Offset Suppression in a Micromachined Lorentz Force Magnetic Sensor by Current Chopping
Mo Li and David A. Horsley, Member, IEEE

Abstract—We present a method to reduce offset in micromachined Lorentz force magnetic sensors by chopping the Lorentz force bias current. By switching the polarity of this current, the sensitivity of the magnetic sensor alternates its sign, whereas the offset remains the same. A residual offset of 31 µT is obtained from the initial offset of 25 nT. The proposed method significantly reduces the long-term drift of the magnetic sensor at the same time. A 9-h measurement shows the maximum drift error is reduced from ±500 µT to ±1 µT. Allan deviation measurements demonstrate that the long-term drift is reduced by a factor of 120. With 0.9 mA rms bias current the white noise is 400 nT/√Hz, limited by the thermal-mechanical noise.

Index Terms—Offset suppression, Lorentz force, magnetic sensor, micromachined resonator, MEMS.

I. INTRODUCTION

Many efforts have been made to improve the sensitivity and resolution of magnetic sensors. The Lorentz force magnetic sensor is an emerging technology for magnetic sensors, and Lorentz force sensors with sensitivity and resolution comparable to, or better than, that of Hall-effect sensors have been reported recently [1]–[3]. Compared to AMR (anisotropic magnetoresistive) and Hall-effect sensors that are commonly used in portable electronics, the Lorentz force sensor has the advantage that it is free of magnetic material (making it potentially CMOS compatible) and therefore free of magnetic hysteresis. A Lorentz force sensor can also be co-fabricated with other MEMS (micro-electromechanical systems) inertial sensors, such as accelerometers and gyroscopes.

Offset is another key parameter that influences the performance of magnetic sensors. It reduces the dynamic range of the system and also results in drift error, which directly transfers to heading error when the sensor is used as a compass. There are three categories of compasses [4]:

A. Basic Principle of a Lorentz Force Magnetometer

Similar to an accelerometer or a gyroscope, a Lorentz force sensor (shown in Figure 1) is a force sensor that measures the displacement resulting from the Lorentz force $F_L$ on a current-carrying proof-mass. The current is modulated at the first flexural mode of the structure at frequency $f_n$, which modulates the low frequency magnetic field $B$ up to near $f_n$, resulting in an amplitude modulated (AM) Lorentz force $F_L$. In a compass application, Earth’s field is constant but varies slowly as the compass heading changes, resulting in displacement amplitude

$$x_L = \frac{F_L Q}{k} = \frac{B i L Q}{k},$$

where $B$ is the magnetic field, $i$ is the Lorentz force bias current, $L$ is the effective length of the current-carrying resonator, $Q$ is the mechanical quality factor and $k$ is the stiffness of the flexures that support the proof-mass.
The reference frequency for the bias current can be provided either by using an external oscillator with its frequency set to the resonance frequency of the magnetometer (open loop), or by using a magnetometer itself as an oscillator (closed loop). In closed-loop operation, a self-sustained oscillation loop is formed using the Lorentz force magnetometer as a resonator [3]. Electrodes DR and S are used for electrostatic driving and capacitive sensing, respectively, and a sustaining amplifier maintains a constant electrostatically-excited displacement amplitude \( x_E \). Operating the magnetometer in closed-loop provides better stability over temperature and also reduces the complexity of the system. However, closed-loop operation introduces a large offset (\( x_E \)) in the oscillation amplitude, which can be 100X larger than the amplitude generated by the Lorentz force (\( x_L \)).

When the electrostatic force is in phase with the Lorentz force, at steady state, the total oscillation amplitude of the magnetometer can be expressed as:

\[
x = x_E + x_L = V_{DC} V_{AC} \frac{dC}{dx} \frac{Q}{k} + BiL \frac{Q}{k},
\]

where \( V_{DC} \) is the dc bias, \( V_{AC} \) is the electrostatic driving voltage, and \( dC/dx \) is the capacitive displacement sensitivity. Note here that both \( V_{AC} \) and \( i \) have frequency of \( f_n \). Equation (2) makes it clear that \( x_E \) appears as offset in the AM output. As a result, the offset from the micromechanical resonator introduces long-term drift in the sensor’s output, including \( 1/f \), \( 1/f^2 \) and \( 1/f^3 \) noise [15].

**B. Basic Principle of Current Chopping**

Conceptually, the current chopping method modulates the magnetic field signal to a frequency where there is no offset and long-term drift from the micromechanical resonator. This can be achieved by periodically reversing the sign of the magnetic field sensitivity, whereas the offset from the electrostatic force remains constant. To do this, the bias current is reversed by introducing a 180° phase shift between the Lorentz current and the electrostatic drive signal, which flips the direction of the Lorentz force relative to the electrostatic force (assuming a constant magnetic field is applied), as illustrated in Figure 2. When the Lorentz force and the electrostatic force have the same sign, as shown in Figure 2(a), the oscillation amplitude is:

\[
x_+ = x_E + x_L = (V_{DC} V_{AC} \frac{dC}{dx} + BiL) \frac{Q}{k},
\]

whereas when the Lorentz force is reversed, as shown in Figure 2(b), the oscillation amplitude is:

\[
x_- = x_E - x_L = (V_{DC} V_{AC} \frac{dC}{dx} - BiL) \frac{Q}{k}.
\]

By sampling the oscillation amplitudes for these two cases and subtracting them from each other at the steady state, the difference in oscillation amplitude is:

\[
x_d = x_+ - x_- = 2BiL \frac{Q}{k}.
\]

The sensitivity is given by:

\[
S_x = \frac{\partial x_d}{\partial B} = 2iL \frac{Q}{k}.
\]

The sensitivity is doubled and the offset from electrostatic force is removed. Note here that when the sensor operates continuously with current chopping, the sensitivity will be half of \( S_x \) due to the dechopping stage, as discussed below.

The offset amplitude \( x_E \), the amplitude from magnetic field signal \( x_L \) and the total oscillation amplitude \( x_E + x_L \) are illustrated in the time domain in Figure 3. It is assumed that the input magnetic field starts at 0 and changes to a constant dc value at an arbitrary time. From Figure 3(a), it can be observed that the sensor offset \( x_E \) and magnetic signal \( x_L \) are not separated from each other without current chopping, thus they cannot be distinguished from each other in the final oscillation amplitude \( x_E + x_L \). However, in Figure 3(b), the magnetic field signal \( x_L \) is a modulated signal with twice the amplitude observed in Figure 3(a). An extra dechopping step
at the sensor’s output separates $x_E$ from $x_L$, returning the chopped magnetic field signal to the baseband with minimal offset.

In the frequency domain, two modulation and two demodulation steps are involved. Before modulation, the magnetic field signal and the $1/f$ noise from the detection electronics are in baseband. The $1/f$, $1/f^2$ and $1/f^3$ noise terms from the micromechanical resonator are centered at the natural frequency of the resonator at $f_n$, as shown in Figure 4(a). The first modulation occurs when the Lorentz force is generated by an ac bias current $icos(2\pi f_n)$, which is a method commonly used in resonant Lorentz force magnetometers [1]. The magnetic signal is free of $1/f$ noise from the front-end detection electronics and the motion resulting from the magnetic signal is amplified by the $Q$ of the resonator, which can be higher than 10000 for resonators in vacuum. The second modulation happens when the ac bias current reverses its sign at a much lower frequency $f_c$, so the magnetic field signal is modulated to $\pm nf_c$ sidebands centered at $f_n$. Note here that $n$ is an odd integer, as shown in Figure 4(b). As discussed later, $n = 1$ is the dominant frequency component in the chopped output spectrum. Since the first two modulation steps are applied to the excitation current, which is independent from the electrostatic force, the offset due to electrostatic excitation is not modulated and remains centered at $f_n$. The output of the magnetometer is then demodulated at $f_n$. During this demodulation, the magnetic signal centered at $f_n$ is demodulated back to $nf_c$ and the offset is demodulated back to baseband, whereas the $1/f$ noise from the detection electronic is modulated to $f_n$, as shown in Figure 4(c). A low-pass filter is used here to filter out the noise and harmonics at higher frequencies. The final chopping stage recovers the magnetic signal to the baseband, whereas the offset of the magnetometer is modulated to $nf_c$, as shown in Figure 4(d). To avoid aliasing, the current chopping method limits the frequency of the input magnetic field to $f_c/2$. The upper bound of the chopping frequency $f_c$ is set by the bandwidth of the mechanical resonator. In this work, the bandwidth of the mechanical resonator is 4 Hz. During our experiment, $f_c$ is set to 1 Hz, which limits the bandwidth of the sensor system to 0.5 Hz. For compass applications, a bandwidth of 5 to 10 Hz is required. In order to achieve this requirement in the future, a mechanical resonator with a bandwidth larger than 20 Hz needs to be used. The bandwidth of the mechanical resonator is given by:

$$BW = \frac{f_n}{2Q} = \frac{b}{4\pi m}, \tag{7}$$

where $b$ is the damping coefficient and $m$ is the modal mass. The bandwidth can be enhanced by increasing $b$ or decreasing $m$. The former can be easily implemented by increasing the pressure inside vacuum package to introduce more squeeze film gas damping, however this would result in smaller sensitivity and worse resolution as a trade-off. The latter can be achieved by optimizing the resonator design.

III. MEASUREMENT

A single-axis Lorentz force magnetometer is used to demonstrate the current chopping method. The magnetometer was fabricated in the epi-seal encapsulation process [16]. The epi-seal encapsulation process was proposed by researchers at the Robert Bosch Research and Technology Center in Palo Alto and then demonstrated in a close collaboration with Stanford University. This collaboration is continuing to develop improvements and extensions to this process for many applications, while the baseline process has been brought into commercial production by SiTime Inc. Device characterization and closed-loop operation are discussed in our previous work [3]. The photograph of the test setup and device under test is shown in Figure 5(top). The block diagram of the current chopping operation is shown in Figure 5(bottom). The transimpedance amplifier and bias current generation circuits were implemented on a PCB. All other blocks were implemented using a digital lock-in amplifier (Zurich Instruments HF2-LI) controlled by a customized Matlab program. The micromachined resonator is first put into an oscillation loop to generate a bias current at the natural frequency $f_n$. Here, we use a phased locked loop (PLL) to sustain the oscillation. The oscillation offset $x_E$ is set by the electrostatic driving voltage $V_{AC}$. The PLL can be replaced by a voltage limiter or a 1-bit ADC for lower power consumption, as discussed below.
The oscillation amplitude at the output of the transimpedance amplifier is demodulated at the natural frequency \( f_n \), and passed through a low-pass filter to filter out the harmonics introduced during demodulation. Without current chopping, the system operates the same as the closed-loop system we demonstrated in our previous work [3].

To perform current chopping and de-chopping, two sets of choppers are included in the system with the control signals \( \phi_1 \) and \( \bar{\phi}_1 \) generated by an additional oscillator. An additional low pass filter is also required after the de-chopping stage. The final output amplitude is sampled and recorded after it reaches steady state. The signal is de-chopped at \( f_c \) and then the magnetic signal is recovered to baseband. During all measurements, a 0.9 mA bias current driven through the 155 \( \Omega \) proof-mass results in 125 \( \mu \)W power consumption. The dc bias \( V_{DC} \) is 4 V.

### A. Sensitivity and Offset

To demonstrate the effect of current chopping on the magnetometer’s sensitivity and offset, the magnetometer is mounted inside a Helmholtz coil. A dc magnetic field ranging from −400 \( \mu \)T to 400 \( \mu \)T is applied to the magnetometer for each data point. The sensitivity of the magnetometer is measured first with \( \phi_1 \) enabled and \( \bar{\phi}_1 \) disabled, which results in a positive sensitivity in the magnetometer, as shown by the blue line in Figure 6. The measurement is repeated with \( \bar{\phi}_1 \) enabled and \( \phi_1 \) disabled, reversing the sensitivity’s sign as indicated by the red line. The offset in both cases is 112 mV, which is equivalent to a 25 mT offset.

The differential sensitivity, obtained by taking the difference between the two sensitivities before current chopping, is indicated with a black line. The offset is suppressed by a factor of 400 to 0.3 mV (equivalent to a 31 \( \mu \)T offset). The 0.3 mV residual offset results mainly from parasitic feedthrough capacitance in the MEMS resonator. Parasitic capacitances from electrodes IB+ and IB− to electrode S both generate feedthrough currents. If these capacitances are equal, the currents are equal and opposite, and no net current results. However, due to imperfections in device fabrication and wire-bonding, the resulting feedthrough current cannot be neglected and creates phase error in the oscillation loop. Each time the bias current reverses its sign, the feedthrough current also reverses its sign, creating phase error in the oscillation loop. Since the oscillation always occurs at the frequency where the phase shift around the loop is equal to 0°, the extra phase introduced by the feedthrough current shifts the oscillation frequency (and therefore the oscillation amplitude) of the MEMS resonator.

### B. Offset Stability

To demonstrate that the current chopping method also reduces the long-term drift error, the magnetometer is placed in a magnetically-shielded environment without temperature regulation for long-term measurement. A thermocouple (NI-USBTC01, National Instruments) with 0.1 °C resolution is mounted next to the magnetometer to measure the temperature. Figure 7 shows the measured sensor’s offset over 9-hour with the mean value subtracted. The maximum drift error before current chopping (blue curve) is ±500 \( \mu \)T, and is reduced to ±1 \( \mu \)T after current chopping (red curve). Before current chopping, the magnetometer’s offset is strongly correlated to the temperature change. This is mainly due to the quality factor (\( Q \)) dependence on temperature [17]. This effect is suppressed by the current chopping technique. The inset of Figure 7 shows a close-up view of the drift after chopping. The noise is 1 \( \mu \) TRMS, resulting from the sensor’s 400 nT/\( \sqrt{\text{Hz}} \).
noise density. The offset, averaged from 50 samples, varies by ±1 μT over 9 hrs.

To quantify the long-term drift over a range of averaging times, the Allan deviation of the same data set is evaluated and shown in Figure 8. For an averaging times shorter than 0.4 s, the magnetometer has the same short-term noise before and after chopping. At longer averaging times, the 1/f noise (slope of ~0.5) dominates the output without chopping, whereas white noise (Brownian noise in this particular magnetometer) is still the dominant noise source for averaging times up to 100 s after chopping. A bias drift of 60 nT is achieved for 171 s averaging time. For longer averaging times, a slope of ~0.5 is also observed after chopping, which is due to the 31 μT residual offset. A fit of the white noise indicates that if the residual offset is reduced by a factor of 10, a bias drift of 10 nT may be achieved.

Fig. 7. Measured sensor’s output before and after current chopping with the mean value of offset subtracted over 9 hours. Inset: Close-up view of the drift after chopping. The average value is shown in black.

Fig. 8. Allan deviation of the data from Figure 7. The current chopping method significantly improves the long-term stability of the magnetic sensor.

Fig. 9. Measured noise spectrum of the magnetic sensor before and after chopping.

IV. DISCUSSION

A. Noise Analysis

Similar to the nested chopping technique discussed in [18], the current chopping technique discussed here uses two pairs of choppers. One pair operates at the natural frequency of the mechanical resonator, which normally ranges from 1 kHz to 100 kHz depending on the resonator design. This chopper removes the 1/f noise from the front-end capacitive sensing electronics and amplifies the signal from the magnetometer by Q. The other pair operates at much lower frequency, which normally ranges from 1 Hz to 20 Hz. Figure 9 shows the measured sensor noise before and after chopping. The chopper suppresses the 1/f noise, 1/f² noise and 1/f³ noise from the mechanical resonator.

We fitted the measured Allan deviation using −0.5, 0, +0.5 and +1 slope lines and extracted Brownian noise, 1/f noise, 1/f² noise and 1/f³ noise components, respectively. The effect of chopping can be analyzed via the power spectral density (PSD) of each noise component. Because the sensor used in experiments had a white noise spectrum that was dominated by mechanical thermal (Brownian) noise, the contribution of electronic thermal noise was neglected. The chopper input PSD, \( S_N(f) \), based on the fitted noise components, is presented in Figure 10. The chopped output PSD is given by [19]:

\[
S_c(f) = \left(\frac{2}{\pi}\right)^2 \sum_{n = -\infty}^{+\infty} \frac{1}{n^2} S_n\left(f - \frac{n}{T}\right)
\]  

where \( T \) is the chopping period, which is 1 s in our measurement. From (8), it can be concluded that the noise floor after chopping is dominated by the \( n = \pm 1 \) terms. The contribution of higher-order harmonics decreases drastically because of the \( n^2 \) term in the denominator. The \( n = \pm 1 \) terms and chopper output PSD are also plotted in Figure 10. Note here that the 1/f³ noise is not plotted in the figure because it is much smaller than the 1/f and 1/f² noise for \( |fT| \geq 0.01 \), and does not contribute to the bias drift (minimum Allan deviation). This conclusion is consistent with the Allan
deviation plotted in Figure 8. Before chopping, the $1/f^3$ noise starts to dominate the Allan deviation for averaging times greater than 200 s. However, after chopping, $1/f^3$ noise is not observed for averaging times up to 4000 s and the minimum Allan deviation is instead determined by $1/f^2$ noise. Also, the $1/f^2$ noise is much lower than the $1/f$ noise for $|fT| ≥ 0.5$ in Figure 10. Therefore, we can use the $1/f$ noise and Brownian noise components to estimate the total residual noise in the chopped output PSD [19]:

$$S_C \approx S_0(1 + 0.8525f_k T) \text{ for } |fT| ≤ 0.5 \quad (9)$$

where $S_0$ is the input white (Brownian) noise PSD and $f_k$ is the $1/f$ noise corner frequency, which is the frequency at which $1/f$ noise is equal to the white noise floor. From the fitted value, $f_k$ is 0.3 Hz in our current setup. The chopped $1/f$ noise introduces approximately 0.27 $S_0$ white noise in the base band, which is equivalent to a 2 dB increase in the white noise floor. The modulation of $1/f$ noise and $1/f^2$ noise can be further reduced by increasing the chopping frequency $f_c$, which would also increase the measurement bandwidth. However, larger $f_c$ requires a larger mechanical bandwidth from the MEMS resonator.

B. Residual Offset and Offset Stability

From (6), the sensitivity $S_x$ is independent of the offset $x_E$. When the magnetometer is operating in closed-loop, the offset $x_E$ can be set via different means: (1) The loop-gain is initially larger than 1 at the start of oscillation, but as $x_E$ increases, the motional impedance of the MEMS resonator also increases due to mechanical or electrical non-linearity. Thus $x_E$ reaches its steady state when the loop-gain drops to 1 [20]. (2) An automatic level control (ALC) can be implemented to adjust the gain of oscillation loop by changing the gain in the detection electronics or by changing the dc bias on MEMS resonator [21]. (3) The oscillation amplitude can be clamped at a certain amplitude by using a hard-limiter or a 1-bit ADC before the MEMS resonator reaches its nonlinearity [22]. The first method is suitable for MEMS oscillators because it is desirable to have large oscillation amplitude to reduce phase noise. However, for Lorentz force magnetometers operating in closed-loop, operating the resonator in its non-linear region decreases the measurement range and stability. The second method is commonly used in modern MEMS oscillators and MEMS gyroscopes. The ALC is more predictable and more stable than the first method. It also provides a direct control of the oscillation amplitude. However, the ALC has greater power consumption, and also modulates $1/f$ noise back to the oscillation loop. In our previous implementation, we used the third method [3], which is power efficient and stable. In this work, we use a PLL to lock to the natural frequency to minimize the phase error in the oscillation loop, and also to provide a convenient way to set the electrostatic oscillation amplitude $x_E$ by adjusting $V_{AC}$. The PLL can be replaced by a hard-limiter or 1-bit ADC.

To reduce the residual offset, both the offset $x_E$ and parasitic capacitance need to be decreased. The former can be easily achieved by reducing the oscillation amplitude set-point. Figure 11 shows the measured Allan deviation for the same magnetometer operating at different oscillation amplitudes by adjusting $V_{AC}$. The oscillation amplitude is adjusted linearly from 1.3% to 13% of the capacitive air gap. Bias current was not applied during this measurement, and the sensor’s output voltage was converted to an equivalent magnetic field strength using the previously measured-scale factor [3]. It can be observed that $1/f$ and $1/f^2$ noise increase linearly as the oscillation amplitude increases, indicating that the drift error of the magnetometer is a constant fraction of the oscillation amplitude. Some measurement error is observed for averaging times greater than 1 s, which is likely due to temperature fluctuations in the lab environment. Since increasing the oscillation amplitude does not improve magnetic sensitivity but does increase drift, Figure 11 indicates that smaller oscillation amplitude helps to reduce residual offset and to improve the offset stability. However, extra care needs to be taken to ensure oscillation at smaller amplitude. Smaller amplitude makes the
oscillation loop more sensitive to the phase error introduced by the capacitive feedthrough current. The residual offset can be further suppressed by removing the feedthrough signal from feedthrough capacitance must be reduced to 1% of its drift error as AMR and GMR sensors, the error resulting from feedthrough capacitance must be reduced to 1% of its present value using feedthrough cancellation and improved symmetry in the layout, as described above. As for resolution, many Lorentz force magnetometers have resolution ranging from 50 nT/√Hz to 1000 nT/√Hz with 1 mA bias current, which is comparable to, or even better than, that of Hall effect devices by continuous spinning current method, [24]. Kyyvänäärinen’s Lorenz force magnetometer [2] achieved 1 nT/√Hz resolution with 1 mA bias current, which is comparable to an AMR magnetometer [25]. However, that device required extra fabrication steps to produce an electrically-isolated metal layer on top of the MEMS structure.

Table I compares the different types of magnetic sensors with offset suppressing methods. The Lorentz force magnetometer presented in this work achieves offset comparable to that of Hall effect sensors, however AMR and GMR sensors have been demonstrated with even lower offset. For a Lorentz force magnetometer to achieve the same residual offset and drift error as AMR and GMR sensors, the error resulting from feedthrough capacitance must be reduced to 1% of its present value using feedthrough cancellation and improved symmetry in the layout, as described above. As for resolution, many Lorentz force magnetometers have resolution ranging from 50 nT/√Hz to 1000 nT/√Hz with 1 mA bias current, which is comparable to, or even better than, that of Hall effect sensors [24]. Kyyvänäärinen’s Lorenz force magnetometer [2] achieved 1 nT/√Hz resolution with 1 mA bias current, which is comparable to an AMR magnetometer [25]. However, that device required extra fabrication steps to produce an electrically-isolated metal layer on top of the MEMS structure.

V. CONCLUSION

We propose a method to reduce the offset of Lorentz force magnetometers by chopping the bias current. The sensitivity is doubled whereas the drift error from the offset is suppressed. The method is verified with a single-axis Lorentz force magnetometer, and a 400X reduction in the offset is achieved. After current chopping, the magnetometer has an offset of 31 μT and a bias drift of 60 nT at 171 s averaging time. The offset can be further reduced by decreasing the parasitic capacitance or with feedthrough compensation. These results demonstrate that Lorentz force magnetic sensors can achieve offset levels suitable for use in electronic compass applications.

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TABLE I

<table>
<thead>
<tr>
<th>Sensor Type</th>
<th>Initial Offset</th>
<th>Suppress Method</th>
<th>Residual Offset</th>
<th>Drift Error</th>
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<tbody>
<tr>
<td>Hall Effect [7]</td>
<td>2–100 mT</td>
<td>Spinning Current</td>
<td>250 μT</td>
<td>Not Reported</td>
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<tr>
<td>Hall Effect [12]</td>
<td>10 mT</td>
<td>Orthogonal Switching</td>
<td>10 μT</td>
<td>Not Reported</td>
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<tr>
<td>Hall Effect [6, 9, 10]</td>
<td>2.42±0.04 mT</td>
<td>Symmetric Vertical Sensor</td>
<td>0.06±0.01 mT</td>
<td>Not Reported</td>
</tr>
<tr>
<td>AMR &amp; GMR [13, 14]</td>
<td>300 μT</td>
<td>AC Bias</td>
<td>1 μT</td>
<td>±10 nT/10 h</td>
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<tr>
<td>This work</td>
<td>25 mT</td>
<td>Current Chopping</td>
<td>31 μT</td>
<td>±1 μT/9 h</td>
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</tbody>
</table>

REFERENCES

Mo Li received the B.S. degree in microelectronics from Peking University, Beijing, China in 2011. He finished his undergraduate thesis in MEMS Lab, University of California, Davis as an exchange student. He is currently a Graduate Student Researcher with the MEMS Lab, University of California, Davis. His research interests include MEMS sensors and analog integrated circuits.

David A. Horsley (M’97) received the B.S., M.S., and Ph.D. degrees in mechanical engineering at the University of California, Berkeley, CA, in 1992, 1994, and 1998, respectively. He is an Associate Professor in the Department of Mechanical and Aerospace Engineering at the University of California, Davis, CA and has been a co-director of the Berkeley Sensor and Actuator Center (BSAC) since 2005. Prior to joining the faculty at UC Davis, he held research and development positions at Dicon Fiberoptics, Hewlett Packard Laboratories, and Onix Microsystems. His research interests include microfabricated sensors and actuators with applications in optical MEMS, communication, displays, and physical and biological sensors. Prof. Horsley is a recipient of the NSF CAREER Award and the UC Davis College of Engineering’s Outstanding Junior Faculty Award.