A 104-dB Dynamic Range Transimpedance-Based CMOS ASIC for Tuning Fork Microgyroscopes

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Abstract—In this paper, the design, implementation and characterization of a continuous time transimpedance-based ASIC for the actuation and sensing of a high-Q MEMS tuning fork gyroscope (TFG) is presented. A T-network transimpedance amplifier (TIA) is used as the front-end for low-noise, sub-atto-Farad capacitive detection. The T-network TIA provides on-chip transimpedance gains of up to 25 M and consumes 15 mW of power. The CMOS interface ASIC uses this TIA as the front-end to sustain electromechanical oscillations in a MEMS TFG with motional impedance greater than 10 MΩ. The TFG is fabricated in a 0.6-μm standard CMOS process with an area of 500 mm² and rate sensitivity of 2 mV/deg/hr—one of the lowest recorded for MEMS gyroscopes to date. This is approximately three orders of magnitude lower than the requirements for automotive grade MEMS gyroscopes, which are similar full-scale ranges but the noise floor on the other. The TIA is fabricated in a 0.6-μm standard CMOS process with an area of 500 mm² and rate sensitivity of 2 mV/deg/hr—one of the lowest recorded for MEMS gyroscopes to date. Finally, microsystem performance is summarized in Section V.

I. INTRODUCTION

MICROMACHINED gyroscopes constitute one of the fastest growing segments of the microsensor market. They are small form-factor and low power consumption coupled with inexpensive IC-like mass production have generated numerous applications for such devices [1]. The application domain of MEMS gyroscopes is quickly expanding from automotive to consumer products, aerospace and personal navigation systems. Examples include anti-skid and safety systems in cars, image stabilization in digital cameras, smart user interfaces in handheld pointing devices for use in gaming/toys, and short-range navigation.

As silicon vibratory gyroscopes attain navigation grade performance [1], the interface electronics that actuate, sense and control these micromechanical structures are key elements in determining the overall performance of the micro-gyro system. Automotive and consumer product applications require rate noise floors in the range of 100°/deg/hr (deg/√Hz) and must be able to sense rotation rates as high as ±500°/sec. Navigation grade gyroscopes have similar full-scale ranges but the noise floor specifications are on the order of 0.1°/sec. These specifications are of the order of magnitude lower than the requirements for the commercial counterparts. Since vibratory gyroscopes, like micromachined accelerometers, are capacitive sensors, they translate to the need for ultra-low-noise front-ends which are able to detect sub-atto Farad capacitance changes [2], [3]. Additionally, while mechanical structures can typically attain dynamic ranges in excess of 120 dB, designing front-end electronics with such large dynamic range is challenging. The dynamic range is limited by the supply voltage on one hand and the noise floor on the other.

This paper presents system and circuit topologies that allow the integration of a high-performance MEMS tuning fork gyroscope (TFG) [4] into inertial measurement units (IMUs) while maintaining large dynamic range and low power operation. Section II gives a brief description of the mode-matched tuning fork gyroscope (M²-TFG) and discusses strategies and tradeoffs involved with interfacing CMOS circuits to microgyroscopes. A low-noise, T-network transimpedance amplifier (TIA) with a dynamic range of 104 dB is the proposed front-end used in this work and is subject to Section III. Section IV presents the interfacing results of the various CMOS system blocks with the micromechanical sensor element. The angular rate sensor exhibits a rate noise floor of 2.7°/hr/√Hz and a bias stability of 1°/hr—one of the lowest recorded for MEMS gyroscopes to date. Finally, microsystem performance is summarized in Section V.

II. MICROSYSTEM IMPLEMENTATION

A. Mode-Matched Tuning Fork Gyroscope—Principle of Operation

Fig. 1 shows the scanning electron micrograph (SEM) view of an in-plane tuning fork gyroscope [4] fabricated on 40-μm-thick silicon-on-insulator (SOI) substrate using a simple two-mask process similar to one used for micro-gravity accelerometers reported in [3]. The gyroscope is comprised of two proof-masses, supported by flexural springs and anchored at a central post. Actuation, sensing, quadrature nulling and tuning electrodes are distributed around the proof-masses and flexures. The sensor structure is maintained at a DC polarization voltage (Vp) to provide the bias for capacitive transduction and to prevent frequency doubling of the drive force [5]. The proof-masses are driven at resonance along the x axis using inter-digitated comb-drive electrodes. When the sensor undergoes a rotation about the z axis the resultant Coriolis acceleration causes the proof masses to vibrate along the y axis [1]. This rotation induced proof-mass motion causes the gap between the sense electrode and the proof-mass to change proportional to the applied rate and is detected electronically. The gyroscope is a resonant sensor and the
drive and sense modes have been designed to yield mechanical quality factors in excess of 40000 [4]. The resonant frequencies are in the range of 10–20 kHz.

B. Gyroscope System Architecture

Fig. 2 shows the complete system block diagram of the implemented microgyroscope system and a close-up SEM of the driving and sensing electrodes. Table I summarizes the key mechanical parameters of the microgyroscope used in this work. The interface electronics can be divided into the following main subsystems:

1) Drive Oscillator: The reference vibrations along the x axis are set up and sustained by placing the TFG in an electromechanical oscillator where the drive mechanical resonance is the frequency determining element. Proof-mass movement is detected by using the comb electrodes that are located symmetrically on the side of each proof-mass. The signal is sufficiently amplified, phase shifted to ensure a loop phase shift of 0° and then applied back to the central drive electrode. Driving the proof-masses at the center ensures that they are driven exactly in phase, preventing lock-in to spurious modes. An automatic level control (ALC) is used to control the amplitude of vibration. An off-chip phase-locked loop (PLL) locks on to the drive signal and is used to provide carefully phased signals for subsequent signal processing. While commercial gyroscopes use drive voltages in the order of 12 Vpp [6], the high drive quality factor of this TFG allows for drive voltages as low as 200 mVpp. This significantly lowers power dissipation and precludes the need for custom high voltage transistors for charge-pumps and off-chip capacitors as in [6]. This will help in reduction of the form-factor of the microsystem and lower cost by allowing the use of standard CMOS interfaces.

2) Quadrature nulling and Mode-matching: Fabrication imperfections of the mechanical structure results in off axis movement of the proof-mass, causing a residual displacement along the sense axis even in the absence of rotation [7]. This is referred to as quadrature error. This quadrature error is minimized by varying the mechanical bias voltages on the electrodes 2, 3, 6, and 7 in Fig. 2. The gyroscope performance is enhanced when the mechanical sense frequency (ω_{\text{SENSE}}) is matched to the drive resonant frequency (ω_{\text{Drive}}) due to the mechanical amplification provided by the effective quality factor (Q_{\text{EFF}}). This mode-matching is achieved by varying the polarization voltage (V_p) until the frequency separation between the drive and the sense mode is reduced to approximately 0 Hz [8] (at matched mode, ω_{\text{Drive}} ≈ ω_{\text{SENSE}} = ω_{\text{O}}). All analysis and measurements reported in this paper are done with the sensor at mode-matched condition (i.e., sensor operating frequency ω_{\text{O}} ~ 15 kHz). Appropriate mechanical design of this gyroscope results in the mode-matching to be extremely stable over both time as well as temperature [4].

3) Sense Channel: When subject to rotation about the z axis, the proof-mass vibrates along the sense (y) axis and the amplitude of vibration is modulated by the applied rate signal. The sense channel detects this proof-mass displacement and extracts the amplitude modulated rate information. Proof-mass displacements due to both Coriolis acceleration and quadrature error take place at the sensor resonant frequency (ω_{\text{S}}). The only distinguishing characteristic is that there exists a 90° phase difference between

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Drive comb gap (g)</td>
<td>7</td>
<td>µm</td>
</tr>
<tr>
<td>DC polarization voltage (V_p)</td>
<td>40</td>
<td>V</td>
</tr>
<tr>
<td>Drive displacement (δ_{\text{VOLS}})</td>
<td>~2.5</td>
<td>µm</td>
</tr>
<tr>
<td>Measured matched-mode quality factor Q_{\text{EFF}}</td>
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<td>—</td>
</tr>
<tr>
<td>Sense capacitor gap (δ_{c})</td>
<td>5</td>
<td>µm</td>
</tr>
<tr>
<td>Sensor resonant frequency (ω_{s})</td>
<td>15</td>
<td>kHz</td>
</tr>
<tr>
<td>Proof-mass dimensions</td>
<td>400 x 400</td>
<td>µm²</td>
</tr>
<tr>
<td>Thickness of the structure (h)</td>
<td>40</td>
<td>µm</td>
</tr>
<tr>
<td>Drive motional impedance</td>
<td>~10</td>
<td>MΩ</td>
</tr>
<tr>
<td>Sense motional impedance</td>
<td>~1</td>
<td>MΩ</td>
</tr>
<tr>
<td>Theoretical sensor Brownian noise floor</td>
<td>0.5</td>
<td>2/√hr/µgHz</td>
</tr>
</tbody>
</table>
the two. This 90° phase difference arises due to the fact that quadrature error is proportional to the proof-mass position, while the Coriolis response is proportional to the proof-mass velocity along the driven axis. The rate and quadrature are distinguished by demodulating the sense output with the 0° and 90° signals from the PLL, respectively. It is for this reason that the sense channel uses a phase sensitive synchronous I–Q demodulation scheme to extract the rotation rate, rather than a simple envelope detection scheme. Finally, the TFG electrode configuration shown allows for fully differential sensing topology, which automatically rejects linear acceleration as common-mode.

C. Challenges and Tradeoffs

The minimum detectable rotation rate depends on the noise floor of the system. Also referred to as the total noise equivalent rotation (TNEΩ), the system noise floor consists of two uncorrelated components (1)—the mechanical noise equivalent rotation (MNE Ω) and the electrical noise equivalent rotation (ENE Ω):

\[ \text{TNEΩ} = \sqrt{\text{MNEΩ}^2 + \text{ENEΩ}^2} \quad (\text{Units: } ^\circ/\text{hr}), \]  

\[ \text{MNEΩ} = \frac{1}{2k_B T} \sqrt{4k_B T \omega_0 M Q_{\text{EFF}}} \sqrt{\text{BW}} \quad (\text{Units: } ^\circ/\text{hr}). \quad (2) \]

Since a fixed DC potential (Vp) has been maintained across the sense gap (d_so), Coriolis-induced y axis displacement of the proof-mass in response to input rotation Ωz changes the sense rest capacitance (Cso), generating a motional current ISensor, given by

\[ I_{\text{SENSOR}} = \frac{2Vp C_s Q_{\text{EFF}} }{d_{so}} \sqrt{\omega_0 Vp} \Omega_z \quad (\text{Units: } A). \quad (3) \]

The ENEΩ of the microgyroscope depends on the minimum detectable capacitance (ΔC_MIN) of the sense channel interface electronics and the mechanical scale factor (F/°/hr):

\[ \text{ENEΩ} = \frac{\Delta C_{\text{MIN}}}{\text{capacitive sensitivity}} \quad (\text{Units: } ^\circ/\text{hr}). \quad (4) \]

For a parallel plate capacitive transducer, the minimum detectable capacitance change (ΔC_MIN) is proportional to the input-referred current noise of the interface electronics integrated over the bandwidth of interest, as given by

\[ \Delta C_{\text{MIN}} = \frac{I_{\text{noise}} \sqrt{\text{BW}}}{\omega_0 Vp} \quad (\text{Units: } F/\sqrt{\text{Hz}}). \quad (5) \]

While the use of high aspect ratio micromachining techniques like HARPS [9] lower the ENEΩ, the focus of this work will be circuit techniques to reduce the total input-referred current noise of the sensor electronics. The theoretical MNEΩ of the sensor used here is 0.5°/hr/√Hz, which means the electronic front-end must be able to detect a proof-mass displacement as small as 0.1 Å to resolve a capacitance change of 0.02 aF/√Hz at the sensor operating frequency (∼15 kHz).

The drive resonant mode of the TFG can be modeled as a two-port series RLC circuit [10], [11]. The reactive elements determine the mechanical frequency of resonance and the motional resistance (R_MOT−DRIVE) represents the transmission loss element. The value of R_MOT is obtained by equating the mechanical energy dissipated per cycle to the electrical energy supplied by the sustaining sources. In order to avoid lateral snap-down and maximize the drive displacement, the gap, g, between
adjacent comb electrodes must be increased [12]. This leads to larger $R_{\text{MOT-DRV}}$ values given by (6), where $K_{\text{DRV}}$ and $Q_{\text{DRV}}$ are the effective mechanical spring constant and quality factor, respectively, of the drive mode, $h$ is the structural thickness, $\varepsilon_0$ the relative permittivity constant ($8.85 \times 10^{-12}$ F/m), and $N$ the number of combs:

$$R_{\text{MOT-DRV}} = \frac{K_{\text{DRV}} \omega_0^2}{\varepsilon_0 Q_{\text{DRV}} (N\varepsilon_0 h)^2 V_P^2}.$$  

To achieve drive amplitudes of about 4–5 $\mu$m, which results in this microgyroscope having a drive motional impedance of about 16 M$\Omega$ in vacuum.

Large motional impedances require a large gain to be provided by the sustaining circuitry in the drive oscillator loop. These also require a higher AC drive voltage to be applied to the comb-drive electrodes, thereby dissipating more power. Circuits that can achieve large on-chip gains for capacitive detection, with low power and area overheads are therefore necessary.

The TFG implemented in this work is fabricated using a bulk-micromachining technology which allows for the fabrication of MEMS structures with narrow capacitive gaps and large inertial mass [4]. The sensor is fabricated on a different substrate and is connected directly to the IC via wire-bonds as shown in Fig. 3. A two-chip implementation allows decoupling of the MEMS design and fabrication from the design of the interface electronics. Sensor performance can be improved considerably, unlike [6], by leveraging the benefits of high aspect-ratio mixed-mode processes [9]. Secondly, standard CMOS processes can be used which significantly lowers cost and allows the electronics to be optimized for low power dissipation, speed and reliability. However, the front-end analog interface must be strategically chosen to ensure that the sub-pico-ampere level motional currents can be detected even in the presence of the increased parasitics.

III. FRONT-END CIRCUIT BLOCKS

Several techniques have been used in electronic front-ends to sense the small capacitive displacements in MEMS gyroscopes. Charge integration using switched capacitor front-ends with correlated double sampling were initially used for static MEMS accelerometers [2], [3], but have recently been used for microgyroscopes [13], [14]. These schemes are best suited for microgyroscopes with low operating frequency (<5 kHz) because of the power budget associated with the switching and clock generation [15]. Secondly, the effects of the capacitive loading of these front-ends on the microgyroscope quality factor have not been studied. The use of such a front end necessitates a switching voltage to be applied to the mechanical structure, which results in significant feedthrough and parasitic electrical coupling.

In [9] a unity gain, CMOS source follower amplifier was used as the front-end to detect capacitance changes in a vibrating polysilicon ring gyroscope. The DC bias at the pick-off electrode was set using a minimum geometry diode. The noise injected by the diode at the input can significantly degrade performance [16]. Special techniques like internal bootstrapping and feedback need to be applied to minimize the capacitance of the input transistor. Finally, the use of a unity gain buffer does not allow independent control of the signal-to-noise ratio (SNR) of the electronic front-end.

Continuous time (CT) charge integrator front-ends are attractive for sensing capacitive displacements in microgyroscopes [6], [7] because at typical operating frequencies, much larger AC impedances can be generated in a standard CMOS process using capacitors rather than resistors. Additionally, since these...
capacitors are not switched, there is no $kT/C$ noise associated with them. However, the CT charge integrator requires the use of a large resistor to bias the input node. Various techniques like the use of controlled impedance field-effect transistors (FETs) [6] and subthreshold MOSFETS [7] have been proposed in literature to implement these feedback resistors. The thermal noise of this resistor forms the dominant noise contributor of the front-end and determines overall performance.

Finally, TIAs that use a resistor for CT sensing of the motional current are described in [6], [7]. While the TIA is the interface architecture of choice for micromechanical resonator-based oscillators [11], [17], its use as a low-noise front-end for capacitive Coriolis detection has yet to be fully explored. In this work, a continuous time, programmable T-network TIA that provides state of the art capacitive resolution is proposed as the interface for the MEMS microgyroscope. This section presents analyses and measurement results which validate that sub-atto-Farad capacitance changes, and hence degree-per-hour rate resolutions can be detected using a CMOS TIA front-end.

A. Transimpedance Front-Ends for Motional Current Detection

Fig. 3 shows a schematic of a CT-TIA interfaced with a microgyroscope. This work differs significantly from [6] and [7] in that here, a TIA that has been optimized for noise is used as the front-end in both the drive loop as well as for sub-atto Farad capacitive detection in the sense channel. Further, the gain of the TIA is variable and the proof-mass is maintained at a constant DC potential unlike [7], [13], [14]. In Fig. 3, $R_F$ is the feedback resistance, $C_F$ is the associated stray capacitance and $C_{TOT}$ is the lumped parasitic capacitance at the inverting terminal of the operational amplifier (op-amp). $C_{TOT}$ is composed of the electrode to substrate capacitance on the MEMS die ($C_{PAD-MEMS}$ $\sim$ 1.5 pF), the interface IC pad capacitance ($C_{PAD-ASIC}$ $\sim$ 1.5 pF) and the gate capacitance of the input differential pair transistors ($C_{GS-INV}$ $\sim$ 0.5 pF) in the op-amp.

The high open loop DC gain of the op-amp ensures that the inverting terminal is a good virtual ground and the shunt-shunt feedback presents low input impedance to the high-impedance sensor pick-off node. This makes the signal path relatively insensitive to the total parasitic capacitance ($C_{TOT} = C_{PAD-MEMS} + C_{PAD-ASIC} + C_{GS-INV}$), preventing significant signal loss.

The low input impedance provided by the shunt-shunt feedback helps reduce the loading that the sustaining electronics will have on the quality factor of the gyroscopes drive mode. When locked into electromechanical oscillations, the loaded drive mode quality factor ($Q_L$) [11] is lower than its unloaded value ($Q_{UL}$)

$$Q_L = \frac{R_{MOT-DRV}}{R_{MOT-DRV} + R_{I-AMP} + R_{O-AMP}} \times Q_{UL}.$$  (7)

In (7), $R_{I-AMP}$ and $R_{O-AMP}$ are the input and output impedances seen by the sensor from the sustaining electronics. Since a TIA front-end provides low $R_{I-AMP}$, there will be minimal $Q$-loading.

The TIA interface allows the proof-masses to be maintained at a constant DC potential unlike [13]–[15] where an AC capacitance bridge configuration is used. Applying a switching signal to the proof-masses as in [14] increases the amount of electronic coupling into the zero rate output of the gyroscope. By maintaining the proof-mass at a constant DC potential, any spurious signal coupling into the sensing electrodes is eliminated and the number of demodulation and filtering stages required are minimized, thereby lowering power consumption as compared to [15].

Fig. 3 shows the main noise contributors in the transimpedance front-end, where $\frac{\alpha^2_{op-amp}}{R_F}$ and $\frac{\alpha^2_{op-amp}}{R_F}$ are the input-referred voltage and current noise of the core op-amp respectively and $I_F = 4 k_BT/R_F$ represents the thermal noise power of the feedback resistor ($R_F$). Since the sensor output is a current proportional to proof-mass displacement, it is the total input-referred current noise of the TIA front-end that ultimately determines the minimum detectable capacitance (5) and hence resolution of the microgyroscope. The equivalent input noise current ($I_{N-TOT}$) [18] for a TIA front-end is given by (8) which includes effects of both, the total parasitic capacitance seen at the input node ($C_{TOT}$) and the input resistance of the core amplifier, $R_{IN-AMP}$. The noise contributions of $\frac{\alpha^2_{op-amp}}{R_F}$ and of $R_{IN-AMP}$ are ignored for succinctness.

$$\frac{\alpha^2_{N-TOT}}{R_F} = \frac{\alpha^2_{op-amp}}{R_F} + \frac{4k_BT}{R_F} + \frac{\alpha^2_{op-amp}}{R_F} \times \left( \frac{1}{R_F} + \frac{1}{R_{IN-AMP}} + \omega^2 C_{TOT} \right)^2$$

$$\frac{\alpha^2_{N-TOT}}{R_F} \approx \frac{4k_BT}{R_F} + \frac{\alpha^2_{op-amp}}{R_F} \times \left( \frac{1}{R_F} + \omega^2 C_{TOT} \right)^2.$$  (8)

In a bandwidth of 10 Hz about the sensor operating frequency, the equivalent input noise spectrum is assumed white and thermal noise of the feedback resistor forms the dominant noise contributor. The electronic noise floor (ENE$\Omega$) for the $M^2$-TFG interfaced with a TIA is given by

$$ENE\Omega = \frac{d_{SO}}{2\nu_F C_{GS} Q_{EFF} Q_{drive}} I_{N-TOT} \sqrt{B.W.}.$$  (9)

The advantage of using a TIA front-end becomes clear when we consider the SNR of the front-end interfaced with a microgyroscope. A TIA with transimpedance gain $R_F$ yields an output signal voltage of $I_{NSN} \times R_F$ (for input motional current $I_{NSN}$) and output noise voltage of $\sqrt{4k_BT/R_F}$. The amount of displacement current $\dot{x}_{Brownian}$ due to the random Brownian motion of the proof-mass along the sense axis is derived by applying the equi-partition theorem [19] to the $M^2$-TFG at resonance and computing the noise displacement $\dot{x}_n$ [9], [11].

$$\frac{\dot{x}_n^2}{\dot{x}_n^2} = \frac{4k_BT Q_{EFF}}{M_o \omega_0^2}.$$  (10)

The sense resonant mode of the microgyroscope can be modeled as a second order system with an equivalent series RLC representation, very similar to that presented for the drive mode. The
Brownian noise displacement is related to the mechanical motional resistance of the sense mode \((R_{\text{MOT}} - \text{SN})\) and the equivalent Brownian noise current is derived:

\[
\tau_{\text{Brownian}} = \frac{1}{2} \omega_0 V^2 \left( \frac{\partial C_{\text{in}}}{\partial F} \right)^2 \tau_n^2 = \frac{4k_B T}{\omega_{\text{MOT}} - \text{SN}}.
\]  

(11)

By using \(i_{\text{SN}} = i_{\text{Brownian}}\), the overall SNR can be derived to be

\[
\text{SNR} = \sqrt{\frac{R_F}{R_{\text{MOT}} - \text{SN}}}.
\]  

(12)

Therefore, increasing \(R_F\) improves the total SNR of an angular rate sensor. From (8), (9), and (12) it is evident that a large \(R_F\) for capacitive detection is beneficial not only in terms of increased transimpedance gain, but also for better SNR and lower input current noise. Therefore, the basis of this work is to focus on strategies that yield large on-chip transimpedance.

In practice, the transimpedance (TZ) for the case of an op-amp with finite DC gain \((A_{\text{UVC}})\), dominant pole \(p_1\) (at frequency \(\omega_{p_1}\)) and input capacitance \(C_{\text{TOT}}\), is given by (13). \(C_{\text{TOT}}\) and \(R_F\) introduce a second pole \((p_2)\) at frequency \(\omega_{p_2} = 1/R_F C_{\text{TOT}}\) in the transfer function. Instead of rolling off close to the UGBW of the op-amp \(\omega_{\text{UGBW}}\), \(C_{\text{TOT}}\) causes the transimpedance gain to roll off much sooner.

\[
\frac{v_{\text{out}}}{i_{\text{in}}} = \frac{-R_F A_{\text{UVC}}}{1 + \frac{s}{\omega_{p_1}} + A_{\text{UVC}}}.
\]  

(13)

Further, since this is quadratic, the gain will peak before rolling off. The frequency at which gain peaking occurs, \(\omega_{\text{PEAK}}\), is given by the geometric mean of the input pole and the UGBW of the op-amp and sets the effective bandwidth of the TIA:

\[
\omega_{\text{PEAK}} = \sqrt{\omega_{p_2} \omega_{\text{UGBW}}}.
\]  

(14)

Gain peaking and the position of poles \(p_1\) and \(p_2\) affect the phase response of the transimpedance amplifier, placing restrictions on the maximum transimpedance that can be used in the microgyroscope. The phase characteristics of the sensor at resonance are key in distinguishing the rate signal from the quadrature error. When operating at matched-mode condition there is a net 360° phase shift due to the fact that there are four poles at the same frequency. To ensure that the precise phase relationship between the sensor quantities is not affected by the electronics, in this work, the TIA front-end has been designed to provide no more than 3.6° of phase shift (100 x lower) at the sensor resonant frequency. This automatically satisfies the stability requirement that the maximum phase shift at the TIA bandwidth be no more than 45°. Care has also been taken to ensure that the gain peaking frequency is higher than the sensor resonant frequency \((\omega_{\text{O}})\), so that bandwidth is not limited.

B. Low-Noise Wide Dynamic Range T-Network TIA

Large transimpedance gains can be implemented on-chip in a number of ways \([6],[7],[20],[21]\). In [6], the transresistance was implemented using a controlled impedance FET. In [20], long MOSFETs biased in the linear regime using a constant voltage were used. MOS-bipolar pseudo-resistors are used in \([21]\) for generating large resistances, but the maximum bandwidth obtained for the neural amplifier was 7.2 kHz. The main disadvantage of these approaches is that real-time control of the transresistance gain is not possible. Variation of the transresistance is possible to some extent using the approach proposed in \([6]\), but it involves variation of the duty cycle used to switch the controlled impedance FET. The strategy adopted in this work is to implement the feedback resistor in a TIA using a T-network of resistors. The implemented T-network TIA provides both, high gain and low-noise for sub-atto-Farad capacitive detection in an area and power efficient manner. Further it allows for a simple analog control of the transimpedance without excessive phase shift.

1) Design Considerations: Fig. 4 shows the complete schematic of the implemented T-network TIA front-end, interfaced for capacitive detection.

The equivalent transimpedance of the T-network TIA \((R_{\text{F EQ}})\) is given by (15), where the voltage divider formed by \(R_2\) and \(R_3\) in the feedback path provides an amplification of the equivalent transresistance:

\[
\frac{v_{\text{out}}}{i_{\text{in}}} \equiv R_{\text{F EQ}} = R_1 \left( 1 + \frac{R_2}{R_3} \right) + R_2.
\]  

(15)

The primary advantage of using the T-network is that it reduces the resistance levels to be placed on-chip, making on-chip integration tractable. In this work, \(R_1\) is implemented as a long MOS transistor operating in the triode or deep-triode regions, and \(R_2\) and \(R_3\) were on-chip poly-resistors. The MOSFET adds a degree of gain control to the transimpedance, which can be used for temperature compensation or for automatic level control applications. The resistances are designed such that \(R_1 \gg R_2\) and \(R_3\). From (15) it might seem that arbitrarily high transimpedance can be obtained by increasing the \(R_2/R_3\) ratio. However, in practice, bandwidth, noise, offset and stability tradeoffs limit the choice of this ratio.

For the case of the T-network TIA interfaced to a capacitive sensor, the SNR of the front-end degrades by a factor of \(\sqrt{(1 + R_2/R_3)}\) as given by

\[
\text{SNR} = \sqrt{\frac{R_{\text{F EQ}}}{R_{\text{MOT}} - \text{SN}}} \left( 1 + \frac{R_2}{R_3} \right).
\]  

(16)

This places a limit on the maximum transimpedance that can be used in the front-end. The SNR degradation can be explained better if one analyzes the noise gain of the T-network TIA. While the noise gain might not limit the absolute value of the input-referred current noise, it will impact the subsequent signal processing stages. The noise gain of the T-network TIA is given by

\[
A_{\text{NET}} = \left( 1 + \frac{R_2}{R_3} \right) \frac{1 + sR_1 (C_{\text{TOT}} + C_F)}{1 + sR_{\text{F EQ}} C_F}.
\]  

(17)

By appropriately sizing the resistor ratio \(R_2/R_3\), it is possible to ensure \(R_{\text{F EQ}} C_F \approx R_1 (C_{\text{TOT}} + C_F)\), thereby ensuring that the zero and pole cancel \([22]\), yielding a constant value.
The relationship to prevent excessive noise increase due to the T-network’s amplification of the op-amp’s voltage noise is therefore given by

$$\frac{R_2}{R_3} \leq \frac{C_{TOT}}{C_F}.$$  \hspace{1cm} (18)

$C_{TOT}$ for the two-chip solution varies between 2–5 pF. The stray feed-through capacitance ($C_F$) between the input and output is typically around 100–200 fF. An $R_2/R_3$ ratio of two was therefore chosen as it allowed for sufficient amplification of the transimpedance gain, without excessively increasing the noise gain. Typically in TIA’s, a shunt capacitance is placed in parallel with $R_F$ to alleviate gain peaking considerations. The use of a feedback T allows the increase of this capacitance to values that are less sensitive to parasitic effects [22].

DC offset restricts the maximum output signal swing thereby determining the upper limit of the dynamic range. The expression for the output DC voltage due to finite offset ($V_{OS}$) for a T-network TIA interfaced directly with a capacitive sensor is given by (19), which interestingly is the value of the noise gain at DC. Optimizing the $R_2/R_3$ ratio for noise gain automatically minimizes effects of DC offset.

$$V_O = \left(1 + \frac{R_2}{R_3}\right) V_{OS}. \hspace{1cm} (19)$$

A two-stage Miller-compensated operational transconductance amplifier (OTA) inherently has lower noise than a folded cascade OTA and was therefore chosen as the core amplifier. The OTA is biased with a 1 µA bias current that is generated by a constant transconductance bias circuit. Transistors M1, M2, M3, and M4 are the primary noise contributors [18]. To minimize the input-referred flicker noise of the designed OTA, the pMOS input transistors (M1, M2) were sized to be 300 µm/3 µm. The transconductance of the input transistors is also designed to be large enough to avoid noise contributions of other transistors. When biased with a current of 1 µA, their transconductance ($g_{m1} = g_{m2}$) is calculated to be 84 µS for the 0.5 µm process. nMOS load transistors (M3, M4) were designed to have a W/L ratio of 1.5 µm/6 µm to ensure that their thermal noise contribution is minimized. The $g_{m}$ of these transistors ($g_{m3} = g_{m4}$) is calculated to be 7.48 µS. The equivalent thermal noise floor of the core OTA was calculated to be about 17 nV/√Hz. Reliable flicker noise parameters ($K_{FP-MOS}$ and $K_{FN-NMOS}$) were not available for the process, but SPICE simulations performed with $K_{FP-MOS} \approx 5 \times 10^{-27}$ and $K_{FN-NMOS} \approx 10^{-26}$ yielded a flicker noise corner frequency between 1 to 10 kHz. Despite careful layout techniques and optimizing the core-amplifier to minimize systematic offset, the amplifier recorded an input-referred offset of about 2 mV. Table II summarizes the measured characteristics of the op-amp.

2) Characterization Results of T-Network TIA Front-End: To measure the transimpedance characteristics, the microgyroscope was replaced with an external resistor that had roughly the same value as the motional resistance along the sense axis ($\sim 1$ MΩ). A voltage signal was applied to this 1-MΩ resistor which was connected in series with the TIA and the gain was
characterized using an Agilent 4395A network analyzer. The input of the TIA saw exactly the same input capacitance ($C_{TOT}$) as it would in the case of interfacing with the gyroscope. The transimpedance gain was characterized for different values of gate control voltage ($V_{CNTRL}$) and plotted in Fig. 5.

At 10 kHz, the transimpedance gain can be varied between 0.2 MΩ to 22 MΩ by varying the gate control voltage of the MOS resistor in the feedback T. A transimpedance as large as 25 MΩ has been implemented on-chip, in a fraction of the area consumed otherwise. Optimizing the $R_2/R_3$ ratio has ensured that there is no gain peaking at the frequencies of interest, as evident from Fig. 5.

It is really never possible to measure the input-referred noise of a circuit! The total output noise of the T-network TIA was measured using the Agilent 4395A spectrum analyzer for different values of $V_{CNTRL}$. The noise measurement set-up was similar to the one used to characterize the gain. For noise measurement, the 1-MΩ series resistor was replaced by a capacitor (0.1 pF) that had roughly the same value as the total sensor rest capacitance. This prevents any noise from shunting to ground. The measured output voltage noise is divided by the transimpedance measured from Fig. 5 to yield the total input-referred current noise of the TIA. Fig. 6 plots the measured total input-referred current noise of the T-network TIA for different values of transimpedance.

Fig. 6 clearly shows that with increasing $R_F$, the current noise floor decreases, as predicted by (8). Therefore, a larger $R_F$ leads to a lower noise floor and hence smaller minimum detectable capacitance. In the region between 1–10 kHz, flicker noise is still significant and accounts for the $1/f$ characteristic.
The noise gain peaking due to the effect of the capacitance at the input node \((C_{\text{TOT}})\) is clearly visible from the plot. However, this noise gain peaking occurs beyond the sensor operating range (10–20 kHz) and there is a minima in the noise floor for the case of \(R_F = 1.6 \, \text{M}Q\) within the sensor operating range. This validates that optimally sizing the \(R_2/R_3\) ratio is effective in preventing excessive noise gain for the microgyroscope with sense motional impedance of 1–2 \(\text{M}Q\).

When interfacing CMOS front-ends with high-\(Q\) narrowband resonant MEMS sensors, the spot noise of the interface at the sensor resonant frequency determines the minimum detectable capacitance. At 15 kHz, the T-network TIA with a \(V_{\text{CNTRL}}\) of 0.96 V has a transimpedance gain of 1.6 \(\text{M}Q\) and an input-referred current noise of 88 fA/\(\sqrt{\text{Hz}}\). This corresponds to a capacitive resolution of 0.02 aF/\(\sqrt{\text{Hz}}\) at 15 kHz \((V_P = 40 \, \text{V})\). This is an order of magnitude better than that reported for the CT integrator of [20] and of the same order as the transcapacitance amplifier of [6]. Further, this is comparable to the capacitive resolution of the chopper stabilized front-end interface of [15] and has been attained without any power-consuming switching and does not require any clock generation.

Fig. 7 plots the measured transimpedance gain and phase characteristics for the case of \(R_F = 1.6 \, \text{M}Q\) \((\text{i.e.}, V_{\text{CNTRL}} = 0.96 \, \text{V})\). At the sensor operating frequency, the maximum phase deviation is found to be 3.9°. This is close to the target value of 3.6°, validating that the front-end T-network TIA provides the large gain and low noise without adversely affecting the phase relationship between the sensor signals.

Fig. 8(a) plots the measured output voltage noise of the T-network TIA for \(R_F = 1.6 \, \text{M}Q\) \((V_{\text{CNTRL}} = 0.96 \, \text{V})\) as well as the measured output voltage noise of the core amplifier. The thermal noise floor of the amplifier is measured to be about 25 nV/\(\sqrt{\text{Hz}}\), which is in excellent agreement with the theoretical value calculated earlier. Also, it is evident from the plot of the measured output noise of the amplifier that the flicker noise corner is between 1–10 kHz.

The maximum dynamic range provided by the front-end T-network TIA for sensing is therefore of interest. The maximum dynamic range is defined as [18]

\[
\text{DR}_{\text{MAX}} = \frac{\text{max output signal}}{\text{noise floor} \times \text{Bandwidth}},
\]

From Fig. 8(a), the measured output spot voltage noise of the T-network TIA at 10 kHz is about 250 nV/\(\sqrt{\text{Hz}}\). This is slightly higher than the thermal noise from an ideal 1.6-MQ resistor, which must be expected because of the noise gain of the T-network. This noise is integrated over a bandwidth of 10 Hz and is used to determine the lower bound of the dynamic range.

While the noise floor determines the lower end of the front-end dynamic range, the upper limit is determined by the maximum output swing of the T-network TIA. DC offset limits the output swing, thereby determining the maximum linear range of the front-end interface. In order to find the maximum (nondistorted) signal-to-noise ratio i.e., signal-to-(noise + distortion) ratio (SNDR) for the circuit, the input voltage level was swept upwards until the output signal was found distorted.

![Fig. 7. Measured transimpedance gain and phase relationship for the T-network TIA with a gain of 1.6 MΩ.](image-url)
Fig. 9. Schematic of the drive oscillator loop with the automatic level control circuit.

The maximum linear output swing of the TIA with TZ gain of 1.6 MΩ at a frequency of 10 kHz is limited to about 0.4 V\textsubscript{RMS}, as shown in Fig. 8(b). Beyond this level the nonlinearity in the output exceeds 2%, which is unacceptable. The distortion in carrier amplitude translates to phase inaccuracies and, hence, it will no longer be possible to distinguish the Coriolis signal from the quadrature error. Therefore, this forms the upper bound of the maximum usable dynamic range (DB\text{MAX}). The maximum dynamic range computed for a 10-Hz bandwidth and is found to be at least 104 dB, at the sensor resonant frequency.

IV. INTERFACING RESULTS

A. Drive Resonant Oscillator

The T-network TIA described earlier is cascaded with a phase shifting buffer to satisfy Barkhausen’s criterion and sustain oscillations in the series resonant electromechanical drive loop, as shown in Fig. 9. An automatic level control (ALC) circuit [23] is used to keep q\textsubscript{Drive} constant, thereby preventing false rate outputs. The diodes in the rectifier circuit of the ALC were implemented using the S/D-to-bulk junctions of a PMOS transistor. The ALC output voltage is passed through an off-chip low-pass ripple filter and used to control the gate of the MOS transistor in the front-end T-network TIA.

Fig. 10 shows the buffered closed loop drive oscillation waveform and the spectrum of the signal when the drive loop was interfaced with a second TFG. The high mechanical Q of the structure forms an excellent narrowband filter at the mechanical resonant frequency and therefore significantly alleviates the linearity requirements on the driving voltage output by the IC.

B. Synchronous Coriolis Detection

A regenerative two-stage comparator [24] converts the sinusoidal drive signal into a square pulse that is used to demodulate the AM Coriolis response and extract the rate signal. In the current implementation, an off-chip PLL locks-in to the drive signal and generates the 0° drv and 90° drv signals that are in phase with the proof-mass velocity and position respectively. The voltage-controlled oscillator (VCO) of the external PLL used here [25] has a maximum center frequency range that is significantly higher than the sensor operating frequency. The effects of PLL jitter are therefore negligible. Since the VCO
frequency is down-converted to generate the I–Q signals, the effects of the jitter are further suppressed. A CMOS Gilbert multiplier with 200-kΩ on-chip poly load resistors is used for the multiplication. Fig. 11 shows the mixer schematic and input and output test waveforms.

The output of the Gilbert multiplier is low-pass filtered to yield an analog signal proportional to the rotation rate. The integrated, active first-order low pass filter uses a 1.5-nF off-chip capacitor to set the cut-off frequency to 100 Hz and has a low-pass gain of about 2. The rate signal from the two channels can be converted to a single ended signal using an off-chip instrumentation amplifier. The M²-TFG was placed on an Ideal Aerosmith rate table and its scale factor was characterized as shown in Fig. 12.

The measured scale factor from one of the channels is 2 mV/°/s, with a maximum nonlinearity of 3% over the measured range. The sources of nonlinearity in the sense channel signal processing chain are the slight difference in the capacitive gaps of the MEMS structure, the nonlinearity of the front-end TIA and more significantly, the incomplete cancellation of the higher order harmonic terms as the Gilbert cell was operated in a single ended configuration. Fig. 12 also shows the sensor response to a 1.5-Hz sinusoidal input rotation as well as the response of the microgyroscope to both positive (CCW) and negative (CW) input step rotations.

C. System Integration

The noise floor and long-term stability of the microgyroscope interfaced with electronics was characterized by performing an Allan variance analysis [26] on the zero rate output (ZRO). For microgyroscopes, the root Allan variance is the preferred means of specifying the noise floor rather than the power spectral density (PSD), and is the method adopted in this work. The ZRO from one of the channels was buffered with an off-chip amplifier with a gain of 10 and sampled every 100 ms for a period of 12 hours using an Agilent 34401A digital multi-meter. The root Allan variance plot obtained without applying any pre-whitening or filtering is shown in Fig. 13 and the inset shows a
time slice of the sampled ZRO. The slope at short cluster times (τ) yields the angle random walk (ARW) which is a measure of the white noise in the system. The ARW is 0.045°/√hr which corresponds to a measured noise floor of 15 μV/√Hz (~96 dBV/√Hz) over the signal bandwidth (10 Hz) for the entire microsystem. The output referred total equivalent noise density (MEMS + electronics) is therefore 2.7°/hr/√Hz. This is about an order of magnitude better than commercially available gyroscopes [6] and is one of the lowest recorded for a silicon vibratory gyroscope.

The second significant performance metric is the bias drift which is a measure of the long-term stability of the microgyro system. Bias drift in a microgyroscope is important since it can accumulate over time, resulting in large errors in angular position. The minimum of the Allan variance plot gives the value of the bias drift of the system [26], which for this case is 1°/hr. This about 50X better than [6] and is one of the lowest recorded for MEMS gyroscopes till date.

The increase in the root Allan variance at large cluster times indicates the presence of a rate random walk component [26], [27]. In this case, it is attributed to the incomplete nulling of the quadrature error in the MEM structure. Despite a low-noise TIA front-end, the measured spot rate noise floor at the output of the microgyroscope system is slightly higher than the theoretical noise floor of the MEM sensor itself (0.5°/hr/√Hz). This is primarily due to the amplification of the TIA noise by the gain of the remaining portion of the sense signal chain. Additionally the noise contribution of the subsequent signal processing stages, which include the multiplier, low pass filters and external buffers adds to the overall output referred noise floor. This can be significantly decreased by the use of bipolar stages [6] or by the use of low-noise chopper stabilization techniques [28] in the final output stage.

Fig. 14 shows the micrograph of the 3-V 0.6-μm CMOS IC that is interfaced to the M^2-TFG using wire-bonds on a custom PCB [29].

V. CONCLUSION

In this paper, a MEMS tuning fork gyroscope is interfaced with a transimpedance-based 0.6-μm two-poly three-metal (2P3M) CMOS IC to yield a low-cost two-chip angular rate sensor. The key parameters are summarized in Table III. Large transimpedances are integrated on-chip by the use of a T-network of resistors in the feedback path. Design considerations to mitigate the effects of the parasitic capacitances while maintaining high SNR are derived for the proposed front-end T-network TIA. By optimizing the core OTA for noise and strategically sizing of the resistor ratios of the T-network, a low-noise front-end with a capacitive resolution of 0.02 aF/√Hz and wide dynamic range of 104 dB in a bandwidth of 10 Hz have been demonstrated. This low-power front-end provides an attractive alternative to switched capacitor front-end interfaces for micromachined resonant sensors. The proposed T-network TIA is used as the core building block for both the drive oscillator and sense channels in the implemented angular rate sensor. The large transimpedance gains achieved are used to set up resonant oscillations in the MEMS structure with motional impedances as large as 10 MΩ. The implemented angular rate sensor system exhibits a rate noise floor of 2.7°/hr/√Hz and a bias stability of 1°/hr.
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REFERENCES


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