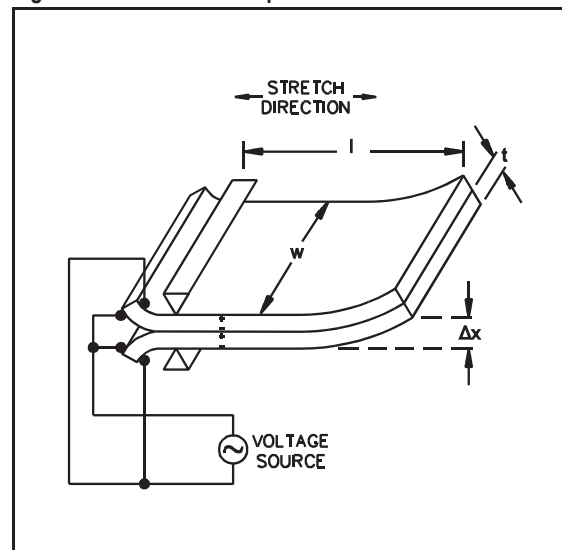
Each design option mentioned above has advantages and disadvantages. For example, scrolled multilayer actuators can generate a higher force but may sacrifice some displacement.

Bimorph

Like a bimetal strip, two sheets of piezo film of opposite polarities, adhered together form a bending element, or "bimorph" (Figure 29). An applied voltage causes one film to lengthen, while the other contracts, causing the unit to bend. An applied voltage of opposite polarity bends the bimorph in the opposite direction.

The bimorph configuration converts small length changes into sizable tip deflections, but producing low force.

Figure 29. Piezo film bimorph



Thicker films and multilayers improve the force developed by the bimorph, but sacrifice displacement unless the unit can be operated at higher fields.

The amount of tip deflection and the force developed are given by:

$$\Delta x = 3/4d_{31}(l^2/t^2)V \quad \text{meters}$$

and

$$F = 3/2Ywd_{31}(t/l)V \quad \text{Newtons}$$

where

- Δx = displacement at dc
- F = generated force
- d_{31} = piezoelectric coefficient in the "1" direction
- l, t, w = length, thickness, and width of piezo film
- V = applied voltage (Volts)
- Y = Young's modulus of piezo film ($2 \times 10^9 \text{ N/m}^2$)

By applying an ac voltage, the bimorph can act as a fan, similar to an insect wing. Although the piezo film bimorph does exhibit a dc response, maximum tip deflections are obtained when the unit is operated at resonance, determined by the length and thickness of the bimorph beam.

EXAMPLE 4:

100 volts are applied across a 2 cm long cantilever bimorph comprised of two strips of $9 \mu\text{m}$ PVDF. The resultant tip displacement Δx is:

$$\Delta x = \frac{3/4Vd_{31}l^2}{t^2}$$

$$\Delta x = \frac{3/4(100\text{ V})\left(23 \times 10^{-12} \frac{\text{m/m}}{\text{V/m}}\right)(2 \times 10^{-2} \text{ m})^2}{(9 \times 10^{-6} \text{ m})^2}$$

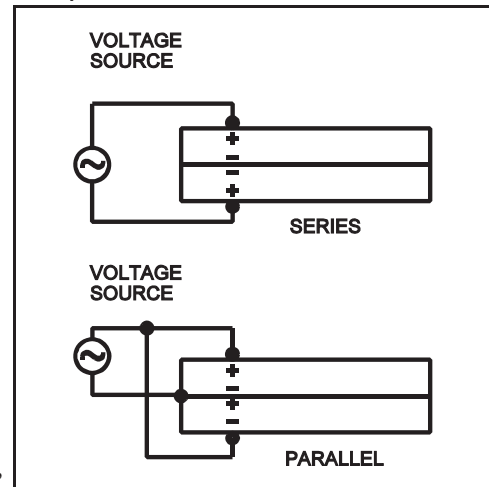
$$\Delta x = 8.52 \text{ mm}$$

As shown in the equations, more displacement can be obtained from a longer bimorph. Larger forces can be obtained from a wider bimorph. The ratio of displacement at a resonance frequency and dc is defined by Q which indicates a mechanical gain. A typical Q value for a piezo film bimorph is 20 to 25.

For example, a 5 mm long $70 \mu\text{m}$ thick bimorph with 120 volts dc creates a displacement of $57 \mu\text{m}$. With the same bimorph, however, displacement can be 1.4 mm at the resonant frequency of 580 Hz. For applications that require a higher force, such as cooling fans, multilayer construction can be considered. The resulting output force is proportionally increased by the number of layers.

In terms of electrical connections to the bimorph, there are two basic methods as shown in Figure 30 — parallel and series connections. In order to generate the same amount of displacement, the parallel connection requires a lower voltage than the series connection. Series connections, on the other hand,

Figure 30. Lead attachment methods for a bimorph



draw less current than parallel connections. For both parallel and series connections, the total electrical power to the actuator is identical. However, it is obvious that the lead attachment of the series connection is much simpler than that of the parallel connection for manufacturing purposes. Typical applications of the bimorph bender are cooling fans, toys, and decoratives.

Scrolled Actuator

The generated force and displacement of a scrolled piezoelectric cylinder in Figure 31 are expressed as follows:

$$x = d_{31}El \quad \text{Meters}$$

$$E = V/t \quad \text{Volts/meters}$$

$$F = Yd_{31}EA \quad \text{Newtons}$$

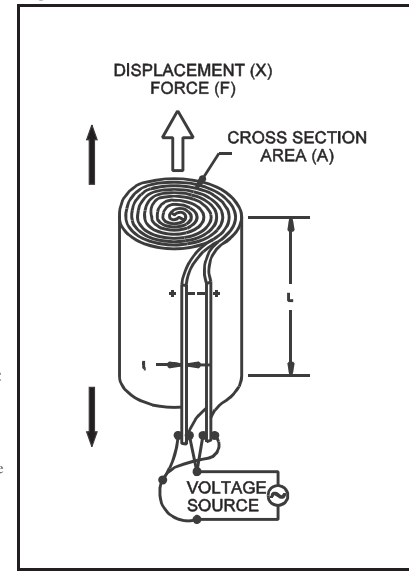
$$f = (1/2\pi)\sqrt{YA/l(M_e + 0.405M_p)}$$

where

- x = displacement at dc (meter)
- F = generated force (Newton)
- f = resonance frequency
- l, t = Length, thickness of piezo film (meters)
- M_e = externally loaded mass (kilograms)
- M_p = piezo actuator mass (kilograms)
- A = cross sectional area (m^2)
- Y = Young's modulus (N/m^2)
- E = electrical field (volt/meter)

As shown in the equations, a scrolled actuator can generate more force and can respond with a higher resonant frequency by increasing the cross sectional area. A longer actuator generates more displacement but reduces the response speed. Note that the actuator output, with $M_e = 0$, will be maximized when the length l is adjusted to satisfy the resonant condition. As an example, the performance of a 12 mm diameter, 25 mm long scrolled actuator can be maximized at 32 KHz operation.

Figure 31. Scrolled piezo film actuator



Folded Actuator

Another design option for a high speed, high force actuator is to fold over a long sheet of piezo film as shown in Figure 32. This design effectively creates a parallel wired stack of piezo film discs. The center hole is used to secure the actuator to a base. Design equations of the scrolled actuator also can be applied to this type of actuator. In the previous equations, d_{31} should be replaced with d_{33} ($-33 \times 10^{-12} C/m^2$) for a folded actuator. An example of specifications for the folded actuator is shown below:

- Displacement: $1 \mu m/1 \text{ mm}$ length
- Generated force: 15 kg/10 mm dia.
- Frequency: dc - 100 kHz
- Drive voltage: 800 volts

Compared to mechanical or piezo ceramic actuators, multilayer piezo film actuators have fewer ringing problems due to their lower Q. Applications of multilayer actuators are micropositioners for industrial equipment, acoustic wave generators and ink jet printers.

Ultrasonic Actuators

Ultrasonic actuators, as discussed in this section, exclude very high frequency (> 1 MHz) transmitter applications. The use of piezo film in these very high frequency applications, like medical ultrasound imaging and nondestructive testing, use thickness mode operation, d_{33} . This section deals with low frequency ultrasound (20-100 KHz) where the piezo film can be used in the length change (d_{31}) mode.

The advantage of piezo film in low frequency ultrasound can be found from the flexibility of the material. Piezo film can be easily curved or formed to make circular transducers as shown in Figure 33. The beam pattern is determined by the number of half circular elements and their diameter. The operating frequency is determined by the diameter of the half circular elements. Note that the difference between Figures 33(a) and 33(b) is their number of active elements and diameters. To widen the beam coverage, the number of active elements should be reduced. With a cylindrical transducer, a 360° beam pattern is obtained.

In ultrasound applications, a narrow beam with minimum side lobes is required for remote distance measurements. On the other hand, a wide beam, as wide as 180° or more, is required for applications like automobile rear bumper proximity sensing. Figure 33 shows design configurations for both narrow beam and wide beam ultrasound transducers. The applications for piezo film in through-air ultrasonic actuators include distance ranging for air pen, air mouse, white board digitizer, collision avoidance, physical security systems, air flow velocity (doppler) sensors, and inter-object communications. Similar constructions can be produced for underwater or fluid sensing, including flow sensors, level sensors, and communications.

Figure 32. Folded piezo film actuator

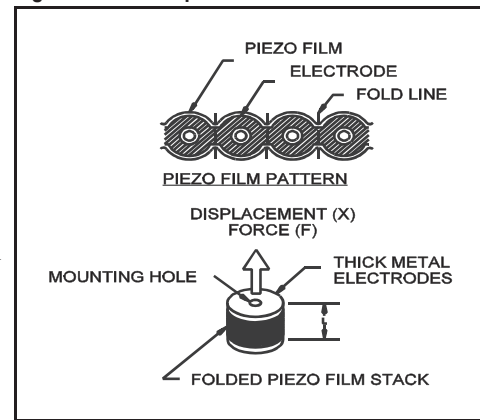
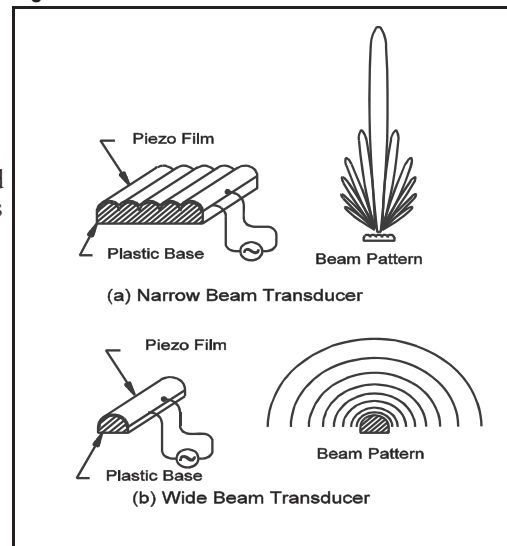


Figure 33. Piezo film ultrasound transducers



PYROELECTRIC BASICS

Piezoelectric polymers, such as PVDF and its copolymers of VF_2/VF_3 , are also pyroelectric. Pyroelectric sensor materials are normally dielectric materials with a temperature-dependent dipole moment. As these materials absorb thermal energy, they expand or contract, thereby inducing secondary piezoelectric signals. As piezo film is heated, the dipoles within the film exhibit random motion by thermal agitation. This causes a reduction in the average polarization of the film, generating a charge build up on the film surfaces. The output current is proportional to the rate of temperature change (ΔT). The amount of electrical charge produced per degree of temperature increase (or decrease) is described by the pyroelectric charge coefficient, ρ .

The charge and voltage produced in a given film of area A permittivity ϵ , and thickness t is given by

$$Q = p\Delta TA$$

$$V = pt\Delta T/\epsilon$$

EXAMPLE 5:

A piezo film pyroelectric detector having a film thickness (t) of $9\mu\text{m}$, a permittivity (ϵ) of $106 \times 10^{-12} \text{ C/Vm}$ and a pyroelectric coefficient (p) of $30 \times 10^{-6} \text{ C/(m}^2\text{K)}$, undergoes a temperature increase (ΔT) of 1°K due to incident IR radiation. The output voltage is given by:

$$V = \frac{(30 \times 10^{-6} \text{ C/m}^2 \text{ } ^\circ\text{K})(9 \times 10^{-6} \text{ m})(1^\circ\text{K})}{(106 \times 10^{-12} \text{ C/Vm})}$$

$$V = 2.55 \text{ volts}$$

The pyroelectric voltage coefficient of piezo film is about an order of magnitude larger than those of Lead Zirconate Titanate (PZT) and Barium Titanate (BaTiO_3). Table 5 compares the pyroelectric properties of these materials, but a far lower figure of merit due to the low capacitance of PVDF.

Table 5. Comparison of pyroelectric materials

Material	TGS	LiTaO ₃	BaTiO ₃	PZT	PbTiO ₃	PVDF	VF ₂ VF ₃
ρ_Q	350	200	400	420	230	30	50
ϵ/ϵ_0	30	45	1000	1600	200	10.7	8.0
α	.16	1.31	1.00	.44	.67	.06	.06
L	225	646	564	374	461	138	138
P_v	1.32	.50	.05	.03	.10	.47	.71
M_l	.53	.16	.02	.01	.03	.20	.31
Pyroelectric Charge Coefficient				$(\rho_Q)\mu\text{Coul}/[\text{m}^2 \cdot ^\circ\text{K}]$			
Dielectric Constant				(ϵ/ϵ_0) , where $\epsilon_0 = 8.85 \text{ pF/m}$			
Thermal Diffusivity				$(\alpha) \text{ m}^2/\text{sec} \cdot 10^{-6}$			
Thermal Diffusion Depth @ 1Hz				(L) μm			
Pyroelectric Voltage Coefficient				$(P_v)\rho_Q/\epsilon, \text{ V}/[\mu\text{m} \cdot ^\circ\text{K}]$			
Figure of Merit				$(M_l)\rho_Q/[C_v \cdot \epsilon], \text{ V} \cdot \text{mm}^2/\text{J}$			

Piezo film advantages including:

- moisture insensitivity (<.02% H₂O absorption)
- low thermal conductivity
- low dielectric constant
- chemical inertness
- large detector sizes

The pyroelectric response of piezo film can also become a noise source for piezo sensor applications at low frequencies. In piezoelectric applications where low frequency strain sensing is desired, there are several convenient methods to “common-mode reject” the pyroelectric response. Examples include:

- ! Two equal sized electrode patterns on one piezo film element; one electrode oriented parallel to the d_{31} and the other electrode pattern is perpendicular to the d_{31} direction. Both develop equal signals in response to pyro, but the electrode area parallel to the d_{31} develops about 10X the perpendicular electrode pattern. Subtracting the signals yields a pure piezo response.
- ! Two equal sized piezo film elements, laminated in a stacking configuration; one film has d_{31} parallel to strain surface, the other has d_{31} perpendicular to strain surface. As above, signals are subtracted to isolate the piezo response from pyro.
- ! Several other common mode rejection techniques can be described by MSI’s applications engineers.

For higher frequencies, where the rate of temperature change seen by the piezo film element is slower than the strain event to be measured, frequency filters readily sort out the unwanted pyro signal.

BASIC CIRCUIT CONCEPTS

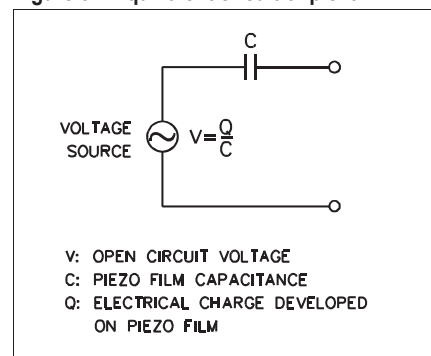
A properly designed interface circuit plays a key role in the optimization of piezo film sensors. The applications of piezo film span from toys to military sensors and interfacing to electronics is highly application dependent. In many cases, piezo film can be directly connected to electronic circuits without special interface considerations. However, for those cases where an interface circuit is required, the following 3 steps are recommended:

1. Consider the frequency range and signal amplitude requirements over the desired dynamic range.
2. Choose a proper load resistance to assure the low end operating frequency and to minimize signal loss due to the loading effect.
3. Select a buffer circuit if the signal level is small. If a high value load resistance is needed (such as 22M Ω or higher value), a low leakage high impedance buffer amplifier is recommended. JFET's or CMOS operational amplifiers are commercially available for a buffer.

Simplified Equivalent Circuits

The first step in an interface circuit design is to understand the piezo film characteristics as part of an electrical equivalent circuit. Figure 34 shows a simplified equivalent circuit of piezo film. It consists of a series capacitance with a voltage source. The series capacitance C_f represents piezo film capacitance which is proportional to the film

Figure 34. Equivalent circuit of piezo film



permittivity and area and inversely proportional to film thickness. The voltage source amplitude is equal to the open circuit voltage of piezo film and varies from microvolts to 100's of volts, depending on the excitation magnitude. This simplified equivalent circuit is suitable for most applications but is of limited value at very high frequencies such as that used in ultrasound transducers.

Figure 35 shows an equivalent circuit as a charge generator. This equivalent circuit has film capacitance C_f and internal film resistance R_f . The induced charge Q is linearly proportional to the applied force as described earlier. The capacitance C_f is proportional to the surface area of film and is inversely proportional to the film thickness. In low frequency applications, the internal film resistance R_f is very high and can be ignored. The open circuit output voltage can be found from the film capacitance; i.e., $V=Q/C_f$.

Input Resistance

The most critical part of an interface circuit is the input resistance. The input resistance affects low frequency measurement capability as well as signal amplitude. This is called the "loading effect".

Piezo film capacitance can be regarded as an equivalent source impedance. It is important to note that this source impedance increases with decreasing film capacitance and decreasing frequency of operation. This source impedance combined with the input resistance produces a voltage divider. As the ratio of input resistance to source impedance is decreased, the overall output voltage is reduced. Therefore, choosing a proper input resistance for the electronic interface is critical in minimizing the loading effect.

Time Constant

In addition to input resistance, the input capacitance of an interface circuit can also affect the output. Figure 36 shows the equivalent circuit of film with input resistance R_i and input capacitance C_i . A typical time domain response of piezo film is shown in Figure 37. The charge developed on the film due to an applied force decays with a time constant which is defined by $R_i(C_f + C_i)$.

This time constant represents the time required for a signal to decay to 70.7% (-3dB) of its original amplitude. The smaller the time constant, the quicker the signal decays. Because of this finite time constant, piezo film is suitable for dynamic measurements rather than static measurement (0.001 Hz minimum).

If a long time constant is desired, a high input resistance and film capacitance can be used. It should be understood, however, that a high input resistance can also produce higher noise, requiring compensation through shielding, etc.

Figure 35. Equivalent circuit for piezo film

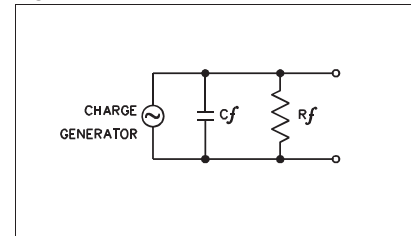


Figure 36. Equivalent circuit of piezo film with input resistance of electronic interface

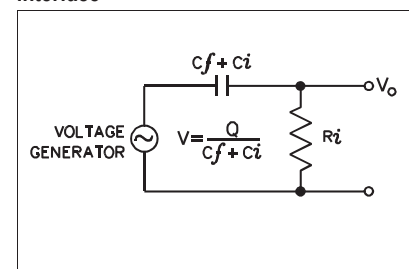
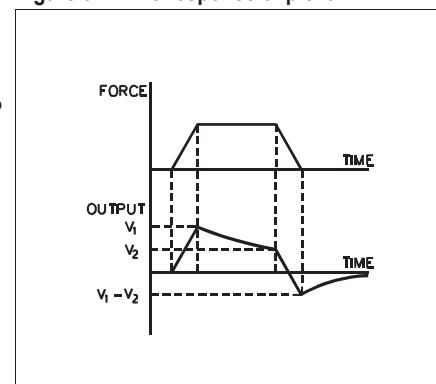


Figure 37. Time response of piezo film



Frequency Response

Another important aspect of the time constant can be seen in the frequency response of the equivalent circuit. The circuit exhibits an RC high-pass filter characteristic as shown in Figure 38. In this figure, the vertical axis implies the ratio of observable output signal to the developed signal (open circuit voltage of the piezo film). Zero dB implies no loss of signal. The cutoff frequency (3 dB down) is inversely proportional to the time constant. When a piezo film sensor is operated below this cut-off frequency, the output signal is significantly reduced. For a low frequency measurement, an input resistance needs to be high enough so that the cut-off frequency is well below the desired operating frequency. This consequence can be verified from consideration of the time constant as well as the loading effect.

As an example, the frequency response of a shielded piezo film sensor (model SDT1) is shown in Figure 39. In this example, the SDT is interfaced with a circuit which contains a 10M Ω load resistor and an FET. The capacitance of the piezo film is 2.4 nF. With 10M Ω load resistance, the time constant becomes 24 msec and thus, the cut-off frequency is 6.6 Hz. For comparison, the cut-off frequency can be reduced to 0.66 Hz if a 100M Ω resistor is used instead of the 10M Ω resistor. This sensor component can be used for any application operating above the cut-off frequency determined by the resistance value.

In applications where the electronic circuit cannot be placed near the sensor, a buffer circuit is recommended close to the sensor. The buffer circuit converts the high output impedance of the piezo film element into a low output impedance and thus minimizes the signal loss and noise through the cable. For large size (i.e., high capacitance) piezo film sensors a buffer may not be required, even with small signals and long cables.

When a high piezo film output impedance is required, a low-leakage, high impedance buffer is necessary. For example, infrared motion sensor and accelerometer applications require up to 50G Ω of input resistance to obtain a very low frequency response. For such cases, the input impedance of the buffer must be much higher than the output resistance of the piezo film in order to maintain the low frequency response. In addition, minimum leakage current of the buffer is critical in order to maximize the measurement accuracy. Some examples of low leakage buffer electronics include: JFET - 4117 (Siliconix, Sprague); Operational amplifiers — LMC660, LF353 (National Semiconductor), OP80 (PMI), and 2201 (Texas Instruments).

Figure 40 shows unity gain buffer circuit examples for general applications. Operational amplifiers offer a great deal of versatility as both buffers and amplifiers. They can be used as either charge-mode or voltage-mode amplifiers. Figure 41 shows basic charge and voltage amplifier configurations. The voltage output of the charge

Figure 38. High pass filter characteristic of piezo film

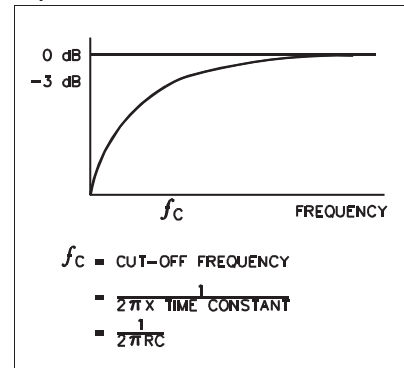


Figure 39. Frequency response of SDT1

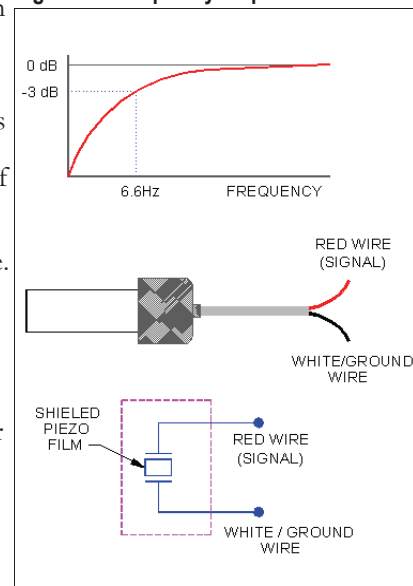
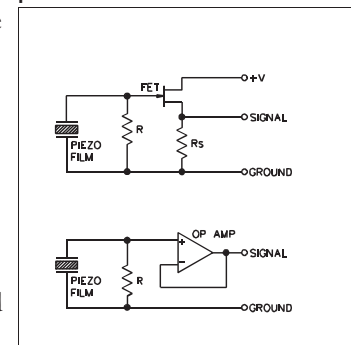


Figure 40. Unity gain buffer for piezo film sensors



amplifier is determined by Q/C_f . Q is the charge developed on the piezo film and C_f is the feedback capacitance of the charge amplifier.

The output voltage of the charge amplifier depends on the feedback capacitance, not the input capacitance. This indicates that the output voltage of a charge amplifier is independent of the cable capacitance. The major advantage of a charge amplifier can be realized when a long cable is used between a piezo film sensor and electronics. In addition, it also minimizes charge leakage through the stray capacitance around the sensor. Otherwise, simple voltage amplifiers are sufficient for most applications. Included in Figure 41 is a typical non-inverting voltage amplifier.

The advantage of a voltage amplifier can be seen when ambient temperature is considered. The voltage sensitivity (g-constant) variation over temperature is smaller than the charge sensitivity (d-constant) variation. Consequently, voltage amplifiers with piezo film exhibit less temperature dependence. In Figure 41, the time constants for the charge amplifier and voltage amplifier are determined by RC_f and RC respectively.

As a design example, a traffic sensor interface is described. Because of its flexibility, piezo cable is an ideal sensor material for traffic measurement applications. MSI's BL traffic sensor is constructed with a piezo cable sheathed in a compressed brass tube, with a variety of signal cable lengths tailored to the installation requirements. The BL is available in sensing lengths of more than 3 meters. In this specific example, the BL sensor is 2 meters long. This electrically shielded sensor has 100 feet of coax cable. The electrical specifications of this sensor include:

Capacitance = 9.5 nF (including piezo cable and signal cable capacitances)
 Output = 500mV (for a wheel load of 800 pounds at 55mph and 70°F)
 Signal : Noise = 10:1

The basic requirements of an interface circuit are:

Low end frequency = 1.6 Hz
 Circuit output = Digital pulse count

An interface circuit to meet these requirements is shown in Figure 42. This circuit works as a comparator. A 10MΩ input resistance is chosen in order to reduce the cut-off frequency to about 1 Hz. The actual cut-off frequency with this resistor can be calculated as 1.6 Hz. A 10MΩ potentiometer is used to adjust the threshold voltage, V and the diode is included to protect the electronics from high voltage damage. Typical piezo film and interface circuit output signals from a passenger car at 55 mph are shown in Figure 42.

Figure 41. Typical amplifiers for piezo film sensors

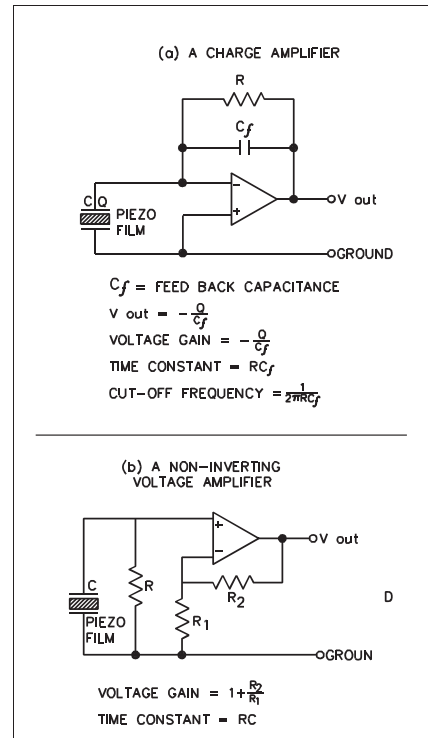
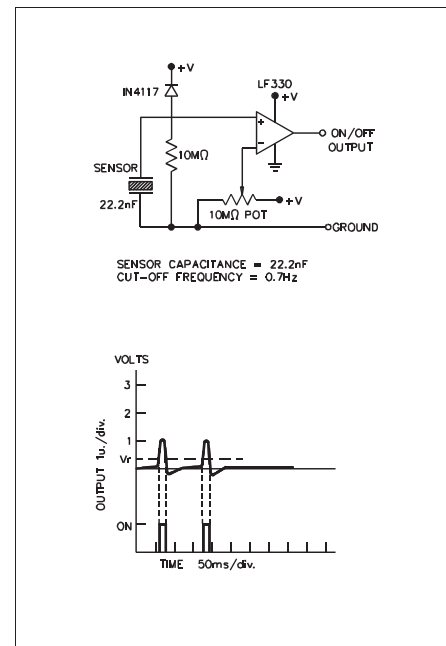


Figure 42. An interface circuit of a traffic sensor



Signal Conditioning

Because piezo film is both piezoelectric and pyroelectric, some provision must be made to eliminate—or at least reduce—the effect of unwanted signals. The primary principles of signal conditioning include:

- **Filtering**—Electrical filters designed to give the desired band-pass and band-rejection characteristics.
- **Averaging**—If the desired signal exhibits periodicity, while the undesired signal is random, signal averaging can increase the signal-to-noise ratio.
- **Common Mode Rejection**—By wiring two equal areas of a piezo film electrode out-of-phase, unwanted common-mode signals can be made to cancel.

Basic Switch Circuitry

A variety of circuits are available to electronically interface with piezo film including field effect transistors (FETs), operational amplifiers (Op Amps), and low-current digital logic (CMOS).

FETs lend themselves to applications of small size since they are readily available in surface mount technology. Important characteristics to consider when using FETs are switching frequency, piezo film capacitance, leakage current of the FET in the off-state, input bias resistance, and shielding from electromagnetic interference (EMI).

Figures 43 and 44 show typical FET circuit configurations for a piezo film switch. Figure 43, the common drain or source follower, applies well in applications where simple buffering is important. Here, the circuit voltage gain is approximately one.

The common source circuit in Figure 44 is suitable for low frequency applications where voltage gain is required. The gain is determined by resistances R_D and R_S . As the gain increases, frequency bandwidth decreases by a factor of one decade per 20 dB of gain.

Operational amplifiers offer a great deal of versatility for piezo film switch applications. Adaptation to a particular application is often as simple as making a few wiring changes. Important op amp circuit characteristics include input bias resistance, film switch capacitance, and EMI shielding.

The op amp circuit of Figure 45, a charge amplifier, suits applications where a detected vibration actuates the switch. It also works well in small signal applications. A charge amplifier eliminates the effects of the time constants of both the piezo film and connecting cable. The charge amplifier is a current operated circuit with zero input impedance, which results in no voltage being generated across the film. The charge amplifier quickly absorbs charges developed by the film. With no charge left on its electrodes, the film exhibits no time constant.

The capacitance of the film and connecting cable have no adverse effect on the circuit's transfer function. Thus tolerances on film size and cable length need not be exceptionally tight. The charge is

Figure 43. High frequency, low gain FET circuit interface

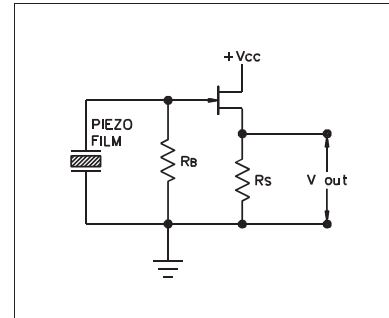


Figure 44. Low frequency, high gain FET circuit interface

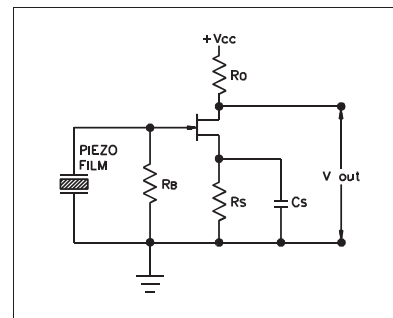
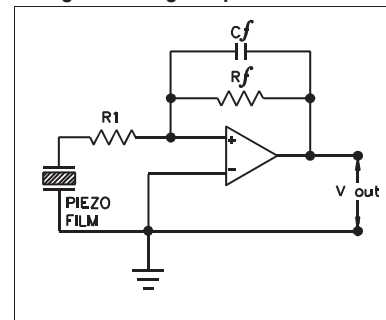


Figure 45. Op Amp Interface circuit acting as a charge amplifier

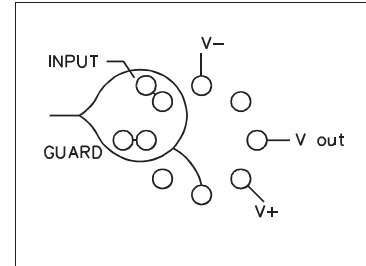


transferred from the film to the capacitor in the amplifier's feedback loop, which determines the output voltage: $V = Q/C_F$.

The charge amplifier requires an op amp having a high input resistance and low bias current. A high input resistance avoids bleed-off of the charge on the feedback capacitor, and low bias current prevents the feedback capacitor from charging and discharging at excessive rates. The layout of the charge amplifier circuit is critical. The op-amp casing must be well grounded and the inputs should be guarded and connected to the same ground as the casing.

A layout with guarded inputs is shown in Figure 46. Also, to prevent leakage noise from being amplified by the op-amp, the input cable should be terminated using a well-insulated stand-off connector.

Figure 46. Layout for guarding inputs

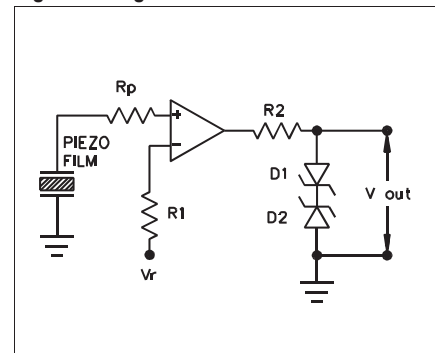


Even with the above precautions, it is likely that the output voltage will drift. To compensate for drift, a reset switch is generally designed into the circuit to manually reset the output to zero at intervals. One technique is to place a reed switch in series with a resistor, which is in parallel with the feedback capacitor C_F . Activating the reed switch closes the switch, discharging the voltage stored in the feedback capacitor.

Another method is to use a MOSFET device in which the maximum output voltage and off-gate voltage determine the minimum gate voltage of the FET. In practice, a supply voltage greater than the amplifier voltage is applied to the gate of the MOSFET, thereby lowering its drain/source resistance and creating a current path for discharge of the feedback capacitor.

The third alternative is to place a bleed resistor across the feedback. This resistor creates a time constant ($C_F R_F$), which is independent of the film capacitance and can be accurately controlled.

Figure 47. Signal level detector

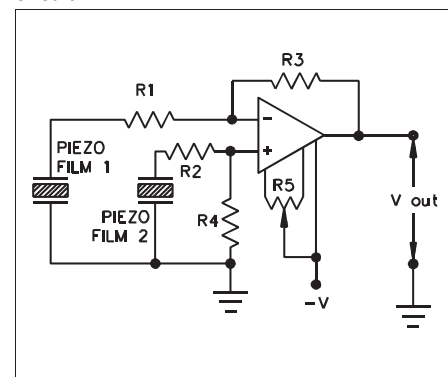


The signal level detector of Figure 47 fits applications where large signal-to-noise ratios are desirable. This circuit is perfect for detecting an impact among low-level vibrations. For situations where signal to noise ratios are low and where impacts or pressures must be discerned from background vibration, the differential amplifier circuit of Figure 48 is appropriate. This circuit consists of two sensors driving a differential amplifier.

This configuration uses a common-mode rejection concept. The two switches are mechanically coupled to cancel unwanted vibrations that stimulate both. An input or pressure on one switch but not the other, will produce an output.

CMOS logic offers a low-cost way to interface with piezo film. As mentioned earlier, low-power circuits implemented with CMOS technology are ideally suited to piezo film switches. CMOS applications for piezo film are generally for low frequency operation. Other characteristics to consider include device input leakage current and input impedance,

Figure 48. Differential Op Amp interface circuit



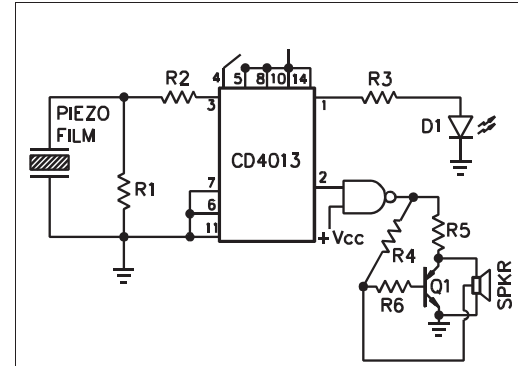
input bias resistance, and the effect of EMI. A CMOS circuit can be used, for example, in applications to sense a single impact or a single pressure.

The D-Flip Flop in Figure 49 indicates the presence of either the impact or pressure to set off an audible alarm.

The circuit in Figure 50, senses multiple impacts or pressures for counting applications.

Many different CMOS circuit configurations are possible to interface with piezo film. Common to all of them is an input bias resistor in parallel with the piezo film, and an input resistor in series with the film. The bias resistor handles leakage current and the series resistor limits current to protect against electrostatic discharge.

Figure 49. CMOS circuit for detecting a single impact

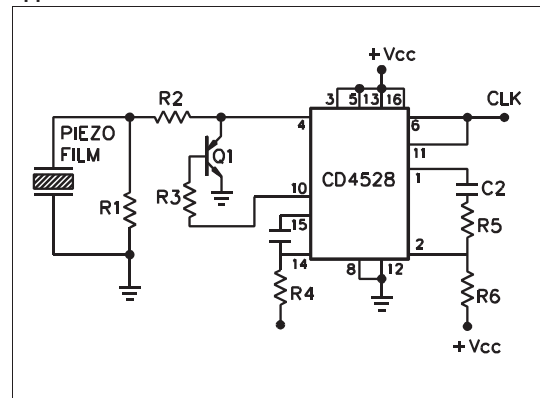


Cables

In applications where it is not possible to place the amplification circuit in close proximity to the piezo film transducer, considerable care must be exercised in selecting the connecting cable that carries the high-impedance signal.

Shielded coaxial cable, while used for noise reduction, can add problems associated with cable leakage and added capacitance. In most cases the cable's primary insulation should consist of highly resistant, non-polar plastics such as high-purity polyethylene or Teflon® (PTFE). It is equally important to make the cable as vibration-free as possible since cable movements generate noise that interferes with signal transmission.

Figure 50. CMOS interface circuit for counting applications



MANUFACTURING

Rolls of piezo film are produced in a clean room environment. The process begins with the melt extrusion of PVDF resin pellets into sheet form, followed by a stretching step that reduces the sheet to about one-fifth its extruded thickness. Stretching at temperatures well below the melting point of the polymer causes chain packing of the molecules into parallel crystal planes, called "beta phase". To obtain high levels of piezoelectric activity, the beta phase polymer is then exposed to very high electric fields to align the crystallites relative to the poling field. Copolymers of PVDF are polarizable without stretching.

Evaporatively deposited metals are typically 500 to 1000 Å in thickness, and almost any metal can be deposited. Popular metals are nickel, aluminum, copper, gold and alloys. Electrode patterns are made by sputtering through masks or by chemical etching continuous metallizations using photoresists. Resolution to 25µm line widths has been achieved. Screen printed electrodes of conductive silver ink are much thicker, about 5-10 µm, and can be applied in complex patterns to form multiple sensors on a single sheet. Foils are adhered with thin adhesive layers and capacitively coupled to the piezo film. Each electrode alternative has advantages and disadvantages.

Generally, sputtered metals are for very high resolution arrays, pyroelectric applications requiring a low thermal mass, or for inertness, as with invasive medical applications. Fully metallized sheets can be carefully cut with a razor blade without shorting across the film thickness. Screened inks are very robust and compliant, withstand very high strains (>10%), can operate at high voltages without breakdown, and are easy to pattern on a continuous basis. However, unmetallized borders are required for cutting elements out of a sheet of screen printed electrodes, since there is a high likelihood of shorting across the films thickness with the thick inks. Foils may mechanically restrict the piezo film from responding to externally applied stresses and strains in the plane of the film, but foils are useful in pure "thickness mode" operation.

After metallization, a wide variety of possible processing steps are followed to produce a packaged sensor. Generally, the piezo film is laminated in a protective carrier film, die cut to size, and packaged with lead wires or crimp connectors and, often, signal conditioning electronics. The wide range of packaged sensors, from a few square millimeters (including an ASIC chip) as a shipping damage sensor, to multiple square meter sensors for sports scoring targets suggests the versatility of this technology.

APPLICATIONS

The sensor applications described below represent a good cross-section of the products now using piezo film sensors.

Switches

The reliability of contact switches is reduced due to contaminants like moisture and dust which foul the contact points. Piezo film offers exceptional reliability as it is a monolithic structure, not susceptible to this and other conventional switch failure modes. One of the most challenging of all switch applications is found in pinball machines.

A pinball machine manufacturer uses a piezo film switch manufactured by MSI as a replacement for the momentary rollover type switch. The switch is constructed from a laminated piezo film on a spring steel beam, mounted as a cantilever to the end of a circuit board.

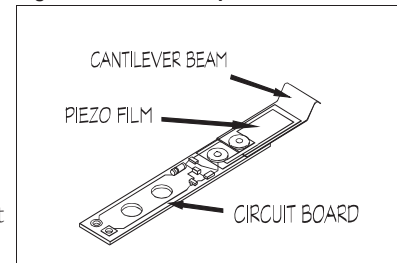
The "digital" piezo film switch features a simple MOSFET circuit that consumes no power during the normally-open state. In response to a direct contact force, the piezo film beam momentarily triggers the MOSFET. This provides a momentary "closure" for up to a 50 V maximum voltage. The output of this low profile contactless switch is well suited to logic-level switching. The unit does not exhibit the corrosion, pitting or bounce that are normally associated with contact switches.

The company has tested these switches in excess of 10 million cycles without failure. The switch solves the nagging problem of fouled contacts in pinball machines, a significant source for machine downtime and lost revenue. The simplicity of the design makes it effective in applications which include:

- ! Counter switches for assembly lines and shaft rotation
- ! Switches for automated processes
- ! Impact detection for machine dispensed products
- ! Panel switches
- ! Foot pedal switches
- ! Door closure switches

The cantilever beam that carries the piezo film can be modified to adjust switch sensitivity for high to low impact forces. Figure 51 shows the construction of the digital switch.

Figure 51. Switch for pinball machine

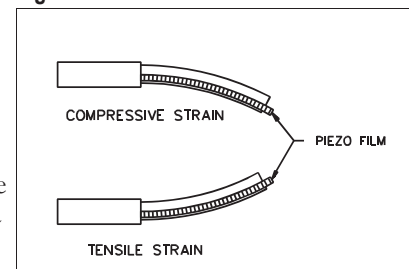


Beam Switch

Piezo film switches can be used to measure the amplitude, frequency and direction of an event and are useful in object detection and recognition, counting, wakeup switches and bidirectional encoding applications. The construction of the beam-type switch is shown in Figure 52.

Note that the piezo film element is laminated to a thicker substrate on one side, and has a much thinner laminate on the other. This moves the neutral axis of the structure out of the piezo film element, resulting in a fully tensile strain in the piezo film when deflected downward, and a fully compressive strain when deflected in the opposite direction. Were the neutral axis in the center of the piezo film element, as would be the case if the two laminae were of equal thickness, the top half of the piezo film would be oppositely strained from the bottom half under any deflection condition, and the resulting signals would therefore be canceled.

Figure 52. Beam switch



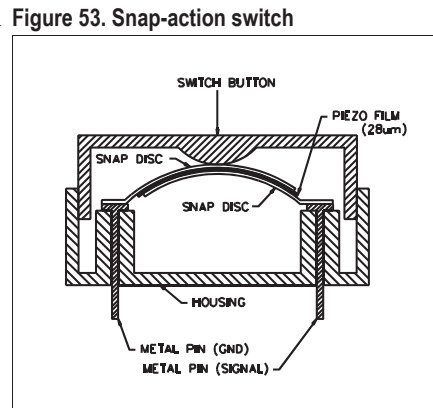
Beam switches are used in shaft rotation counters in natural gas meters and as gear tooth counters in electric utility metering. The beam switch does not require an external power source, so the gas meter is safe from spark hazard. Other examples of applications for the beam switch include a baseball target that detects ball impact, a basketball game where a hoop mounted piezo film sensor counts good baskets, switches inside of an interactive soft doll to detect a kiss to the cheek or a tickle (and the sensor is sewn into the fabric of the doll), coin sensors for vending and slot machines and as digital potentiometer for high reliability.

Snap-Action Switches

Piezoelectric materials do not have a true dc response. Very slow events, 0.0001 Hz, for example, are not normally possible to detect with piezoelectric film.

In switch applications where dc response is required, piezo film in combination with a snap dome provides a high voltage pulse.

When the snap device actuates, the film is rapidly strained, typically generating a 10 volt pulse into a one megohm circuit as shown in Figure 53. This concept is especially well suited for wakeup switches, where an electronic device can be dormant for long periods without power consumption until the snap action device is actuated. The piezoelectric pulse turns on the electronics. Battery operated parking meters, where battery life is very critical, are an example of a piezo snap action switch application. A thermal snap action device also employs this principle.



Impact Sensors

Impact Printers

High speed impact printers require very accurate print head timing. Impact must occur the instant that a high speed revolving steel band, embossed with print characters, is properly positioned in front of the print hammer. Any advance or delay in energizing the print hammer will result in an offset print of the desired character.

Piezo film sensor strips, built into the printer platen, monitor the impact timing and force of the bank of print heads, and transmit the information to the controller. Automatic adjustment is made in the actuator timing to accommodate any minor change in print head timing. The very high speed of the embossed steel ribbon, about 300 inches per second, requires a very fast switch response. Alternative impact switches are quickly destroyed by the large impact forces of the print head. Piezo film switches have been in use in this application for more than five years without failure.

Sports Scoring

Piezo film sensors can be used to measure impact time, location (accuracy) and force. These parameters are desirable in several sports scoring applications. The energy of a 90 mph pitch has instantaneous power of about 50,000 watts! The great challenge in this application is target ruggedness without the introduction of severe bounceback into the design.

A second sports scoring application is electronic dartboards, where piezo film monitors the many impact zones in the game. Scoring is electronically recorded.

Musical Instruments

The popularity of electronics for musical instruments presents a special problem in drums and pianos. The very high dynamic range and frequency response requirements for drum triggers and piano keyboards are met by piezo film impact elements. Laminates of piezo film are incorporated in foot pedal switches for bass drums, and triggers for snares and tom-toms. Piezo film impact switches are force sensitive, faithfully duplicating the effort of the drummer or pianist. In electronic pianos, the piezo film switches respond with a dynamic range and time constant that is remarkably similar to a piano key stroke.

Traffic Sensors

The U.S. Government is actively studying "smart highways" as an alternative to major new highway construction. The idea is that existing highways can accommodate greater vehicle densities if electronically managed. In addition to conventional traffic monitoring for highway studies and enforcement, the Intelligent Vehicle/Highway System (IVHS) programs create the need for new classes of "smart highway" high speed sensors to count and classify vehicles, provide lane control, and to monitor weight and speed. IVHS also requires "smart car" sensors, and advanced vehicle surveillance, communications, and software.

Futuristic programs like the IVHS, and more contemporary projects like the Strategic Highway Research Program (SHRP), require traffic data collection to provide the necessary information required by the Federal Highway Administration on highway structures. Recent advancements in signal processing open the door to greatly improved real-time vehicle data analysis, provided that inexpensive reliable sensor technologies are developed.

Pneumatic road tubing has long been the workhorse of traffic data collection. Road tubes provide a pneumatic pulse to a piezoelectric membrane, which triggers nearby electronics when an axle is detected.

The evaluation of alternative sensor technologies has shown piezo cable provides the necessary sensitivity, linearity, noise immunity and environmental stability for high traffic interstate vehicle classification and weight-in-motion systems. Piezo cable BL sensors are used for traffic data collection from Saskatchewan to Florida.

Piezo cable traffic sensor constructions are shown in Figures 54 and 55. There are two basic categories of traffic sensor ... permanent and temporary. Generally,

permanent sensors are mounted in the road with the top of the sensor flush to the highway surface, while temporary units are adhesively applied to the road surface for shorter monitoring periods. Permanent sensors, used for toll booths and interstate highway data collection, are flush mounted to a road surface and must withstand the rigors of years of high density traffic, snow plows, salt, sand, water and dragging mufflers.

Figure 54. Permanent In-The-Road Traffic Sensor

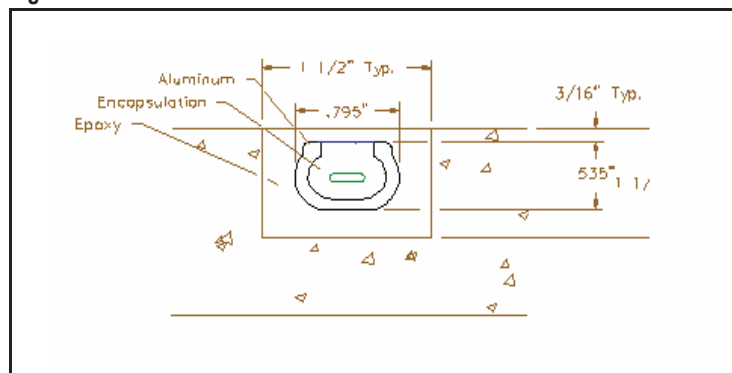
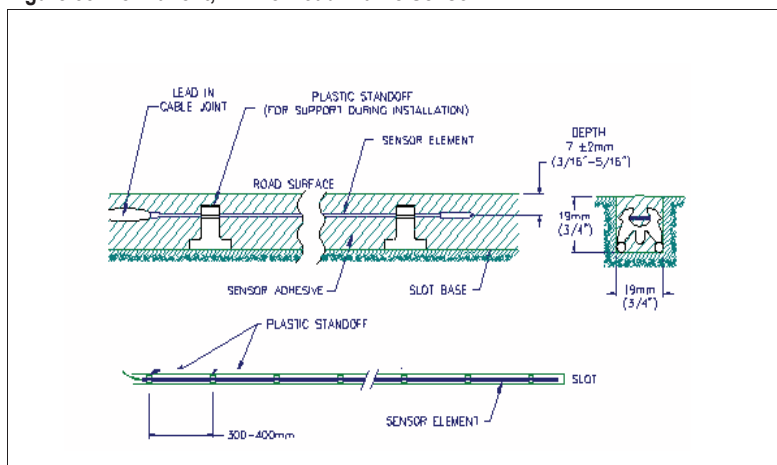
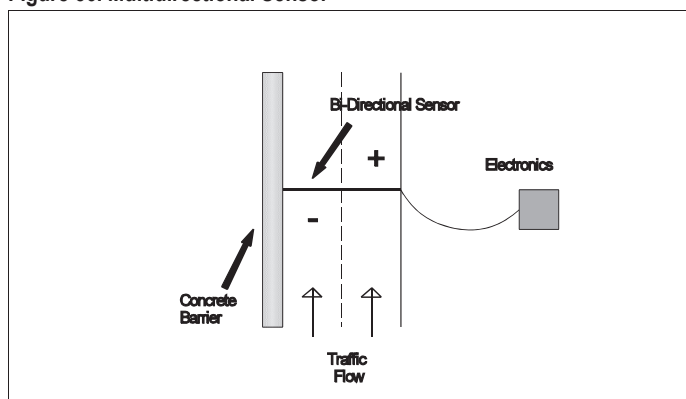


Figure 55. Permanent, In-The-Road Traffic Sensor



"New Jersey barriers", the modern concrete barriers that separate opposing lanes on highways, introduce problems for multilane sensing of four lane highways. Piezo cable can solve this problem with a single sensor that has opposite polarities corresponding to each lane. Vehicles crossing the near lane produce a signal of opposite sign from vehicles in the far lane (Figure 56). This ability to provide lane activity in a single sensor is a significant development.

Figure 56. Multidirectional Sensor



Traffic sensors can monitor vehicle speed, count axles, weigh vehicles, provide direction, and vehicle classification. Recently, these sensors have also proven valuable on airport taxiways. From the output, one can discern the ground speed of an aircraft (time lag between two sensors), its direction, weight (fueled), number of axles, and the span of the aircraft (determined from the speed and the known fixed distance between sensors). This information can be used to classify the aircraft and provides taxiway traffic control and safety information at airports.

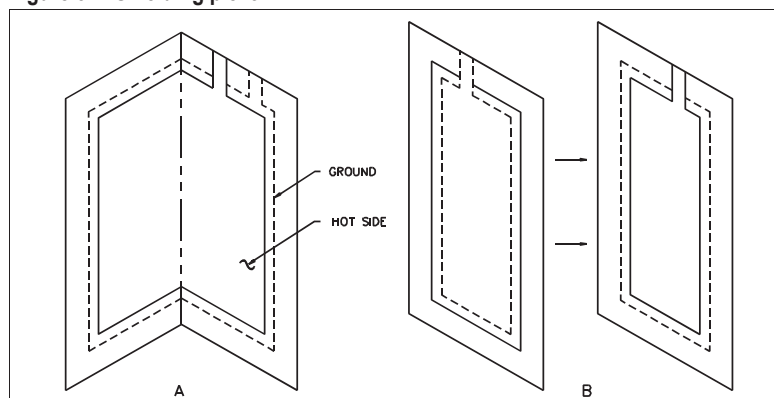
VIBRATION SENSING

One of the first applications for piezo film was as an acoustic pickup for a violin. Later, piezo film was introduced for a line of acoustic guitars as a saddle-mounted bridge pickup, mounted in the bridge. The very high fidelity of the pickup led the way to a family of vibration sensing and accelerometer applications.

Music Pickups

Piezo film is used today in several guitar pickup designs; one is a thick film, compressive (under the saddle) design; another is a low cost accelerometer, while another is an after market pickup design that is taped to the instrument. Because of the low Q of the material, these transducers do not have the self-resonance of hard ceramic pickups. Shielding can be achieved by a foldover design as shown in Figure 57. The hot side is the slightly narrower electrode on the inside of the fold. The foldover technique provides a more sensitive pickup than alternative shielding methods because the shield is formed by piezoelectric material. Conventional shielding laminates can be easily fabricated by a multilayer laminate of piezo film, adhesive and shielding foil.

Figure 57. Shielding piezo film



Machine Monitoring

The fidelity of a shielded piezo film sensor in musical instruments led to the development of vibration sensors for machines. In its simplest mode, piezo film vibration sensors behave essentially like dynamic strain gages. The film does not require an external power source, yet typically generates signals greater than strain gages **after** amplification. A typical piezo film sensor produces four orders of magnitude higher voltage signal than a foil-type strain gage, and two orders higher than semiconductor types. The frequency response of the piezo film strain gage is also superior.

The extreme sensitivity is due to the form of the piezo film material. The low thickness of the film results in a very small cross sectional area. Thus very small longitudinal forces create very large stresses within the material.

Piezo film sensors can be affixed to a vibrating surface and monitor the amplitude and frequency of the vibrating structure. The sensors can cover larger areas than normal strain gages so any direct comparisons should be performed in **uniform** strain fields for meaningful results. Obviously, point-type transducers may be used where required, although the low capacitance of the small sensor area will require additional consideration. Operation down to fractions of Hz can be achieved by either conventional charge amplifiers or, since signal levels are relatively high, simple high impedance FET buffer circuits.

Bearing Wear Sensors

A shielded piezo film sensor has been used to monitor bearings for wear and evidence of spall. The sensors are permanently affixed to the outer surface of the bearing race with epoxy. The low mass and thin profile allow its use as a built-in nondestructive testing sensor, rather than the time consuming use of accelerometers for periodic fault-condition checks.

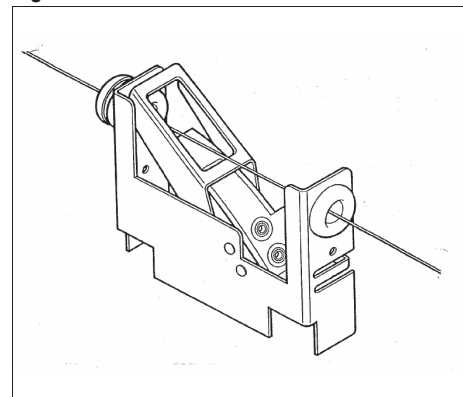
Fan Flow Sensor

A laminated beam type sensor is used in ducted airflow as a centrifugal fan failure sensor. The presence of air flow is detected by the vibrations in the sensor caused by the turbulence of air flow at about 100 Hz. The absence of this signal is used for trigger electronics. The sensor and switching electronics are based around a TL084 quad bi-FET op amp, with typical input signals of 80 mV. Sensor reliability is the key feature. Since the sensor is subjected to virtually no operating stresses, it has an indefinite working life.

Thread Break Sensor

Textile plants require the continuous monitoring of often thousands of lines of thread for breakage. An undetected break event can require that a large volume of material be discarded, as the labor costs to recover the material exceed the manufacturing cost. Drop switches, where switch contact closure occurs when the thread breaks, are very unreliable. Lint fouls the contact points, resulting in no output signal. A piezoelectric film vibration sensor, mounted to a thin steel beam, monitors the acoustic signal caused by the abrasion of the thread running across the beam, analogous to a violin string (Figure 58). The absence of the vibration instantly triggers the machinery to stop. Thread tension and defects in man-made fibers are also monitored with piezo film textile sensors.

Figure 58. Threadbreak sensor



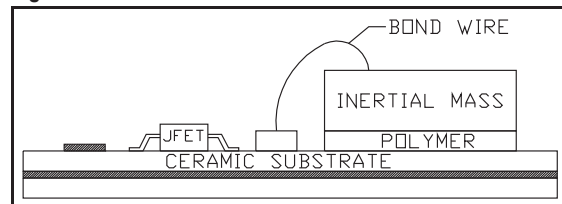
Vending Sensors

Shielded dynamic strain gages of piezo film are affixed to the underside of a vending product delivery tray to verify that product was properly vended. The absence of the impact induced vibration triggers an "Out of Order" warning. In a second application, slot machine coin counting is provided by a piezo film element. The sensors confirm delivery of coins won, discouraging gamblers from falsely claiming equipment defects. A ticket dispensing machine counts tickets delivered with a piezo beam design. Coin sensors also trigger or wake-up vending machines and coin changer electronics to verify coin authenticity.

ACCELEROMETERS

A logical outgrowth of the many vibration sensor applications of MSI's piezoelectric technology are accelerometers. These accelerometer designs are based on more traditional piezoelectric ceramic, as well as piezoelectric polymer materials. The choice of base materials allows the product to be tailored for specific applications. Table 6 lists the key specifications for the MSI Accelerometer product family.

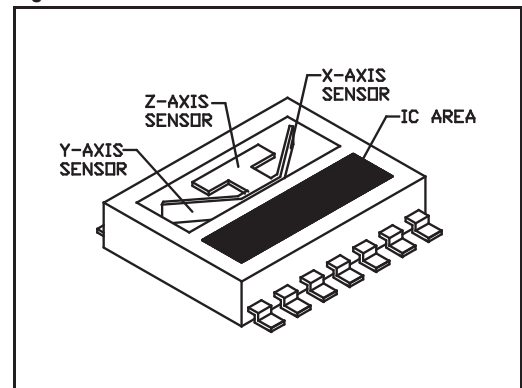
Figure 59. ACH-01-XX internal view



Like more conventional sensors, these accelerometers are configured as either compression-design type or beam-design type. Compression-design accelerometers typically have higher resonant frequencies providing wide useful frequency ranges. An internal view of MSI's ACH-01-XX compression-design accelerometer is shown in Figure 59.

Beam-design accelerometers tend to have lower resonant frequencies and useful frequency ranges. Beam-design accelerometers also have another very interesting feature: They can be oriented to sense acceleration in multiple-axes with one monolithic sensing element using MSI's patented "Origami" beam technology ("Origami" is the Japanese word for the art of paper folding). An internal view of the ACH-04-08-05 three axis beam-design accelerometer, with its origami sensing element, is shown in Figure 60.

Figure 60. ACH-04-08-05 internal view



To reduce system costs as well as simplify use, all of MSI's accelerometers include buffers. JFET biasing and signal processing is implemented external to the device.

The ACH-01 family of products is typically used in applications which require broad frequency capability, high sensitivity, low noise, and low cost. Such applications include: speaker feedback and control systems, automotive anti-theft systems, acoustic pick-ups, machine-health and pump and centrifuge monitoring systems, and medical body motion monitoring.

The ACH-04-08 product family is used in a very broad range of applications such as speaker feedback and control systems, appliance fault monitoring, virtual reality systems, automotive systems, medical body motion monitoring, shipment damage and material-handling monitoring systems, vibration switches and earthquake shut-off switches. OEM applications that require acceleration or vibration measurements in more than one axis are perfect for the ACH-04-08-05.

MSI is constantly developing and upgrading its accelerometer product line. Please contact MSI for further details on these products or on customizing one of our other products.

Table 6. Accelerometer Family

Production Qualified Accelerometers				
		ACH-01-XX	ACH-04-08-05	ACH-04-08-09
Key Features		-Wide Frequency Range -Wide Dynamic Range -High Sensitivity -Low Noise -JFET Buffer	-Low Frequency Operation -3 Simultaneous Analog Outputs -Low current generation - JFET Buffer	-Low Frequency Operation -Low Current Operation - JFET Buffer
Sensitive Axes	X-Axis	---	X	--
	Y-Axis	---	X	X
	Z-Axis	X	X	--
Sensitivity (Nominal)		10 mV/g	1.8 mV/g	6mV/g
Frequency Range (± 3 dB)		1.0 Hz-20 kHz	0.5 Hz to 4 kHz	0.2 to 1500 Hz
Dynamic Range		± 250 g	± 250 g	± 40 g
Resolution (@ 100 Hz)		$40\mu\text{g}/\sqrt{\text{Hz}}$	$200\mu\text{g}/\sqrt{\text{Hz}}$	$40\mu\text{g}/\sqrt{\text{Hz}}$
Resonant Frequency		>35 kHz	9.2 kHz	3.4 kHz
Resonant Q (Hz/Hz)		30	10	10
Transverse Sensitivity		5%	15%	<20%
Linearity		0.1%	0.1%	0.1%
Operating Temperature		-40 °C to +85 °C	-40 °C to +85 °C	-40 °C to +85 °C
Storage Temperature		-40 °C to +85 °C	-40 °C to +85 °C	-40 °C to +85 °C
Maximum Shock		1000 g	1000 g	1000 g
Supply Voltage		3 V to 40 V	3 V to 40 V	3 V to 28 V
Supply Current		2 μ A Typical	6 μ A Typical	2 μ A Typical
Weight		3 grams	0.35 grams	0.35 grams
Size (mm)		13 x 19 x 6	11 x 10 x 1.8	11 x 10 x 1.8
Mounting Method		Adhesive	Hand Solder to PCB	Hand Solder to PCB

Table 7. Accelerometer Applications

Industry	ACCELEROMETER PRODUCTS			
	Application	ACH-01-XX	ACH-04-08-05	ACH-04-08-09
<i>Aerospace & Defense Electronics</i>	Anti-Tamper Sensors	X	X	X
	Surveillance	X		
	Modal Analysis	X	X	
<i>Automotive</i>	Antitheft	X	X	X
	Skid/Rain Sensing	X		
<i>Computers & Peripherals</i>	Computer Mouse Sensor		X	
	Virtual Reality Sensor		X	
<i>Household Appliances</i>	Out-of-Balance Sensor	X	X	X
	Spray Arm Jam Sensor		X	X
	μ wave Acoustic Sensor	X		
<i>Consumer Electronics</i>	Speaker Feedback	X		
	Acoustic Pick-ups	X		
	Security		X	X
<i>Industrial</i>	Machine Health Monitor	X	X	
	Bearing Monitor	X	X	
<i>Instruments & Measuring Equipment</i>	Active Vibration Damping	X	X	X
	Predictive Maintenance	X	X	X
<i>Medical</i>	Motion Sensor	X	X	
	Pacemaker	Consult Factory for further Information		
<i>Power & Utilities</i>	Earthquake Shut-Off		X	X
	Machine Monitoring	X	X	
<i>Transport & Material Handling</i>	Shipment Monitoring	X	X	
	Railroad Systems	X	X	X

ULTRASOUND APPLICATIONS

The wide frequency response and physical attributes of its polymeric construction makes piezo film a material of choice in certain medical probes and in nondestructive testing applications. Additionally, the film sensors are found in applications in ultrasonic based sensing devices, like air-ranging ultrasound for distance measurement, in fluid level sensors, and in-flow measurement instruments using doppler shift of sound velocity perturbations which are proportional to fluid flow.

Medical Imaging Ultrasound

Piezoelectric ceramic materials are used in medical imaging transducers because of their high sensitivity and broad bandwidth. The d_{33} constant, strain developed for an applied voltage, is about an order of magnitude higher for piezo ceramics than for piezo polymer. A disadvantage of piezo ceramic is its high acoustic impedance, about 30 MRayls (1 MRayl = 10^6 kg/m²s) in contrast to about 1.5 MRayls for body tissue. This impedance mismatch can be compensated by quarter wavelength matching layers, but these can degrade the ultrasonic pulse due to adhesive layers and construction methods. The acoustic impedance of piezo film is about 4 MRayls, a much better match. Additionally, in higher frequency applications requiring very thin piezo elements, ceramics are too fragile, and cannot be shaped to desired geometries.

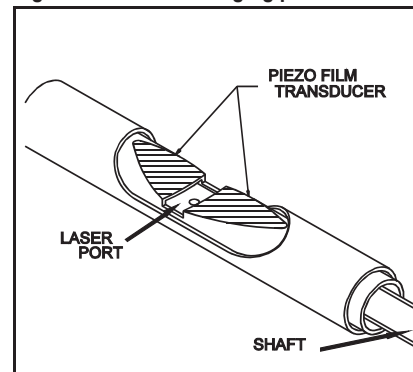
Invasive imaging requires lower powered devices than external probes. Resolution of the image is considerably improved at the higher frequencies of invasive catheters. A medical imaging company has developed an invasive imaging probe with piezo film for a therapeutic laser prostate catheter (Figure 61). The piezo film sensor is about 30 microns thick, and is located near the catheter tip. The unit operates at frequencies of 7 MHz and higher.

Steered *in-vivo* phased-array images using piezo polymer film have been produced for the first time by researchers at Duke University. A 32 element array of 11 mm x .56 mm elements was fabricated and tested with a well matched circuit designed to optimize the transducer. The result was 28 dB lower sensitivity than PZT transducers at 2.5 MHz operating frequency.

However, the piezo film array had improved axial resolution, better angular response (6 dB pulse-echo response at 30 degrees), and a low interelement cross-coupling of -35 dB. It is exceptionally difficult to diamond blade saw PZT ceramic into these small elements; while, for piezo film, complex patterns are readily etched into the surface gold electrode. PZT must be diced due to the severe interelement coupling problem. Duke University researchers plan to improve the polymer probe by expanding the number of array elements to 128.

Very high resolution arrays have been traditionally formed by etching an electrode pattern on the surface of a piezo film. Newer techniques include deposition of the copolymer directly onto silicon wafers. The wafers are etched to minimize interelement coupling, then the copolymer is applied by spin-coating, followed by poling. Then a top ground electrode is applied and inter-connections made. This advance results in a very high resolution imaging. Capacitively coupling copolymer film to a dense array of conductive traces on a PCB has achieved remarkable performance as a Tx/Rx array.

Figure 61. Invasive imaging probe



NonDestructive Testing (NDT)

Advanced composite materials are very desirable as structural members. Light weight, high strength, corrosion resistance, and non-magnetic are among the advantages for these materials. The need for very routine nondestructive testing of such structures to prevent catastrophic failure due to delamination, is one of technology's greatest concerns. Flexible sheets of piezoelectric polymer transducer arrays, acoustically well matched to the composites, are desirable for use for non-destructive testing. One example is as an NDT array for testing rocket motor housings prior to launch. These arrays can be applied to the surfaces of composite fuel housings, and each element sequentially activated to provide a pulse-echo response. An array element size of about 0.5 to 1 square inch is sufficient for this application, as well as most large area NDT. Center frequencies of 3-10 MHz and -6 dB fractional bandwidths exceeding 100% are typical with such transducer arrays.

Systems and Instrumentation, Ltd. personnel use piezo film for NDT of aerospace engine parts. NDT transducers capable of detecting flaws down to 1/64th inch are now required. Further, the frequency response range of these new materials are broader than the bandwidth of conventional transducers. S&I, Ltd. find that a single broadband transducer covers the bandwidth of interest. Their transducers are also used in near-surface NDT applications, where high resolution and short pulse duration are required. Defects of 0.8 mm in size, lying within 1 mm of the surface, have been detected with the S&I probes.

Critical points or inaccessible test areas within a composite structure, like support strut mounts, where delamination or other damage is especially likely, can have custom fabricated NDT arrays permanently affixed for in-service testing and monitoring. It is possible to achieve uniformity of ± 1 dB between the elements of a multi-element transducer array. Special shaped transducers, providing special focal characteristics, have also been built with these polymer transducers.

Acoustic Emission

Acoustic emission of materials including fiber-reinforced composites, aluminum, steel and glass can be performed with contact microphones of piezo film, or, as with NDT, by large area arrays. These arrays can continuously monitor structures for 0.1 - 1.0 MHz acoustic emission, the precursor to structural failure. Piezo film, being broad band, responds well at these frequencies. This capability is especially necessary for critical application like tank rail cars carrying toxic products, underground fuel storage tanks, nuclear plants, etc.

Fluid Level Sensor

There are a variety of fluid level sensing transducer technologies available to the designer. A float arm, attached to a sliding potentiometric device is still widely used in automobiles. Ultrasonic pulse-echo devices that measure the distance from a fixed transducer to the fluid surface from above through air, or from below through the fluid, are popular. Newer capacitance types, where the fluid becomes the dielectric, are also used. Each of these technologies represent tradeoffs in system cost, performance and reliability. A new ultrasound level sensor, in development by MSI, holds promise as a digital, solid state ultrasonic level sensor.

The novel construction is a level sensor with ultrasonic through-transmission with multiple transmitters and a single, common receiver. The sensor is fabricated by attaching an unmetallized strip of piezo film to a printed circuit board containing electrode patterns, conductors, and interconnections to circuitry on the opposite side of the board. The electrode patterns are capacitively coupled to the piezo film layer, becoming the multiple transmitter elements. A second conductor bar, parallel to the patterned elements becomes the common receiver.

The presence of fluid couples the transmit signal to the receiver to a much greater extent (60dB) than when the ultrasound energy is coupled by the air above the fluid. The excitation signal for the transmitter is a 1.1 MHz sine wave tone burst with an amplitude of 20 volts peak to peak. The required circuitry consists of a high frequency oscillator and clock, an array of analog switches, a single receiver amplifier with input gate, and a threshold detector. These electronics can be reduced to the chip level, and are incorporated on the backside of the circuit board.

Resolution of the level sensor is determined by the resolution of the patterned transmit electrodes on the circuit board. Parallel elements of 2 mm width and 0.5 mm spacing between elements is a representative capability. The ground electrode for the transmitters is a fully gold metallized surface on the fluid side of the piezo film transmitter array. The receiver is formed by the same piece of piezo film, capacitively coupled to the signal electrode which is a separate conductor trace on the printed circuit board (PCB). Again, the ground is the backside electrode on the film.

The new level sensor has several unique advantages. The spacing between transmitter elements need not be uniform. For tanks that do not have a uniform volume throughout the tank height, a simple PCB layout can linearize the nonlinear tank volume by setting the transmitter element spacing accordingly. The output of the device is digital—no expensive A/D conversion is required. The level sensor is small in width, less than 1 inch, so it can be inserted into a small diameter tube. The tube confines the motion of the fluid, reducing large swings in fluid height readings caused by motion, as with an automobile fuel tank during cornering. Reliability is greatly improved. The level sensor is self diagnostic to the extent that the transmitter/receiver pair must be operational to deliver a meaningful signal. The absence of the signal indicates a fault condition. For a detailed discussion on Ultrasonic Ink Level Sensing, see Appendix C.

Air Ranging Ultrasound

Ultrasonic devices used in pulse-echo modes are used in robotics, vehicle safety and control system, object recognition systems and other remote distance measurement devices. The sensors provide high resolution in the targeted direction, and can be used to measure the elapsed time from transmit to receive to determine the distance to an object. Unlike piezo ceramic and electrostatic devices, piezo film can deliver a very short pulse (due to its low Q), allowing the same device to be used as both transmitter and receiver, even in the near field of the transducer.

Multiple piezo film elements can be easily fabricated, as shown in Figure 62. The geometries of these cylindrically shaped elements (length, radius of curvature, number of elements) can be designed to control the directivity pattern and acoustic properties. Transducers with operating frequencies from 40-200 KHz have been made. Average values of transducer sensitivity are 0.1-1 mV/Pa in the receive mode (noise was < 1 μ V) and 15-75 mPa/Vcm² in the transmit mode for 1 m of distance. The minimum distances measured in pulse-echo mode was 30 mm. Distances to 15 meters have been measured with a main beam width of less than 10 degrees, and maximum side lobe amplitudes that are 12 dB down at 60 KHz. Examples of directivity patterns for single and multiple element transducers are shown in Figure 63. Multiple

Figure 62. Air ranging ultrasound transducers

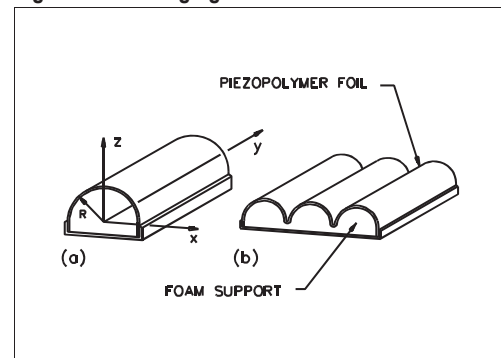
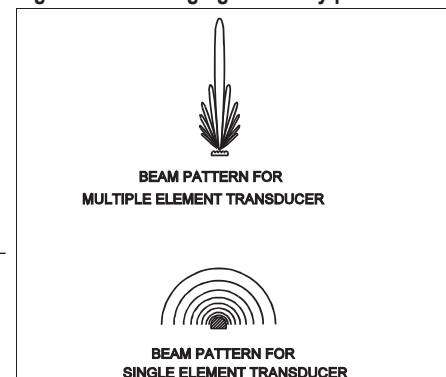


Figure 63. Air-ranging directivity patterns



elements can also be used for scanning of objects without physically moving the transducer. Each element within the transducer can be activated sequentially, as with ultrasonic arrays.

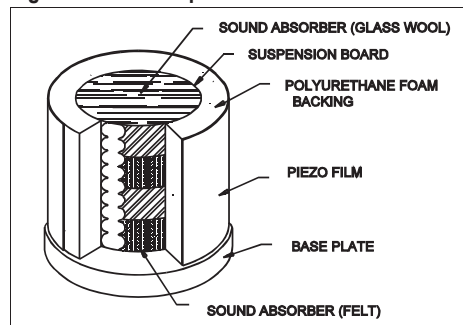
AUDIO

Speakers

One of the earliest applications for piezo film was in stereo tweeters (Figure 64) and headset speakers developed by Pioneer Electronics. There is strong renewed interest in these applications as a result of the improvements in the reliability of the electrodes and lead attachment and packaging techniques. Gallo Acoustics has developed a high fidelity omnidirectional tweeter using a cylinder of 52 μm thick piezo film. The tweeter rolls off at frequencies below 2 KHz, and features:

- 330 degrees of horizontal dispersion at high frequencies, which is as much as ten times the dispersion of conventional tweeters,
- very wide dynamic range,
- linear frequency response,
- very fast impulse response, faithfully reproducing the highest frequencies.

Figure 64. Audio speaker



Novelty audio speakers have also been developed. These devices make use of the thin, light weight, conformal nature of the piezo film. Examples include speakers for inflatables (like balloons and air inflatable toys), speakers in apparel (including headgear) and paper thin speakers for magazine advertising, greeting cards and posters.

Microphones

A diaphragm of piezo film, affixed in a retaining ring or mounted over a hole in a plate, makes an excellent microphone. Vacuum formed domes on a support can be introduced into the design to take the membrane slightly out of its neutral axis with a foam backing, a small post, bar or structure to give the film membrane a slight radius of curvature. A self-supporting, cylindrically curved film also achieves the mechanical bias. A typical radius of curvature for piezo film microphones which optimizes sensitivity and electroacoustic efficiency is $R_0 = 25 \text{ mm}$, similar to that of an electrostatic microphone construction.

Sennheiser reports a frequency response for a typical foam backed piezo film microphone of 25 mm diameter, having $R_0 = 25 \text{ mm}$. The free field sensitivity of the device measured at 1 KHz, for sound pressure incident on the membrane perpendicularly, was -58 dB re 1 V/Pa. Harmonic distortion approaches 1% only at sound pressure levels exceeding 122 dB, and are not significantly higher for the range of higher frequencies.

Microphones built with piezo film are low cost, but more importantly, are inherently immune to moisture, unlike electrostatic types.

Electrostatics dominate the market due to the low cost that has been achieved through very high volume manufacturing. Nonetheless, piezo film microphones are finding application in designs where

environmental stability is critical. Waterproof microphones are being supplied for divers, withstanding total immersion in salt water without damage.

SONAR

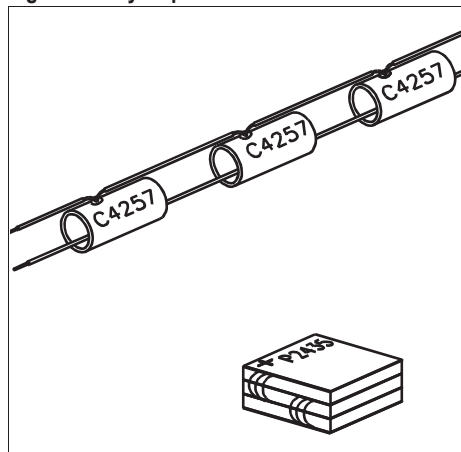
Piezo ceramic materials have been used in hydrophones for SONAR since the 1940's. Ceramics have many desirable features as hydrophones including high hydrostatic sensitivity, high capacitance, ability to fabricate into shapes, and availability in thick cross-section (sensitivity is proportional to piezo element thickness). However, the weight of ceramic is much greater than water, so buoyancy must often be added to structures as compensation.

The current trend in SONAR is toward higher numbers of hydrophone sensors per vehicle. Longer towed arrays and larger and more numerous hull mounted hydrophone panels are being employed by the fleet. This trend requires the advancement of much lower weight hydrophones making piezo film an excellent choice for these applications.

New piezo polymer hydrophones of thick film (1200 μ m) have been made into sheet and cylinder form (Figure 65).

The cylinders, about 2-3 cm in length and <1 cm in diameter, are strung together into a continuous hydrophone line array. The hydrophones are very light weight (0.60 g/cm³), yet have excellent hydrostatic sensitivity.

Figure 65. Hydrophone



Piezo polymer hull mounted hydrophones in panels, of about one square meter in area and greater than 1000 μ m in piezo polymer thickness, are used by the British and French Navies for submarines. Called "Flank Arrays", these units work in combination with towed arrays in advanced SONAR systems. Piezo film sheets are not fragile, conform to the radius of curvature of the hull, and are low cost. Typical sensitivities for the piezo film thick sheet hydrophones are similar to ceramic type (-192 to -195 dB re 1 V/ μ Pa). The lower capacitance of the piezo film hydrophones is compensated by the large area and number of parallel hydrophones used.

New hydrophone technologies, including underwater ultrasound imaging for divers and related applications are also in development.

FUTURE APPLICATIONS

Piezo film research is underway into an exciting new array of applications. A sampling of this R&D activity is highlighted below:

Active Vibration Damping

Piezo film sensor and actuator pairs are in development for active vibration damping. In this application, a piece of piezo film is employed as a strain gauge to detect vibration and another piece of piezo film is employed as an actuator to dampen the vibrational noise by applying a 180 degree phase-

shifted signal. Wide coverage of piezo film, with distributed electrode patterns, can be used to create individual sensor/actuator pairs. Critical damping has been achieved with this piezo film laminate by researchers at Massachusetts Institute of Technology and elsewhere.

The applications for this technology include the reduction in harmful vibrations in space-based structures, fuselage for aircraft to cancel engine noise, quiet cars, quiet appliances, and a wide range of other possibilities.

Sensors on Silicon

Piezoelectric polymers can be solution cast (spin coated) onto silicon, polarized in place, metallized in pattern arrays, and interconnected with the integrated circuits on the chip. Minor modifications to wafer processing make the silicon suitable for the piezo Polymer Sensor-On-Silicon (PSOS). Early work in PSOS technology at Stanford University was frustrated by the need to adhesively bond thin sheet piezo film onto silicon. The adhesive layer was difficult to apply to the silicon, introducing air bubbles, wrinkles, nonuniform thickness layers, etc. The advent of new piezo polymer processing that eliminates the adhesive makes the PSOS technology practical. R&D work at MSI and elsewhere includes the development of pyroelectric arrays for infrared cameras, fingerprint readers, and thermal imaging devices, ultrasound arrays for high resolution NDT and invasive medical imaging, and integrated accelerometers for microminiature components.

Smart Skin

Piezoelectric film can both sense surface energy and can microdeflect the surface. These capabilities may be of interest in active surfaces for sound cancellation, extension of laminar to turbulent flow boundaries, etc. Considerable work is under way at university and government laboratories in these and related applications.

Appendix A – Applications of Piezo Film

APPLICATIONS OF PIEZO FILM	
<u>COMPUTER INPUT/OUTPUT</u>	
<i>Keypad arrays</i> Digitizer Air Mouse Joystick Pen (Signature verification; Handwriting Recognition) <i>Printers</i> Impact Flight Time Ink Drop Generation and Detection Toner and Ink Jet Level Toner Activation <i>Business Equipment</i> Antitamper for ATM Machine Coin Counters Copiers Switches Paper Path Switches Toner Level and Activation <i>Disc Drives</i> Shock Sensing Accelerometers	<i>Physical Security & Energy Management</i> Glass Break Detectors Floor/Mat Sensor Penetration Detection Contact Microphone Piezo Cable Perimeter Protection Pyrometer/Flame Sensor <i>Flow/Level</i> Vortex Fluidic Oscillator Air Flow Doppler Ultrasound Solid State Fluid Level Laminar/Turbulent Boundary Layer Fan Failure
<u>INDUSTRIAL</u>	<u>INSTRUMENTATION</u>
<i>Switches</i> Solid State Momentary Snap Action Cantilever Beam Keypad Vandal-Proof Intrinsically Safe CMOS Wake-up Low-Deflection Singing Switch (a.c. switch) Coin Counter Acoustic Switch Shaft Rotation Counter <i>Robotics</i> Tactile Sensor Micropositioner Safety Mats & Switches Bumper Impact	<i>Machine Health Monitor</i> Accelerometers Contact Microphones Hi-Strain Dynamic Strain Gages <i>Weather Sensors</i> Rain Intensity Hail Detection Wind Velocity <i>Active Vibration Damping</i> Strain Gages Sensor Arrays Actuator Arrays <i>Non Destructive Engineering</i> Flexible Contact NDT Probes NDT Arrays Acoustic Emission Sensors <i>Air Ranging Ultrasound</i> Safety Distance <i>Adaptive Optics</i> Fiber Optic Shutters/Modulators Deformable Mirrors Laser Scanners <i>Oil Exploration</i> Hydrophones Seismic Geophones

<u>MEDICAL</u> <i>Diagnostics</i> - Apnea Monitor - Ambulatory/Gait Monitors - Blood Pressure Cuff - Pulse Counter - Stethoscope - Sleep Disorder Monitors - Respiratory Air Flow - Isokinetics - Patient Bed Monitor <i>Ultrasound</i> - Near Field Imaging - Prostate - Transdermal - Transluminal - Coronary Arterial - Breast - Lithotripter - Hydrophone Calibration Probes <i>Handicapped Aides</i> - Switches - Braille Reader - Hearing Aid - Speech Intensification <i>Implantables</i> - Pacemaker Activity Monitor - Implantable Switch - Vascular Graft Monitor - Micropower Source <i>Instrumentation</i> - Intravenous Drop Counter - IV Air Bubble Detection - Laser Switch/Modulator <u>AUTOMOTIVE</u> - Accelerometers - Occupancy Seat Sensor <i>Switches</i> - Passenger Compartment Switches - Horn Switch - Control Panel <i>Fuel Level, Tire Rotation, Security</i> - Keyless Entry - Motion (Theft) Sensor	<u>CONSUMER</u> <i>Musical Instruments</i> - Piano Keys - Pick-up - Drum Trigger <i>Sports Equipment</i> - Target Location - Reaction Time - Foul Line - Force (Karate, Impact) - Sweet Spot <i>Toys/Games</i> - Switches - Proximity (Air Ranging Ultrasound, Pyro) - Novelty Speakers (Microphones) - Target Scoring <i>Audio</i> - Tweeter - Balloon Speakers - Novelty Speakers (Visor, Poster) - Microphone - Speaker distortion Feedback Accelerometer <i>Appliance</i> - Washer Imbalance - Vacuum Soil Sensing - Dishwasher Spray Arm - Level Sensing Switches <u>MILITARY/GOVERNMENT</u> <i>Hydrophones</i> - Towed Cable Array - Hull Mounted Arrays - Sonobuoys - Active Noise Suppression <i>Ballistics</i> - Safety and Arming Fuses - Shock Wave Gages - Seismic Accelerometers <i>Physical Security</i> - Perimeter Security Cable (Buried or Fence) - Seismic/Geophones - Covert Microphones <i>Traffic Sensors</i> - Vehicle Classification - Weight-In-Motion - Speed, Red Light Enforcement - Lane Designation - Toll Booth
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Appendix B - INDEX OF PIEZO FILM ARTICLES

[NOTE: THIS LIST IS ONLY A REFERENCE. COPIES OF THESE ARTICLES ARE NOT SUPPLIED BY MSI, INC.]

CODE	AUTHOR AND REFERENCE	TITLE
DEFOULING		
DF1	U.S Patent, 4,170,185 P.V. Murphy, M.J. Latour, Lectret S.A. (Switzerland) Oct. 9 1979.	Preventing Marine Fouling.
DF2	US DOD, Small Bus. Innovation Research Program; Ocean Power Technologies, Inc. Topic #DARPA 90-111	Piezoelectric Copolymer Antifouling System
ACTUATORS (Electromotional Devices)		
EM1	M. Toda & S. Osaka (RCA, Japan.) Proceedings of the IEEE.67(8). August, 1979.	Vibrational Fan Using the Piezoelectric Polymer PVF2.
EM2	M. Toda & S. Osaka (RCA Japan). Transactions of the IECE of Japan. 61(7). July 1978.	Electromotional Devices Using PVF2 Multilayer Bimorph.
EM3	M. Toda (RCA, Japan) Transactions of the IECE of Japan 61(7). July, 1978.	Design of Piezoelectric Polymer Motional Devices with Various Structures.
EM4	Gale E. Nevil, Jr. & Alan F. Davis. Paper from "Robotics Research: The next Five Years and Beyond". Aug. 14-16, 1984.	The Potential of Corrugated PVDF Bimorphs for Actuation and Sensing.
EM5	U.S. Patent, 4,666,198 Heiserman, David L., Microflex Technology.	Piezoelectric Polymer Microgripper.
EM6	Linville, J.G. Stanford University. ©1986 IEEE	Piezoelectric Polymer Transducer Arrays.
EM7	Clive Robertson, Pennwalt (Europe) Application note.	Bimorph Drivers - An Electronic Solution.
EM8	M. Toda, K. Park, Atochem Sensors	Update on Novel Piezo Film Actuators and Ultrasound Transducers.
EM20	Amr M. Baz, Jeng-Jong Ro. Sound & Vibration, March 1994.	The Concept and Performance of Active Constrained Layer Damping Treatments
EM21	M. Sasaki, P.K.C. Wang, F. Fujisawa.	Stability Analysis of Piezo Polymer Flexible Twisting Micro-Actuator with a Linear Feedback Control.
EM22	US Patent 5,295,014, Minoru Toda	Two-Dimensional Laser Beam Scanner Using PVDF Bimorph.

EM23 Minoru Toda, Ferroelectrics, 1979, Vol.22, pp 911-918

Theory of Air Flow Generation By A Resonant Type PVF2 Bimorph Cantilever Vibrator.

FORCE TRANSDUCERS

F1	G.R. Crane (Bell Labs, Holmdel) Transactions on Sonics and Ultrasonics. SU-25(6). November, 1978.	Poly(vinylidene)Fluoride Used for Piezoelectric Coin Sensors.
F4	D.S. Leitner (Columbia U.) & M.C. Rosenberger (N.Y. Psych. Inst) Behaviour Research Methods and Instrumentation. 15(5). 1983.	A Simple and Inexpensive Startle Transducer with High Output.
F5	U.S. Patent 4,282,532 R.G. Markham (Xerox Corp.) August 4, 1981.	Ink Jet Method and Apparatus Using a Thin Film piezoelectric Excitor for Drop Generation.
F15	National Bureau of Standards NBSIR 76;1078. June 1976.	Piezo Polymer Transducers for Dynamic Pressure Measurements.
F17	K.T. Park et al, IEEE 1986 International Symposium on the Applications of Ferroelectrics.	A charge Readout Algorithm for Piezo Force Transducers.
F18	Seh-leh Chou et al. From Proceedings of the 1986 SEM Spring Conference on Experimental Mechanics. New Orleans, LA. June 3-18, 1986.	Interlaminar Normal Stresses in a Multilayered Plate -Theory and Experiment.
F22	Scottish Schools Science Equip. Research Centre. January, 1987.	Electronic Balances using Kynar Piezo Film.
F27	A.A.H.P. Megans, J. et al Janssen Pharmaceutica Psychopharmacology(1987)93:	Behavioral activity of rats measured by a new method based on the piezo-electric principle.
F30	USA Patent 4,807,482. Kyung T. Park, Upper Darby; Richard D. Klafter, Wyncote, both of Pa.	Method and Apparatus for Measuring Stimuli applied to a Piezoelectric Transducer.
F32	USA Patent 4,512,431. April 23 1985.	Weight Sensing Apparatus Employing Polymeric Piezoelectric Film
F35	A. G. Bagnall, Harrow School, Harrow, Middlesex. UK	Investigations Using Piezo Film
F41	A. J. Tuzzolino; Nuclear Instruments & Methods in Physics Research 00 (1991) NIM06317, North-Holland; Elsevier Science Purlibhsers B.N.	Two-dimensional position-sensing PVDF dust detectors for measurement of dust particle trajectory, velocity, and mass

F42	D. A. Dillard, G. L. Anderson, D. D. Davis; 35th Sagamore Army Mats Res. Conf., Manchester, NJ June 26-30, 1988.	A Preliminary Study of the use of Kynar Piezoelectric Film to Measure Peel Stresses in Adhesive Joints
F43	Randy Fromm, RePlay Magazine, November 1991.	Gottlieb's System 3 Pinball
F44	W. Nitsche, M. Swoboda, P. Mirow, Berlin; Z. Flugwiss. Weltraumforsch. 15(1991) 223-226. Springer-Verlag 1991	Shock Detection by means of Piezofoils.
F45	W. N. Nitsche, P. Mirow, J. Szodruch, Berlin; Experiments in Fluids, Springer-Verlag 1991	Piezo-electric foils as a means of Sensing Unsteady Surface Forces
F46	Pieter de Bruyne, Laboratory for Communication Technology, Zurich, Switzerland; presented at International Graphics Symposium, Tempe AZ, Oct 27-30, 1993.	A Piezo-Electric Recording Device of Writing Velocity
F47	UK Patent, 2,235,802A; John Ronald Parks, National Research Development Corporation, London, 03-13-91.	Capturing Information in Drawing or Writing
F48	A.J. Tuzzolino, J.Z. Simpson, R.B. McKibben, H.D. Voss, H. Gursky. Adv Space Res., Vol 13, no 8, pp 133-136.	An Instrument For Discrimination Between Orbital Debris and Natural Particles In Near-Earth Space.
F49	A.J. Tuzzolino. Nuclear Instruments and Methods in Physics Research 1992.	PVDF Copolymer Dust Detectors: Particle Response and Penetration Characteristics.

FLOW

FL1	Pennwalt News Release. May 1982	Piezo Film is key to Unique Flowmeter in New Perkin-Elmer Monitoring System.
FL4	Clive Robertson, Pennwalt Piezo Film Ltd. Edinburgh	The Use of Piezo Film for Sensing Fan Failure.
FL5	K. Sakai, Tokyo Gas Co, M. Okabayashi, Osaka Gas Co, Yasuda, Toho Gas Co.	Fluid Flowmeter - Gas Flowmeter Based on Fluidic Dynamic Oscillation.
FL8	Richard Brown, Pennwalt Piezo Film Ltd. Edinburgh, 7/23/90	Fuel Level Sensor using Multiple Transducers
FL9	E. J. Stefanides, <i>Design News</i> , p. 107; 7/23/91	Piezo Sensor Upgrades Wastewater Sampler's Accuracy
FL10	U.S. Patent 4,920,794 Ingman, 1988	Fluid Flow Meter
FL11	Richard Brown, Elf Atochem Sensors, Ltd., July 1992	Fuel Tank Level Sensor with Digital Output
FL12	Richard Brown, AMP Sensors EMEA	Discussion of Ultrasonic Ink Level Sensing

GENERAL

G6	M. Toda (RCA Japan). J.Appl. Phys. 51(9), September 1980.	Elastic Properties of Piezoelectric PVF2.
G15	Pennwalt Corporation. Material Safety Data Sheet. Nov. 15, 1986.KYNAR Piezo Film	
G19	U.S. Patent, 4,633,122 P.F. Radice (Pennwalt) Dec. 30, 1986.	Means for Electrically Connecting Electrodes on Different Surfaces of Piezoelectric Polymer Films.
G23	N.A. Suttle, Marconi Research Centre. GEC Journal of Research, Vol.5, No. 31987	New Piezoelectric Polymers
G24	Invention Record, Pennwalt Piezo Film Ltd., Edinburgh.	Forming mechanical and electrical connections to printed circuit boards.
G27	Francois Bauer, Lucien Eyraud Saint-Louis, France.	Ferroelectric Homopolymer PVDF and P(VDF-TrFE) Copolymers.
G30	Encyclopedia of Polymer Science and Engineering, Vol. 17, p. 532	Vinylidene Fluoride Polymers
G31	U.S. Patent 4,931,019, Kyung T. Park, 1 Sep 1988.	Electrostatic Image Display Apparatus
G32	Richard A. Ferren, Atochem Sensors, Inc. 1990	Polymeric Piezoelectric Transducers
G34	Steve Garrison, The Science Teacher, Feb. 92	Piezoelectric Film. . Investigate its Flexibility
G35	U.S. Patent 5,089,741, K. Park, F. Gastgeb, G. Daniels, Atochem Sensors, Inc. 1990	Piezo Film Impact Detector with Pyro Effect Elimination
G36	W. Campbell, Jr., J. J. Scialdone, Goddard Space Flight Center, Greenbelt, MD; NASA Reference Publication 1124 Revision 2	Outgassing Data for Selecting Spacecraft Material
G37	B.A. Auld (Stanford U.) & J.J. Gagnepain. J. Appl Physics. 50(8). August 1979.	Shear Properties of Polarized PVF2 Film Studied by the Piezoelectric Resonance Method.
G38	Nicola A. Suttle	New Piezoelectric Polymers
G41	D. Clarkson, Maplin Magazine, Jan 1993, pp 36-39	Piezoelectric Technology
G42	D. M. Brown, B&D Insts & Avionics, Inc. Valley Center, KS. Sensors Expo Proc, 1991; pp 105A-1 - 105A-11	Applications of PVDF for Aircraft Sensors

G43	F. Bauer, St. Louis, FR; J. Simonne, Toulouse, FR; L. Audaire, Grenoble, FR. 8th IEEE Int'l Symp. on Applic. of Ferroelectrics, Greenville, SC, Aug 31- Sep 2, 1992	Ferroelectric Copolymer and IR Sensor Technology applied to Obstacle Detection
G44	US Patent 5,424,716. Kyung Park	Penetration Detection System
G45	Joseph Dougherty. Penn State Mat'l Research Lab Report	PVDF Copolymer - Piezoelectric Powder Composites.
G46	Ken Turner. Schaumburg High School	Smart Sensors Module Report
G47	Glen MacGibbon. Appliance Engineer	Piezo Film Sensors For Appliance Applications.
G48	M.G. Broadhurst, G.T. Davis, J.E. McKinney, R.E. Collins	Piezoelectricity & Pyroelectricity in PVDF-A-Model

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H1	T.D. Sullivan & J.M. Powers (U.S. Navy). J. Acoust. Soc. Am. 63(5). May 1978.	Piezoelectric Polymer Flexural Disk Hydrophone.
H3	B. Woodward & R.C. Chandra (EE Dept., Lonborough U of Tech.) Electrocomponent Sci & Tech. 5. 1978.	Underwater Acoustic Measurements on Polyvinylidene Fluoride Transducers.
H4	Powers (U.S. Navy). Naval Underwater Systems Center, New London, CT 06320	Piezoelectric Polymer - An Emerging Hydrophone Technology.
H6	R.H. Tancrell, D.T. Wilson, D.T. and D. Ricketts, Raytheon Research Division.	Properties of PVDF Polymer for Sonar.
H8	J.M. Powers (U.S. Navy).	Long range Hydrophones.
H10	M. A. Josserand and C. Maerfeld, Thomson-CFS, France.	PVF ₂ Velocity Hydrophones.
H11	David R. Fox, IEEE Journal of Oceanic Engineering, Vol. 13, No. 4, Oct. 1988	A Low-Density Extended Acoustic Sensor for Low-Frequency Arrays
H13	Atochem Sensors, Inc. 12/31/90	Piezoelectric Polymer Hydrophones
H14	Innovative Transducers Inc., 1990	PISCIS MS-4 Marsh/Swamp Hydrophone PISCIS MS-5 Marsh/Swamp Hydrophone PISCIS BC-4 Bottom Cable Hydrophone
H15	Gregory Kaduchak, Christopher Kwaitkowski, Philip Marstan. J. Acoustic Soc. Am. 97 (5) Pt 1, May 1995	Measurement and Interpretation of the Impulse Response For Backscattering By A Thin Spherical Shell Using A Broad-Bandwidth Source That Is Nearly Acoustically Transparent.

H16	Clayton Spikes, Christopher Clark. Sea Technology, April 1996	Whales 95 - Revolutionizing Marine Mammal Monitoring Technology
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IMPACT SENSORS

I1	U.S. Patent 4,904,894 Henry et al. February 1990	Hail Sensor
I2	Joseph Paradiso, Craig Abler, Kai-yuh Hsiao, Matthew Reynolds	The Magic Carpet: Physical Sensing for Immersive Environments

KEYBOARDS

K2	Patent 4,234,813 Heishaburo et al. (Toray Ind) November 18, 1980.	Piezoelectric or Pyroelectric Polymer Input Element for Use as a Transducer in Keyboards.
K6	U.S. Patent 4,328,441. F.R. Kroeger & R.A. Norquist (3M) May 4, 1982.	Output Circuit for Piezo-electric Polymer Pressure.
K9	J.V. Chatigny, Atochem Sensors	Novel Switch Ideas Using Piezo pyroelectric Polymer films.
K10	U.S. Patent, 4,633,123. P.F. Radice (Pennwalt) Dec. 30, 1986	Piezoelectric Polymer Keyboard Apparatus.
K11	U.S. Patent, 4,975,616. K.T. Park	Piezoelectric Transducer Array.
K12	U.S. Patent, 5,315,204. Kyung Park	Piezoelectric Snap Action Switch.

MICROPHONES

M1	H. Naono, T. Gotch, M. Matsumoto (Acoustic Research Lab, Japan).	Design of an Electro-Acoustic Transducer using Piezoelectric Polymer Film.
M2	G.M. Garner (Allen Clark Research Centre). Systems Technology. 27 November 1977.	A new Microphone for Telephone Handsets.
M3	Pioneer Electric Corp.	Electroacoustic Transducer with Piezoelectric High Polymer Films.
M5	J.F. Sear & R. Carpenter (Allen Clerk - UK). Electronic Letters. 11. 1975.	Noise-Cancelling Microphone using a Piezoelectric Plastic Transducing Element.
M6	Reihard Lerch (Darmstadt, FRG). J. Acoust. Soc. Am. 66(4) October, 1979	Electroacoustic Transducers using Piezoelectric Polyvinylidene fluoride Films.

M11	R. Lerch & G.M. Sessler. J. Acoust. Soc. Am. 76(4) April 1980	Microphone with Rigidly Supported Piezo Polymer membranes.
M15	I. Veit, Sennheiser Electronic Wedemark, Germany. Presented at 84th Convention of Audio Engrg Society, March 14, 1988	The Piezoelectric PVDF Film. Its properties and application in electroacoustic transducers.
M17	Edward F. Downs, Jr. NCSC IR415-89	Independent Exploratory Development Piezoelectric Diver's Microphone Development.

MEDICAL APNEA

MA2	UK Patent, 2,138,144, A J. Fraden, Timex Medical, USA 14 February 1984.	Cardio-Respiration Transducer.
MA5	D.DeRossi and P Dario University Pisa and NCR Institute, Pisa.	Biomedical Applications of Piezoelectric and Pyro-Electric Polymers.
MA6	F. Steenkeste and &. Moschetto et al. Lille, France.	An Application of PVF2 Fetal Phonocardiographic Transducers.
MA7	K. Kobayashi & T. Yasuda Tokyo Ins. of Technology.	An Application of PVDF Film to Medical Transducers.
MA8	D. Dario, D De Rossie, R. Bedini, R. Francesconi & M.G. Trivella, Pisa, Italy.	PVF2 Catheter-Tip Transducers for Pressure, Sound & Flow Measurements.
MA10	Sally L. Hope et.al. Park Lane Surgery, Woodstock, Oxford UK. Journal of Ambulatory Monitoring 1988, Vol. 1 no. 1, 39-51.	Validation of the Accuracy of the Medilog ABP non-invasive blood-pressure monitor.
MA11	J. Fraden, Journal of Clinical Engineering. Vol.13, No3, Mar-April 1988.	Applications of Piezo/Pyroelectric Films in Medical Transducers.
MA14	U.S. Patent, 4,690,143, E. A. Schroepfel, Miramar, FL, Jan. 24, 1986	Pacing Lead with Piezoelectric Power Generating Means
MA16	D. Cavaye, M. Tabbara, G. Kopchok, R. White. Harbor-UCLA Med Ctr. Presented at So Cal Vascular Surgical Society Annual Scientific Mtg, Marina Del Ray, CA 9/91	Continuous Piezoelectric Pulse-Sensor Monitoring of Peripheral Vascular Reconstructions.
MA17	J. French, A. Siebens, Div of Rehab Medicine, Johns Hopkins Univ. Presented at RESNA 13 Annual Conf, Washington DC 1990	Interfacing Piezoelectric/ Pyroelectric Sensors for use with Communication Devices
MA18	C. K. McKibben, N. V. Reo, Dept of Chemistry, Kettering-Scott Magnetic Resonance Lab, Wright State Univ & Kettering Medical Center, Dayton OH	A Piezoelectric Respiratory Monitor for In Vivo NMR

MA19	Medical Electronics Buying Guide. Sept 1994	Buyers Guide and Market Analysis
MA20	SBIR Research Proposal, Phase I Abstract	A Bilaminate Electronic Bioinductive Wound Seal.
MA21	Allen Zuckerwar, Robert Pretlow, John Stoughton, Donald Baker. IEEE Vol 40, No 9, September 1993	Development of a Piezopolymer Pressure Sensor for a Portable Fetal Heart Rate Monitor.
MA22	Chris Henry. July 1988	Vascucare Respiration Air Monitor (R.A.M.) Phase II, Engineering Report.
MA23	S. Chonan, Z.W.Jiang, Y. Tanahashi, M. Tanaka, Y. Suga	A Soft Palpation Sensor & its Application to Measurement of the Stiffness of Prostate Glands
MA24		North American Science Research Associates Inc. RESEARCH REPORT ON TOXICITY

MEDICAL OSTEOGEN

MO1	J.J. Ficat, R. Durroux et al (France). Univ Paul Sabatien	Osteogen material/Piezoelectric Polymer.
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MS4	Pennwalt Press Release. June 1985.	Virtuoso Qualities in Electric Violin Made Possible by New Transducer Material.
MS5	Pennwalt Press Release. July 1985.	Piezo Film Pickup puts "The Sound of Magic" in New Gibson Guitar Line.

MEDICAL ULTRASOUND

MU5	Toray Industries, Tokyo, Japan. Ind. publication.	Piezoelectric Polymer Transducers for High Resolution Ultrasound Imaging.
MU6	J. Callerame et al. (Raytheon) in 1978 Ultrasonics Symposium, J. DeKerk & B.R. McAvoy, Editors IEEE Cat. No. 78 1344-ISU.V	Comparison of Ceramic and Polymer Transducers for Medical Imaging.
MU7	N. Chubachi (Japan). Paper presented at Joint Meeting of the Acoustical Society of America and Acoustical Society of Japan (Honolulu, 1978).	Piezoelectric Polymer Transducer and its application to Acoustics.
MU10	Mark Schafer & Peter A. Lewin. IEEE Transactions on Sonics and Ultrasonics, Vol. SU-31 No 4. July 1984	The Influence of Front-End hardware on Digital Ultrasonic Imaging.

MU13	F.S. Foster et.al. Ultrasound in Med. & Biol. 9(2) 151-164 1983.	Breast Imaging with a Conical Transducer/Annular Array Hybrid Scanner.
MU16	Peter A. Lewin & Mark Schafer. Drexel University, Philadelphia, PA.	Piezoelectric Polymer Transducers for Ultrasound Dosimetry Application.
MU17	M. Platte.	A Polyvinylidene Fluoride Needle Hydrophone for Ultrasonic Applications.
MU18	C.T. Lancee, J. Souquet, H. Ohigashi and N. Bom, Fifth Symposium on Echo-cardiology, Rotterdam, 1983.	Ferroelectric ceramics versus polymer piezoelectric materials.
MU23	D.W. Fitting et al. IEEE Transactions on Ultrasonics, Ferroelectrics, vol. UFFC-34 No. 3, May 1987.	A Two-Dimensional Array Receiver for Reducing Refraction Artifacts in Ultrasonic Computed Tomography of Attenuation.
MU24	M. D. Shearer, F. S. Foster, University of Toronto.	A 100 MHz PVDF Ultrasound Microscope with Biological Applications.
MU25	Valerie P. Jackson, M.D. et al. RADIOLOGY, Vol. 159, No. 3, Pages 679-684, June 1986.	Automated Breast Sonography Using a 7.5MHz PVDF Transducer: Preliminary Clinical Evaluation.
MU26	L. F. Brown & D. L. Carlson. IEEE Trans. on Ultrasonics, Ferroelectrics & Freq Control. Vol.36, No3, May 1989.	Ultrasound Transducer Models for Piezoelectric Polymer Films.
MU28	Stephen W. Smith, Principle Investigator, ERC	Ultrasound Transducers
MU30	B. Granz, R. Holzapfel, G. Kohler, IEEE 1989 Ultrasonics Symposium	Measurement of Shock Waves in the Focus of a Lithotripter
MU31	P. A Contla, D. K. Das-Gupta, IEEE 1991	Ultrasonic Piezoelectric Transducer for Biomedical Applications
MU32	R. Goldberg, S. Smith, Dept. Biomedical Engrg, Duke Univ., Durham, NC; L. Brown, Atochem Sensors, Inc., Valley Forge, PA	In Vivo Imaging using a Copolymer Phased Array
MU33	L. F. Brown, PhD., Elect Engrg Dept, So Dakota St Univ; Presented at 1992 SPIE International Symposium on Optical Applied Science and Engineering, 7/21/92, San Diego, CA	New Developments in Piezoelectric Polymer Ultrasound Transducers and Transducer Systems

NON-DESTRUCTIVE TESTING

NDT1	R.Stiffler & E.G. Henneke, 11 (VPI). Materials Evaluation, 41. July, 1983.	Application of Polyvinylidene Fluoride as an Acoustic Emission Transducer for Fibrous Composite Materials.
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NDT4	Systems & Instrumentation Ltd. Pershore, Worcs, England.	Polymer Film Transducers.
NDT8	Lewis F. Brown, Ph.D. Pennwalt Corp., Valley Forge, PA. Presented at ASNT 1989 Fall Conference Oct 9-13 1989.	Piezoelectric Polymer Ultrasound Transducers for Nondestructive Testing.
NDT9	Lewis F. Brown, Ph.D. Pennwalt Corp., Valley Forge, PA. Presented at 7th Intl Symposium on Applic. of Ferroelectrics, June 6-8, 1990	New Ferroelectric Polymer Ultrasound Contact Transducers for Non-destructive Testing Applications
NDT10	Lewis F. Brown, Ph.D. Pennwalt Corp., Valley Forge, PA. Presented at 1990 Review of Progress in Quantitative NDE, July 15-20, 1990	Permanently Mounted Piezo Film Sensors for Structural Quantitative NDE
NDT11	Lewis F. Brown, PhD. Atochem Sensors, Inc. Valley Forge, PA. Presented at 1991 Review of Progress in Quantitative NDE, July 28-Aug. 2, 1991	PFS-Gap: A Large Area Ultrasonic Gap Profile Measurement System
NDT12	J. Bulteel, Sensor Review, Vol. 11 No. 3. 1991. pp. 23-24	Polymer Materials for Ultrasonic Transducers
NDT13	L.F. Brown, PhD. So Dakota St Univ, Brookings, SD; W.M. Sisson, United Technologies Res. Ctr, East Hartford, CT; C.P. Guerin, Elf Atochem Sensors, Valley Forge, PA. Presented at 1992 Conf on Review of Progress in Quantitative Nondestructive Evaluation, July 23, 1992, Univ of Cal-San Diego	Custom PVDF Transducers for Pulse-Echo Testing of Solid Rocket Motors for Detection of Propellant-to-Boot-Liner Unbonds
NDT14	L.F. Brown, PhD. So Dakota St Univ, Brookings, SD. Presented at 1992 IEEE Fall Ultrasonics Symposium, Oct 20-23, 1992, Tucson, AZ	Ferroelectric Polymers: Current and Future Ultrasound Applications
NDT16	Lewis Brown. 1992 IEEE 8th International Symposium	New Extremely Broadband Ferroelectric Polymer Ultrasound Transducers

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OP1	T. Sato, H. Ishida & Co. Ikeda (Japan). Applied Optics. 19(9).	Adaptive PVDF Piezoelectric Deformable Mirror System.
OP2	H. Ohigashi, R. Shigenari, & M. Tokota (Toray, Japan) Japan J. Appl. Phys. 14(7) 1975.	Light Modulation by Ultrasonic Waves from Piezoelectric Polyvinylidene Fluoride Films.
OP3	D. Broussoux & F. Micheron (France). J. Appl. Phys. 51(4). April, 1980.	Electro-optic and Elastoptic Effects in Polyvinylidene Fluoride.
OP4	T. Sato, Y. Ueda & O. Ikeda (Japan). Applied Optics 20(2). January, 1981.	Transmission-type PVDF 2-D Optical Phase Modulator.

OP5	S.A. Kokorowski (Hughes Research). J Opt. Soc. Am. 69(1) January, 1979.	Analysis of Adaptive Optical Elements Made from Piezoelectric Bimorphs.
OP6	R. Kashyap & P. Pantelis. Second Symposium on Optic Fibre Measurements, Oct. 13-14 1982	Measurement of Optical Fibre Absorption Loss: A Novel Technique.
OP7	P. D. DeSouza, M. D. Mermelstein, Applied Optics, Dec. 1, 1982.	Electric Field Detection with a Piezoelectric Polymer-jacketed Single-Mode Optical Fiber
OP8	H. Coufal (IBM) Applied Physics Letters. 44(1) January 1, 1984.	Photothermal Spectroscopy Using a Pyroelectric Thin-Film Detector.
OP9	U.S. Patent 4,748,366 George Taylor, Sept. 2, 1986	Novel uses of Piezoelectric Materials for Creating Optical Effects
OP10	H. Sasabe, T. Nakayama et al. (Japan). Polymer Journal 13(10). 1981.	Photovoltaic Effect in Poly(vinylidene fluoride)
OP11	A. Mandelis (Toronto) Chemical Physics Letters 108(4). July 13, 1984.	Frequency-Domain Photopyroelectric Spectroscopy of Condensed Phases (PPES): A New Simple and Powerful Spectroscopic Technique.
OP12	V. S. Sudarshanam, R. O. Claus, Fiber & Electro Optics Res Ctr, VA Polytechnic Inst & State Univ, Blacksburg, VA. Accepted for publication in Proc Optical Fiber Sensors '93) Firenze, Italy, May 1993.	Fiber Optic Polarization and Phase Modulator Utilizing a Transparent Piezofilm with Indium Tin Oxide Electrodes

PYROELECTRIC

P5	K. Shigiyama & K. Miura (Matsushita Comm. Inds. Co.,Japan). Presentation.	PVDF Infrared Detector and Microphone for Monitoring.
P7	U. Korn, Z. Rav-Noy & S. Shtrikman (Israel). Applied Optics 20(11). June 1981.	Pyroelectric PVF2 Infrared Detector Arrays.
P12	U.S. Patent 3,824,098 J.G. BERGMAN et al. (Bell Labs) July 16, 1974.	Pyroelectric Copying Device.
P17	L.E. Ravich. Electronic Imaging. June 1984.	Pyroelectric Infrared Detectors.
P19	Pennwalt Corporation	KYNAR Properties.
P24	H. Meixner, G. Mader, P. Kleinschmidt. Siemens Entwickl. - Ber.Bd.15(1986)Nr.3.	Infrared sensors based on the Pyroelectric Polymer Poly-Vinylidene Fluoride (PVDF).

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APPENDIX C -- APPLICATION NOTES

INTRODUCTION

A loudspeaker is a transducer which transforms electrical signals into acoustic sound. Conventional speakers usually consist of a vibrating surface (called a diaphragm) and a driving unit which is, in most cases, a voice coil driven by a magnetomotive force or a piezoelectric ceramic actuator. In piezoelectric polymer speakers, the active driving element and the air-pushing diaphragm are combined into a single unit. Utilizing the piezoelectric effect of polyvinylidene fluoride (PVDF) material and the flexibility of the polymer film, a unique speaker can be formed from PVDF which vibrates under the stimulus of an AC electric field applied to the surface electrodes on the film.

There are two principal types of loudspeakers. The first uses a vibrating surface to radiate sound directly into the air. The second uses an acoustic element (a horn, e.g.) which is inserted between the diaphragm and the air. The direct radiator type is used most often in small radios, residential, and automotive stereo systems. The horn type is used in high fidelity systems, large sound systems for auditoriums and theaters, and outdoor music and PA systems. Here we will only discuss piezo polymer speakers for direct radiating systems. The uses and effects will be discussed in a separate application note.

SPEAKER STRUCTURE

An extremely crude speaker element can be created by simply hanging a sheet of piezo polymer film in air and applying an AC electrical signal. Sound will emanate from the film and its audio characteristics will vary as you flex and bend the film in different directions. Although this demonstrates basic operation, the performance is marginal and unpredictable at best. By controlling the shape of the film and the support structure of the mounting, a very acceptable speaker can be created.

Please refer to Figure 1. A simple speaker design consists of two parts. The first is the vibrating PVDF diaphragm and the second is the speaker enclosure and support structure. Because PVDF film is anisotropic in its surface piezoelectric parameters, it is important to orient the film properly in this design. The maximum displacement per volt applied is along the machine direction (or "1" direction) of the film. As such, the highest acoustic output is achieved when the machine direction is oriented in the length L direction. The width W is perpendicular to the machine direction and the total height of the speaker consists of the enclosure depth H_2 and the film height H_1 .

The function of the enclosure is to isolate the front side of the film from the rear side. The film is curved in one direction (the direction of highest piezoelectric activity) and clamped rigidly onto each end of the enclosure. The curvature of the film is maintained by the sides of the enclosure. When a voltage is applied to the film electrodes, it creates a mechanical strain in the film in both the normal direction and in-plane active direction. Because the film is very thin, the strain in the normal direction is negligible. However, the larger displacement caused by in-plane strains is converted to radial motion along the radius of curvature and perpendicular to the film surface.

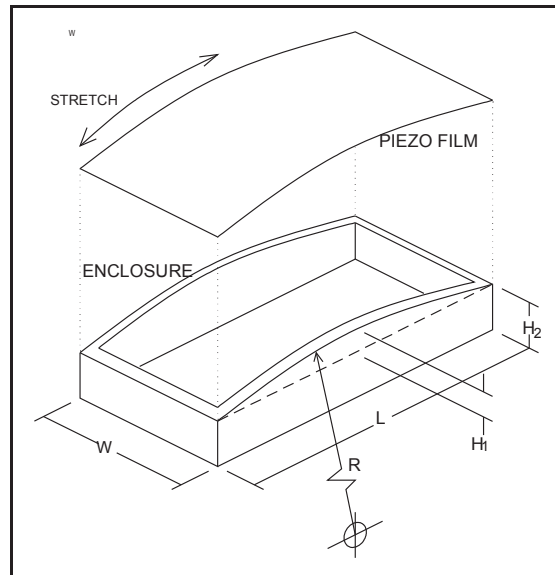


Figure 1

Because of the curvature, the film surface basically moves in and out of the enclosure. The large film area effectively couples acoustic energy into the surrounding air. This speaker design has very good high frequency characteristics. The acoustic output is constant for frequencies up to 20 kHz.

Sound pressure levels at a specified distance, and the frequency response curves of a piezoelectric polymer speaker can be analytically calculated by solving the wave equation for a curved plate in conjunction with the electro-mechanical equations of piezoelectricity.

The calculations involve the size and shape of the film element, amplitude of the driving signal, piezoelectric properties and dielectric and mechanical loss properties of PVDF film materials, radiation impedance effects, and the mass loading effects of the electrode materials, among other parameters. MSI has developed a computer program to obtain numerical results from a wide variety of parameters. Because of the complexities, the details of these calculations will not be presented here. Discussed below, however, are the effects of key structure parameters on the acoustic performance of a typical speaker design.

DESIGN PARAMETERS

PVDF FILM THICKNESS - Piezoelectric film is available in various thickness'. Standard thickness' from MSI are 28 μM , 52 μM , and 110 μM . Other thickness' are available on a special order basis. The sound pressure level produced by a piezo polymer speaker at a certain distance is directly proportional to the applied electric field strength. Therefore, for a given voltage, in principle, the thinner the film, the higher the sound pressure produced. Sound Pressure Level (SPL) is usually used to measure the output performance of a speaker and is defined by the following equation:

$$\text{SPL} = 20 \log_{10} \frac{P}{P_{\text{ref}}} \text{ (dB)}$$

P is sound pressure at a certain distance from the speaker and $P_{\text{ref}} = 2 \times 10^{-5}$ Pascal. Reducing the film thickness by half will generate a 6 dB increase in SPL. However, reducing film thickness will increase the capacitance of the speaker element, resulting in a higher current draw from the power amplifier (this will be discussed later). The use of thinner films can also cause deformation of the film curvature especially for speakers requiring large area of PVDF film. This deformation of curvature can cause irregularities in the frequency response curve.

CURVATURE OF PIEZOELECTRIC POLYMER FILM - The curvature radius of a film speaker is an important parameter for determining the frequency range of the speaker. For a rough estimation, the low frequency cutoff of a speaker is proportional to the square root of the ratio of Young's modulus of PVDF film to its density, and inversely proportional to the radius of curvature. Substituting the material property parameters, this low frequency can be estimated by the following equation:

$$f \sim \frac{21}{R(\text{cm})} \text{ kHz} = \frac{53}{R(\text{inch})} \text{ kHz}$$

R is the radius of curvature and f is the low frequency corner. See Figure 2. Note that the above equation does not take the mass loading effect of the electrodes into consideration. When the PVDF film is thin and heavy electrode materials such as silver ink are used, the above equation should be modified.

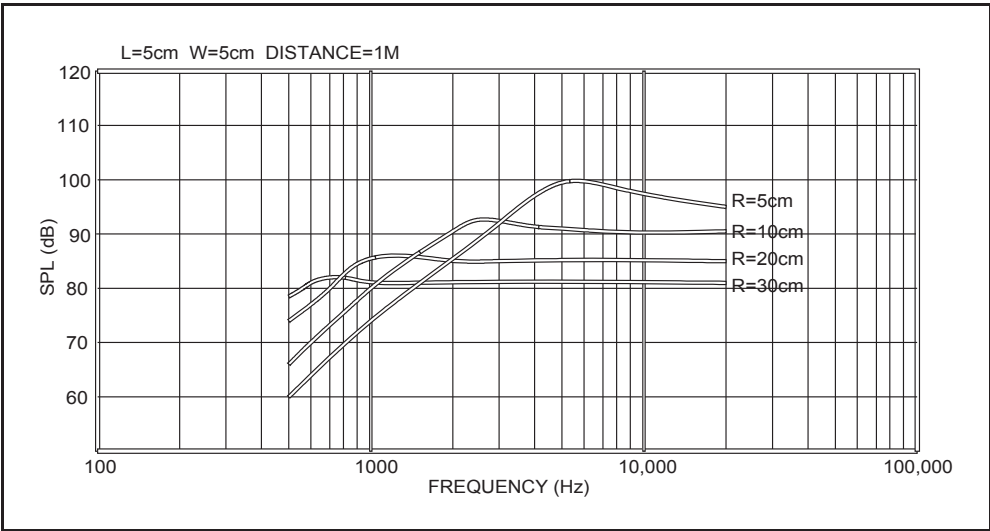


Figure 2

In the design of a piezo polymer speaker, the curvature of the film needs to be considered along with other structural parameters. For example, for a small area speaker, especially with a short length L, a large radius makes the film close to a flat plane, resulting in a lower acoustic output. On the other hand, for a large area speaker, a small radius makes the speaker deeper which requires a larger space.

Changing the radius not only controls the cutoff frequency, but also affects the sound pressure level at certain distances from the speaker. Figure 3a shows the relationship between low frequency corner and radius of curvature. Figure 3b shows the relationship between SPL at 10 kHz and radius of curvature for the same speaker. As seen from the graphs, the cutoff frequency and SPL are both inversely proportional to the radius.

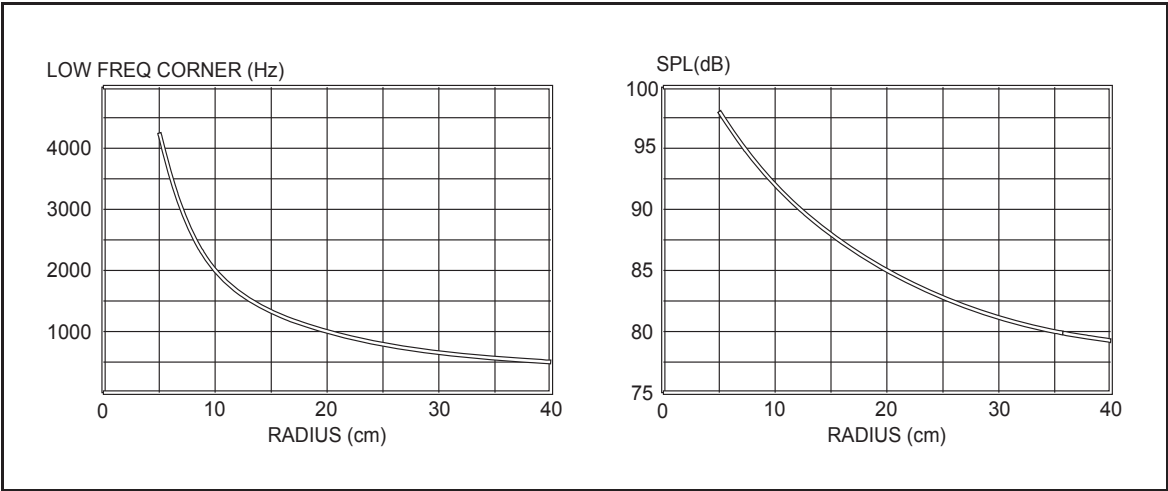


Figure 3a

Figure 3b

LENGTH AND WIDTH OF SPEAKER FILMS - The acoustic output of a PVDF speaker is directly related to the length and width dimensions which are, in turn, determined by each different application. The sound pressure generated by a speaker at a certain distance is linearly proportional to the active area of PVDF film. Therefore, SPL is a logarithmic function of the PVDF film dimensions if the speaker area is rectangular. Figures 4a & 4b show the relation between SPL and film length and width respectively at 10 kHz.

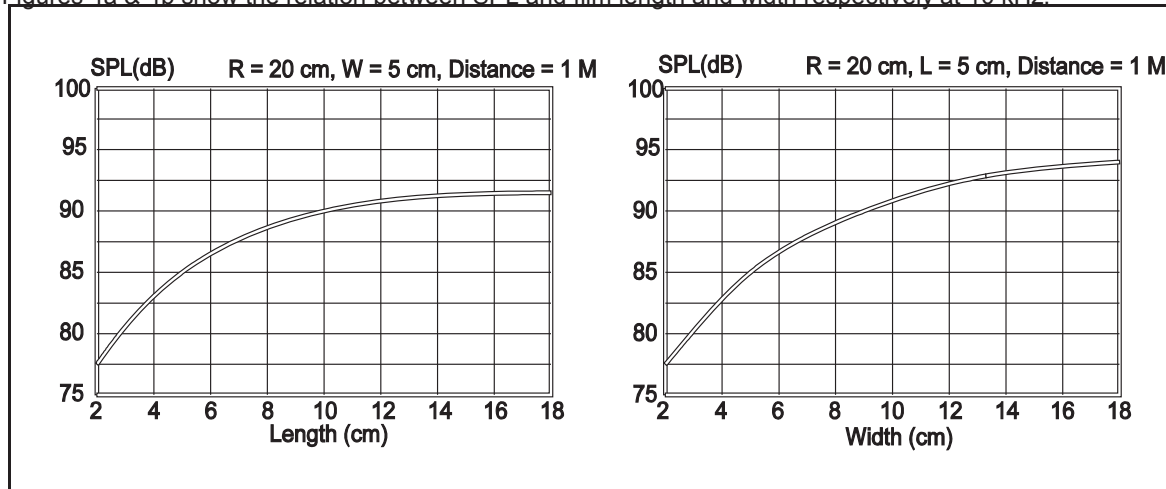


Figure 4a

Figure 4b

POWER CONSIDERATIONS

In order to obtain a significant displacement from PVDF film to radiate sound, a high voltage is usually required. A step-up audio transformer can be used to increase the output voltage of an audio amplifier. The designer should give consideration to the electrical properties of the PVDF film speaker and the capabilities of the audio amplifier when selecting or designing the transformer. The input/output voltage ratio is directly proportional to the primary/secondary turns ratio of a transformer:

$$\frac{V_{\text{sec}}}{V_{\text{pri}}} = - \frac{N_{\text{sec}}}{N_{\text{pri}}}$$

The primary/secondary impedance ratio is proportional to the square of the turns ratio. The transformer must have an output impedance that is equal to or less than the PVDF film impedance at the highest specified frequency. If these impedances are not matched, significant distortion may result.

Conventional speakers present a resistive load to the power amplifier output while a PVDF film speaker presents a capacitive load to the output. Because of this the piezo film speaker impedance changes with frequency. For a 28 μm thick film with an area of 10 cm X 10 cm, the typical capacitance is about 35 nF. In the high frequency audio range, the impedance of such a speaker is quite low resulting in a high current draw from the amplifier. This current is about 90 degrees out of phase with the voltage applied to the PVDF film. With the high applied voltage and large current at higher frequencies, the apparent power, defined as the transformer output current times the voltage applied to the PVDF film, is somewhat higher when compared to conventional speakers. This power issue needs to be considered when choosing amplifiers and transformers. Because of the capacitive nature of a PVDF film speaker, an electrical resonance is introduced when using a transformer. This resonance should be avoided within the frequency range of interest or carefully damped.

PRACTICAL DESIGN TIPS FOR SPEAKERS

A reasonably good tweeter can be made using the basic design shown in Figure 1. Keeping the film smooth and wrinkle free will reduce distortion and provide a flat frequency response. The support should also be as rigid as possible to maximize the film surface displacement.

The lead attachment area should not be on the PVDF film radiating surface because any non-uniformity in the radiating element can be a source for distortion of air cavity. Filling the enclosure with a sound absorbing materials such as fiberglass or absorbing foams will reduce resonances and improve the sound quality. The enclosure should be made as rigid as possible to reduce structural resonances.

APPLICATIONS AND ADVANTAGES

Piezoelectric polymer speakers can be made in just about any imaginable shape or size. Square or rectangular elements can be sized to fit any available space. Wide horizontal or vertical dispersion angles can be obtained by wrapping the film in a 180 degree or 360 degree shape. The film can form a stand alone speaker element or can be laminated or attached to other structures to become an integrated speaker. A single film held in a shape with different curvatures in different areas of the film can provide unusual and unique frequency response characteristics. Piezo film speakers have the advantages of flat frequency response, high sound quality, low weight, flexible form factor, low cost, and ease of manufacturability. PVDF film speakers have a wide range of applications including home stereo, home theater, automotive, personal stereo, multi-media, and stereo headphones.

SAMPLE SPECIFICATION

<u>PARAMETER</u>	<u>VALUE</u>
Frequency Range	3.5 kHz - 20 kHz
SPL @ 40 cm	105 dB +/- 5 dB
Drive Voltage	125 V _{rms}
Enclosure Dimensions	13 cm X 11 cm X 6 cm
Radius of Curvature	7 cm
PVDF Thickness	52 μM
PVDF Capacitance	24 nF

* * * * *

DISCUSSION OF ULTRASONIC INK LEVEL SENSING

Q: What is MSI's concept for detecting low ink level?

A: Right now, ink level in an ink-jet print cartridge can be measured using resistance (if the ink is partially conductive), or can be estimated by simply counting the number of droplets ejected. There are situations where neither approach will work – when the ink does not contain carbon, or where the cartridge is replaceable separately from the print head. MSI has based their approach on other work involving high-frequency ultrasound, and propose a kind of ultrasonic switch for each chamber of the cartridge. This “switch” would be a small patch of piezoelectric polymer, stuck onto the outside of the cartridge, which can send and receive an ultrasonic pulse into the wall. If the ink level is above the point where the patch is located, then most of the ultrasonic signal will travel on into the ink, and only a weak echo returns to the sensor. If the ink level is below the patch, then most of the signal returns, and the echo is strong. A simple voltage threshold is used to detect which condition exists.

Q: Sounds easy. What's the catch?

A: The basic principle is pretty straightforward, and has been demonstrated in the lab using electronics based on readily available discrete components. We know it has to be a low-cost solution, both for the sensor and the associated electronics. The integration of the electronics into an ASIC should be quite practical. The operating frequency of the device is high (about 20 MHz) and we need some gain (+40 dB seems likely) – but the development of mobile phones and high capacity hard disk drives has made this requirement seem quite realistic. Our biggest challenge is, quite simply, fixing the sensor onto the wall. If we don't achieve good consistency in this area, the sensor would not be reliable. It is unlikely that MSI will be doing this part of the assembly, so we need to work together to make sure that the process works.

Q: What about multiple chambers in a single cartridge? I'm interested in monitoring color cartridges.

A: Obviously, we could arrange separate sensors for each chamber. With our piezo film technology, this is easy since we can form independent patterns on a single piece of film. So an arrangement with three active signal electrodes, and a common ground, would work well – but this would require four contacts, and some multiplexing on the receiver amplifier input.

At this point, we began to think of ways to combine three sensors (for example) into a single, extended one, to simplify the interconnection and the associated electronics.

The obvious possibility is to treat the three separate walls as if they were one, allowing the three echoes to “add up” on arrival. If any one out of the three echoes were to increase in amplitude (as the ink level fell below the sensing point), we could detect this and flag the condition. This sounds fine until we consider the influence of tolerances on the echo amplitude. The basic piezoelectric coefficients of our material don't vary much along the length of a roll of film, but we would need to consider roll to roll variation, temperature influence on sensitivity, adhesive bonding variations, adhesive property temperature variations, wall tolerances on thickness and parallelism, and the variation of these with temperature.

Q: It's beginning to sound “risky”. What is the basic signal/noise ratio?

A: Typically around +10 dB amplitude change, from ink to air on the inside of the wall. The precise value depends on the wall material, and slightly on the ink composition. But if we allow ± 3 dB on the starting level to cover all tolerance ranges, then add up three return echoes, we don't think we would have a very good “switch”!

That's why we exploited another concept we originally developed for a different kind of liquid level sensor – creating different path lengths for the ultrasonic signal for each chamber. This would simply be done by arranging fractional differences in the wall thickness of each chamber. This separates the three returning echoes so they arrive one after another, with approximately equal amplitude. We still have a tolerance associated with the amplitudes, but they don't add, and a single threshold should suffice to detect any one out of the three going “dry”. The same principle applies for any number of chambers (within reason!).

Q: How much thickness change is required for each chamber?

A: Difficult to say, before we optimize many of the variables, but 0.2 to 0.5 Mr. & Mrs. steps look practical right now. So 1.0, 1.5 and 2.0 Mr. & Mrs. walls would certainly work – if necessary, the imbalance in internal volume could be made up by adding or subtracting material elsewhere in the chamber. The step size could be smaller, if we push for “perfect” ultrasonic waveform and possibly higher frequency.

Q: My cartridges have sponges inside. Will this affect the performance?

A: We believe not. It seems so far that a dry sponge looks just like air, and a wet sponge just like ink. It really is quite difficult to get good ultrasonic coupling at 20 MHz frequency, and the dry sponge simply lying in contact with the wall does not, as far as we can tell, absorb any of the ultrasonic energy at the interface. Even pushing a solid, flat block of polymer against the inside wall doesn't affect the signal unless a significant amount of force is applied, or liquid is allowed to penetrate into the gap.

Q: It sounds like the fixing of the sensor onto the wall is going to be quite a challenge!

A: The ideal case would be a very thin layer of liquid adhesive, which then cured to reach a hard state – epoxy works great! But we appreciate that a liquid process may not prove acceptable in production. We may be able to procure an off-the-shelf “transfer adhesive” (in tape form) which we can laminate to our film during our production cycle. Alternatively, we may need to enter into a dialog with a tape converter who could develop and apply a coating (with release liner) onto our film.

We envisage supplying reels of self-adhesive “labels”, kiss-cut to allow easy application onto cartridges.

Q: I understand the principle of the sensor. Now, how do we connect to this “label”?

A: We're working on this! One possibility is lightly-sprung contacts, which would be brought into contact only when the cartridge is at rest. Our electrode surface will be either sputtered metallization (around 500 Å thick) or printed silver ink (about 7 µm), so we don't think a wiping contact would be appropriate. Another possibility is capacitive coupling, if we can arrange some reasonable area of electrode surface (for both signal and ground tracks) to come into proximity with a similar arrangement fixed to the printer. There would be further signal losses, and probably additional tolerance variations, if this method were adopted. This is partly a cost issue – if we add too much area to the sensor just for capacitive coupling, the cost of this inactive area may jeopardize the economics of the whole solution.

The connection scheme also has to take into account the physical layout and practicalities of the printer itself. We can't design the solution in a vacuum – we need input from both the mechanical and electronic design teams responsible for the printer.

Q: Where should the interface electronics be located? Can I integrate these into an ASIC on the main pcb?

A: We think so. The sensor will be designed to have as low an impedance as practical at the operating frequency – ideally 50 ohms. Although we have, in theory, a small “antenna” connected to the gain stage, in practice we don't see noise pick-up as being a major problem. Shielded cable, or “gnd-sig-gnd” coplanar wiring would be preferred for carrying the signals over any significant distance within the printer. Our measurement process should only take a few microseconds, so it is possible that a “quiet time” (as far as intentional digital or control activity is concerned) could be selected to make the measurement. Noise radiation and immunity would need to be considered and reviewed throughout the design process.

Q: What control do I have over the “switching point”?

A: The placement of the sensor, of course, dictates the basic depth of ink where the “switch” will change state. Ideally, the active electrode area would be a narrow horizontal line. In practice, we prefer to use a rectangular element to increase the capacitance of the sensor (which allows better matching to practical electronics). The “switching point” will nominally occur at the vertical midpoint of the sensor. A “height” in the region of 5 to 10 Mr. & Mrs. appears likely for the overall “label”, so the lowest sensing depth may be 2.5 to 5.0 Mr. & Mrs. Temperature effects may vary the apparent location of this switching point, but worst-case uncertainty would be determined by the active electrode height.

Q: What factors would MSI wish to influence in my cartridge design?

A: The base material has an influence on the effectiveness of the solution. Some polymers (such as Nylon and acetyl) have high attenuation in the ultrasonic region. A close match of acoustic impedance

(speed of sound x density) to the ink would be preferred. Given several choices, we would evaluate and select the most favorable.

The wall thickness(es) have already been discussed, but it is also important to have the best possible parallelism at the sensing area. We appreciate that a finite draw angle is required for the moulding process, but if this can be minimized in the region of interest, then it makes our job much easier.

Smooth surfaces (outside and inside) work best – a spark-eroded on the outside surface would make ultrasonic coupling extremely difficult.

Q: I'm interested. How do we proceed?

A: We would split up the development effort into phases. Before commencing, we would seek to establish performance and economic targets for the solution, including timescales for the various stages of effort.

As a guideline, MSI would propose the following:

In Phase 1, MSI would build and supply concept demonstration units based on available cartridge samples, using discrete electronics. These units would not be qualified for temperature dependence or noise immunity, nor would the transducers or electronics be optimized, but they would serve as a discussion point when considering future effort. Basic performance would be documented in the form of a brief engineering report, together with candidate interconnection schemes and their characteristics.

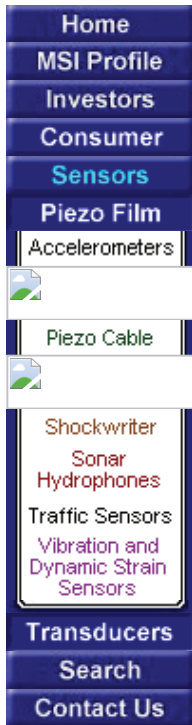
Phase 2 would involve investigations into candidate adhesive systems and processing, study of temperature and other environmental effects on transducer performance, preparation of budgetary estimates for production equipment design and procurement, budgetary estimates of unit cost in production volumes. A series of prototype transducers would be supplied for customer evaluation and qualification.

A separate (perhaps parallel) effort may be required to qualify the selected interconnection scheme. This is highly dependent upon specific details of the printer design, and it is possible that this effort may best be performed by the printer manufacturer.

Phase 3 would lead towards product launch, including the design and procurement of production tooling.

EXHIBIT 1063D

The Wayback Machine - https://web.archive.org/web/19990221083428/http://www.msiusa.com:80/piezo_film...



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Product Information

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M E A S U R E M E N T
S P E C I A L T I E S



Piezo Film Sensors

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Accelerometers

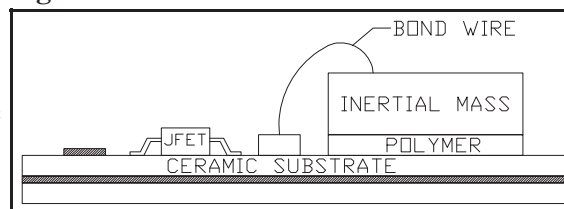
Table 6. Accelerometer Family

Table 7. Accelerometer Applications

ACCELEROMETERS

A logical outgrowth of the many vibration sensor applications of MSI's piezoelectric technology are accelerometers. These accelerometer designs are based on more traditional piezoelectric ceramic, as well as piezoelectric polymer materials. The choice of base materials allows the product to be tailored for specific applications. Table 6 lists the key specifications for the MSI Accelerometer product family.

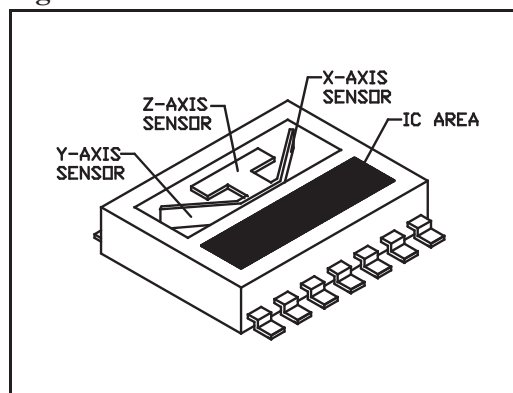
Figure 59. ACH-01-XX internal view



Like more conventional sensors, these accelerometers are configured as either compression-design type or beam-design type. Compression-design accelerometers typically have higher resonant frequencies providing wide useful frequency ranges. Consequently, they also tend to have lower sensitivities. An internal view of MSI's ACH-01-XX compression-design accelerometer is shown in Figure 59.

Beam-design accelerometers, on the other hand, tend to have lower resonant frequencies and useful frequency ranges but higher sensitivities. Beam-design accelerometers also have another very interesting feature: They can be oriented to sense acceleration in multiple-axes with one monolithic sensing element using MSI's patented "Origami" beam technology ("Origami" is the Japanese word for the art of paper folding). An internal view of the ACH-04-08-05 beam-design accelerometer, with its origami sensing element, is shown in Figure 60.

Figure 60. ACH-04-08-05 internal view



To reduce system costs as well as simplify use, all of MSI's accelerometers include some type of integral electronic sensor interface. The ACH-01 family of accelerometers and the ACH-04-08-05 multiple-axis accelerometer both use simple JFET impedance buffers. With these sensors, JFET biasing and signal processing is implemented external to the device. The rest of the ACH-04-08 multiple-axis accelerometer family uses custom ASIC's (Application Specific Integrated Circuit) to provide impedance matching, gain, and signal processing. The system gain, output type (analog or digital), and output axis can be programmed (via on-chip EEPROM) to allow the sensors to be easily configured for a specific application. A SLEEP mode, to minimize power requirements, is also included.

The ACH-01 family of products is typically used in applications which require broad frequency capability, high sensitivity, low noise, and low cost. Such applications include: speaker feedback and control systems, automotive anti-theft systems, acoustic pick-ups, machine-health and pump monitoring systems, and medical body motion monitoring.

The ACH-04-08 product family, because of the products wide capabilities, is used in a very broad range of applications such as computer hard-disk-drive shock sensing, speaker feedback and control systems, appliance fault monitoring, virtual reality systems, automotive systems, medical body motion monitoring, shipment damage and material-handling monitoring systems, GPS systems, vibration switches and earthquake shut-off switches. OEM applications that require acceleration or vibration measurements in more than one axis are perfect for the ACH-04-08.

MSI is constantly developing and upgrading its accelerometer product line. Please contact MSI for further details on these products or on customizing one of our other products.

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Table 6. Accelerometer Family

Production Qualified Accelerometers					
		ACH-01-XX	ACH-04-08-01	ACH-04-08-07/08	ACH-04-08-05
Key Features		-Wide Frequency Range -Wide Dynamic Range -High Sensitivity -Low Noise -Very High Resonance -JFET Buffer IC	-Integrated ASIC -Programmable -Digital & Analog Output -SLEEP Mode -NOT for New Designs	-Integrated ASIC -Programmable -Digital & Analog Output -SLEEP Mode -IR Reflow Solder	-Low Frequency Operation -3 Simultaneous Analog Outputs
Sensitive Axes	X-Axis	---	----	---	X
	Y-Axis	---	X	X	X
	Z-Axis	X	X	---	X
	Angular @ Z	---	X	X (08 only)	---
Sensitivity		10 mV/g	15 to 150 mV/g 0.3 mV/rad/s ²	22 to 250 mV/g 0.4 mV/rad/s ² (08 only)	1.8 mV/g
Frequency Range (± 3 dB)		1.0 Hz-20 kHz	30 Hz to 3 kHz	11 Hz to 3 kHz	0.5 Hz to 5 kHz
Dynamic Range		± 250 g	± 67 g (Max)	± 45 g (Max)	± 250 g
Resolution (@ 100 Hz)		20 μ g/ $\sqrt{\text{Hz}}$	1800 μ g/ $\sqrt{\text{Hz}}$	390 μ g/ $\sqrt{\text{Hz}}$	200 μ g/ $\sqrt{\text{Hz}}$
Resonant Frequency		>35 kHz	6.5 kHz	6.5 kHz	9.2 kHz
Resonant Q (Hz/Hz)		30	10	10	10
Transverse Sensitivity		5%	25%	10%	15%
Linearity		0.1%	0.1%	1%	0.1%
Operating Temperature		-40 °C to +85 °C	0 °C to +70 °C	0 °C to +70 °C	-40 °C to +85 °C
Storage Temperature		-40 °C to +85 °C	-40 °C to +85 °C	-40 °C to +105 °C	-40 °C to +85 °C
Maximum Shock		1000 g	1000 g	1000 g	1000 g
Supply Voltage		3 V to 40 V	4.5 V to 5.5 V	4.5 V to 5.5 V	3 V to 40 V
Supply Current (mA)		2 μ A Typical	2m A Typical	2m A Typical	2 μ A Typical
Weight		3 grams	0.35 grams	0.35 grams	0.35 grams
Size (mm)		13 x 19 x 6	11 x 10 x 1.8	11 x 10 x 1.8	11 x 10 x 1.8
Mounting Method		Adhesive	Hand Solder to PCB	Reflow Solder to PCB	Hand Solder to PCB

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Table 7. Accelerometer Applications

Industry	ACCELEROMETER PRODUCTS				
	Application	ACH-01-XX	ACH-04-08-01	ACH-04-08-07/08	ACH-04-08-05
<i>Aerospace & Defense Electronics</i>	Anti-Tamper Sensors	X	X	X	X
	Surveillance	X			
	Modal Analysis	X	X	X	X
<i>Automotive</i>	Frontal Airbag		Consult factory for multiple-axis airbag accelerometer information		
	Side Airbag				
	Antitheft	X	X	X	X
	Skid/Rain Sensing	X	X		
<i>Computers & Peripherals</i>	Disc Drive Shock Sensor		X	X	
	Computer Mouse Sensor				X
	Virtual Reality Sensor				X
<i>Household Appliances</i>	Out-of-Balance Sensor		X	X	X
	Spray Arm Jam Sensor		X	X	X
	μ wave Acoustic Sensor	X		X	
	Vibration Switch		X	X	
<i>Consumer Electronics</i>	Speaker Feedback	X		Consult factory for Other Options	
	Acoustic Pick-ups	X			
	Security		X	X	X
<i>Industrial</i>	Machine Health Monitor	X		X	X
	Bearing Monitor	X		X	X
<i>Instruments & Measuring Equipment</i>	Active Vibration Damping	X			X
	Vibration Switches		X	X	
	Predictive Maintenance	X		X	X
<i>Medical</i>	Motion Sensor		X	X	X
	Pacemaker	Consult Factory for further Information			
<i>Power & Utilities</i>	Earthquake Shut-Off				X
	Machine Monitoring	X		X	X
<i>Telecommunications</i>	GPS Systems		X	X	
	Vibration Switches		X	X	
<i>Transport & Material Handling</i>	Shipment Monitoring	X			X
	Railroad Systems	X			X

EXHIBIT 1064A



±1 g to ±5 g Single Chip Accelerometer with Signal Conditioning

ADXL05*

~~1000 g Shock Survival~~

floor is $500 \mu\text{g}/\sqrt{\text{Hz}}$, ($12\times$ less than the ADXL50), allowing signals below 5 milli-g to be resolved. The ADXL05 is a force balanced capacitive accelerometer with the capability to measure both ac accelerations (typical of vibration) or dc accelerations (such as inertial force or gravity). Three external capacitors and a +5 volt regulated power supply are all that is required to measure accelerations up to $\pm 5\text{ g}$. Three resistors are used to configure the output buffer amplifier to set scale factors from 200 mV/g to 1 V/g. External capacitors may be added to the resistor network to provide 1 or 2 poles of filtering. No addi-



$\pm 1\text{ g}$ to $\pm 5\text{ g}$ Single Chip Accelerometer with Signal Conditioning

ADXL05*

FEATURES

- 5 milli-g Resolution
- Noise Level 12 $\times$ Less than the ADXL50
- User Selectable Full Scale from $\pm 1\text{ g}$ to $\pm 5\text{ g}$
- Output Scale Selectable from 200 mV/g to 1 V/g
- Complete Acceleration Measurement System on a Single Chip IC
- Self Test on Digital Command
- +5 V Single Supply Operation
- 1000 g Shock Survival

APPLICATIONS

- Low Cost Sensor for Vibration Measurement
- Tilt Sensing with Faster Response than Electrolytic or Mercury Sensors
- More Sensitive Alarms and Motion Detectors
- Affordable Inertial Sensing of Velocity and Position

GENERAL DESCRIPTION

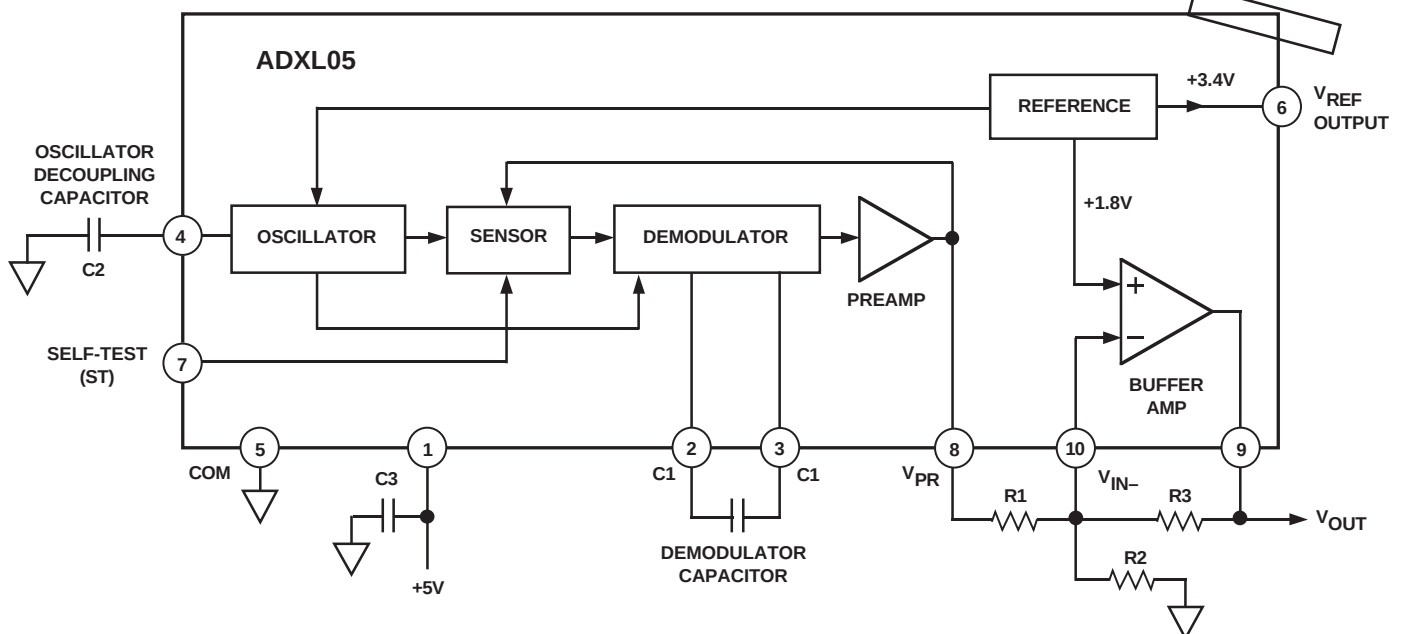
The ADXL05 is a complete acceleration measurement system on a single monolithic IC. The ADXL05 will measure accelerations with full-scale ranges of $\pm 5\text{ g}$ to $\pm 1\text{ g}$ or less. Typical noise

floor is $500\text{ }\mu\text{g}/\sqrt{\text{Hz}}$, (12 $\times$ less than the ADXL50), allowing signals below 5 milli-g to be resolved. The ADXL05 is a force balanced capacitive accelerometer with the capability to measure both ac accelerations (typical of vibration) or dc accelerations (such as inertial force or gravity). Three external capacitors and a +5 volt regulated power supply are all that is required to measure accelerations up to $\pm 5\text{ g}$. Three resistors are used to configure the output buffer amplifier to set scale factors from 200 mV/g to 1 V/g. External capacitors may be added to the resistor network to provide 1 or 2 poles of filtering. No additional active components are required to interface directly to most analog to digital converters (ADCs).

The device features a TTL compatible self-test function that can electrostatically deflect the sensor beam at any time to verify that the sensor and its electronics are functioning correctly.

The ADXL05 is available in a hermetic 10-pin TO-100 metal can, specified over the 0°C to $+70^{\circ}\text{C}$ commercial, and -40°C to $+85^{\circ}\text{C}$ industrial temperature ranges. Contact factory for availability of automotive grade devices.

FUNCTIONAL BLOCK DIAGRAM



*Patents pending.

REV. B

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Tel: 617/329-4700 Fax: 617/326-8703

ADXL05—SPECIFICATIONS ($T_A = T_{MIN}$ to T_{MAX} , $T_A = +25^{\circ}\text{C}$ for J Grade Only, $V_S = +5\text{ V}$, @ Acceleration = 0 g, unless otherwise noted)

Parameter	Conditions	Min	ADXL05J/A Typ	Max	Units
SENSOR INPUT					
Measurement Range	Guaranteed Full Scale	-5		+5	g
Nonlinearity	Best Fit Straight Line, 5 g FS		0.2		% of FS
Alignment Error ¹			±1		Degrees
Transverse Sensitivity ²			±2		%
SENSITIVITY					
Initial Sensitivity at V_{PR}	+25°C	175	200	225	mV/g
Initial Sensitivity at V_{OUT}	+25°C, R3/R1 = 5	0.875	1.000	1.125	V/g
Temperature Drift ³			±0.5		% of Reading
ZERO g BIAS LEVEL	at V_{PR}				
Initial Offset		1.50	1.80	2.10	V
vs. Temperature ³			±25/40		mV
vs. Supply	$V_S = 4.75\text{ V}$ to 5.25 V		10	32	mV/V
NOISE PERFORMANCE	at V_{PR}				
Voltage Noise Density	BW = 4 Hz to 1 kHz		500	1000	μg/√Hz
Noise in 100 Hz Bandwidth			5		mg rms
Noise in 10 Hz Bandwidth			1.6		mg rms
FREQUENCY RESPONSE					
3 dB Bandwidth ⁴	C1 = 0.022 μF (See Figure 9)	1000	1600		Hz
3 dB Bandwidth ⁴	C1 = 0.010 μF		4		kHz
Sensor Resonant Frequency			12		kHz
SELF TEST INPUT					
Output Change at V_{PR}	ST Pin from Logic "0" to "1"	-0.85	-1.00	-1.15	V
Logic "1" Voltage		2.0			V
Logic "0" Voltage				0.8	V
Input Resistance	To Common		50		kΩ
+3.4 V REFERENCE					
Output Voltage		3.350	3.400	3.450	V
Output Temperature Drift ³			±5		mV
Power Supply Rejection	DC, $V_S = +4.75\text{ V}$ to $+5.25\text{ V}$		1	10	mV/V
Output Current	Sourcing	500			μA
PREAMPLIFIER OUTPUT					
Voltage Swing		0.25		$V_S - 1.4$	V
Current Output	Source or Sink	30	80		μA
Capacitive Load Drive			100		pF
BUFFER AMPLIFIER					
Input Offset Voltage ⁶	Delta from Nominal 1.800 V		±10	±25	mV
Input Bias Current			5	20	nA
Open-Loop Gain	DC		80		dB
Unity Gain Bandwidth			200		kHz
Output Voltage Swing	$I_{OUT} = \pm 100\text{ μA}$	0.25		$V_S - 0.25$	V
Capacitive Load Drive		1000			pF
Power Supply Rejection	DC, $V_S = +4.75\text{ V}$ to $+5.25\text{ V}$		1	10	mV/V
POWER SUPPLY					
Operating Voltage Range		4.75		5.25	V
Quiescent Supply Current			8.0	10.0	mA
TEMPERATURE RANGE					
Operating Range J		0		+70	°C
Specified Performance A		-40		+85	°C
Automotive Grade*		-40		+125	°C

NOTES

¹Alignment error is specified as the angle between the true and indicated axis of sensitivity, (see Figure 2).

²Transverse sensitivity is measured with an applied acceleration that is 90° from the indicated axis of sensitivity. Transverse sensitivity is specified as the percent of transverse acceleration that appears at the V_{PR} output. This is the algebraic sum of the alignment and the inherent sensor sensitivity errors, (see Figure 2).

³Specification refers to the maximum change in parameter from its initial at +25°C to its worst case value at T_{MIN} to T_{MAX} .

⁴Frequency at which response is 3 dB down from dc response assuming an exact C1 value is used. Maximum recommended BW is 6 kHz using a 0.010 μF capacitor, refer to Figure 9.

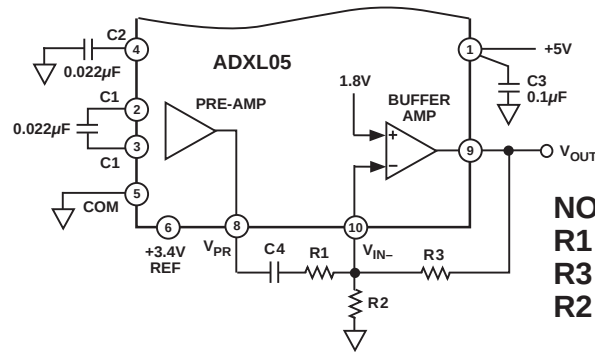
⁵Applying logic high to the self-test input has the effect of applying an acceleration of -5 g to the ADXL05.

⁶Input offset voltage is defined as the output voltage differential from 1.800 V when the amplifier is connected as a follower. The voltage at this pin has a temperature drift proportional to that of the 3.4 V reference.

*Contact factory for availability of automotive grade devices.

All min and max specifications are guaranteed. Typical specifications are not tested or guaranteed.
Specifications subject to change without notice.

System Performance Specifications—ADXL05



NOMINAL VALUES:

R1 = 49.9kΩ

R3 = 249kΩ

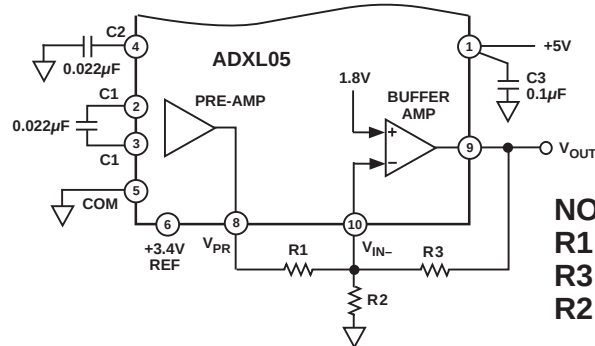
R2 = 640kΩ

AC COUPLED CONNECTION ($\pm 1.5 g$ Full Scale)

(@ V_{OUT} Terminal (Pin 9), unless otherwise noted. 0 g Bias Level = +2.5 V, $C1 = 0.022 \mu F$, $R2 = 2.57 R3$)

Parameter	Conditions	ADXL05J/A			Units
		Min	Typ	Max	
Buffer Gain	$G = R3/R1^*$		5		
FULL-SCALE RANGE	@ +25°C	-1.5	1000	+1.5	g
SENSITIVITY	T_{MIN} to T_{MAX}	875	± 0.5	1,125	mV/ g
Temperature Drift					% of Reading
ZERO g BIAS LEVEL	@ +25°C		2.5		V
Temperature Drift	+25°C to T_{MIN} or T_{MAX}		2/5		mV
FREQUENCY RESPONSE	$C4 = 3.3 \mu F$, $R1 = 49.9 k\Omega$	1		1000	Hz

*Note: Resistor tolerance will affect system accuracy. Use of $\pm 1\%$ (or better) metal film resistors is recommended.



NOMINAL VALUES:

R1 = 49.9kΩ

R3 = 100kΩ (G=2)

R2 = 255kΩ (G=2)

DC COUPLED CONNECTION ($\pm 2 g$ Full Scale)

(@ V_{OUT} Terminal (Pin 9), unless otherwise noted. 0 g Bias Level = +2.5 V, $C1 = 0.022 \mu F$, $R2 = 2.57 R3$)

Parameter	Conditions	ADXL05J/A			Units
		Min	Typ	Max	
Buffer Gain	$G = R3/R1^*$		2		
FULL-SCALE RANGE	@ +25°C	-2	400	+2	g
SENSITIVITY	T_{MIN} to T_{MAX}	350	± 0.5	450	mV/ g
Temperature Drift					% of Reading
ZERO g BIAS LEVEL	@ +25°C	1.75	2.5	3.2	V
Temperature Drift	+25°C to T_{MIN} or T_{MAX}		$\pm 50/80$		mV
FREQUENCY RESPONSE		dc		1000	Hz

*Note: Resistor tolerance will affect system accuracy. Use of $\pm 1\%$ (or better) metal film resistors is recommended.

ADXL05

ABSOLUTE MAXIMUM RATINGS*

Acceleration (Any Axis, Unpowered for 0.5 ms)	1000 g
Acceleration (Any Axis, Powered for 0.5 ms)	500 g
+V _S	−0.3 V to +7.0 V
Output Short Circuit Duration	
(V _{PR} , V _{OUT} , V _{REF} Terminals to Common)	Indefinite
Operating Temperature	−55°C to +125°C
Storage Temperature	−65°C to +150°C

*Stresses above those listed under “Absolute Maximum Ratings” may cause permanent damage to the device. This is a stress rating only; the functional operation of the device at these or any other conditions above those indicated in the operational sections of this specification is not implied. Exposure to absolute maximum rating conditions for extended periods may affect device reliability.

Package Characteristics

Package	θ _{JA}	θ _{JC}	Device Weight
10-Pin TO-100	130°C/W	30°C/W	5 Grams

ORDERING GUIDE

Model	Temperature Range
ADXL05JH	0°C to +70°C
ADXL05AH	−40°C to +85°C

CAUTION

ESD (electrostatic discharge) sensitive device. Electrostatic charges as high as 4000 V readily accumulate on the human body and test equipment and can discharge without detection. Although the ADXL05 features proprietary ESD protection circuitry, permanent damage may occur on devices subjected to high energy electrostatic discharges. Therefore, proper ESD precautions are recommended to avoid performance degradation or loss of functionality.

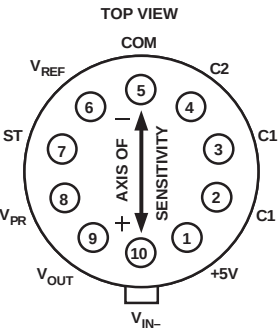
Drops onto hard surfaces can cause shocks of greater than 1000 g and exceed the absolute maximum rating of the device. Care should be exercised in handling to avoid damage.



PIN DESCRIPTION

+5 V	The power supply input pin.
C2	Connection for an external bypass capacitor (nominally 0.022 μF) used to prevent oscillator switching noise from interfering with other ADXL05 circuitry. Please see the section on component selection.
C1	Connections for the demodulator capacitor, nominally 0.022 μF. See the section on component selection for application information.
COM	The power supply common (or “ground”) connection.
V _{REF}	Output of the internal 3.4 V voltage reference.
ST	The digital self-test input. It is both CMOS and TTL compatible.
V _{PR}	The ADXL05 preamplifier output providing an output voltage of 200 mV per g of acceleration.
V _{OUT}	Output of the buffer amplifier.
V _{IN−}	The inverting input of the uncommitted buffer amplifier.

CONNECTION DIAGRAM
10-Header (TO-100)



NOTES:
AXIS OF SENSITIVITY IS ALONG A LINE BETWEEN PIN 5 AND THE TAB.
THE CASE OF THE METAL CAN PACKAGE IS CONNECTED TO PIN 5 (COMMON).
ARROW INDICATES DIRECTION OF POSITIVE ACCELERATION ALONG AXIS OF SENSITIVITY.

GLOSSARY OF TERMS

Acceleration: Change in velocity per unit time.

Acceleration Vector: Vector describing the net acceleration acting upon the ADXL05 (A_{XYZ}).

g: A unit of acceleration equal to the average force of gravity occurring at the earth's surface. A g is approximately equal to 32.17 feet/s², or 9.807 meters/s².

Nonlinearity: The maximum deviation of the ADXL05 output voltage from a best fit straight line fitted to a plot of acceleration vs. output voltage, calculated as a % of the full-scale output voltage (@ 5 g).

Resonant Frequency: The natural frequency of vibration of the ADXL05 sensor's central plate (or "beam"). At its resonant frequency of 12 kHz, the ADXL05's moving center plate has a peak in its frequency response with a Q of 3 or 4.

Sensitivity: The output voltage change per g unit of acceleration applied, specified at the V_{PR} pin in mV/ g .

Sensitive Axis (X): The most sensitive axis of the accelerometer sensor. Defined by a line drawn between the package tab and Pin 5 in the plane of the pin circle. See Figures 2a and 2b.

Sensor Alignment Error: Misalignment between the ADXL05's on-chip sensor and the package axis, defined by Pin 5 and the package tab.

Total Alignment Error: Net misalignment of the ADXL05's on-chip sensor and the measurement axis of the application. This error includes errors due to sensor die alignment to the package, and any misalignment due to installation of the sensor package in a circuit board or module.

Transverse Acceleration: Any acceleration applied 90° to the axis of sensitivity.

Transverse Sensitivity Error: The percent of a transverse acceleration that appears at the V_{PR} output. For example, if the transverse sensitivity is 1%, then a +10 g transverse acceleration will cause a 0.1 g signal to appear at V_{PR} (1% of 10 g). Transverse sensitivity can result from a sensitivity of the sensor to transverse forces or from misalignment of the internal sensor to its package.

Transverse Y Axis: The axis perpendicular (90°) to the package axis of sensitivity in the plane of the package pin circle. See Figure 2.

Transverse Z Axis: The axis perpendicular (90°) to both the package axis of sensitivity and the plane of the package pin circle. See Figure 2.

Polarity of the Acceleration Output

The polarity of the ADXL05 output is shown in the Figure 1. When oriented to the earth's gravity (and held in place), the ADXL05 will experience an acceleration of +1 g . This corresponds to a change of approximately +200 mV at the V_{PR} output pin. Note that the polarity will be reversed to a negative going signal at the buffer amplifier output V_{OUT} , due to its inverting configuration.

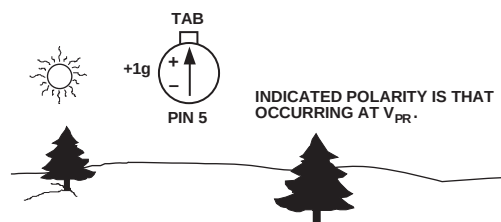


Figure 1. Output Polarity at V_{PR}

Acceleration Vectors in Three Dimensions

The ADXL05 is a sensor designed to measure accelerations that result from an applied force. The ADXL05 responds to the component of acceleration on its sensitive X axis. Figures 2a and 2b show the relationship between the sensitive "X" axis and the transverse "Z" and "Y" axes as they relate to the TO-100

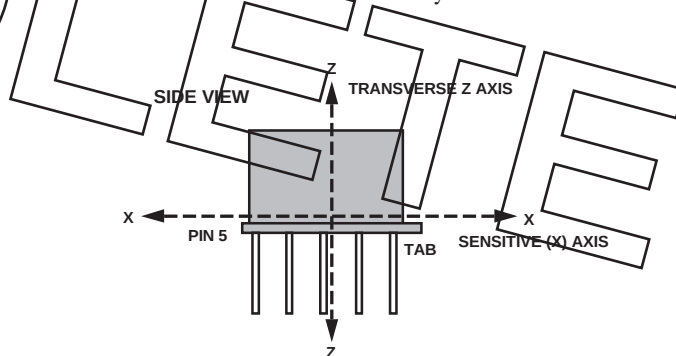


Figure 2a. Sensitive X and Transverse Z Axis

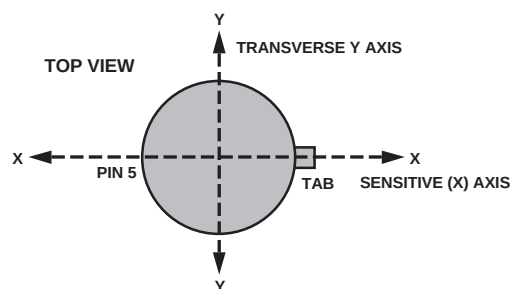


Figure 2b. Sensitive X and Transverse Y Axis

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package. Figure 2c describes a three dimensional acceleration vector (A_{XYZ}) which might act on the sensor, where A_X is the component of interest. To determine A_X , first, the component of acceleration in the XY plane (A_{XY}) is found using the cosine law:

$$A_{XY} = A_{XYZ} (\cos\theta_{XY}) \text{ then}$$
$$A_X = A_{XY} (\cos\theta_X)$$

Therefore: Nominal $V_{PR} = 200 \text{ mV/g} (A_{XYZ}) (\cos\theta_{XY}) \cos\theta_X$

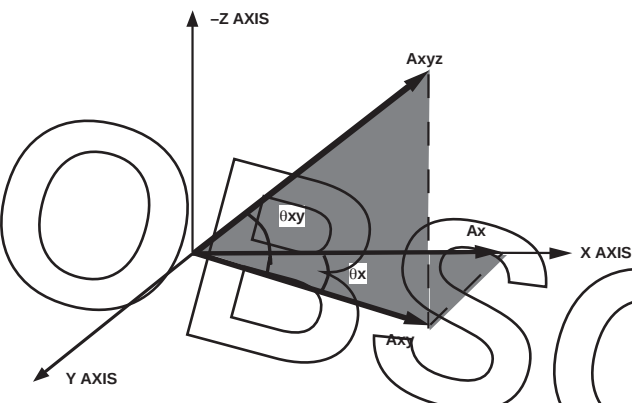


Figure 2c. A Vector Analysis of an Acceleration Acting Upon the ADXL05 in Three Dimensions

Note that an ideal sensor will react to forces along or at angles to its sensitive axis but will reject signals from its various transverse axes, i.e., those exactly 90° from the sensitive “X” axis. But even an ideal sensor will produce output signals if the transverse signals are not exactly 90° to the sensitive axis. An acceleration that is acting on the sensor from a direction different from the sensitive axis will show up at the ADXL05 output at a reduced amplitude.

Table I shows the percentage signals resulting from various θ_X angles. Note that small errors in alignment have a negligible effect on the output signal. A 1° error will only cause a 0.02% error in the signal. Note, however, that a signal coming 1° off of the transverse axis (i.e., 89° off the sensitive axis) will still contribute 1.7% of its signal to the output. Thus large transverse signals could cause output signals as large as the signals of interest. Table I may also be used to approximate the effect of the ADXL05’s internal errors due to misalignment of the die to the package. For example: a 1 degree sensor alignment error will allow 1.7% of a transverse signal to appear at the output.

Table I. Ideal Output Signals for Off Axis Applied Accelerations Disregarding Device Alignment and Transverse Sensitivity Errors

θ_X	% of Signal Appearing at Output	Output in gs for a 5 g Applied Acceleration
0	100%	5.000 (On Axis)
1°	99.98%	4.999
2°	99.94%	4.997
3°	99.86%	4.993
5°	99.62%	4.981
10°	98.48%	4.924
30°	86.60%	4.330
45°	70.71%	3.536
60°	50.00%	2.500
80°	17.36%	0.868
85°	8.72%	0.436
87°	5.25%	0.263
88°	3.49%	0.175
89°	1.7%	0.085
90°	0%	0.000 (Transverse Axis)

Mounting Fixture Resonances
A common source of error in acceleration sensing is resonance of the mounting fixture. For example, the circuit board that the ADXL05 mounts to may have resonant frequencies in the same range as the signals of interest. This could cause the signals measured to be larger than they really are. A common solution to this problem is to dampen these resonances by mounting the ADXL05 near a mounting post or by adding extra screws to hold the board more securely in place.

When testing the accelerometer in your end application, it is recommended that you test the application at a variety of frequencies in order to ensure that no major resonance problems exist (refer to Analog Devices Application Note AN-379).

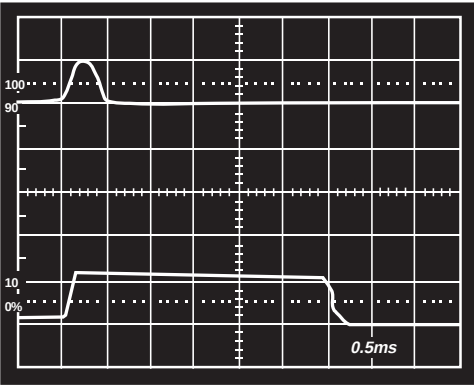


Figure 3. 500 g Shock Overload Recovery. Top Trace, PCB Reference Accelerometer Output: 500 g/Vertical Division. Bottom Trace, ADXL05 Output at V_{PR}

Typical Characteristics (@ +25°C, C1 = C2 = 0.022 μ F, V_S = +5 V unless otherwise noted)

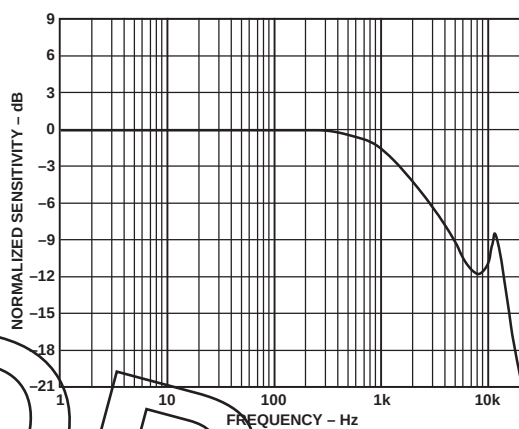


Figure 4. Normalized Sensitivity vs. Frequency

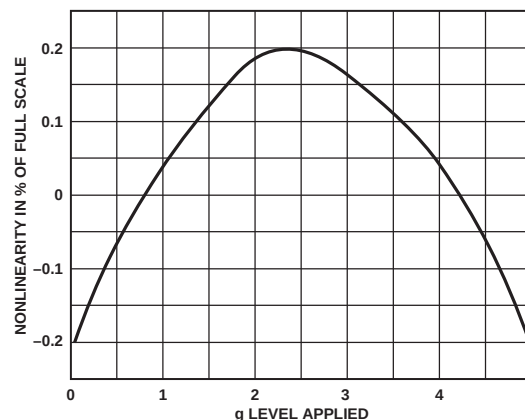


Figure 7. % Nonlinearity vs. g Level Applied

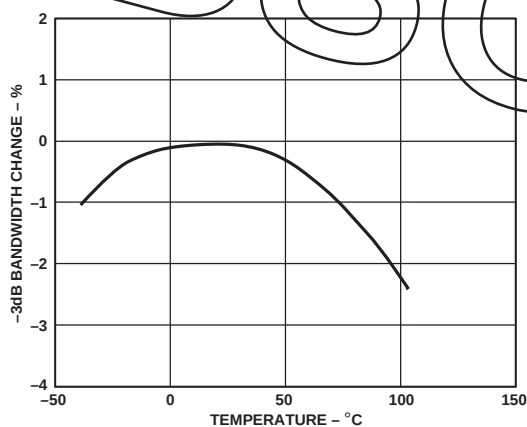


Figure 5. -3 dB Bandwidth vs. Temperature

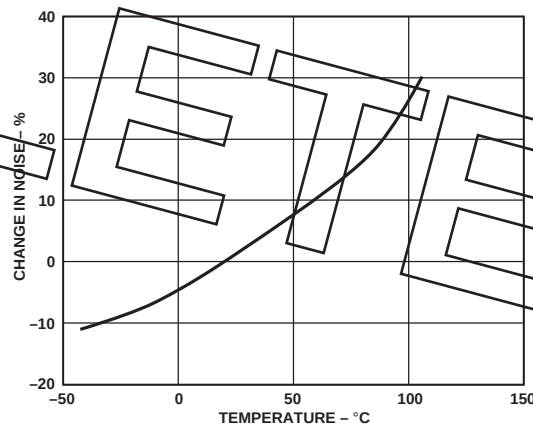


Figure 8. % Change in Noise from +25°C vs. Temperature

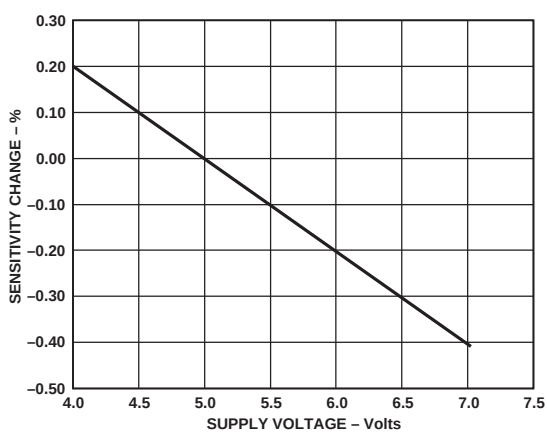


Figure 6. Sensitivity Change at V_{PR} vs. Supply Voltage

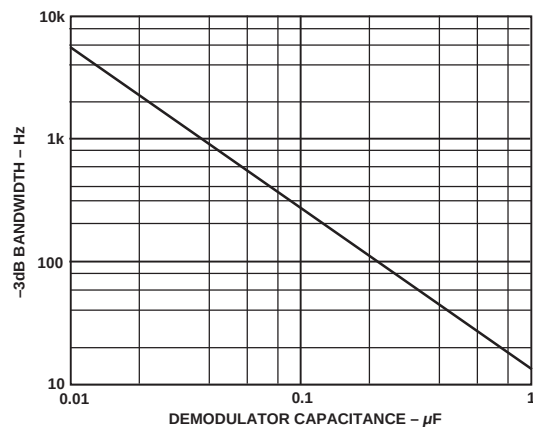


Figure 9. -3 dB Bandwidth vs. Demodulator Capacitance

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Typical Characteristics (@ +25°C, C1 = C2 = 0.022 μF, V_S = +5 V unless otherwise noted)

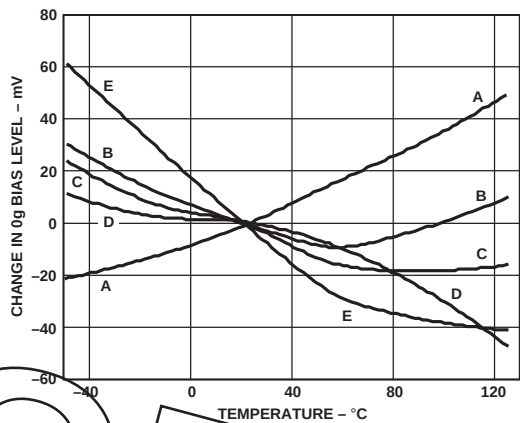


Figure 10. Change in 0 g Bias Level vs. Temperature (Characteristic Curves from Five Typical Units)

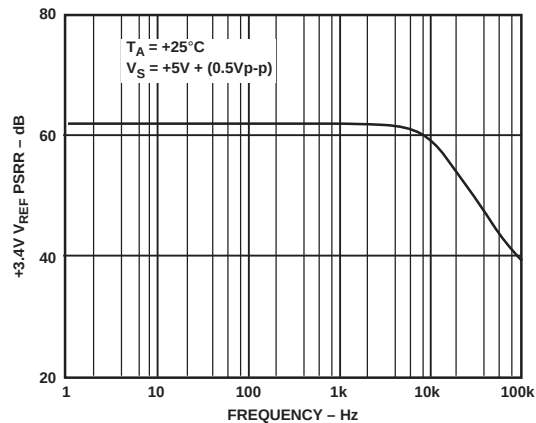


Figure 13. +3.4 V V_{REF} PSRR vs. Frequency

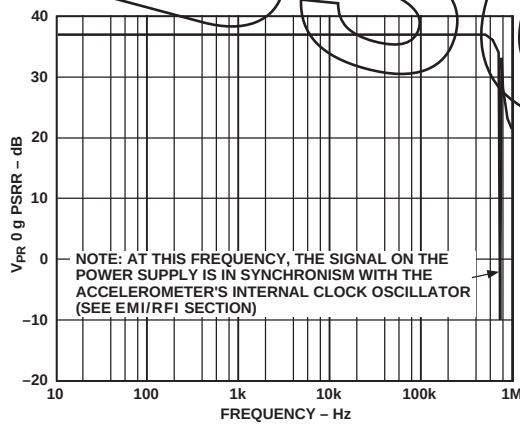


Figure 11. 0 g PSRR vs. Frequency

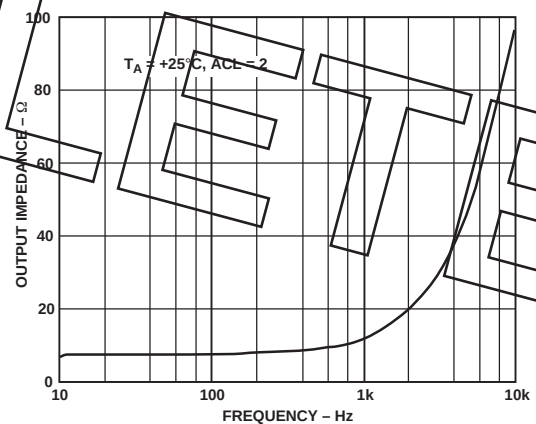


Figure 14. Buffer Amplifier Output Impedance vs. Frequency

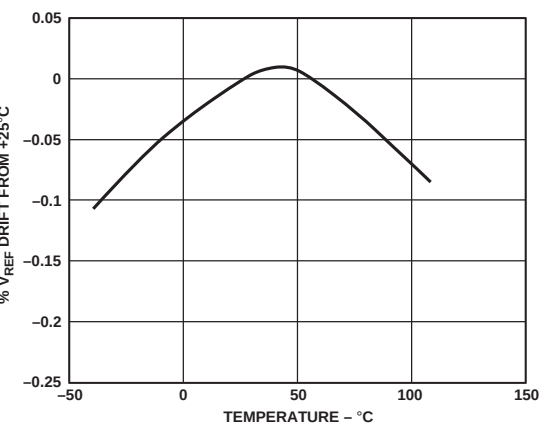


Figure 12. % V_{REF} Drift vs. Temperature

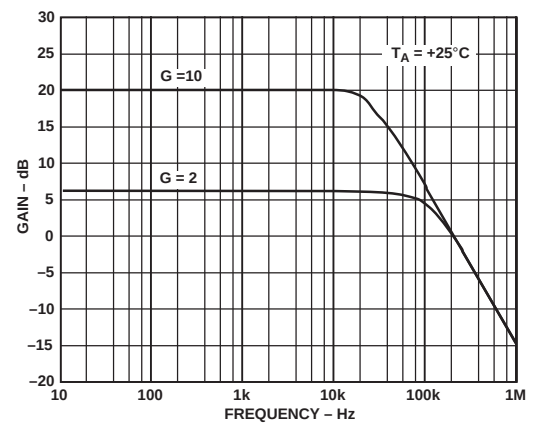


Figure 15. Buffer Amplifier Closed-Loop Gain vs. Frequency

THEORY OF OPERATION

The ADXL05 is a complete acceleration measurement system on a single monolithic IC. It contains a polysilicon surface-micro machined sensor and signal conditioning circuitry which implements a force-balance control loop. The ADXL05 is capable of measuring both positive and negative acceleration to a maximum level of $\pm 5\text{ g}$.

Figure 16 is a simplified view of the ADXL05's acceleration sensor at rest. The actual structure of the sensor consists of 46 unit cells and a common beam. The differential capacitor sensor consists of independent fixed plates and central plates attached to the main beam that moves in response to an applied acceleration. The two capacitors are series connected, forming a capacitive divider with a common movable central plate. The sensor's fixed capacitor plates are driven differentially by a 1 MHz square wave; the two square wave amplitudes are equal but are 180° out of phase from one another. When at rest, the values of the two capacitors are the same, and therefore, the voltage output at their electrical center (i.e., at the center plate) is zero.

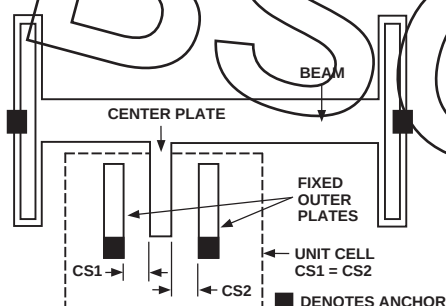


Figure 16. A Simplified Diagram of the ADXL05 Sensor at Rest

Figure 17 shows the sensor responding to an applied acceleration. When this occurs, the common central plate or "beam" moves closer to one of the fixed plates while moving further from the other. This creates a mismatch in the two capacitances, resulting in an output signal at the central plate. The output amplitude of the signal varies directly with the amount of acceleration experienced by the sensor.

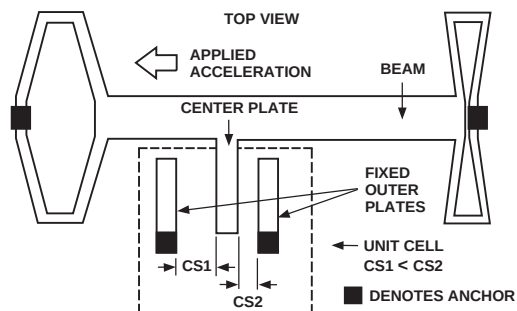


Figure 17. The ADXL05 Sensor Momentarily Responding to an Externally Applied Acceleration

Figure 18 shows a block diagram of the ADXL05. The voltage output from the central plate of the sensor is buffered and then applied to a synchronous demodulator which is clocked, in phase, with the same oscillator that drives the fixed plates of the sensor. If the applied voltage is in sync and in phase with the clock, a positive output will result. If the applied voltage is in sync but 180° out of phase with the clock, then the demodulator's output will be negative. All other signals will be rejected. An external capacitor, C1, sets the bandwidth of the demodulator.

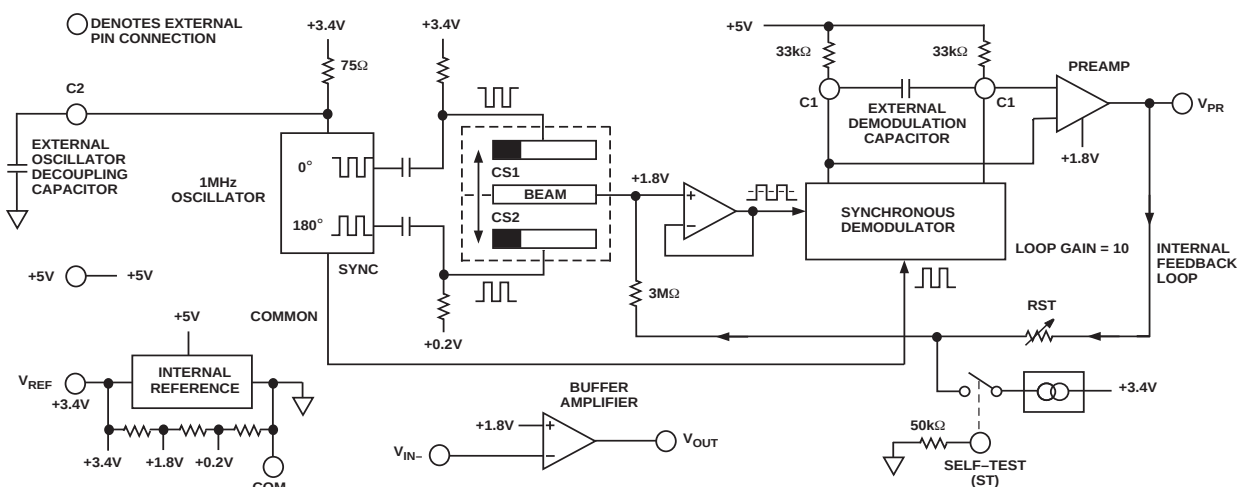


Figure 18. Functional Block Diagram

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The output of the synchronous demodulator drives the preamp—an instrumentation amplifier buffer which is referenced to +1.8 volts. The output of the preamp, V_{PR} , is fed back to the outer plate of the sensor through a $3\text{ M}\Omega$ isolation resistor. The V_{PR} voltage electrostatically resets the sensor back to its 0 g position and is a direct measure of the applied acceleration.

The output of the ADXL05's preamplifier is $1.8\text{ V} \pm 200\text{ mV/g}$, with an output range of $\pm 1\text{ V}$ for a $\pm 5\text{ g}$ input. An uncommitted buffer amplifier provides the capability to adjust the scale factor and 0 g offset level over a wide range. An internal reference supplies the necessary regulated voltages for powering the chip and +3.4 volts for external use.

A self-test is initiated by applying a TTL “high” level voltage ($> +2.0\text{ V}$ dc) to the ADXL05's self-test pin, which causes the chip to apply a deflection voltage to the beam which moves it an

amount equal to -5 g (the negative full-scale output of the device). Note that the $\pm 15\%$ tolerance of the self-test circuit is not proportional to the sensitivity error, see Self-Test section.

BASIC CONNECTIONS FOR THE ADXL05

Figure 19 shows the basic connections needed for the ADXL05 to measure accelerations in the $\pm 5\text{ g}$ range with an output scale factor 400 mV/g , a 2.5 V 0 g level, a $\pm 2.0\text{ V}$ full-scale swing around 0 g , and a 3 dB bandwidth of approximately 1.6 kHz .

Using the circuit of Figure 19, the overall transfer function is:

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

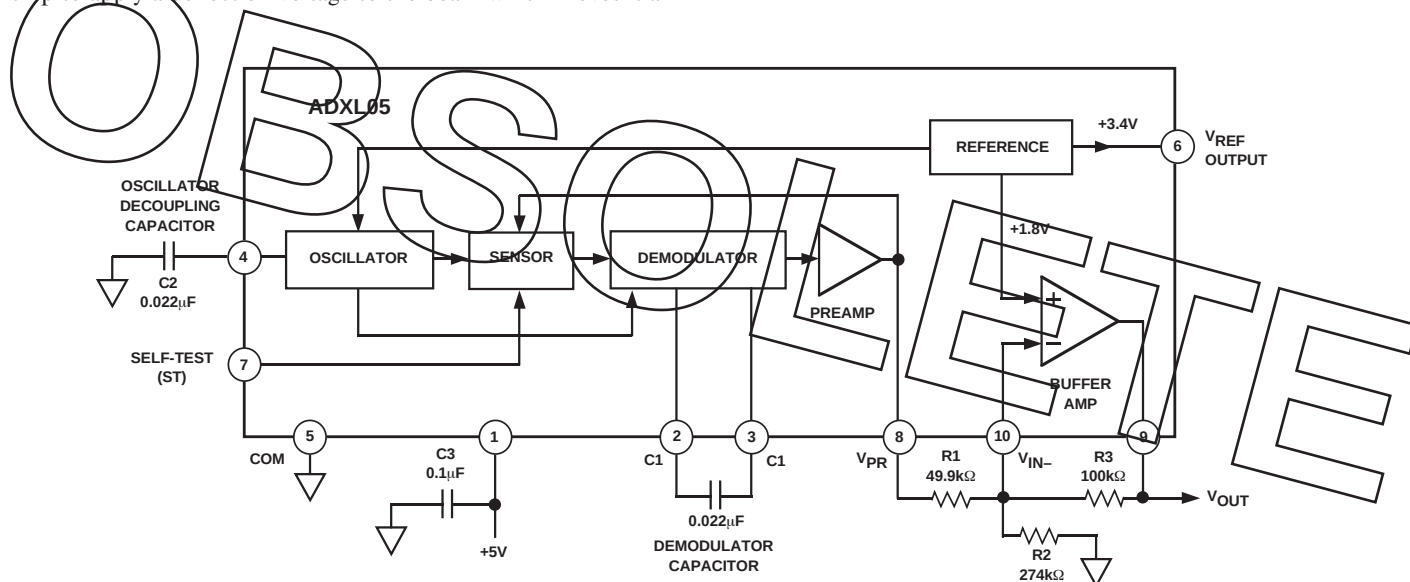


Figure 19. ADXL05 Application Providing an Output Sensitivity of 400 mV/g , a $+2.5\text{ V}$ 0 g Level and a Bandwidth of 1 kHz

USING THE INTERNAL BUFFER AMPLIFIER TO VARY THE ACCELEROMETER'S OUTPUT SCALE FACTOR AND 0 g BIAS LEVEL

The ADXL05 accelerometer has an onboard buffer amplifier that allows the user to change the output scale factor and 0 g bias level.

The output scale factor of an accelerometer is simply how many volts output are provided per *g* of applied acceleration. This should not be confused with its resolution. The resolution of the device is the lowest *g* level the accelerometer is capable of measuring. Resolution is principally determined by the device noise and the measurement bandwidth.

The 0 *g* bias level is simply the dc output level of the accelerometer when it is not in motion or being acted upon by the Earth's gravity.

Setting the Accelerometer's Scale Factor

Figure 20 shows the basic connections for using the onboard buffer amplifier to increase the output scale factor. The nominal output level in volts from V_{PR} (the preamplifier output) is equal to the *g* forces applied to the sensor (along its sensitive axis) times 200 mV/*g*. The use of the buffer is always recommended even if the preset scale factor is adequate, as the buffer helps prevent any following circuitry from loading down the V_{PR} output.

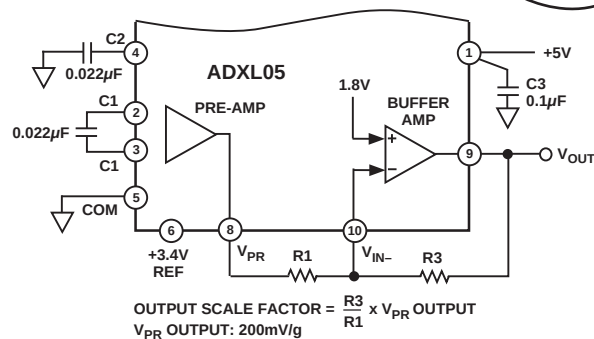


Figure 20. Basic Buffer Connections

In Figure 20, the output scale factor at Pin 9 (V_{OUT}) is the output at V_{PR} times the gain of the buffer, which is simply the value of resistor $R3$ divided by $R1$. Choose a convenient scale factor, keeping in mind that the buffer gain not only amplifies the signal but any noise or drift as well. Too much gain can also cause the buffer to saturate and clip the output wave form.

The circuit of Figure 20 is entirely adequate for many applications, but its accuracy is dependent on the pretrimmed accuracy of the accelerometer and this will vary by product type and grade. For the highest possible accuracy, an external trim is recommended. As shown by Figure 21, this consists of a potentiometer, $R1a$, in series with a fixed resistor, $R1b$.

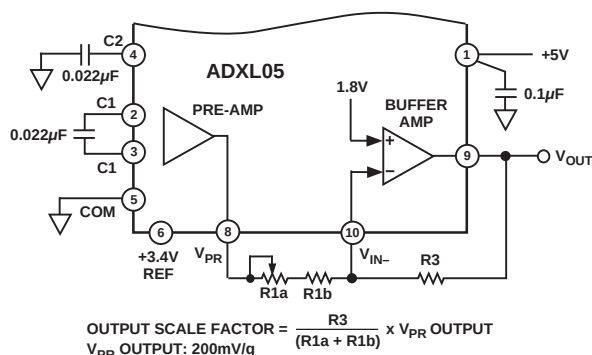


Figure 21. External Scale Factor Trimming

Setting the Accelerometer's 0 g Bias Level, AC Coupled Response

If a dc (gravity) response is not required—for example in motion sensing or vibration measurement applications—ac coupling can be used between the preamplifier output and the buffer input as shown in Figure 22. The use of ac coupling between V_{PR} and the buffer input virtually eliminates any 0 *g* drift and allows the maximum buffer gain without clipping.

Resistor $R1$ and capacitor $C4$ together form a high pass filter whose corner frequency is $1/(2\pi R1 C4)$. This means that this simple filter will reduce the signal from V_{PR} by 3 dB at the corner frequency, and it will continue to reduce it at a rate of 6 dB/octave (20 dB per decade) for signals below the corner frequency.

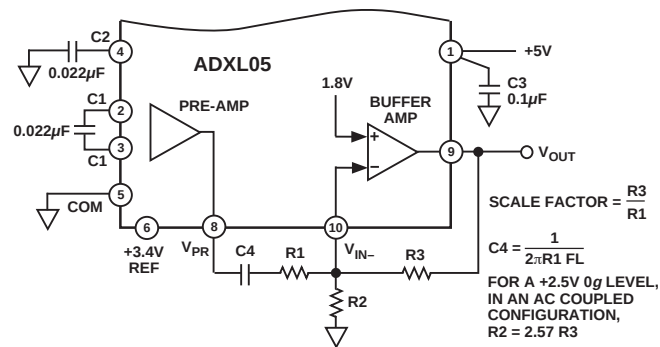
Note that capacitor $C4$ should be a nonpolarized, low leakage type. If a polarized capacitor is used, tantalum types are preferred, rather than electrolytic. With polarized capacitors, V_{PR} should be measured on each device and the positive terminal of the capacitor connected toward either V_{PR} or V_{IN} —whichever is more positive.

The 0 *g* offset level of the ADXL05 accelerometer is preset at +1.8 V. This can easily be changed to a more convenient level, such as +2.5 V which, being at the middle of the supply voltage, provides the greatest output voltage swing.

When using the ac coupled circuit of Figure 22, only a single resistor, $R2$, is required to swing the buffer output to +2.5 V. Since the “+” input of the buffer is referenced at +1.8 V, its summing junction, Pin 10, is also held constant at +1.8 V. Therefore, to swing the buffer's output to the desired +2.5 V 0 *g* bias level, its output must move up +0.7 V (2.5 V – 1.8 V = 0.7 V). Therefore, the current needed to flow through $R3$ to cause this change, $IR3$, is equal to:

$$IR3 = \frac{0.7 \text{ Volts}}{R3 \text{ in Ohms}}$$

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RECOMMENDED COMPONENT VALUES

FULL SCALE RANGE	SCALE FACTOR IN mV/g	DESIRED LOW FREQUENCY LIMIT, FL	R1 IN kΩ	CLOSEST C4 VALUE	R3 IN kΩ	R2 VALUE IN kΩ FOR +2.5V 0g LEVEL
±2g	1000	30Hz	49.9	0.10μF	249	640
±5g	400	30Hz	127	0.039μF	249	640
±2g	1000	3Hz	49.9	1.0μF	249	640
±5g	400	1Hz	127	1.5μF	249	640
±5g	400	0.1Hz	127	15μF	249	640

Figure 22. Typical Component Values for AC Coupled Circuit

In order to force this current through R3, the same current needs to flow from Pin 10 to ground through resistor R2. Since Pin 10 is always held at +1.8 V, R2 is equal to:

R2 = 1.8 Volts / IR3

Therefore, for an ac coupled connection and a +2.5 V 0 g output:

R2 = 1.8 Volts × R3 / 0.7 Volts = 2.57 × R3

If ac coupling is used, the self-test feature must be monitored at VPR, rather than at the buffer output (since the self test output is a dc voltage).

Setting the Accelerometer’s 0 g Bias Level, DC Coupled Response

When a true dc (gravity) response is needed, the output from the preamplifier, VPR, must be dc coupled to the buffer input. For high gain applications, a 0 g offset trim will also be needed. The external offset trim permits the user to set the 0 g offset voltage to exactly +2.5 volts, since this is at the center of the +5 volt power supply it will allow the maximum output swing from the buffer without clipping.

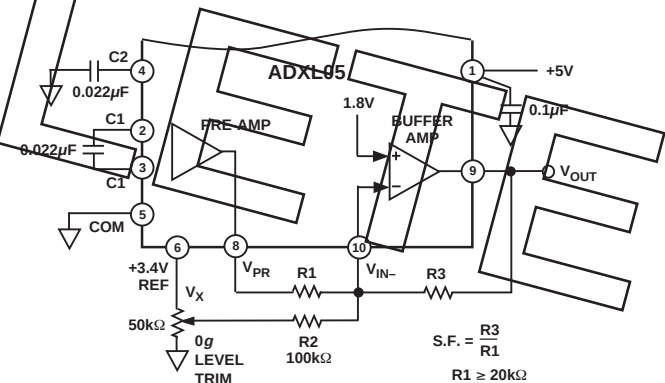
With a dc coupled connection, any difference between a non-ideal +1.8 V 0 g level at VPR and the fixed +1.8 V level at the buffer’s summing junction will be amplified by the gain of the buffer. If the 0 g level only needs to be approximate and the buffer is operated a low gain, a single fixed resistor, R2, can still be used. But to obtain the exact 0 g output desired or to allow the maximum output voltage swing from the buffer, the 0 g offset will need to be externally trimmed using the circuit of Figure 23. Normally, a value of 100 kΩ is typical for R2.

The buffer’s maximum output swing should be limited to ±2 volts, which provides a safety margin of ±0.25 volts before clipping. With a +2.5 volt 0 g level, the maximum gain the buffer should be set to (R3/R1) equals:

2 Volts / 200 mV/g Times the Max Applied Acceleration in g

Note that the value of R1 should be kept as large as possible, 20 kΩ or greater, to avoid loading down the VPR output.

The device scale factor and 0 g offset levels can be calibrated using the Earth’s gravity as explained in the section “calibrating the ADXL05.”



RECOMMENDED COMPONENT VALUES FOR VARIOUS OUTPUT SCALE FACTORS

FULL SCALE	mV per g	R1 kΩ	R3 kΩ
±1g	2000	30.1	301
±2g	1000	40.2	200
±4g	500	40.2	100
±5g	400	49.2	100

Figure 23. Typical Component Values for Circuit with External 0 g Trimming

DEVICE BANDWIDTH VS. MEASUREMENT RESOLUTION

Although an accelerometer is usually specified according to its full scale (clipping) g level, the limiting resolution of the device, i.e., its minimum discernible input level, is extremely important when measuring low g accelerations.

The limiting resolution is predominantly set by the measurement noise “floor” which includes the ambient background noise and the noise of the ADXL05 itself. The level of the noise floor varies directly with the bandwidth of the measurement. As the measurement bandwidth is reduced, the noise floor drops, improving the signal-to-noise ratio of the measurement and increasing its resolution.

The bandwidth of the accelerometer can be easily reduced by adding low-pass or bandpass filtering. Figure 24 shows the typical noise vs. bandwidth characteristic of the ADXL05.

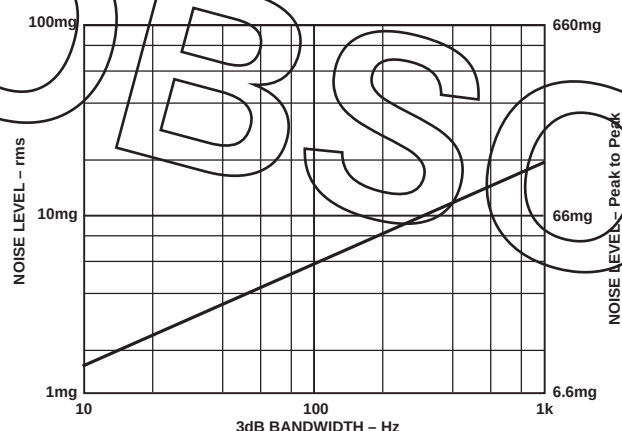


Figure 24. Noise Level vs. 3 dB Bandwidth

The output noise of the ADXL05 scales with the square root of the measurement bandwidth. With a single pole roll-off, the equivalent rms noise bandwidth is π divided by 2 or approximately 1.5 times the 3 dB bandwidth. For example, the typical rms noise of the ADXL05J using a 100 Hz one pole post filter is:

$$\text{Noise (rms)} = 500 \mu\text{g}/\sqrt{\text{Hz}} \times \sqrt{100(1.5)} = 6,124 \mu\text{g} \text{ or } 6.1 \text{ mg rms}$$

For the bandpass filter of Figure 27 where both ac coupling and low pass filtering are used, the low frequency roll-off, F_L , is determined by C4 and R1 and the high frequency roll-off, F_H , is determined by the 1-pole post filter R3, C5.

The equivalent rms noise of the bandpass filter is equal to $500 \mu\text{g}/\sqrt{\text{Hz}} \times \sqrt{(1.5 F_H) - (F_L/1.5)}$.

For example, the typical rms noise of the ADXL05 using 1 pole ac coupling with a bandwidth of 10 Hz and 1 pole low-pass filter of 100 Hz is:

$$\begin{aligned} \text{Noise (rms)} &= 500 \mu\text{g}/\sqrt{\text{Hz}} \times \sqrt{1.5(100) - (10/1.5)} \\ &= 5,987 \mu\text{g rms or } \approx 5.9 \text{ mg rms} \end{aligned}$$

Because the ADXL05'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 II is useful for estimating the probabilities of exceeding various peak values, given the rms value.

Table II.

Nominal Peak-to-Peak Value	% of Time that Noise Will Exceed Nominal Peak-to-Peak Value
$2.0 \times \text{rms}$	32%
$4.0 \times \text{rms}$	4.6%
$6.0 \times \text{rms}$	0.27%
$6.6 \times \text{rms}$	0.1%
$8.0 \times \text{rms}$	0.006%

RMS and peak-to-peak noise (for 0.1% uncertainty) for various bandwidths is estimated in Figure 24. As shown by the figure, device noise drops dramatically as the operating bandwidth is reduced. For example, when operated in a 1 kHz bandwidth, the ADXL05 typically has an rms noise level of 19 mg. With $\pm 5 g$ applied accelerations, this 19 mg resolution limit is normally quite satisfactory; but for smaller acceleration levels the noise is now a much greater percentage of the signal. As shown by the figure, when the device bandwidth is rolled off to 100 Hz, the noise level is reduced to approximately 6 mg, and at 10 Hz it is down to less than 2 mg.

Alternatively, the signal-to-noise ratio may be improved considerably by using a microprocessor to perform multiple measurements and then compute the average signal level. When using this technique, with 100 measurements, the signal-to-noise ratio will be increased by a factor of 10 (20 dB).

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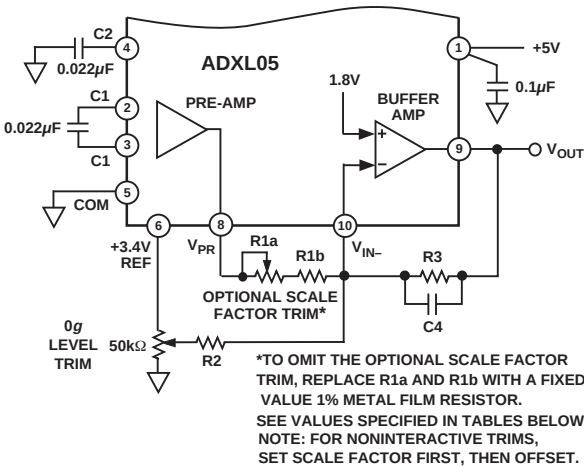
Low-Pass Filtering

The bandwidth of the accelerometer can easily be reduced by using post filtering. Figure 25 shows how the buffer amplifier can be connected to provide 1-pole post filtering, 0 g offset trimming, and output scaling. The table provides practical component values for various full-scale g levels and approximate circuit bandwidths. For bandwidths other than those listed, use the formula:

C4 = 1 / (2 * pi * R3 * Desired 3 dB Bandwidth in Hz)

or simply scale the value of capacitor C4 accordingly, i.e., for an application with a 50 Hz bandwidth, the value of C4 will need to be twice as large as its 100 Hz value. If further noise reduction is needed while maintaining the maximum possible bandwidth, then a 2- or 3-pole post filter is recommended. These provide a much steeper roll-off of noise above the pole frequency. Figure 26 shows a circuit that uses the buffer amplifier to provide 2-pole post filtering. Component values for the 2-pole filter were selected to operate the buffer at unity gain. Capacitors C3 and C4 were chosen to provide 3 dB bandwidths of 10 Hz, 30 Hz, 100 Hz, and 300 Hz.

In this configuration, the nominal buffer amplifier output will be +1.8 V ± the 200 mV/g scale factor of the accelerometer. An AD820 external op amp allows noninteractive adjustment of 0 g offset and scale factor. 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.

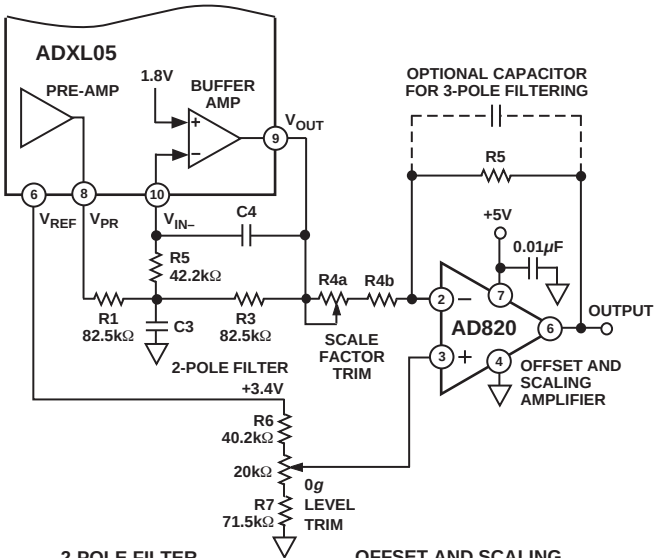


COMPONENT VALUES FOR VARIOUS FULL-SCALE RANGES AND BANDWIDTHS

FULL SCALE	mV per g	3dB BW (Hz)	R1a kΩ	R1b kΩ	R3 kΩ	R2 kΩ	C4 μF
±1g	2000	10	10	24.9	301	100	0.056
±2g	1000	100	10	35.7	200	100	0.0082
±4g	500	200	10	35.7	100	100	0.0082
±5g	400	300	10	45.3	100	100	0.0056

3dB BW = 1 / (2 * pi * R3 * C4)

Figure 25. Using the Buffer Amplifier to Provide 1-Pole Post Filtering Plus Scale Factor and 0 g Level Trimming



2-POLE FILTER COMPONENT VALUES

3dB BW (Hz)	C3 μF	C4 μF
300	0.027	0.0033
100	0.082	0.01
30	0.27	0.033
10	0.82	0.1

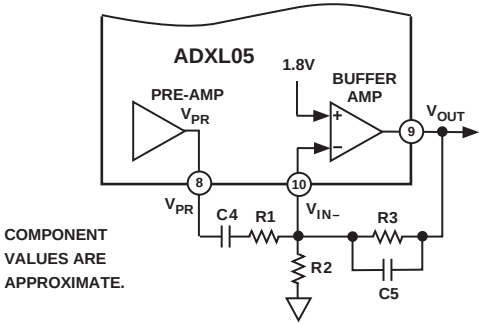
OFFSET AND SCALING AMPLIFIER COMPONENT VALUES

FULL SCALE	mV per g	GAIN	R4a kΩ	R4b kΩ	R5 kΩ
±1g	2000	10.00	10	24.9	301
±2g	1000	4.98	10	35.7	200
±4g	500	2.50	10	35.7	100
±5g	400	2.00	10	45.3	100

Figure 26. Two-Pole Filtering Circuit with Gain and 0 g Offset Adjustment

Bandpass Filtering

Figure 27 shows how the combination of ac coupling and low-pass filtering together form a bandpass filter that provides an even greater improvement in noise reduction.



SCALE FACTOR IN mV/g	DESIRED LOW FREQUENCY LIMIT, F _L	R1 VALUE IN kΩ	CLOSEST C4 VALUE	DESIRED HIGH FREQUENCY LIMIT, F _H	R3 IN kΩ	CLOSEST C5 VALUE	VALUE OF R2 FOR +2.5V 0g LEVEL
1000	30	49.9	0.10 μF	300	249	0.002 μF	640kΩ
200	30	249	0.022 μF	300	249	0.002 μF	640kΩ
1000	3	49.9	1.0 μF	100	249	0.0068 μF	640kΩ
200	1	249	0.68 μF	100	249	0.0068 μF	640kΩ
200	0.1	249	6.8 μF	10	249	0.068 μF	640kΩ

Figure 27 AC Coupling and Low-Pass Filtering Used Together to Provide a Bandpass Function

Additional Noise Reduction Techniques

Shielded wire should be used for connecting the accelerometer to any circuitry that is more than a few inches away—to avoid 60 Hz pickup from ac line voltage. Ground the cable's shield at only one end and connect a separate common lead between the circuits; this will help to prevent ground loops. Also, if the accelerometer is inside a metal enclosure, this should be grounded as well.

Methods for Reducing 0 g Offset Drift

When using any accelerometer with a dc (gravity sensing) response, the 0 g offset level will exhibit some temperature drift. For very high accuracy applications, one very straightforward approach is to use a low cost crystal oven to maintain the accelerometer at a constant temperature. These ovens are available in a variety of different temperatures. After the circuit has been built and is operating correctly, the crystal oven can be mounted over the accelerometer and powered off the same +5 V power supply.

The ovens may be purchased from Isotemp Research, Inc., P.O. Box 3389, Charlottesville, VA 22903, phone 804-295-3101. For more details on crystal oven compensation, refer to application note AN-385.

Other methods for 0 g drift compensation include using a low cost temperature sensor such as the AD22100 to supply a microprocessor with the device temperature. If the drift curve of the accelerometer is stored in the μP , then a software program can be used to subtract out the drift. Alternatively, a simple 1st order (straight line) correction circuit can be used to subtract out the linear portion of the accelerometer's drift by using a temperature sensor and op amp to supply a small compensation current. For more details on software and hardware drift compensation, refer to application note AN-380.

ACCELEROMETER APPLICATIONS

Popular applications for low g accelerometers tend to fall into three categories: measurement of tilt and orientation, inertial measurement of acceleration, velocity and distance, and vibration or shock measurement.

The ADXL05 is a "dc" accelerometer, meaning that it is capable of measuring static accelerations such as the Earth's gravity. The ADXL05 differs from other acceleration measurement technologies such as piezoelectric and piezofilm sensors which can only respond to ac signals greater than approximately 1 Hz. This dc capability is required for tilt and inertial measurement. For ac shock or vibration the ADXL05 can measure frequencies of up to 4 kHz and has the added benefit of measuring all the way down to dc.

Using the ADXL05 in Tilt Applications

The ADXL05's precision dc characteristics make it suitable for tilt measurement. It can directly measure the Earth's gravity and use this constant force as a position reference to determine inclination. As shown in Figure 28, the accelerometer should be mounted so that its sensitive axis is perpendicular to the force of gravity, i.e., parallel to the Earth's surface. In this manner, it will be most sensitive to changes in orientation (when it is oriented 90° to the force of gravity). Its output can be then described by the sine function; a tilt occurring at an angle θ will cause a voltage output equal to:

$$V_{OUT} = \left[\text{Accelerometer Scale Factor} \left(\frac{V}{g} \right) \times \sin(\theta) \times 1g \right] + \text{zero } g \text{ output}(V)$$

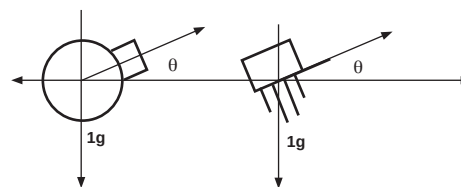


Figure 28. Two Possible Orientations for Tilt Measurement

Conversely, for a given acceleration signal and assuming no other changes in the axis or interfering signals, the tilt angle is proportional to the voltage output as shown in Figure 29. The angle, θ can be calculated using:

$$\theta = \arcsin \left(1g \times \left[\frac{V_{OUT} - \text{zero } g \text{ output}(V)}{\text{Scale Factor}(V/g)} \right] \right)$$

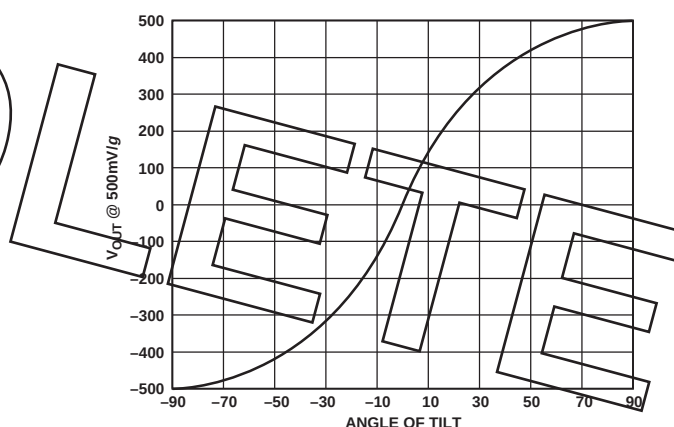


Figure 29. V_{OUT} vs. Tilt Angle

The use of an accelerometer in tilt applications has several advantages over the use of a traditional tilt sensor. A traditional tilt sensor consists of glass vial filled with a conductive liquid, typically a mercury or electrolytic solution. Besides being larger than an XL05, it requires additional signal conditioning circuitry. The settling time and frequency response is limited by the amount of time required for the liquid to stop sloshing around in the vial. In high vibration environments, or where high lateral accelerations may be present, it may not be possible to resolve the tilt signal above the "slosh" noise. The accelerometer has faster frequency (up to 50 x) response and settling time. Interfering vibrations may be filtered out if necessary, an impossibility with a liquid tilt sensor, since one cannot filter the liquid. Finally, in the presence of lateral accelerations, an accelerometer provides more useful information, i.e., an acceleration signal, which if cleverly signal processed, can provide both a tilt and an acceleration output. A single accelerometer can be used to measure tilt over a 180° range; two accelerometers gives a complete 360° of measurement.

An important characteristic for an accelerometer used in a tilt application is its 0 g offset stability over temperature. The ADXL05 typically exhibits offsets that deviate no more than 0.1 g over the 0°C to +70°C temperature range, corresponding

ADXL05

to a 5° tilt error over the entire temperature range. Straight-forward calibration schemes discussed in this data sheet may be used to reduce or compensate for temperature drift to improve the absolute accuracy of the measurement.

Using the ADXL05 in Inertial Measurement Applications

Inertial measurement refers to the practice of measuring acceleration for the purpose of determining the velocity of an object and its change in position, or distance traveled. This technique has previously required expensive inertial guidance systems of the type used in commercial aircraft and military systems. The availability of a low cost precision dc accelerometer such as the ADXL05 enables the use of inertial measurement for more cost sensitive industrial and commercial applications.

Inertial measurement makes use of the fact that the integral of acceleration is velocity and the integral of velocity is distance. By making careful measurements of acceleration and mathematically integrating the signals, one can determine both velocity and the distance traveled. The technique is useful for applications where a traditional speed and distance measurement is impractical, or where a non-contact, relative position measurement must be made.

A practical inertial measurement system uses multiple accelerometers to measure acceleration in three axes, and gyroscopes to measure rotation in three axes, the requirement for a 6 degree of freedom system. For simpler systems where one or more of the axes can be constrained, it is possible to build a system with fewer accelerometers and gyros.

The measurement system must take the acceleration sensor and calibrate out all static errors including any initial inaccuracy or temperature drift. A mathematical model is used to describe the performance of the sensor in order to calibrate it. If these errors are not removed, then the process of double integration will quickly cause any small error to dominate the result. Most practical systems use microprocessors for error correction and a temperature sensor for temperature drift compensation. Another approach is to maintain all of the sensors at a controlled temperature. The microprocessors have the additional advantage of providing a low cost method of performing the single and double integration of the acceleration signal.

The stability and repeatability of the accelerometer is the most important specification in an inertial system. The ADXL05 is "well behaved" that is, its response and temperature characteristics are easy to model and correct, and once modeled they are very repeatable. For example, temperature performance can be adequately modeled using first order, (straight line) approximations for most applications, and other errors such as on-axis and pendulous rectification are minimal. This greatly simplifies the math required to correct the sensor.

Vibration and Shock Measurement Applications

The ADXL05 can measure shocks and vibrations from dc to 4 kHz. Typical signal processing for vibration signals includes fast Fourier transforms, and single and double integration for velocity and displacement. It is possible to build a single integrator stage using the ADXL05's output buffer amplifier in order to provide a velocity output.

The sensitivity of the accelerometer will typically vary only $\pm 0.5\%$ over the full industrial temperature range, making it one of the most stable vibration measurement devices available. In vibration measurement applications, mechanical mounting and

control of system and mounting resonances are critical to proper measurement. Refer to the application note AN-379, available from Analog Devices.

CALIBRATING THE ADXL05

If a calibrated shaker is not available, both the 0 g level and scale factor of the ADXL05 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 30 shows how gravity and package orientation affect the ADXL05's output. 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 ADXL05 should register a 1 g acceleration, either positive or negative, depending on orientation. With the axis of sensitivity in the horizontal plane, no acceleration (the 0 g bias level) should be indicated.

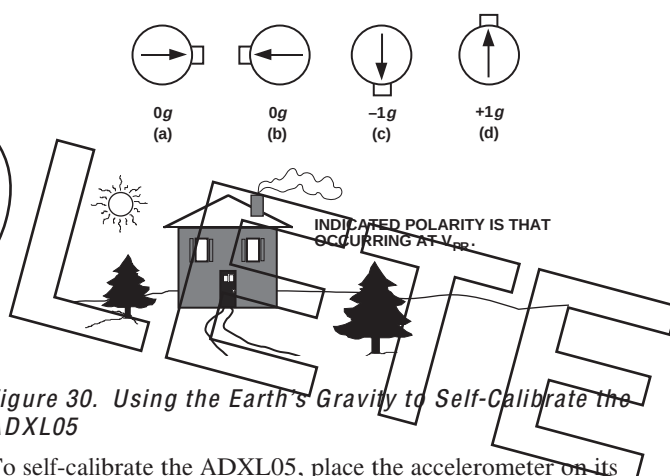


Figure 30. Using the Earth's Gravity to Self-Calibrate the ADXL05

To self-calibrate the ADXL05, 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 (see Figure 25).

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 gs in acceleration. For example, if the circuit scale factor at the buffer output is 400 mV per g, then the scale factor trim should be adjusted so that an output change of 800 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.

REDUCING POWER CONSUMPTION

The use of a simple power cycling circuit provides a dramatic reduction in the ADXL05's average current consumption. In low bandwidth applications such as shipping recorders, this simple, low cost circuit can provide substantial power reduction.

If a microprocessor is available, only the circuit of Figure 31 is needed. The microprocessor supplies a TTL clock pulse to gate buffer transistor Q1 which inverts the output pulse from the μP

so that the duty cycle is correct when the pulse is re-inverted again by transistor, Q2, which cycles the accelerometer's supply voltage on and off.

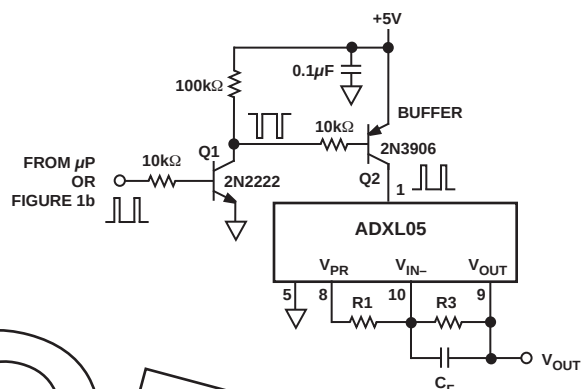


Figure 31. Basic Power Cycling Circuit

Figures 32 and 33 show typical waveforms of the accelerometer being operated with a 10% duty cycle: 1 ms on, 9 ms off. This reduces the average current consumption of the accelerometer from 8 mA to 800 μ A, providing a power reduction of 90%. The μ P should sample acceleration during the interval between the time the 0 g level has stabilized (≈ 400 μ s using a 0.022 μ F demod cap) and the end of the pulse duration. The measurement bandwidth of a power-cycled circuit will be set by the clock pulse rate and duty cycle. In this example, 1 sample can be taken every 10 ms which is 100 samples per second or 100 Hz. As defined by the "Nyquist criteria," the best case measurement bandwidth is $F_s/2$ or half the clock frequency. Therefore 50 Hz signals can be processed if adequate filtering is provided.

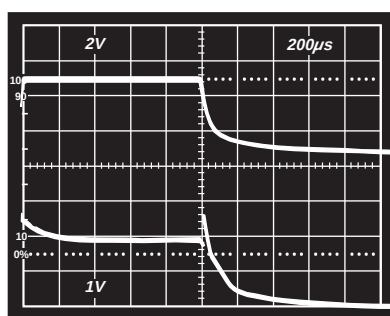


Figure 32. Top Trace: Voltage at Pin 1; Bottom Trace: Output at V_{PR}

Higher measurement bandwidths can be achieved by reducing the size of the demodulation capacitor below 0.022 μ F and increasing the pulse frequency. A 0.01 μ F capacitor was connected across the feedback resistor of the ADXL05 buffer to improve its transient characteristics. The optimum value for this capacitor will change with buffer gain and the cycling pulse rate. For more details, refer to application note AN-378.

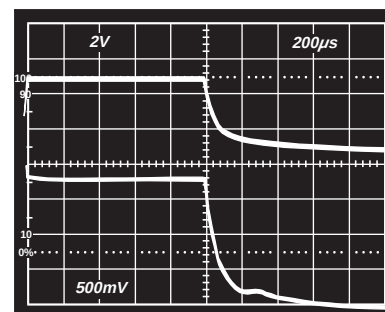


Figure 33. Top Trace: Voltage at Pin 1; Bottom Trace: Buffer Output With $R1 = R3 = 100$ k Ω , $C_F = 0.01$ μ F

COMPONENT SELECTION

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 ± 5 g applications, the resistor R1 from V_{PR} to V_{IN-} is recommended to have a value greater than 20 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

ADXL05

output change at V_{PR} . If the ADXL05 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 $\pm 15\%$ 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 ADXL05's buffer amplifier at Gains > 2 , to provide full-scale outputs of less than ± 5 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 ADXL05 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 ADXL05 power supply should be decoupled with a $0.1 \mu\text{F}$ ceramic capacitor from +5 V pin of the ADXL05 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 ADXL05. A ceramic capacitor with a minimum value of $0.022 \mu\text{F}$ is recommended from the oscillator decoupling capacitor pin to common. Small amounts of capacitor leakage due to a dc resistance greater than $1 \text{ M}\Omega$ 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 $0.022 \mu\text{F}$ is a good typical value, it may be increased for reasons of convenience, but doing this will not improve the noise performance of the ADXL05.

Demodulator Capacitor, C1

The demodulator capacitor is connected across Pins 2 and 3 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 ADXL05 exhibits a single pole roll-off response, see Figure 4.

A nominal value of $0.022 \mu\text{F}$ 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, to preserve stability C1 should be kept $> 0.01 \mu\text{F}$.

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 $0.022 \mu\text{F}$ capacitor across the demodulator pins and perform any additional low-pass filtering using the buffer amplifier. 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 ADXL05 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 $30 \text{ M}\Omega$ 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.

MINIMIZING EMI/RFI

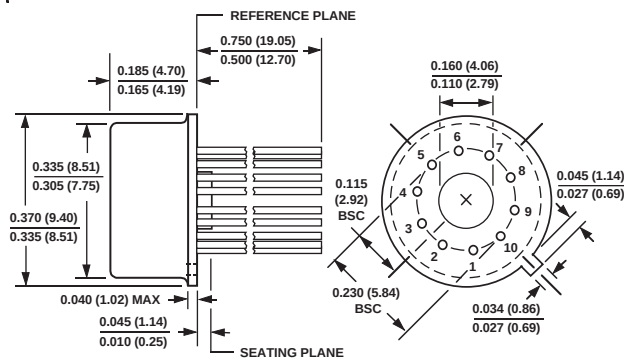
The architecture of the ADXL05 and its use of synchronous demodulation make the device immune to most electromagnetic (EMI) and radio frequency (RFI) interference. The use of synchronous demodulation allows the circuit to reject all signals except those at the frequency of the oscillator driving the sensor element. However, the ADXL05 does have a sensitivity to RFI that is within ± 5 kHz of the internal oscillator's nominal frequency of 1 MHz and also to any odd harmonics of this frequency. The internal oscillator frequency will exhibit part to part variation in the range of 0.5 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 or at odd harmonics. 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. Good grounding practices include having separate analog and digital grounds (as well as separate power supplies or very good decoupling) on the printed circuit boards. A single ground line shared by both the digital and analog circuitry can lead to digital pulses (and clock signals) interfering with the sensor's onboard oscillator. In extreme cases, a low cost radio frequency choke ($\approx 10 \mu\text{H}$) may be needed in series with the accelerometer's power supply pin. This, together with the recommended $0.1 \mu\text{F}$ power supply bypass capacitor, will form an effective RF filter. The use of an RF choke is preferred over a resistor since any series resistance in the power supply will "unregulate" the device from the supply, degrade its power supply rejection and reduce its supply voltage.

OUTLINE DIMENSIONS

Dimensions shown in inches and (mm).



OBSOLETE

OBSOLETE

EXHIBIT 1064B



Multi-Purpose Accelerometer Measures Low g Forces with 0.005g Resolution [ADXL05]

NORWOOD, Mass. (May 1, 1995) -- Analog Devices' ADXL05 is the industry's first single-chip integrated low-g accelerometer which resolves minute changes in acceleration (from 0 g to ± 5 g full scale) with 0.005 g resolution-ten times more precise than the ADXL50 ± 50 g airbag version. More than a sensor, the ADXL05 contains all the signal processing blocks of a complete acceleration system on a single chip. It combines a tiny micromachined sensor, modulator, demodulator, voltage reference, signal conditioning, amplification, and on-command self-test circuitry on one polysilicon die.

The ADXL05 is designed to simplify system complexity, improve performance, and reduce engineering costs in automotive, industrial, and commercial applications where tilt, inertial, shock, vibration, velocity and acceleration characteristics must be measured. Unlike piezoelectric devices, the ADXL05 provides a true dc (gravity) response for precision tilt and inertial measurement systems. Piezoresistive devices are less precise and more expensive.

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Analog Devices' ± 50 g accelerometer, the ADXL50, is now used in a variety of 1995 General Motors cars as the crash sensor in airbag deployment systems. The same high-volume, proven reliability benefits of this 50 g device are now available in the new 5 g ADXL05 version.

The ADXL05 is packaged in a hermetic 10-pin TO-100 metal can with guaranteed specifications over the 0 to +70°C commercial, and -40° to +85°C industrial temperature ranges; delivery is from stock. Contact the factory for the availability of devices for operation over the -40° to +105°C automotive and -55° to +125°C military temperature ranges.

With fiscal 1994 revenues of \$773 million, Analog Devices is a Fortune 500 company that designs, manufactures, and markets precision high-performance integrated circuits used in analog and digital signal processing applications. The company employs approximately 5,500 people worldwide and has manufacturing facilities in Massachusetts, California, North Carolina, Ireland, Japan, the Philippines and Taiwan.

Revised March 4, 1997

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EXHIBIT 1064C

ADXL05

Description

Datasheets

Press Releases

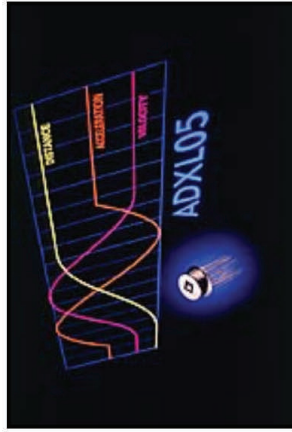
ADXL05

General Description

The ADXL05 is a complete acceleration measurement system on a single monolithic IC. The ADXL05 will measure accelerations with full-scale ranges of ± 5 g to ± 1 g or less. Typical noise floor is $500 \mu\text{g}/(\text{root})\text{Hz}$, (12x less than the ADXL50), allowing signals below 5 milli-g to be resolved. The ADXL05 is a force balanced capacitive accelerometer with the capability to measure both ac accelerations (typical of vibration) or dc accelerations (such as inertial force or gravity). Three external capacitors and a ± 5 volt regulated power supply are all that is required to measure accelerations up to ± 5 g. Three resistors are used to configure the output buffer amplifier to set scale factors from 200 mV/g to 1 V/g . External capacitors may be added to the resistor network to provide 1 or 2 poles of filtering. No additional active components are required to interface directly to most analog to digital converters (ADCs).

The device features a TTL compatible self-test function that can electrostatically deflect the sensor beam at any time to verify that the sensor and its electronics are functioning correctly.

The ADXL05 is available in a hermetic 10-pin TO-100 metal can, specified over the 0°C to $+70^{\circ}\text{C}$ commercial, and -40°C to $+85^{\circ}\text{C}$ industrial temperature ranges.



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ADXL05

- Description
- Datasheets
- Press Releases

the sensor and its electronics are functioning correctly.

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Datasheets

Listed below is the datasheet for this product. This datasheet is available in Adobe Acrobat PDF format. To view Acrobat files you must install Adobe Acrobat Reader. Acrobat Reader is available free of charge. To download Adobe Acrobat visit their [website](#).



- [±1g to ±5g Single Chip Accelerometer with Signal Conditioning](#)

Press Releases

Listed below are any Analog press releases that pertain to this product. Press Releases contain the latest news and information about specific products.

- [Analog Devices Dedicates New FAB to Micromachined Products](#)
- [Multi-Purpose Accelerometer Measures Low g Forces with 0.005g Resolution](#)
- [ADXL Accelerometer Evaluation Kits Simplify System Development, Reduce Design Time](#)



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EXHIBIT 1065A

Monolithic Accelerometer With Signal Conditioning

ADXL50*

Complete Acceleration Measurement System

on a Single Monolithic IC

Full-Scale Measurement Range: ± 50 g

Self-Test on Digital Command

+5 V Single Supply Operation

Sensitivity Precalibrated to 19 mV/g

Internal Buffer Amplifier for User Adjustable Sensitivity and Zero-q Level

Frequency Response: DC to 10 kHz

Post Filtering with External Passive Components

For convenience, the ADXL50 has an internal buffer amplifier with a full 0.25 V to 4.75 V output range. This may be used to offset the zero-g level and change the output sensitivity by using external resistors. External capacitors may be added to the resistor network to provide 1 or 2 poles of filtering. No external active components are required to interface directly to most analog-to-digital converters (ADCs) or microcontrollers.

The ADXL50 uses a capacitive measurement method. The analog output voltage is directly proportional to acceleration, and is fully scaled, referenced and temperature compensated, resulting



ADXL50—SPECIFICATIONS ($T_A = T_{MIN}$ to T_{MAX} , $T_A = +25^{\circ}\text{C}$ for J Grade Only, $V_S = +5\text{ V}$, @ Acceleration = 0 g , unless otherwise noted)

Parameter	Conditions	ADXL50J/A			Units
		Min	Typ	Max	
SENSOR INPUT					
Measurement Range	Guaranteed Full Scale	–50		+50	g
Nonlinearity	Best Fit Straight Line, 50 g FS		0.2		% of FS
Alignment Error ¹			± 1		Degrees
Transverse Sensitivity ²			± 2		%
SENSITIVITY					
Initial Sensitivity at V_{PR}	+25°C	16.1	19.0	21.9	mV/g
Temperature Drift ³			0.75/1.0		% of Reading
ZERO g BIAS LEVEL					
Initial Offset	at V_{PR}	1.55/1.60	1.80	2.05/2.00	V
vs. Temperature ³			$\pm 15/35$		mV
vs. Supply	$V_S = 4.75\text{ V}$ to 5.25 V		10	32	mV/V
NOISE PERFORMANCE					
Voltage Noise Density	at V_{PR}		6.6	12	$\text{mg}/\sqrt{\text{Hz}}$
Noise in 100 Hz Bandwidth	BW = 10 Hz to 1 kHz		66		mg rms
Noise in 10 Hz Bandwidth			20		mg rms
FREQUENCY RESPONSE					
3 dB Bandwidth ⁴	$C1 = 0.022\text{ }\mu\text{F}$ (See Figure 22)	800	1300		Hz
3 dB Bandwidth ⁴	$C1 = 0.0068\text{ }\mu\text{F}$		10		kHz
Sensor Resonant Frequency			24		kHz
SELF TEST INPUT					
Output Change at V_{PR} ⁵	ST Pin from Logic “0” to “1”	–0.85	–1.00	–1.15	V
Logic “1” Voltage		2.0			V
Logic “0” Voltage				0.8	V
Input Resistance	To Common		50		k Ω
+3.4 V REFERENCE					
Output Voltage		3.350	3.400	3.450	V
Output Temperature Drift ³			± 10		mV
Power Supply Rejection	DC, $V_S = +4.75\text{ V}$ to $+5.25\text{ V}$		1	10	mV/V
Output Current	Sourcing	500			μA
PREAMPLIFIER OUTPUT					
Voltage Swing		0.25		$V_S - 1.4$	V
Current Output	Source or Sink	30	80		μA
Capacitive Load Drive			100		pF
BUFFER AMPLIFIER					
Input Offset Voltage ⁶	Delta from Nominal 1.800 V		± 10	± 25	mV
Input Bias Current			5	20	nA
Open-Loop Gain	DC		80		dB
Unity Gain Bandwidth			200		kHz
Output Voltage Swing	$I_{OUT} = \pm 100\text{ }\mu\text{A}$	0.25		$V_S - 0.25$	V
Capacitive Load Drive		1000			pF
Power Supply Rejection	DC, $V_S = +4.75\text{ V}$ to $+5.25\text{ V}$		1	10	mV/V
POWER SUPPLY					
Operating Voltage Range		4.75		5.25	V
Quiescent Supply Current			10	13	mA
TEMPERATURE RANGE					
Operating Range J		0		+70	°C
Specified Performance A		–40		+85	°C
Automotive Grade*		–40		+125	°C

NOTES

¹Alignment error is specified as the angle between the true and indicated axis of sensitivity, (see Figure 2).

²Transverse sensitivity is measured with an applied acceleration that is 90° from the indicated axis of sensitivity. Transverse sensitivity is specified as the percent of transverse acceleration that appears at the V_{PR} output. This is the algebraic sum of the alignment and the inherent sensor sensitivity errors, (see Figure 2).

³Specification refers to the maximum change in parameter from its initial at +25°C to its worst case value at T_{MIN} to T_{MAX} .

⁴Frequency at which response is 3 dB down from dc response assuming an exact C1 value is used. Maximum recommended BW is 10 kHz using a 0.007 μF capacitor, refer to Figure 22.

⁵Applying logic high to the self-test input has the effect of applying an acceleration of –52.6 g to the ADXL50.

⁶Input offset voltage is defined as the output voltage differential from 1.800 V when the amplifier is connected as a follower (i.e., Pins 9 and 10 tied together). The voltage at Pin 9 has a temperature drift proportional to that of the 3.4 V reference.

*Contact factory for availability of automotive grade devices.

All min and max specifications are guaranteed. Typical specifications are not tested or guaranteed.
Specifications subject to change without notice.

ABSOLUTE MAXIMUM RATINGS*

Acceleration (Any Axis, Unpowered for 0.5 ms)	2000 g
Acceleration (Any Axis, Powered for 0.5 ms)	500 g
+V _S	−0.3 V to +7.0 V
Output Short Circuit Duration	
(V _{PR} , V _{OUT} , V _{REF} Terminals to Common)	Indefinite
Operating Temperature	−55°C to +125°C
Storage Temperature	−65°C to +150°C

*Stresses above those listed under “Absolute Maximum Ratings” may cause permanent damage to the device. This is a stress rating only; the functional operation of the device at these or any other conditions above those indicated in the operational sections of this specification is not implied. Exposure to absolute maximum rating conditions for extended periods may affect device reliability.

Package Characteristics

Package	θ _{JA}	θ _{JC}	Device Weight
10-Pin TO-100	130°C/W	30°C/W	5 Grams

ORDERING GUIDE

Model	Temperature Range
ADXL50JH	0°C to +70°C
ADXL50AH	−40°C to +85°C

CAUTION

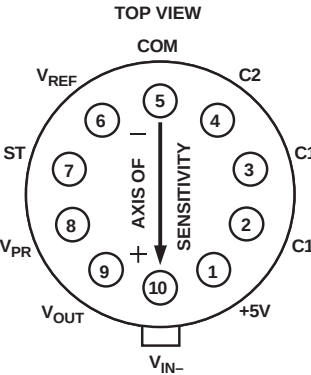
ESD (electrostatic discharge) sensitive device. Electrostatic charges as high as 4000 V readily accumulate on the human body and test equipment and can discharge without detection. Although the ADXL50 features proprietary ESD protection circuitry, permanent damage may occur on devices subjected to high energy electrostatic discharges. Therefore, proper ESD precautions are recommended to avoid performance degradation or loss of functionality.



PIN DESCRIPTION

+5 V	The power supply input pin.
C2	Connection for an external bypass capacitor (nominally 0.022 μF) used to prevent oscillator switching noise from interfering with other ADXL50 circuitry. Please see the section on component selection.
C1	Connections for the demodulator capacitor, nominally 0.022 μF. See the section on component selection for application information.
COM	The power supply common (or “ground”) connection.
V _{REF}	Output of the internal 3.4 V voltage reference.
ST	The digital self-test input. It is both CMOS and TTL compatible.
V _{PR}	The ADXL50 preamplifier output providing an output voltage of 19 mV per g of acceleration.
V _{OUT}	Output of the uncommitted buffer amplifier.
V _{IN−}	The inverting input of the uncommitted buffer amplifier.

CONNECTION DIAGRAM
10-Header (TO-100)



NOTES:

AXIS OF SENSITIVITY IS ALONG A LINE BETWEEN PIN 5 AND THE TAB.

THE CASE OF THE METAL CAN PACKAGE IS CONNECTED TO PIN 5 (COMMON).

ARROW INDICATES DIRECTION OF POSITIVE ACCELERATION ALONG AXIS OF SENSITIVITY.

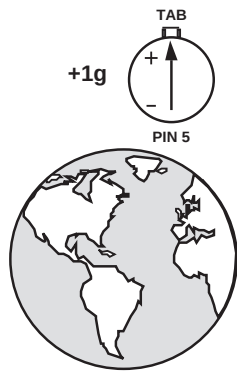


Figure 1. Output Polarity at V_{PR}

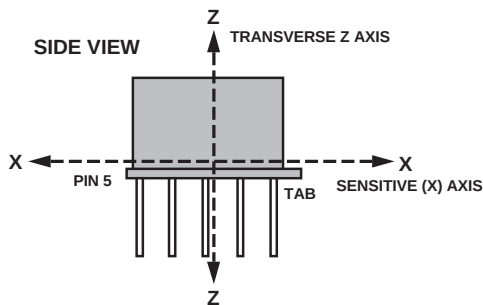


Figure 2a. Sensitive X and Transverse Z Axis

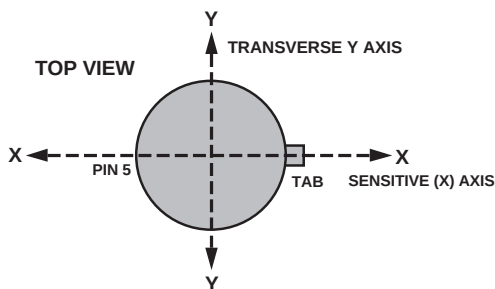


Figure 2b. Sensitive X and Transverse Y Axis

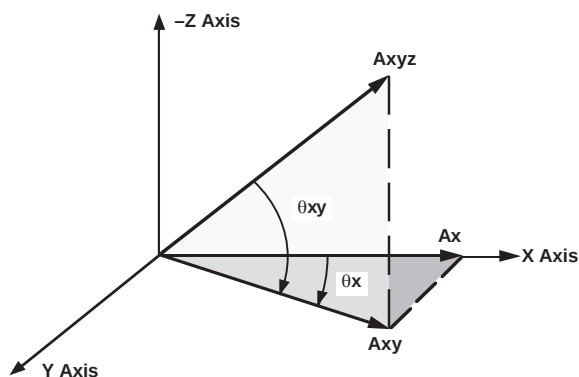


Figure 2c. A Vector Analysis of an Acceleration Acting Upon the ADXL50 in Three Dimensions

Polarity of the Acceleration Output

The polarity of the ADXL50 output is shown in the Figure 1. When oriented to the earth's gravity (and held in place), the ADXL50 will experience an acceleration of +1 g. This corresponds to a change of approximately +19 mV at the V_{PR} output pin. Note that the polarity will be reversed to a negative going signal at the buffer amplifier output V_{OUT} , due to its inverting configuration.

Mounting Considerations

There are three main causes of measurement error when using accelerometers. The first two are alignment and transverse sensitivity errors. The third source of error is due to resonances or vibrations of the sensor in its mounting fixture.

Errors Due to Misalignment

The ADXL50 is a sensor designed to measure accelerations that result from an applied force. Because these forces act on the sensor in a vector manner, the alignment of the sensor to the force to be measured may be critical.

The ADXL50 responds to the component of acceleration on its sensitive X axis. Figures 2a and 2b show the relationship between the sensitive "X" axis and the transverse "Z" and "Y" axes as they relate to the TO-100 package.

Figure 2c describes a three dimensional acceleration vector (A_{XYZ}) which might act on the sensor, where A_X is the component of interest. To determine A_X , first, the component of acceleration in the XY plane (A_{XY}) is found using the cosine law:

$$A_{XY} = A_{XYZ} (\cos\theta_{XY}) \text{ then}$$

$$A_X = A_{XY} (\cos\theta_X)$$

$$\text{Therefore: Typical } V_{PR} = 19 \text{ mV/g } (A_{XYZ}) (\cos\theta_{XY}) \cos\theta_X$$

Note that an ideal sensor will react to forces along or at angles to its sensitive axis but will reject signals from its various transverse axes, i.e., those exactly 90° from the sensitive "X" axis. But even an ideal sensor will produce output signals if the transverse signals are not exactly 90° to the sensitive axis. An acceleration that is acting on the sensor from a direction different from the sensitive axis will show up at the ADXL50 output at a reduced amplitude.

Table I. Ideal Output Signals for Off Axis Applied Accelerations Disregarding Device Alignment and Transverse Sensitivity Errors

θ_X	% of Signal Appearing at Output	Output in gs for a 50 g Applied Acceleration
0	100%	50 (On Axis)
1°	99.98%	49.99
2°	99.94%	49.97
3°	99.86%	49.93
5°	99.62%	49.81
10°	98.48%	49.24
30°	86.60%	43.30
45°	70.71%	35.36
60°	50.00%	25.00
80°	17.36%	8.68
85°	8.72%	4.36
87°	5.25%	2.63
88°	3.49%	1.75
89°	1.7%	0.85
90°	0%	0.00 (Transverse Axis)

Table I shows the percentage signals resulting from various θ_x angles. Note that small errors in alignment have a negligible effect on the output signal. A 1° error will only cause a 0.02% error in the signal. Note, however, that a signal coming 1° off of the transverse axis (i.e., 89° off the sensitive axis) will still contribute 1.7% of its signal to the output. Thus large transverse signals could cause output signals as large as the signals of interest.

Table I may also be used to approximate the effect of the ADXL50's internal errors due to misalignment of the die to the package. For example: a 1 degree sensor alignment error will allow 1.7% of a transverse signal to appear at the output. In a nonideal sensor, transverse sensitivity may also occur due to inherent sensor properties. That is, if the sensor physically moves due to a force applied exactly 90° to its sensitive axis, then this might be detected as an output signal, whereas an ideal sensor would reject such signals. In every day use, alignment errors may cause a small output peak with accelerations applied close to the sensitive axis but the largest errors are normally due to large accelerations applied close to the transverse axis.

Errors Due to Mounting Fixture Resonances

A common source of error in acceleration sensing is resonance of the mounting fixture. For example, the circuit board that the ADXL50 mounts to may have resonant frequencies in the same range as the signals of interest. This could cause the signals measured to be larger than they really are. A common solution to this problem is to dampen these resonances by mounting the ADXL50 near a mounting post or by adding extra screws to hold the board more securely in place.

When testing the accelerometer in your end application, it is recommended that you test the application at a variety of frequencies in order to ensure that no major resonance problems exist.

GLOSSARY OF TERMS

Acceleration: Change in velocity per unit time.

Acceleration Vector: Vector describing the net acceleration acting upon the ADXL50 (A_{XYZ}).

g: A unit of acceleration equal to the average force of gravity occurring at the earth's surface. A g is approximately equal to 32.17 feet/s², or 9.807 meters/s².

Nonlinearity: The maximum deviation of the ADXL50 output voltage from a best fit straight line fitted to a plot of acceleration vs. output voltage, calculated as a % of the full-scale output voltage (@ 50 g).

Resonant Frequency: The natural frequency of vibration of the ADXL50 sensor's central plate (or "beam"). At its resonant frequency of 24 kHz, the ADXL50's moving center plate has a peak in its frequency response with a Q of 3 or 4.

Sensitivity: The output voltage change per g unit of acceleration applied, specified at the V_{PR} pin in mV/g.

Sensitive Axis (X): The most sensitive axis of the accelerometer sensor. Defined by a line drawn between the package tab and Pin 5 in the plane of the pin circle. See Figures 2a and 2b.

Sensor Alignment Error: Misalignment between the ADXL50's on-chip sensor and the package axis, defined by Pin 5 and the package tab.

Total Alignment Error: Net misalignment of the ADXL50's on-chip sensor and the measurement axis of the application. This error includes errors due to sensor die alignment to the package, and any misalignment due to installation of the sensor package in a circuit board or module.

Transverse Acceleration: Any acceleration applied 90° to the axis of sensitivity.

Transverse Sensitivity Error: The percent of a transverse acceleration that appears at the V_{PR} output. For example, if the transverse sensitivity is 1%, then a +10 g transverse acceleration will cause a 0.1 g signal to appear at V_{PR} (1% of 10 g). Transverse sensitivity can result from a sensitivity of the sensor to transverse forces or from misalignment of the internal sensor to its package.

Transverse Y Axis: The axis perpendicular (90°) to the package axis of sensitivity in the plane of the package pin circle. See Figure 2.

Transverse Z Axis: The axis perpendicular (90°) to both the package axis of sensitivity and the plane of the package pin circle. See Figure 2.

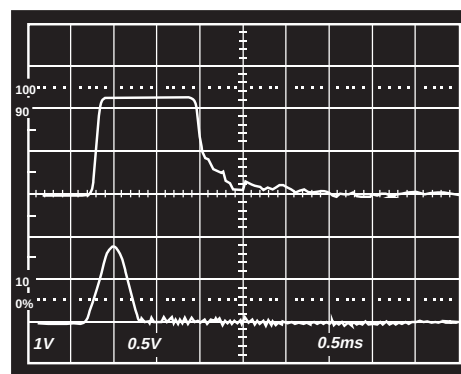


Figure 3. 500 g Shock Overload Recovery. Top Trace: ADXL50 Output. Bottom Trace: Reference Accelerometer Output

ADXL50—Typical Characteristics

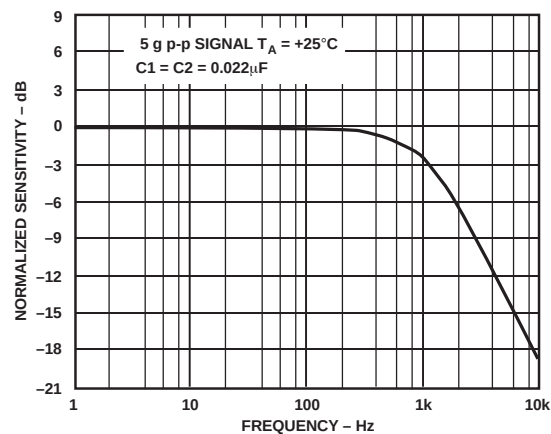


Figure 4. Normalized Sensitivity vs. Frequency

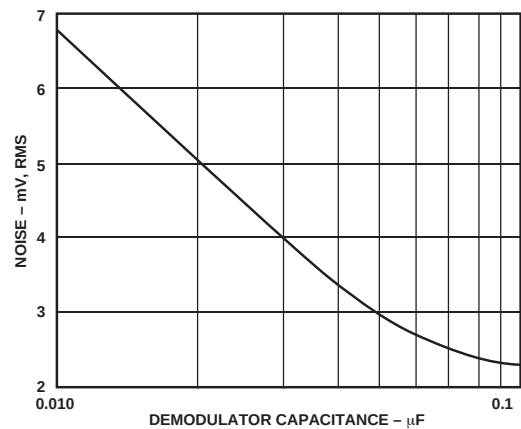


Figure 7. RMS Noise vs. Value of Demodulator Capacitor, C1

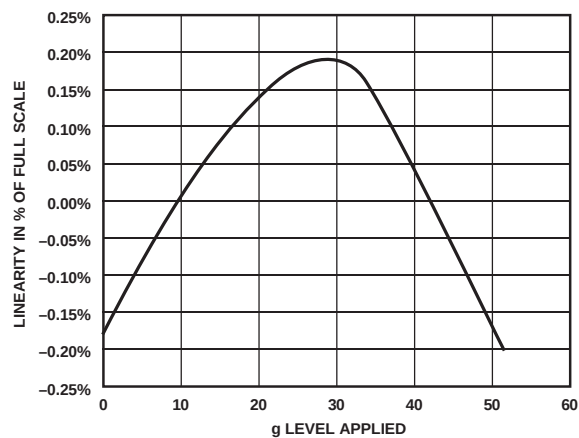


Figure 5. Linearity in Percent of Full Scale

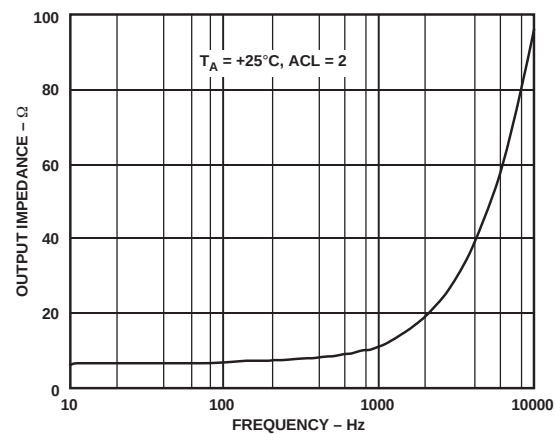


Figure 8. Buffer Amplifier Output Impedance vs. Frequency

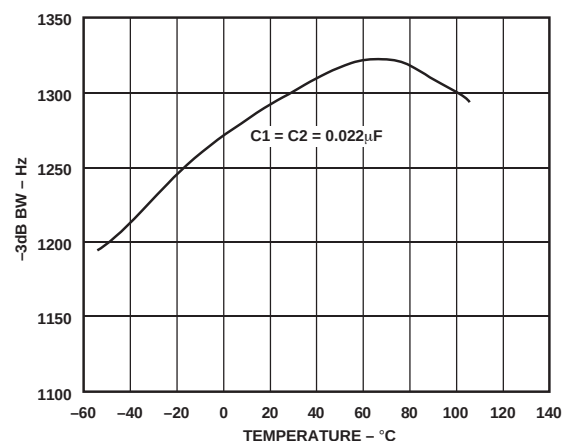


Figure 6. -3 dB Bandwidth vs. Temperature at V_{PR}

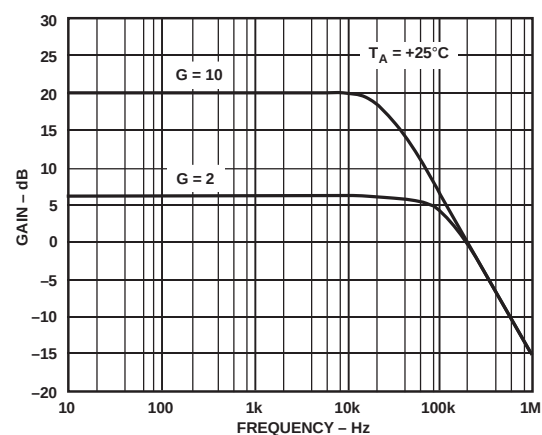


Figure 9. Buffer Amplifier Closed-Loop Gain vs. Frequency

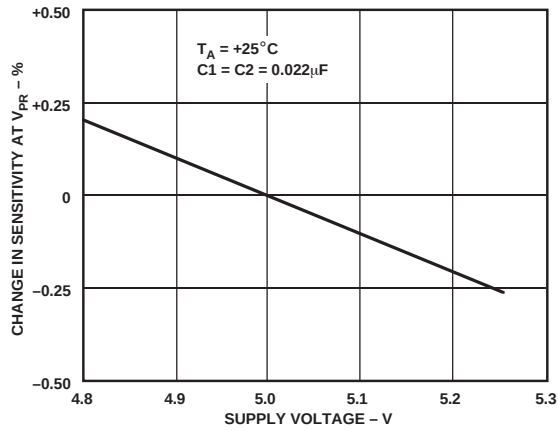


Figure 10. Change in Sensitivity vs. Supply Voltage

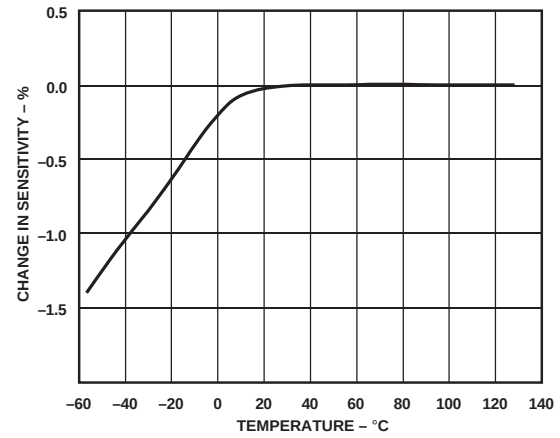


Figure 13. Percent Change in Sensitivity at V_{PR} vs. Temperature

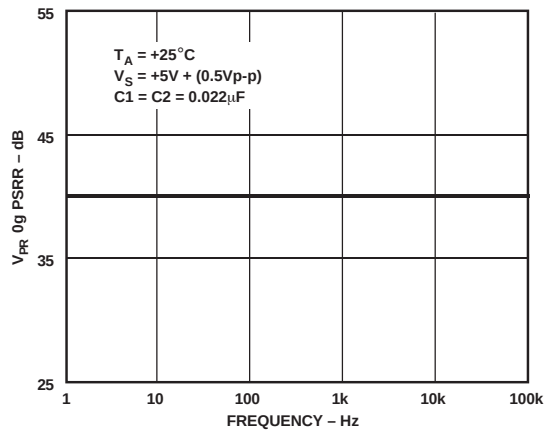


Figure 11. V_{PR} 0 g PSRR vs. Frequency

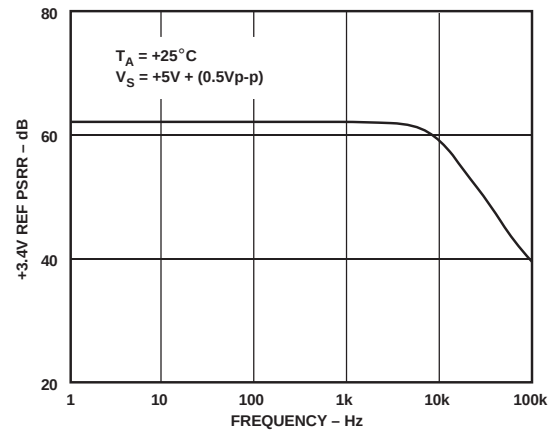


Figure 14. +3.4 V REF PSRR vs. Frequency

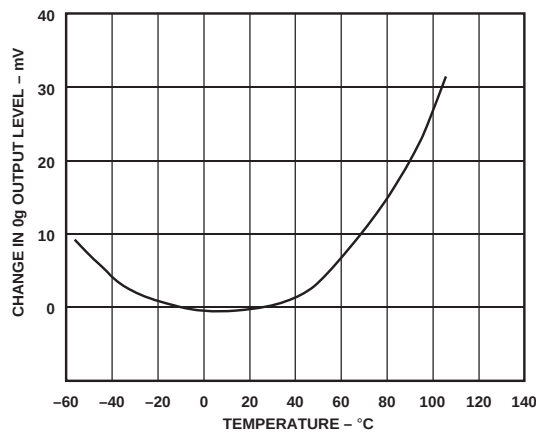


Figure 12. 0 g Bias Level vs. Temperature

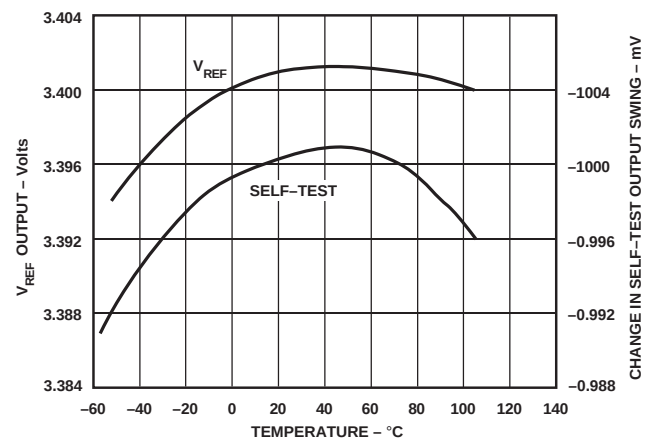


Figure 15. V_{REF} Output and Change in Self-Test Output Swing vs. Temperature

ADXL50

THEORY OF OPERATION

The ADXL50 is a complete acceleration measurement system on a single monolithic IC. It contains a polysilicon surface-micro machined sensor and signal conditioning circuitry. The ADXL50 is capable of measuring both positive and negative acceleration to a maximum level of ± 50 g.

Figure 16 is a simplified view of the ADXL50's acceleration sensor at rest. The actual structure of the sensor consists of 42 unit cells and a common beam. The differential capacitor sensor consists of independent fixed plates and a movable "floating" central plate which deflects in response to changes in relative motion. The two capacitors are series connected, forming a capacitive divider with a common movable central plate. A force balance technique counters any impeding deflection due to acceleration and serves the sensor back to its 0 g position.

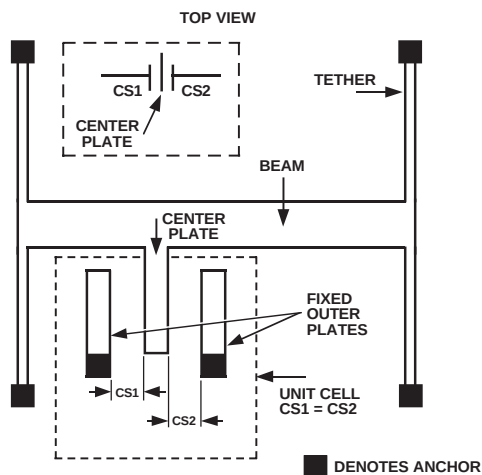


Figure 16. A Simplified Diagram of the ADXL50 Sensor at Rest

Figure 17 shows the sensor responding to an applied acceleration. When this occurs, the common central plate or "beam" moves closer to one of the fixed plates while moving further from the other. The sensor's fixed capacitor plates are driven differentially by a 1 MHz square wave: the two square wave amplitudes are equal but are 180° out of phase from one another. When at rest, the values of the two capacitors are the same and therefore, the voltage output at their electrical center (i.e., at the center plate) is zero.

When the sensor begins to move, a mismatch in the value of their capacitance is created producing an output signal at the central plate. The output amplitude will increase with the amount of acceleration experienced by the sensor. Information concerning the direction of beam motion is contained in the phase of the signal with synchronous demodulation being used to extract this information. Note that the sensor needs to be positioned so that the measured acceleration is along its sensitive axis.

Figure 18 shows a block diagram of the ADXL50. The voltage output from the central plate of the sensor is buffered and then applied to a synchronous demodulator. The demodulator is also supplied with a (nominal) 1 MHz clock signal from the same oscillator which drives the fixed plates of the sensor. The

demodulator will rectify any voltage which is in sync with its clock signal. If the applied voltage is in sync and in phase with the clock, a positive output will result. If the applied voltage is in sync but 180° out of phase with the clock, then the demodulator's output will be negative. All other signals will be rejected. An external capacitor, C1, sets the bandwidth of the demodulator.

The output of the synchronous demodulator drives the preamp—an instrumentation amplifier buffer which is referenced to +1.8 volts. The output of the preamp is fed back to the sensor through a 3 MΩ isolation resistor. The correction voltage required to hold the sensor's center plate in the 0 g position is a direct measure of the applied acceleration and appears at the V_{PR} pin.

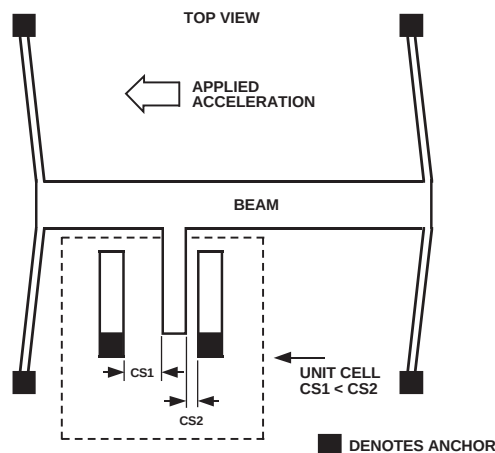


Figure 17. The ADXL50 Sensor Momentarily Responding to an Externally Applied Acceleration

When the ADXL50 is subjected to an acceleration, its capacitive sensor begins to move creating a momentary output signal. This is signal conditioned and amplified by the demodulator and preamp circuits. The dc voltage appearing at the preamp output is then fed back to the sensor and electrostatically forces the center plate back to its original center position.

At 0 g the ADXL50 is calibrated to provide +1.8 volts at the V_{PR} pin. With an applied acceleration, the V_{PR} voltage changes to the voltage required to hold the sensor stationary for the duration of the acceleration and provides an output which varies directly with applied acceleration.

The loop bandwidth corresponds to the time required to apply feedback to the sensor and is set by external capacitor C1. The loop response is fast enough to follow changes in g level up to and exceeding 1 kHz. The ADXL50's ability to maintain a flat response over this bandwidth keeps the sensor virtually motionless. This essentially eliminates any nonlinearity or aging effects due to the sensor beam's mechanical spring constant, as compared to an open-loop sensor.

An uncommitted buffer amplifier provides the capability to adjust the scale factor and 0 g offset level over a wide range. An internal reference supplies the necessary regulated voltages for powering the chip and +3.4 volts for external use.

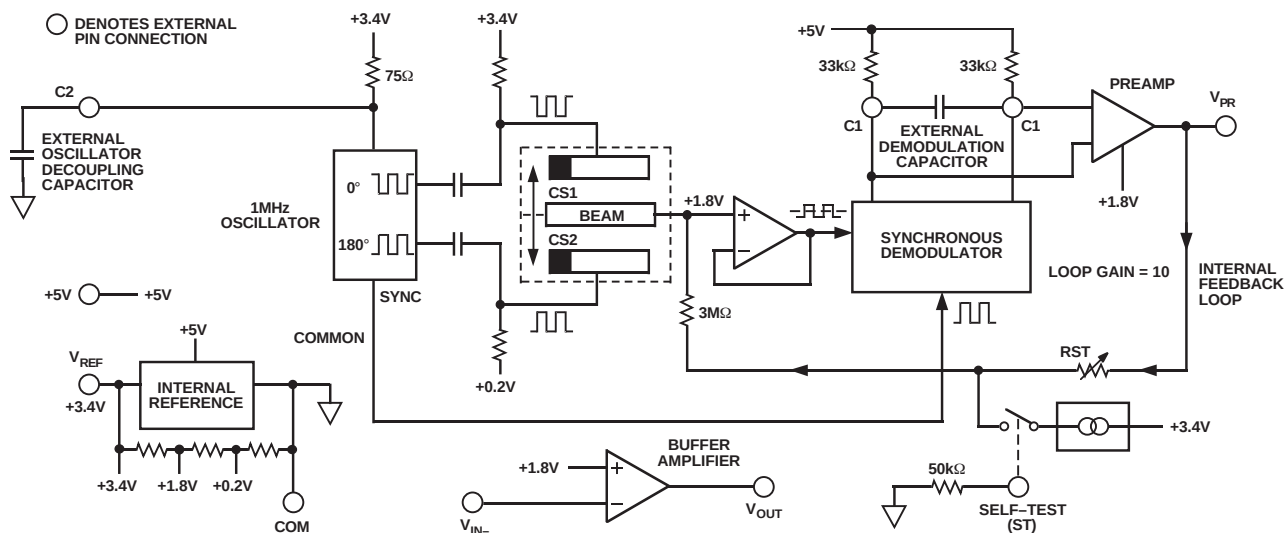


Figure 18. Functional Block Diagram

The sensor's tight mechanical spacing allows it to be electrostatically deflected to full scale while operating on a 5 volt supply. A self-test is initiated by applying a TTL "high" level voltage ($>+2.0$ V) to the ADXL50's self-test pin which causes the chip to apply a deflection voltage to the beam which moves it an amount equal to -50 g (the negative full-scale output of the device). Note that the $\pm 10\%$ tolerance of the self-test circuit is not proportional to the sensitivity error, see Self-Test section.

The output of the ADXL50's preamplifier is 1.8 V at 0 g acceleration with an output range of ± 0.95 V for a ± 50 g input, i.e., 19 mV/g. An uncommitted buffer amplifier has been included on-chip to enhance the user's ability to offset the 0 g signal level and to amplify and filter the signal. Access is provided to both

the inverting input and the output of this amplifier via pins V_{OUT} and V_{IN-}, while the noninverting input is connected internally to a +1.8 V reference. The +1.8 V is derived from a resistor divider connected to the 3.4 V reference.

BASIC CONNECTIONS FOR THE ADXL50

Figure 19 shows the basic connections needed for the ADXL50 to measure accelerations in the ± 50 g range with an output scale factor 40 mV/g corresponding to a 2.5 V 0 g level, a ± 2.0 V full-scale swing around 0 g and a 3 dB bandwidth of approximately 1 kHz.

In general, the designer will need to take into account the initial zero g bias when designing circuits. For the ADXL50J this offset is $1.8 \text{ V} \pm 250 \text{ mV}$. When microprocessors and software

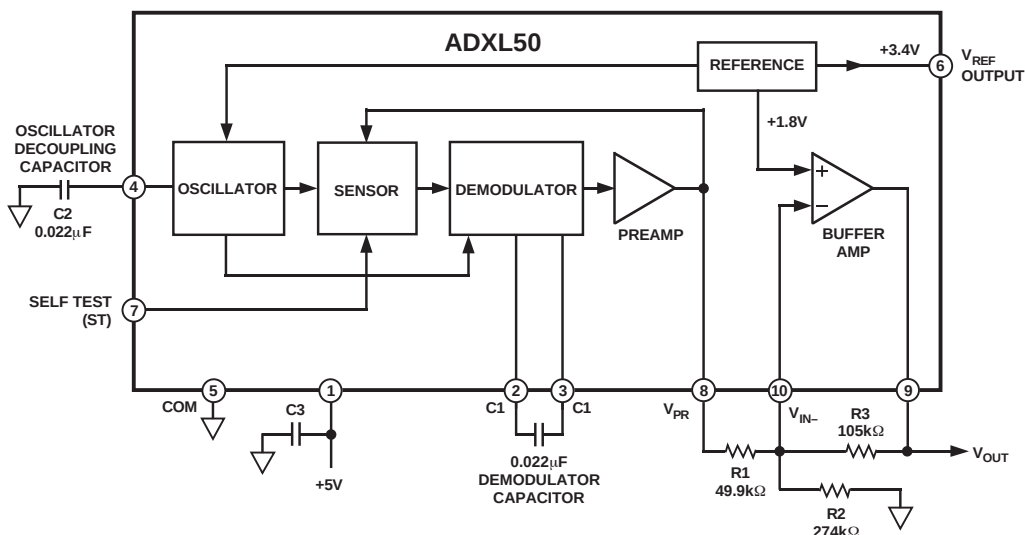


Figure 19. ADXL50 Application Providing an Output Sensitivity of 40 mV/g, a +2.5 V 0 g Level and a Bandwidth of 1 kHz

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

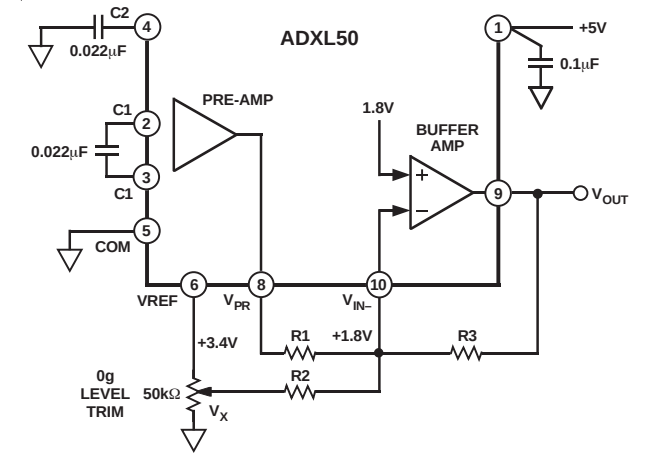


Figure 20. ADXL50 Circuit Using the Buffer Amplifier to Set the Output Scaling and 0 g Offset Level

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

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

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.

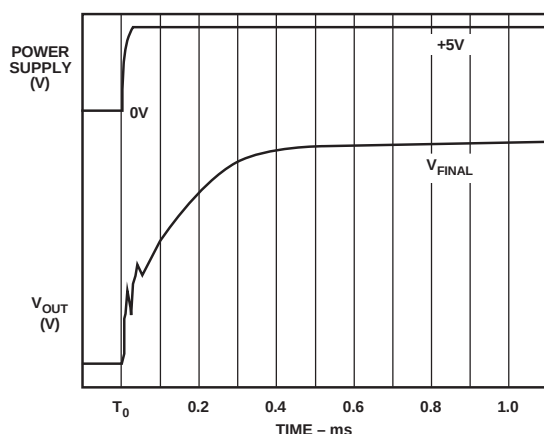


Figure 21. Power-On Settling Time when Power Cycling

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.

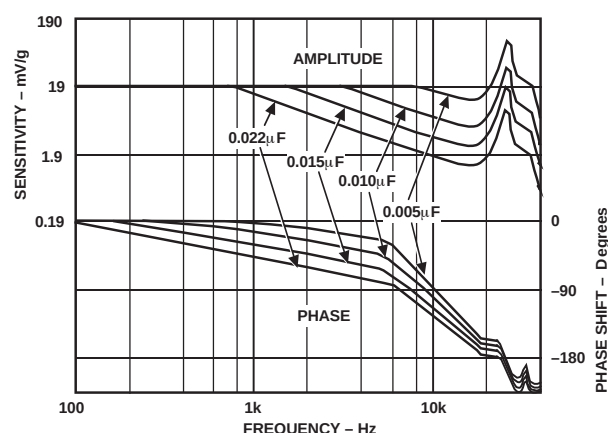


Figure 22. Frequency Response of the ADXL50 for Various Demodulator Capacitors

FULL SCALE	mV per g	3dB BW (Hz)	R1a k Ω	R1b k Ω	R3 k Ω	R2 k Ω	C4 μ F
± 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

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

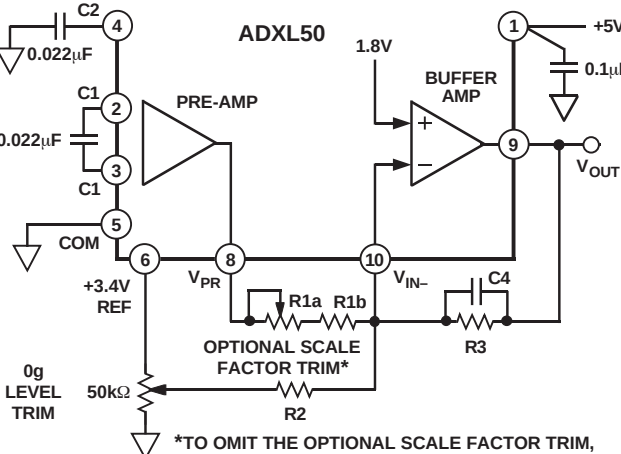


Figure 23. Using the Buffer Amplifier to Provide One Pole Post Filtering Plus Scale Factor and 0 g Level Trimming

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:

$$\text{Capacitor } C4 \text{ in Farads} = 1/(2 \pi \times R3 \text{ in Ohms} \times 3 \text{ dB BW in Hertz})$$

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 $0.012\text{ }\mu\text{F} \times 2 = 0.024\text{ }\mu\text{F}$. The closest standard value of $0.022\text{ }\mu\text{F}$ 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.

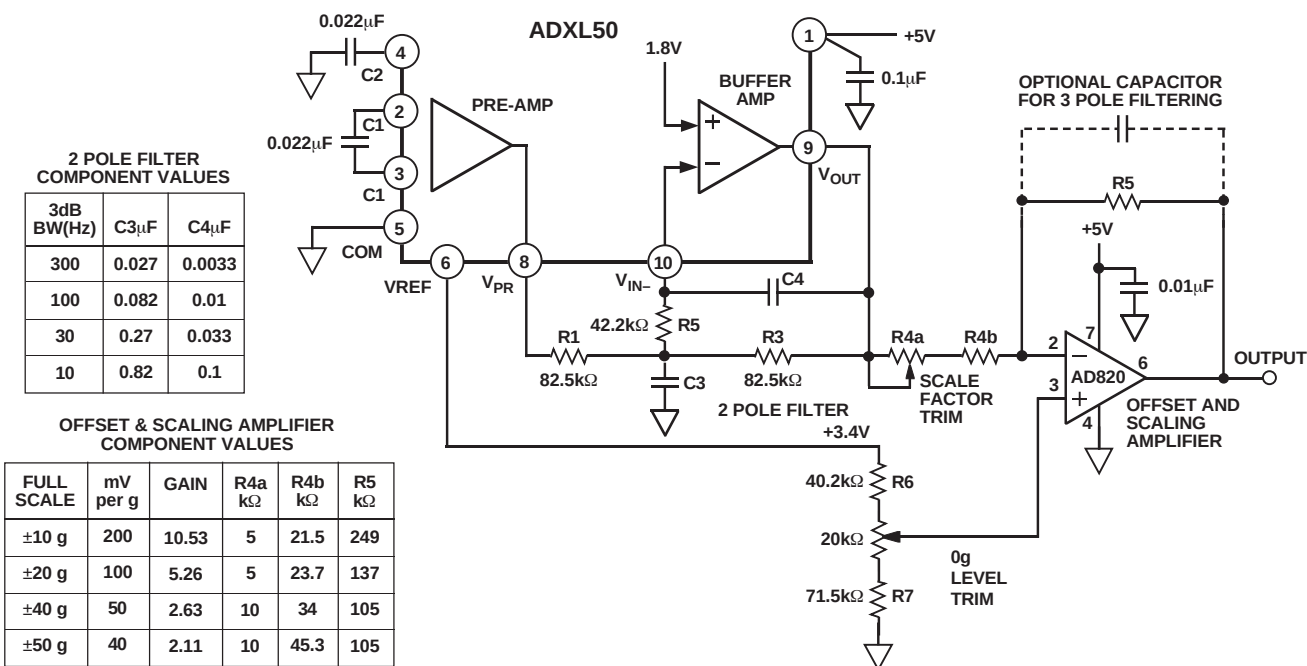


Figure 24. Circuit Providing Two Pole Post Filtering and 0 g Offset and Scale Factor Trimming

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.

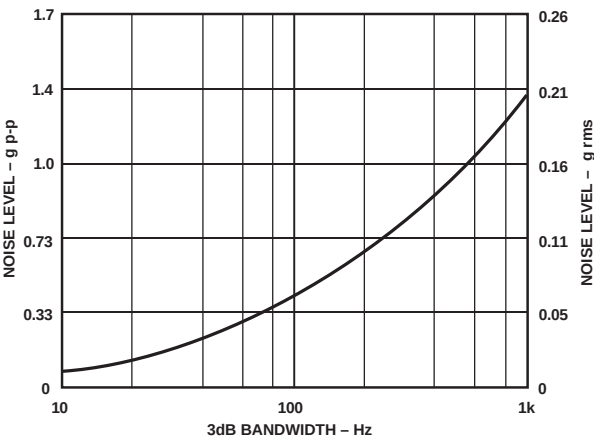


Figure 25. ADXL50 Noise Level and Resolution vs. -3 dB Bandwidth

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.

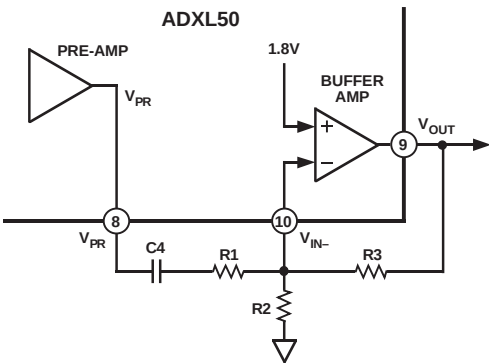


Figure 26. AC Coupling the VPR Output to the Buffer Input

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.

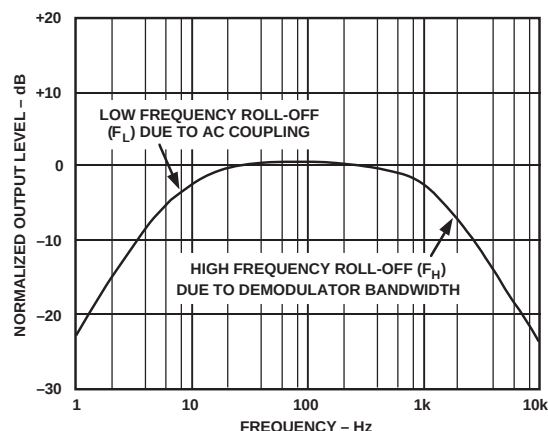


Figure 27. Normalized Output Level vs. Frequency for a Typical Application Using AC Coupling Between V_{PR} and Buffer Amplifier

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.

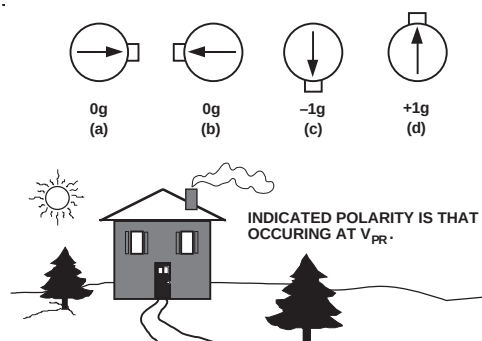


Figure 28. Using the Earth's Gravity to Self-Calibrate the ADXL50

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.

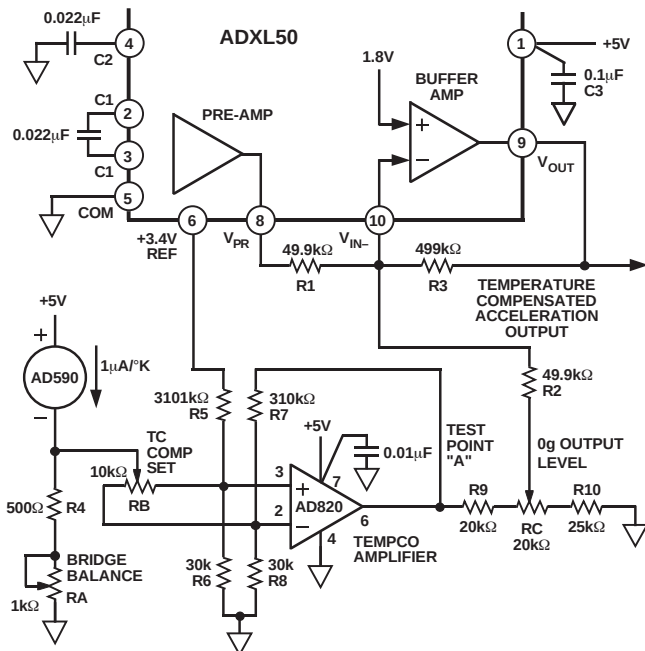
Adjusting the circuit's scale factor will have some effect on its 0 g level so this should be readjusted, as before, but this time checked in both positions "a" and "b." If there is a difference in the 0 g reading, a compromise setting should be selected so that the reading in each direction is equidistant 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

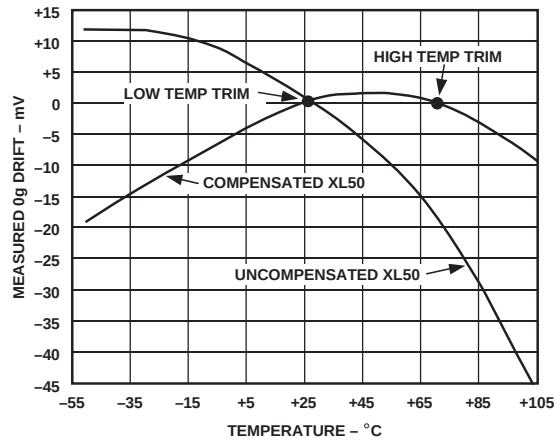


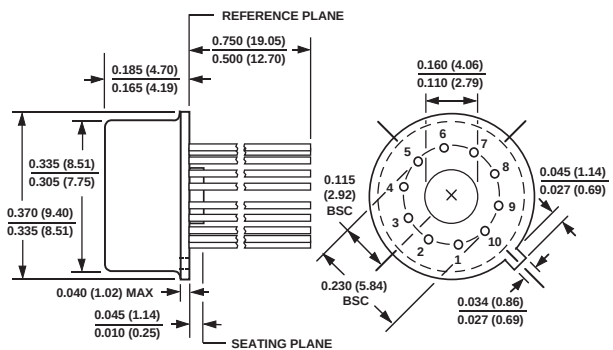
Figure 30. ADXL50 0 g Drift With and Without the Compensation Circuit of Figure 29

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).



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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:

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