AUTOMATIC MODE MATCHING FOR HIGH-Q VIBRATORY RATE GYROSCOPES

C.D. Ezekwe^{1,2} and B.E. Boser¹

¹University of California, Berkeley, California, USA

²Robert Bosch Research and Technology Center, Palo Alto, California, USA

ABSTRACT

We demonstrate automatic mode-matching for high-Q micromachined vibratory rate gyroscopes. Unlike previously proposed techniques, our approach continuously tunes the sense resonance frequency to match the drive frequency and achieves a stable scale factor and more than 50Hz of bandwidth regardless of fabrication and drift.

INTRODUCTION

In a vibratory rate gyroscope, the readout interface senses the Coriolis acceleration, from which the angular rate is inferred. The Coriolis acceleration is sensed by detecting the capacitance between the moving proof mass and fixed electrodes. In converting the capacitance to either voltage or current, the front-end amplifier, invariably, introduces circuit noise, which limits the minimum detectable signal. Matching the drive and sense modes increases the acceleration-to-capacitance sensitivity by the Q-factor of the sense mode, which can be in excess of 1000 in vacuum packaged devices. With *mode matching*, the same Coriolis acceleration produces a much larger capacitance change, significantly reducing the impact of the circuit noise. This paper describes the techniques used to exploit mode matching effectively to achieve a Brownian motion limited noise floor of $0.004^\circ/\text{s}/\sqrt{\text{Hz}}$ over a 50Hz band while dissipating 1mW.

MODE MATCHING CHALLENGES

While mode matching amplifies the Coriolis acceleration induced motion by the Q-factor of the sense mode, it gives rise to several challenges that must be addressed in a practical system. With mode matching, the Coriolis signal is centered at the sense mode frequency, and, thus, the sensor bandwidth is fixed by the sense mode frequency and Q-factor. The resulting difficulty is illustrated by the system in [1], where the use of mode matching,

together with a traditional open-loop readout interface, limits the sensor bandwidth to below 1Hz, much less than the 50Hz required in automotive and consumer applications [2]. Furthermore, the bandwidth and scale factor in that system are poorly controlled due to the normally substantial variation of the Q-factor with the manufacturing process and the ambient. Figure 1 illustrates this problem. Besides, due to the limited accuracy of any practical mode-matching scheme, a residual frequency mismatch will invariably exist. The uncertainty in the residue due to process and ambient variations gives rise to further uncertainty in the scale factor. With the very abrupt change in the phase response near the sense mode frequency, the uncertainty exacerbates the task of synchronously demodulating the Coriolis signal with the correct phase, which is necessary for rejecting quadrature error. Figure 2 illustrates this problem.

Due to these difficulties, many gyroscope implementations avoid mode-matching altogether and, instead, operate away from the sense mode frequency, obtaining a larger bandwidth and better defined scale factor and phase response at the expense of sensitivity [3]. A practical readout interface exploiting the sense resonance must overcome the problems arising from mode matching in a way that neither interferes with gyroscope performance nor negates the advantage derived from mode matching.

FORCE FEEDBACK

Feedback is widely used in electronics to obtain precise characteristics, for example precise gains, from imprecise elements. It has been used in sensors to improve bandwidth, dynamic range, linearity, and drift [4]. Especially in high-Q vibratory gyroscopes with matched drive and sense modes, feedback is imperative to ensure proper operation.

Figure 1: Scale factor (gain) variation with Q-factor.

Figure 2: Scale factor and phase response variations with residual mismatch.

Figure 3 shows the sense dynamics enclosed in a force feedback loop. A compensator and a force transducer are added to the basic *open-loop* readout interface to form a *closed-loop* readout interface. Based on the motion sensed by the front-end, the compensator produces an estimate of the Coriolis force which the force transducer applies with opposite polarity on the proof mass to null its motion. Perfect nulling of the proof mass motion implies that the feedback force is exactly equal and opposite to the Coriolis force. While this is impossible to achieve over all frequencies in practice, adequate nulling is possible within a limited frequency band where the force feedback open-loop gain is sufficiently high. Within that frequency band, the output of the closed-loop interface is an accurate representation of the Coriolis acceleration.

Figure 4 compares the frequency responses of the open-loop sensor and that of a closed-loop interface that has a high open-loop gain over a frequency range that extends beyond the sense mode frequency. Electronic circuits implementing the compensator provide the open-loop gain necessary for a closed-loop response that is flat and stable over a much wider frequency range, regardless of the variations of the sensor parameters. This obviates the traditional tradeoff of mechanical sensitivity for larger bandwidth and better defined scale factor and phase response.

In this work, the feedback force is applied by reusing the sense combs as differential electrostatic actuators. Electrostatic actuators also produce spring softening, which, if not handled with care, results in the parasitic tuning of the sense mode frequency by the feedback voltage, with the unintended consequence of increasing the uncertainty in the residual mismatch.

Figure 3: Basic force feedback loop.

Figure 4: Illustration of the closed-loop response versus open-loop response of the sensor.

As an illustration of this problem, consider the schematic shown in Figure 5, where the proof mass is grounded, and the top and bottom electrodes are biased at V_{bias} and driven differentially by the feedback voltage v_{fb} . The feedback force applied on the proof mass for displacements that are small relative to the gap is

$$F_{fb} = \underbrace{2 \frac{C_{s0}}{x_g} V_{bias} v_{fb}}_{\text{actuator gain}} + \underbrace{2 \frac{C_{s0}}{x_g^2} (V_{bias}^2 + v_{fb}^2) x}_{\text{signal dependent stiffness}} \quad (1)$$

where x_g is the nominal gap and C_{s0} is the nominal sense capacitance between the proof mass and each pair of connected electrodes. In addition to the desired voltage controlled force with the indicated actuator gain, the actuator produces an unwanted stiffness term that also depends on the feedback voltage. With proportional control, this problem results in about 1% dynamic variation of the sense mode frequency in the experimental prototype, whereas an error of much less than 0.3% is required.

Bang-bang control, a feedback control strategy in which the feedback voltage is restricted to just two levels, overcomes this problem. With bang-bang control, the feedback voltage toggles between V_{bias} and $-V_{bias}$ in such a way that its time-average approximates the feedback voltage under proportional control. The technique converts the dynamic frequency variation into a static error since, regardless of the spectral content of the feedback voltage, the square is constant, V_{bias}^2 . The resulting static error is removed by calibration. Though bang-bang control results in a highly nonlinear force feedback loop, we use linear system techniques in the subsequent sections since they lead to useful results when used to analyze certain aspects of the loop behavior.

AUTOMATIC FREQUENCY CALIBRATION

Calibration is also necessary to guarantee resonance frequency matching in the presence of fabrication tolerances and ambient variations. Previously proposed calibration schemes determine the frequency mismatch by monitoring properties of the open-loop sensor [1], and therefore are incompatible with force feedback, which is imperative in a mode-matched gyroscope to stabilize the scale factor and widen the bandwidth. In this work, we inject a calibration signal into the force feedback loop to monitor the mismatch between the drive and sense resonance frequencies. Figure 6 shows the force feedback loop with the calibration input. The transfer function from the calibration input to the output assuming a high open-loop gain is

Figure 5: Schematic diagram of a differential electrostatic actuator.

$$G_{cal}(s) = \frac{1}{K_{v-f}K_{x-v}H_s(s)} \propto \frac{1}{H_s(s)}. \quad (2)$$

$G_{cal}(s)$, which is simply the inverse transfer function of the sense element, has a frequency response that, as shown in Figure 7, vertically mirrors the second-order response of the sensor and, thus, serves as a good proxy for $H_s(s)$. In fact, $G_{cal}(s)$ is preferable over $H_s(s)$ because it has zeros in place of the high-Q poles in $H_s(s)$ that severely limit the tracking bandwidth of conventional open-loop sensing based frequency calibration techniques such as the one in [1].

Two pilot tones serve as the calibration signal. The tones are at approximately $\pm 250\text{Hz}$ offset from the drive frequency to avoid interference with the Coriolis signal, enabling continuous background calibration. Their output amplitudes are nearly equal when the drive and sense resonance frequencies match. When the sense resonance frequency is *higher* (or *lower*) than the drive frequency, the amplitude of the higher frequency tone is *smaller* (or *larger*) than that of the lower frequency tone. Thus, the amplitude difference indicates the magnitude and direction of frequency mismatch. Figure 8 illustrates the estimation principle.

Figure 9 shows a simplified diagram of the complete calibration loop. The same tones that are applied at the calibration

Figure 6: Force feedback loop with added calibration input.

Figure 7: Comparison of $H_s(j\omega)$ and $G_{cal}(j\omega)$.

input are used to demodulate the amplitude information. The mean of the demodulated signal is, to first order, linearly proportional to the difference between the drive and sense resonance frequencies. Adjustment of the sense resonance frequency is accomplished using electrostatic spring softening obtained through a set of tuning combs dedicated to frequency tuning. A digital loop filter with infinite DC gain drives the mismatch to zero.

EXPERIMENTAL RESULTS

The frequency calibration technique was implemented as part of the readout interface described in [5]. Figure 10 shows the

Figure 8: Frequency mismatch estimation using two out-of-band pilot tones. The dashed curves represent $G_{cal}(j\omega)$, whose minimum occurs at the sense mode frequency.

Figure 9: Complete frequency calibration loop.

Figure 10: Measured output spectrum showing misplaced noise notch in the uncalibrated system.

output spectrum measured with and without calibration in the presence of an angular rate sinusoidally varying at 25Hz with an amplitude of 5.3°/s. Calibration prevents the misplacement of the noise notch inherent in the uncalibrated system. Figure 11 shows the measured electrostatic tuning voltage during startup. Due to improvements in the calibration algorithm, the settling time of the calibration loop has been reduced to 80ms from the 300ms we reported in [5]. With calibration, the readout interface achieves a Brownian motion limited noise floor of 0.004°/s/√Hz over a 50Hz band while dissipating 1mW. Figure 12 shows a micrograph of the readout interface ASIC and gyroscope.

CONCLUSION

This work combines several techniques to enable the matching of the drive and sense modes of a vibratory rate gyroscope. The system uses force feedback to overcome the bandwidth, scale factor, and phase response problems that arise with mode matching, and uses bang-band control to avoid the parasitic tuning of the sense resonance by the feedback voltage. Automatic background calibration guarantees resonance frequency matching over fabrication tolerances and ambient variations. The proposed calibration technique estimates the frequency mismatch by injecting out-of-band pilot tones into an appropriate point in the force feedback, and uses electrostatic spring softening to adjust the sense mode frequency.

Figure 11: Measured electrostatic tuning voltage during startup.

Figure 12: Micrograph of readout interface ASIC and gyroscope.

ACKNOWLEDGEMENTS

Help from C. Lang, V. P. Petkov, J. P. Vanderhaegen, X. Xing and others at Bosch RTC, and S. Lanzisera is gratefully acknowledged. The authors thank Robert Bosch LLC for supporting this work.

REFERENCES

- [1] A. Sharma, M. Zaman, F. Ayazi, "A 0.2°/hr micro-gyroscope with automatic CMOS mode matching," *ISSCC Dig. Tech. Papers*, pp.386-387, 2007.
- [2] N. Yazdi, F. Ayazi, K. Najafi, "Micromachined inertial sensors," *Proc. IEEE*, vol. 86, pp. 1640-1659, Aug. 1998.
- [3] M.S. Weinberg and A. Kourepenis, "Error sources in in-plane silicon tuning-fork MEMS gyroscopes," *Journal of Microelectromechanical Systems*, vol. 15, pp. 479-491, June 2006.
- [4] W. Yun, R. Howe, P. Gray, "Surface micromachined, digitally force-balanced accelerometer with integrated CMOS detection circuitry," *Tech. Dig. Solid-State Sensor and Actuator Workshop*, pp. 126-131, 1992.
- [5] C.D. Ezekwe, B.E. Boser, "A Mode-Matching $\Delta\Sigma$ Closed-Loop Vibratory-Gyroscope Readout Interface with a 0.004°/s/√Hz Noise Floor over a 50Hz Band," *ISSCC Dig. Tech. Papers*, pp.580-581, 2008.