Low-Noise MEMS Vibration Sensor for Geophysical Applications

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ABSTRACT

The need exists for high sensitivity, low noise vibration sensors for various applications such as geophysical data collection, tracking vehicles, intrusion detectors, and underwater pressure gradient detection. In general, these sensors differ from classical accelerometers in that they require no DC response, but must have a very low noise floor over a required bandwidth. Theory indicates a capacitive micromachined silicon vibration sensor can have a noise floor on the order of $0.1\mu\text{g}/\sqrt{\text{Hz}}$ over 1 kHz bandwidth while reducing size and weight 10-fold compared to existing magnetic geophones. With early prototypes we have demonstrated Brownian-limited noise floor at $1.5\mu\text{g}/\sqrt{\text{Hz}}$, orders of magnitude more sensitive than surface micromachined devices such as the industry standard ADXL05.

INTRODUCTION

A vibration sensor can be thought of as a very high sensitivity accelerometer with no DC output requirement. With no drift or bias stability spec., the design can be optimized to give the lowest noise floor. Applications for these devices include geophysical sensing, machinery vibration and failure prediction, tracking and identification of vehicles or personnel, and underwater pressure gradient sensing.

Traditional vibration sensors using permanent magnets and fine wire coils are called geophones [1], and measure *velocity* above the fundamental resonance. This is in contrast to capacitive accelerometers which measure *acceleration* below their fundamental resonance. Piezoelectric and ferroelectric accelerometers are also used for these applications. Micromachined sensors can offer size and weight advantages over traditional sensors.

Previous efforts to make micromachined high resolution vibration sensors or accelerometers have included capacitive [2-4], tunneling [5], piezoresistive, optical and piezoelectric sensors. Capacitive sensors have the advantage of no exotic materials, low noise and compatibility with CMOS readout electronics. Tunneling sensors have a low noise floor, but due to the small dynamic range at the tip require a very stiff feedback loop which reduces the useful bandwidth and dynamic range.

THEORY

The sensor is modeled as a spring-mass-damper system with capacitive pickoff. Because no DC output is required, it was decided for initial tests to build a single-capacitor sensor rather than a differential capacitor design. Noise sources modeled include Brownian mechanical noise from air damping and electronic noise from the readout circuit [6].

The Brownian force is $F_B = \sqrt{4kTD}$ (N/ $\sqrt{\text{Hz}}$), which causes Brownian motion of the proof mass x_B

$$x_B = \frac{\sqrt{4kTD}}{k_{sp} + j\omega D - \omega^2 M} \quad (\text{m}/\sqrt{\text{Hz}}) \quad (1)$$

where D is the damping coefficient of the proof mass M supported by spring constant k_{sp} . Solving for the acceleration which generates the same motion, and substituting $Q = \omega_0 M / D$, $\omega_0 = \sqrt{k_{sp} / M}$, and $g = 9.8 \text{ m/s}^2$, gives for Brownian equivalent acceleration noise in $g/\sqrt{\text{Hz}}$:

$$g_{n,B} = \frac{\sqrt{4kTD}}{Mg} = \frac{1}{g} \sqrt{\frac{4kT\omega_0}{MQ}} \quad (g/\sqrt{\text{Hz}}) \quad (2)$$

From Eq. 2 we see that a large mass and high Q (low damping) are helpful to achieve a low noise floor. In order to use a sensor with a high Q , it must be force rebalanced to prevent ringing at the resonant frequency. In the work presented here, the sensors are over-damped, with a Q of about 0.3.

The sensitivity of the device is calculated for the simple case of a DC bias voltage with a high input impedance buffer amplifier. A bias voltage of $1/2$ the snap-down voltage (V_{SD}) is assumed. The sensitivity at low frequency is the V/m in the sense gap times the m/g of the proof mass below resonance:

$$M_0 = \frac{V_{SD}}{2x_0} \frac{Mg}{k_{sp}} = \frac{V_{SD}}{2x_0} \frac{g}{\omega_0^2} \quad (\text{V/g}) \quad (3)$$

The snap-down voltage is $V_{SD} = \sqrt{\frac{8k_{sp}x_0^3}{27\epsilon_0 A}}$. Substituting for the mass M and V_{SD} gives for sensitivity M_0 :

$$M_0 = \frac{g}{\omega_0} \sqrt{\frac{2t\rho_{Si}x_0}{27\epsilon_0}} \quad (\text{V/g}) \quad (4)$$

where t is the proof mass thickness, and ρ_{Si} is the density of silicon. From this we see that sensitivity is inversely proportional to the fundamental resonant frequency.

We consider also an electronic readout noise component, which has in general both equivalent input current noise and voltage noise components. The total equivalent front-end noise is called $v_{n,e}$. This voltage noise can be converted to an equivalent g -noise by dividing by the transducer sensitivity in V/g :

$$g_{n,e} = \frac{v_{n,e}}{M_0} \quad (g/\sqrt{\text{Hz}}) \quad (5)$$

The total noise is the r.m.s. combination of electrical and Brownian contributions.

$$g_{n,T} = \sqrt{g_{n,B}^2 + g_{n,e}^2} \quad (g/\sqrt{\text{Hz}}) \quad (6)$$

ELECTRONICS DESIGN

Voltage noise for a low-noise CMOS front end is typically in the 5 to 20 nV/ $\sqrt{\text{Hz}}$ range above the 1/f corner. Lower noise can be obtained but at the cost of increased current consumption.

To achieve a noise level of 10 ng, a vacuum package must be used to reduce Brownian motion noise.

Fig. 1 is a contour map of combined g-equivalent noise as a function of electrical noise (1nV to 100 nV/ $\sqrt{\text{Hz}}$) and Q (1 to 10^4). To achieve a noise level of 10 ng/ $\sqrt{\text{Hz}}$ requires a Q of 3000 and a front end equivalent noise of under 2 nV/ $\sqrt{\text{Hz}}$. For this calculation, the silicon proof mass is assumed to be 4 mm X 4 mm X 0.38 mm with a 3 μm sense gap, a resonant frequency of 1 kHz, and sensitivity of 0.233 V/g.

Fig. 1. Total noise in dB // $\lg/\sqrt{\text{Hz}}$, as a function of Q and front-end electrical noise.

Table 1. Design Specification for Vibration Sensor

Resonant Frequency	Units	500 Hz Sensor	1 kHz Sensor	10 kHz Sensor
Self Noise Floor * (Combined Brownian and electronic)	ng/ $\sqrt{\text{Hz}}$	25	43	338
Sensitivity with unity gain buffer	V/g	0.49	0.245	0.025
Maximum g-level (open loop)	g's	0.3	1.2	120
Displacement per g	m	10^{-6}	2.5×10^{-7}	2.5×10^{-9}

* Assumes 8 nV/ $\sqrt{\text{Hz}}$ electrical noise, Q = 100

To evaluate the vibration sensor prototypes two types of electronics were used: an open-gate JFET source follower, and a custom CMOS ASIC. The JFET buffer allowed quick, reasonably low noise measurements to be made with a very compact circuit (inside the sensor package) although 1/f and current noise are high at low frequencies. The ASIC uses a carrier to reduce 1/f noise, resulting in better low frequency performance. Fig. 2 shows the open-gate JFET buffer circuit used. This circuit has a DC input impedance of several T Ω , which combined with a typical sensor capacitance of 50 pF gives an RC time constant of minutes. It was sometimes necessary to shine light in the package to allow the JFET to reach bias equilibrium (photonic reset).

Fig. 2. Open-gate JFET buffer circuit.

A custom CMOS mixed signal ASIC was designed in-house and fabricated at Orbit Semiconductor. The design implements synchronous modulation / demodulation using square waves and a novel RF rebalance technique. The block diagram of the system is shown in Fig. 3. The circuit rebalances the proof-mass below the frequency range of interest, while allowing higher frequency vibration signals to pass through open-loop.

The ASIC applies equal and opposite 100kHz square waves to the sensor and a reference capacitor. Vibrations cause a mismatch between the fixed capacitor and the time varying sensor. The charge amplifier converts the capacitance mismatch into an output voltage which is then amplified by the AC gain stage.

The vibration signal is modulated on a 100KHz carrier. The amplifier stages process this signal at 100KHz to avoid 1/f noise in the CMOS transistors. The signal is demodulated after AC amplification to recover the vibration signal.

Application of RF voltage to the sensor applies a force proportional to the square of the applied voltage amplitude. The integral rebalance controller adjusts the carrier amplitude and tunes the time average sensor capacitance to match the fixed reference capacitor. Open loop maximum g's are set by proof mass motion of about 10% of the sense gap. Closed-loop maximum g's are determined by the available rebalance voltage.

The bandwidth of the rebalance loop is designed to be low (<1Hz). The rebalance loop nulls DC and sub 1 Hertz accelerations maintaining signal null with changes in temperature and sensor orientation. This allows high gain in dynamic range limited CMOS (5V) electronics. Vibration above 1Hz are not rebalanced and are sensed open-loop.

Fig. 3 CMOS ASIC Block Diagram

SENSOR FABRICATION

The sensors are fabricated on 0.38 mm thick double-side polished wafers using the Bosch process in an STS etcher. A recess 3 μm deep is etched into the wafer to define anchors and create the sense gap. 30 μm deep damping-relief trenches are then etched to reduce squeeze-film damping. A 10 μm thick boron diffusion is used to create an etch stop layer on both faces of the wafer. After electrostatic bonding to a glass wafer with readout electrodes, the STS etcher is used to trench through the wafer. A brief anisotropic etch in EDP (Ethylene-Diamine-Pyrocatechol-Water) then undercuts the springs.

Figs. 4 and 5 show a sensor chip after the deep ICP (Inductively Coupled Plasma) etch and before the anisotropic etch. A central proof mass is supported by springs attached to 4 anchors on a glass substrate. Damping-relief trenches are visible in Fig. 5 facing the glass substrate. Fig. 6 shows a completed prototype sensor.

Devices were fabricated with 3 proof mass sizes ((3 mm)², (4 mm)², and (5 mm)²) and 3 resonant frequencies (500 Hz, 1kHz, and 10 kHz) to cover various applications. Kovar flat-packs were used to house the sensors.

Figure 5. Corner of device showing electrodes and damping-reduction trenches.

Figure 4. Vibration sensor etched through the wafer (380 μm), before anisotropic etch. Proof mass is 3 mm X 3 mm.

Figure 6. Vibration sensor after anisotropic etching, leaving thin flexures supporting large proof mass.

TEST RESULTS AND DISCUSSION

Chip level testing includes capacitance-voltage curves and current - voltage curves to measure leakage resistance (typically greater than $1\text{ T}\Omega$). Frequency response with a commercial JFET buffer is shown in Fig. 7. A sensitivity of 72 mV/g was obtained at a bias voltage of 22 V . The noise floor (measured on a vibration isolated platform) was as low as $1.5\text{ }\mu\text{g}/\sqrt{\text{Hz}}$ which is the Brownian limit for this device (Fig. 8). A battery powered low noise pre-amp using OP 37 op-amps was used to raise the signal level above the noise floor of the dynamic signal analyzer (HP 3563A).

Fig. 9 shows the advantage of the RF carrier readout chip over a baseband CMOS readout. Orders of magnitude reduction in $1/f$ noise are achieved using the ASIC.

Fig. 7. Frequency Response of $3\text{ mm} \times 3\text{ mm}$ device, design resonance at 1 kHz , $Q = 0.3$.

Fig. 8. Plot of acceleration noise in $\text{g}/\sqrt{\text{Hz}}$ using source-follower buffer circuit. Device is same as Fig. 8.

CONCLUSIONS

A bulk micromachined vibration sensor has been fabricated and tested with a novel, custom CMOS ASIC which removes $1/f$ noise while requiring only a single sense capacitor and one dummy capacitor. Vibration noise at the Brownian noise limit ($1.5\text{ }\mu\text{g}/\sqrt{\text{Hz}}$) was achieved using an overdamped sensor with 1 kHz fundamental resonance. Theory indicates that orders of magnitude improvements can be made over existing MEMS sensors in both bandwidth and noise floor by the use

of wafer-thick bulk micromachining technology with optimized low noise electronics. These sensors should be useful for geophysical work, underwater sensing, machinery monitoring, and sensing of vehicles and people.

Fig. 9. Baseband vs. synchronous modulation / demodulation noise with Orbit CMOS ASIC readout chip.

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