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# A 19.8-mW Eddy-Current Displacement Sensor Interface With Sub-Nanometer Resolution

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**Abstract**—This paper presents an eddy-current sensor (ECS) interface intended for sub-nanometer (sub-nm) displacement sensing in hi-tech applications. The interface employs a 126-MHz excitation frequency to mitigate the skin effect, and achieve high resolution and stability. An efficient on-chip sensor offset compensation scheme is introduced which removes sensor offset proportional to the standoff distance. To assist in the ratiometric suppression of noise and drift of the excitation oscillator, the ECS interface consists of a highly linear amplitude demodulation scheme that employs passive capacitors for voltage-to-current (V2I) conversion. Using a printed circuit board-based pseudo-differential ECS, stability tests were performed which demonstrated a thermal drift of  $<7.3$  nm/°C and long-term drift of only 29.5 nm over a period of 60 h. The interface achieves an effective noise floor of  $13.4$  pm/ $\sqrt{\text{Hz}}$  which corresponds to a displacement resolution of 0.6 nm in a 2-kHz noise bandwidth. The ECS interface is fabricated in TSMC 0.18- $\mu\text{m}$  CMOS technology and dissipates only 19.8 mW from a 1.8-V supply.

**Index Terms**—Amplitude demodulation, chopping, displacement, eddy current, inductive, oscillation, sensor.

## I. INTRODUCTION

**SUB-NANOMETER** (sub-nm) displacement sensing is one of the most important requirements in advanced metrology and high-tech industries, e.g., wafer scanners and relay mirror experiment (RME) [1]. Incremental sensors, such as linear encoders and interferometers, possess high resolution, and therefore, are frequently used in these applications. However, they are bulky and costly. Absolute sensors, such as capacitive and inductive sensors, can compete with the incremental sensors in these applications because the typical measurement range ( $\Delta x_{\text{max}}$ ) is small: in the order of few micrometers.

Capacitive sensors [2], [3], though compact with high resolution, require electrical access to the target, which is often an issue. They are also sensitive to environmental conditions such as humidity, dirt, and so on. Eddy-current sensors (ECSs), on the other hand, are immune to these disadvantages, but their stability is limited by the skin effect [4], [5]. Since

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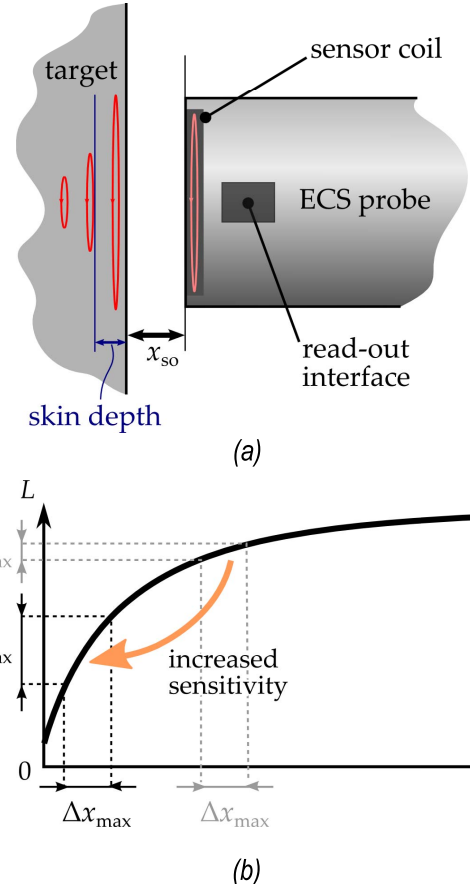


Fig. 1. Sub-nm ECS interface. (a) Readout chip integrated in the sensor probe. (b) Small  $x_{so}$  improves sensitivity and sensor-to-offset ratio.

eddy current is induced deep within the target, the effective displacement sensed by an ECS is dependent on the skin depth  $\delta (= 1/(\pi\mu\sigma f_{\text{exc}})^{1/2})$ , where  $\mu$  is the magnetic permeability,  $\sigma$  is the electrical conductivity, and  $f_{\text{exc}}$  is the excitation frequency of the ECS. Hence, a change in skin depth can be taken as an apparent displacement. For instance, a temperature change of 1 °C in a copper target can cause displacement errors of  $\approx 100$  nm, when  $f_{\text{exc}}$  is 1 MHz. To attain sub-nm measurements, the temperature has to be controlled with an accuracy of 1 mK which is often not possible. A practical way to obviate this instability issue is to use a higher excitation frequency ( $f_{\text{exc}} > 100$  MHz) in the ECS interface [5]–[7].

However, it is not trivial to increase the excitation frequency in the present ECS interfaces. This is largely due to the: 1) use of large ECS coils ( $L > \text{few } \mu\text{H}$ ); 2) large parasitic

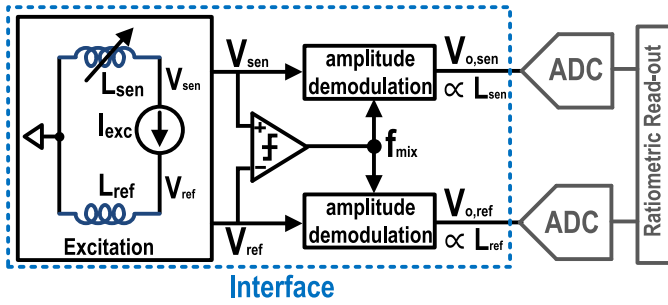


Fig. 2. Conceptual block diagram of the ECS interface.

inductance ( $L_{\text{par}}$ ) of the cable between the ECS coil and readout circuitry; and 3) high power dissipation in the readout circuitry [8], [9]. Large ECS coils typically have a low stability and a low self-resonance frequency (SRF). The parasitic inductance  $L_{\text{par}}$  of the cable, between the ECS coil and the readout circuitry, degrades the sensitivity and exacerbates the dynamic range (DR) by adding offset inductance. Parasitic inductance also degrades the stability of the ECS interface. The readout circuitry is typically not monolithic and dissipates a large amount of power ( $>500$  mW). This makes it difficult to remove the cable between the ECS coil and the readout circuitry because self-heating can cause large displacement errors due to thermal expansion [10]–[12].

A sub-nm ECS must use a stable flat sensing coil (with high SRF) in close proximity to the ECS interface, the power dissipation of which must then be low enough ( $<20$  mW) to avoid self-heating and displacement errors due to thermal expansion. A monolithic solution for the readout circuitry is attractive for a compact ECS interface with low power dissipation. The conceptual diagram of such an ECS interface is shown in Fig. 1(a), where a power efficient monolithic readout interface is integrated into the sensor probe to obviate a large  $L_{\text{par}}$ .

Mechanical assembly poses another challenge for ECSs [5]. Although sub-nm resolution must be achieved over a displacement range ( $\Delta x_{\text{max}}$ ) of only a few micrometers, the nominal standoff distance ( $x_{\text{so}}$ ) to the target is often a few hundreds of micrometers due to mechanical assembly tolerances. There are two issues with a large standoff distance: 1) low sensitivity due to non-linear characteristics of the sensor [Fig. 1(b)] which necessitates lower noise from the readout circuitry and 2) large sensor-offset ( $\propto x_{\text{so}}$ ) which exacerbates both the DR and non-linearity of the ECS interface. Thus, a smaller  $x_{\text{so}}$  is attractive as it improves the sensitivity and the signal-to-offset ratio of the system.

A conceptual block diagram of a pseudo-differential ECS interface is shown in Fig. 2 which performs three important functions: excitation, demodulation, and ratiometric readout. ECS coils, a sensor coil  $L_{\text{sen}}$  and a reference coil  $L_{\text{ref}}$ , are excited by a high frequency excitation current  $I_{\text{exc}}$ . This produces two voltages  $V_{\text{sen}}$  and  $V_{\text{ref}}$ , the amplitudes of which are directly proportional to  $L_{\text{sen}}$  and  $L_{\text{ref}}$ , respectively. Oscillator outputs are amplitude demodulated to generate baseband voltages,  $V_{o,\text{sen}}$  and  $V_{o,\text{ref}}$ , using a synchronous clock ( $f_{\text{mix}} = f_{\text{exc}}$ ). These voltages are digitized and their ratio is calculated

to compute  $D_{\text{out}} = L_{\text{sen}}/L_{\text{ref}}$ , which gives the measure of the displacement. The amplitude is chosen as the carrier of the displacement information due to higher sensitivity when compared to the frequency or losses in the  $LC$  resonator [13]–[15].

Self-oscillation is found to be a power-efficient way to generate  $I_{\text{exc}}$  and to excite ECS coils at a high  $f_{\text{exc}}$  [16]. However, noise and drift of the front-end excitation oscillator are important concerns. Ratiometric measurement helps in suppressing these because of their correlated multiplicative (CM) nature. However, the efficacy of this suppression is limited by the presence of any non-linearity, e.g., due to  $g_m$  stages, in the demodulator [4], [17]. It is also more difficult to attain high linearity ( $>70$  dB total harmonic distortion) at a high  $f_{\text{exc}}$ . The non-linearity of the demodulator is further exacerbated due to the presence of large input offsets caused by a large  $x_{\text{so}}$  and  $L_{\text{par}}$ . To maintain the efficiency of the ratiometric measurements in suppressing the noise and drift of the excitation oscillator, linear amplitude demodulation and on-chip sensor offset compensation are indispensable.

In this paper, we present an ECS interface with displacement resolution of 0.6 nm in a 2-kHz noise bandwidth, with a power consumption of only 19.8 mW [18]. This is achieved by: 1) exciting ECS coils with a high  $f_{\text{exc}}$  ( $=126$  MHz) to alleviate the performance degradation caused by the skin effect; 2) utilizing capacitors for voltage-to-current (V2I) conversion to obviate the use of non-linear  $g_m$  blocks in the demodulator and perform linear amplitude demodulation; and 3) introducing an on-chip sensor offset compensation scheme to remove the sensor-offset ( $\propto x_{\text{so}}$ ), without degrading the efficacy of the ratiometric suppression of the excitation oscillator's noise and drift.

This paper is organized as follows. Section II discusses the architecture of the ECS interface, along with the proposed sensor offset compensation scheme. Section III represents the circuit implementation of important blocks of the interface. Section IV discusses experimental results achieved from the ECS interface prototype. Section V summarizes the performance and concludes the paper.

## II. ECS INTERFACE

### A. Architecture Overview

Fig. 3 depicts the detailed system architecture of the ECS interface. The interface consists of two channels: a sensor channel and a reference channel. These process voltage amplitudes directly proportional to the sensor coil  $L_{\text{sen}}$  ( $= L_{\text{so}} \pm \Delta L(x)$ ) and the reference coil  $L_{\text{ref}}$  ( $= L_{\text{so}}$ ), respectively, where  $L_{\text{so}}$  is the value of  $L_{\text{ref}}$  (and  $L_{\text{sen}}$ ) at the nominal standoff  $x_{\text{so}}$ . The excitation of ECS coils is achieved with a pMOS-based self-oscillating front-end [16] which comprises of  $L_{\text{ref}}$ ,  $L_{\text{sen}}$  and a tank capacitance  $C$ . The excitation frequency generated by them is given as  $f_{\text{exc}} = 1/2\pi((L_{\text{sen}} + L_{\text{ref}}) * C)^{1/2}$ . The voltage swings of the excitation oscillator outputs  $V_{\text{sen}}$  and  $V_{\text{ref}}$  are controlled by a digitally programmable tail current source  $I_{\text{ss,osc}}$ .

The excitation oscillator outputs  $V_{\text{sen}}$  and  $V_{\text{ref}}$  are converted into currents  $I_{\text{sen}}$  and  $I_{\text{ref}}$  using capacitors  $C_{\text{in}}$  and  $C_{\text{ref}}$ . The

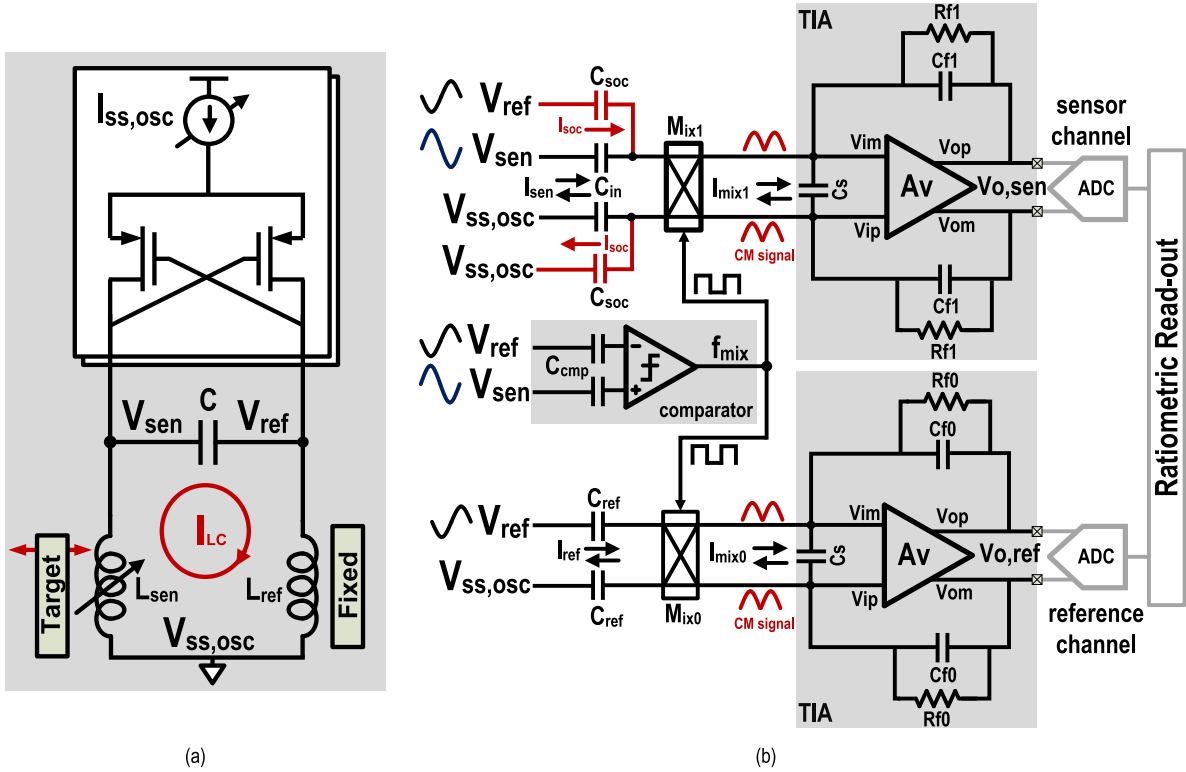


Fig. 3. Architecture of the ratiometric ECS interface with (a) excitation oscillator and (b) two channel TIA amplitude demodulation.

low input impedance of the transimpedance amplifier (TIA) is exploited for this. A clock  $f_{\text{mix}}$  synchronous to  $I_{\text{sen}}$  and  $I_{\text{ref}}$  (but in quadrature to  $V_{\text{sen}}$  and  $V_{\text{ref}}$ ) is generated using a fast continuous-time comparator. Currents  $I_{\text{sen}}$  and  $I_{\text{ref}}$  are demodulated to zero intermediate frequency (IF) using passive current mixers  $M_{\text{ix}1}$  and  $M_{\text{ix}0}$  [19]. TIAs amplify and filter demodulated currents  $I_{\text{mix}1}$  and  $I_{\text{mix}0}$  to generate baseband voltages  $V_{o,\text{sen}}$  and  $V_{o,\text{ref}}$ . These voltages are digitized and a ratiometric readout  $D_{\text{out}} = V_{o,\text{sen}}/V_{o,\text{ref}}$  is performed to suppress the drift and noise of the excitation oscillator.

### B. Sensor Offset Compensation

As shown in Fig. 1(b), the maximum sensor inductance change  $\Delta L_{\text{max}}$  is usually a small fraction of the standoff inductance  $L_{\text{so}}$  ( $\propto x_{\text{so}}$ ). Hence,  $I_{\text{sen}}$  consists of a large signal proportional to  $L_{\text{so}}$  (and  $x_{\text{so}}$ ), along with the useful displacement information in  $\Delta L(x)$ . This offset signal degrades the linearity of the demodulator, and therefore, must be compensated for before being amplified.

The proposed ECS interface leverages the anti-phase relationship between the sensor and reference coil voltages for sensor offset compensation. To mitigate sensor offset,  $V_{\text{ref}}$  is used to generate a compensation current  $I_{\text{soc}}$  through capacitor  $C_{\text{soc}} (= C_{\text{in}})$ . This current is added to  $I_{\text{sen}}$  at the input of  $M_{\text{ix}1}$ . The TIA output  $V_{o,\text{sen}}$  then depends only on  $\Delta L(x)$ . The signal gain in the reference channel is scaled down by a factor  $\beta (= \Delta x_{\text{max}}/x_{\text{so}})$  with  $C_{\text{ref}} = \beta * C_{\text{in}}$  so that the reference channel output  $V_{o,\text{ref}}$  corresponds to only the maximum inductance change  $\Delta L_{\text{max}}$  and not  $L_{\text{so}}$ . The modified ratiometric readout

$D_{\text{out}}^*$  is given by

$$D_{\text{out}}^* = \frac{V_{o,\text{sen}}}{V_{o,\text{ref}}} = \frac{|V_{\text{sen}}| - |V_{\text{ref}}|}{\beta \cdot |V_{\text{ref}}|} = \frac{\Delta L(x)}{\beta \cdot L_{\text{so}}} = \frac{\Delta x}{\Delta x_{\text{max}}} \quad (1)$$

where  $V_{o,\text{sen}}$  and  $V_{o,\text{ref}}$  are the output of the sensor and the reference channel, respectively,  $|V_{\text{sen}}|$  and  $|V_{\text{ref}}|$  are the amplitudes of the two outputs of the oscillator. Due to the proposed sensor offset compensation technique, the sensor channel output  $V_{o,\text{sen}}$  is proportional to  $|V_{\text{sen}}| - |V_{\text{ref}}|$ , i.e.,  $\Delta L(x)$  and does not depend on the standoff inductance  $L_{\text{so}}$ .

The choice of using  $V_{\text{ref}}$  for the sensor offset compensation is important. Current  $I_{\text{soc}}$ , generated from  $V_{\text{ref}}$  using capacitor  $C_{\text{soc}}$ , contains all CM terms of the excitation oscillator, and hence, preserves the functionality of the ratiometric readout in suppressing them after sensor offset compensation. This would not be possible if the compensation current is not generated from the excitation oscillator itself. Fig. 4 depicts the working principle of the sensor channel, with sensor offset compensation, for the case  $L_{\text{sen}} > L_{\text{ref}}$ , i.e., when the target is farther from the nominal standoff position.

## III. CIRCUIT IMPLEMENTATION

### A. Front-End Excitation Oscillator

This section discusses the implementation of the excitation oscillator of the ECS interface. As shown in Fig. 5, a pMOS-based cross-coupled LC oscillator structure is employed. Off-chip ECS coils  $L_{\text{sen}}$  and  $L_{\text{ref}}$  form a resonator along with the on-chip capacitor  $C$ . The voltage swing can be calculated as

$$|V_i| = j \cdot 2\pi \cdot f_{\text{exc}} \cdot L_i \cdot I_{\text{LC}} \quad (2)$$

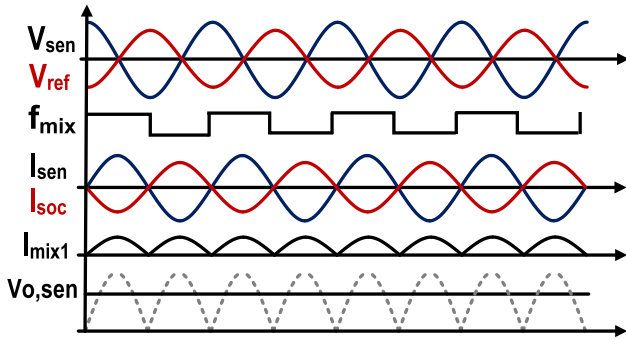


Fig. 4. Working principle of the sensor channel in the ECS interface.

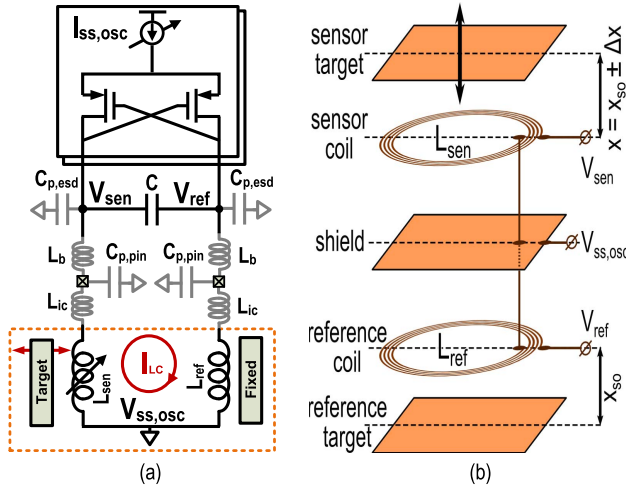


Fig. 5. Self-oscillating front-end to excite ECS coils. (a) Excitation oscillator schematic. (b) Conceptual sensor structure.

where  $V_i$  is the excitation oscillator output voltage corresponding to the ECS coil  $L_i$ ,  $f_{exc}$  is the excitation frequency generated by the  $LC$  resonator and  $I_{LC}$  is the resonance current flowing in the  $LC$  tank.  $I_{LC}$  can be derived from the excitation oscillator bias current  $I_{ss,osc}$ , tank quality factor  $Q$  and switching efficiency  $\eta$  [20], [21]. The oscillator bias current  $I_{ss,osc}$  is digitally programmable to provide currents from  $520 \mu\text{A}$  to  $2.16 \text{ mA}$ . The tail current source is cascoded to prevent modulation due to oscillator output swing. One important point to note is that the effective oscillator amplitude noise also flows through both the ECS coils (like  $I_{LC}$ ) and is suppressed with the ratiometric measurement [4].

Since ECS coils are off-chip, care must be taken to minimize various parasitic elements between coils and interface circuitry. Fig. 5(a) shows the important parasitic components, namely: electrostatic discharge (ESD), cell parasitic capacitance  $C_{p,esd}$ , package pin parasitic capacitance  $C_{p,pin}$ , bond-wire inductance  $L_b$ , and coil-to-pin interconnect parasitic inductance  $L_{ic}$ . ESD cells with 1-KV human body model ESD protection and a small parasitic capacitance ( $\approx 100 \text{ fF}$ ) are chosen so that  $C_{p,esd}$  is much smaller than the tank capacitance  $C$  ( $\approx 8 \text{ pF}$ ). A QFN package with a small  $L_b$  ( $\approx 1.5 \text{ nH}$ ) and  $C_{p,pin}$  ( $\approx 0.3 \text{ pF}$ ) is used. The matching between two  $L_b$ s is attempted by connecting nodes  $V_{sen}$  and  $V_{ref}$  to pins of the package which are at same angles. One important concern is that  $L_b$  can make

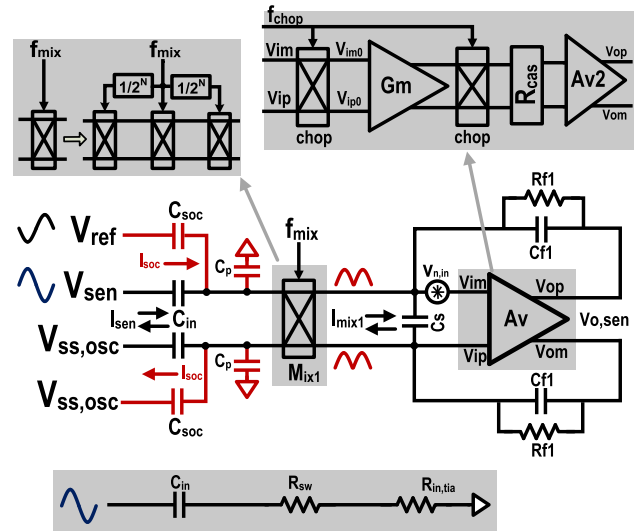


Fig. 6. Architecture of the sensor channel demodulator.

a high-frequency parasitic resonance loop along with  $C_{p,pin}$  and  $C$ . Then, the front-end can start to oscillate at a frequency which is independent of the sensor and reference coils. This is obviated by ensuring that the primary  $LC$  resonator has a higher  $Q$  than parasitic resonators. This also limits the lower bound of the excitation frequency [4].

Fig. 5(b) illustrates the conceptual structure of the sensor head. It consists of two coils  $L_{sen}$  and  $L_{ref}$  of identical geometry. The matching between the sensor and the reference coils is important for the efficacy of the sensor offset compensation technique. The coils are shielded from each other using a grounded middle layer to obviate mutual inductance. The thickness of the shield should be at least six times higher than the skin depth at the excitation frequency, to prevent interaction between the two coils. The nominal value of  $L_{ref}$  is achieved by positioning a fixed reference target at a nominal standoff  $x_{so}$  from the reference coil. The nominal standoff for the sensor coil is chosen to be same as  $x_{so}$ . This approach has three important advantages: 1) the nominal inductance of both the ECS coils is very close at  $x = x_{so}$  which helps in accurate sensor offset compensation; 2) drift in ECS coils, e.g., due to thermal expansion, is suppressed by the ratiometric readout; and 3) the real reference quantity used in the system is the distance between the reference target and  $L_{ref}$ . This is attractive because we are sensing displacement and using the reference same as the measurand makes the system inherently more stable [22].

### B. Amplitude Demodulation

This section discusses the implementation of the amplitude demodulation block in the ECS interface. Fig. 6 depicts the details of demodulator in the sensor channel. The demodulation in the reference channel is identical to that of the sensor channel, the only difference being  $C_{ref} = \beta * C_{in}$ . As explained in Section II, capacitor  $C_{in}$  ( $= 500 \text{ fF}$ ) is used to generate current  $I_{sen}$ . Capacitor  $C_{soc}$  ( $= C_{in}$ ) generates  $I_{soc}$  from  $V_{ref}$  to compensate for the sensor offset. For this reason,  $I_{sen}$  and

$I_{\text{SOC}}$  are added at the input of  $M_{\text{ix}1}$  where the impedance is low around  $f_{\text{exc}}$  [19].

A synchronous clock  $f_{\text{mix}} (= f_{\text{exc}})$  is used for the amplitude demodulation. Direct down-conversion to zero IF is chosen to reduce the complexity of the demodulator. However, this puts additional constraints on the offset and flicker noise, which are dealt with in this paper by employing chopping at various levels [23]. Downconversion mixers and amplifiers in both TIAs are chopped to mitigate their low frequency errors. Unbalanced inputs to the demodulator causes common-mode signals at  $2 \cdot f_{\text{exc}}$  at the virtual ground of the TIA. Capacitor  $C_s$  provides low impedance and assists the TIA in suppressing this and higher harmonics of  $f_{\text{exc}}$ .

1) *Capacitors for V2I*: As shown in Fig. 6, the demodulation scheme uses passive capacitors ( $C_{\text{in}}$  and  $C_{\text{soc}}$ ) as V2I blocks. The effective transconductance gain can be calculated as the product of transfer gain of the V2I block and the downconversion mixer, and is given by

$$G_{m,\text{eff}} = 2\pi \cdot f_{\text{exc}} \cdot C_{\text{in}} \cdot \frac{2}{\pi} = 4 \cdot f_{\text{exc}} \cdot C_{\text{in}}. \quad (3)$$

With  $f_{\text{exc}} = 126$  MHz and  $C_{\text{in}} = 500$  fF,  $G_{m,\text{eff}}$  evaluates to  $252 \mu\text{A/V}$ . This is equivalent to a switched-capacitor resistor value of  $\approx 4$  k $\Omega$ . As an example, for  $|V_{\text{sen}}| = 400$  mV and  $\beta = 0.1$ , the maximum current ( $\propto \Delta x_{\text{max}}$ ) input to the TIA is calculated as  $10 \mu\text{A}$ . Hence, for  $\Delta x_{\text{max}} = 10 \mu\text{m}$  and  $\Delta x_{\text{min}} = 1$  nm, the minimum current to be detected by the TIA is  $1$  nA.

Employing capacitors as V2I has a number of advantages. First, it improves the linearity and reduces power consumption of the demodulator. Passive resistors [24] can also be employed for this purpose, however, they degrade the quality factor of the LC resonator. Another issue with using resistors is that the common-mode voltage of the excitation oscillator outputs is too low for the TIAs to function properly. Using capacitors as V2I also relaxes the mixer size due to the quadrature relationship between the switch resistance  $R_{\text{sw}}$  and capacitive reactance (Fig. 6). This helps in reducing the parasitic capacitance ( $C_p$  in Fig. 6) and improves noise performance of the demodulator. Capacitors also assist in increasing the excitation frequency as they present higher transconductance gain with frequency whereas more current and area has to be spent to improve the transconductance gain of  $g_m$  cells and resistors, respectively.

2) *Downconversion Mixer*: Current-driven passive mixers are attractive due their low flicker noise, and hence they are often employed for demodulation to zero IF [19], [25], [26]. The proposed sensor offset compensation scheme further reduces the mixer's noise contribution by removing the excess offset current flowing through them [27]. Due to the high frequency operation of downconversion mixers, the charge injection of mixers  $\text{Mix}_1$  and  $\text{Mix}_0$  can cause significant residual offset. This is mitigated by a nested-chopping scheme operating at  $f_{\text{mix}}/4096$  [28], [29].

However, for mixers to work in current mode, the mixer switch resistance  $R_{\text{sw}}$  and TIA input impedance ( $R_{\text{in,tia}}$ ) should be much smaller than the capacitive impedance at  $f_{\text{exc}}$  ( $R_{\text{sw}} + R_{\text{in,tia}} \ll (1/j \cdot 2\pi \cdot f_{\text{exc}} \cdot C_{\text{in}})$ ). A small  $R_{\text{sw}}$  can be achieved by increasing the size of switching transistors.

However, this causes an increase in the parasitic capacitance  $C_p$  which deteriorates the noise performance of the demodulator. Assuming the noise of the amplifier  $A_v$  is the dominant source, the noise figure (NF) of the demodulator can be evaluated as

$$\text{NF} = \frac{C_{\text{in}} + C_p}{C_{\text{in}}}. \quad (4)$$

In this paper, mixers are sized to add the total parasitic capacitance  $C_p$  of only 50 fF which degrades noise of TIA by only 10%. Sensor offset compensation capacitor  $C_{\text{soc}}$  does affect the noise of the sensor channel TIA adversely, hence NF has a minimum value of 3 dB. As per simulations, the effective  $R_{\text{sw}}$  is  $150 \Omega$  (increased three times due to nested chopping) and the capacitive reactance at 126 MHz is  $\approx 2.5$  k $\Omega$ . The quadrature relationship between these two impedances further helps in defining currents predominantly by the capacitance.

3) *Transimpedance Amplifier*: The TIA is one of the most important block in the proposed demodulation scheme. For capacitors to perform efficiently as V2I blocks, the TIA must present a low input impedance  $R_{\text{in,tia}}$  within the bandwidth of the input. The impedance transformation property of current-driven mixers shifts this low impedance to the input of mixers around  $f_{\text{exc}}$  [19]. The amplifier  $A_v$  must possess high gain at low frequency to achieve this. Also, the negative feedback in the TIA should still be present at  $2 \cdot f_{\text{exc}}$  because the input to the TIA consists predominantly of a baseband signal around dc and  $2 \cdot f_{\text{exc}}$  ( $I_{\text{mix}1}$  in Fig. 4). To ensure the proper functioning of capacitors as V2I,  $A_v$  is designed for a  $>110$  dB dc gain and a  $>1$  GHz unity gain frequency  $f_u$  ( $\approx 8 \cdot f_{\text{mix}}$ ).

The schematic of the amplifier  $A_v$  is shown in Fig. 7. The amplifier consists of a two-stage Miller-compensated amplifier. The first stage is a telescopic amplifier and the second stage  $A_{v2}$  is a common-source (CS) stage. To achieve high dc gain without deteriorating high-frequency behavior of the amplifier, gain boosting is employed in the telescopic amplifier [30]–[32]. Auxiliary (gain boosting) amplifiers, for both pMOS and nMOS cascodes, are also realized as two-stage amplifiers to achieve sufficient gain. Auxiliary amplifiers consist of a folded cascode amplifier followed by a CS amplifier, and they are Miller-compensated. Although the use of a differential signal path reduces charge injection errors, unbalanced inputs to the TIA lead to common-mode signals at  $2 \cdot f_{\text{mix}}$ . These are suppressed by common-mode feedback (CMFB) circuitry and capacitor  $C_s$ . Two auxiliary amplifiers and the main amplifier have their own CMFB control.

The noise of the amplifier  $A_v$  is the dominant source in the demodulator. The input-referred noise  $v_{n,\text{in}}$  required from the amplifier in the TIA can be evaluated as

$$v_{n,\text{in}} \leq \Delta v_{\text{sen,min}} \cdot \text{NF} \quad (5)$$

where  $\Delta v_{\text{sen,min}}$  is the minimum detectable change in the amplitude of  $V_{\text{sen}}$ . For  $|V_{\text{sen}}| = 400$  mV,  $x_{\text{so}} = 100 \mu\text{m}$  and  $\Delta x_{\text{min}} = 1$  nm,  $\Delta v_{\text{sen,min}}$  is  $4 \mu\text{V}$ . Hence,  $v_{n,\text{in}} < 2 \mu\text{V}$  is required from the amplifier in the TIA, for  $\text{NF} = 3$  dB. For a bandwidth of 2 kHz, the amplifier  $A_v$  should possess an effective input thermal noise floor smaller than  $45 \text{ nV}/\sqrt{\text{Hz}}$ .

Low-frequency errors of the amplifier are very important for the zero IF demodulation scheme. We have mitigated them

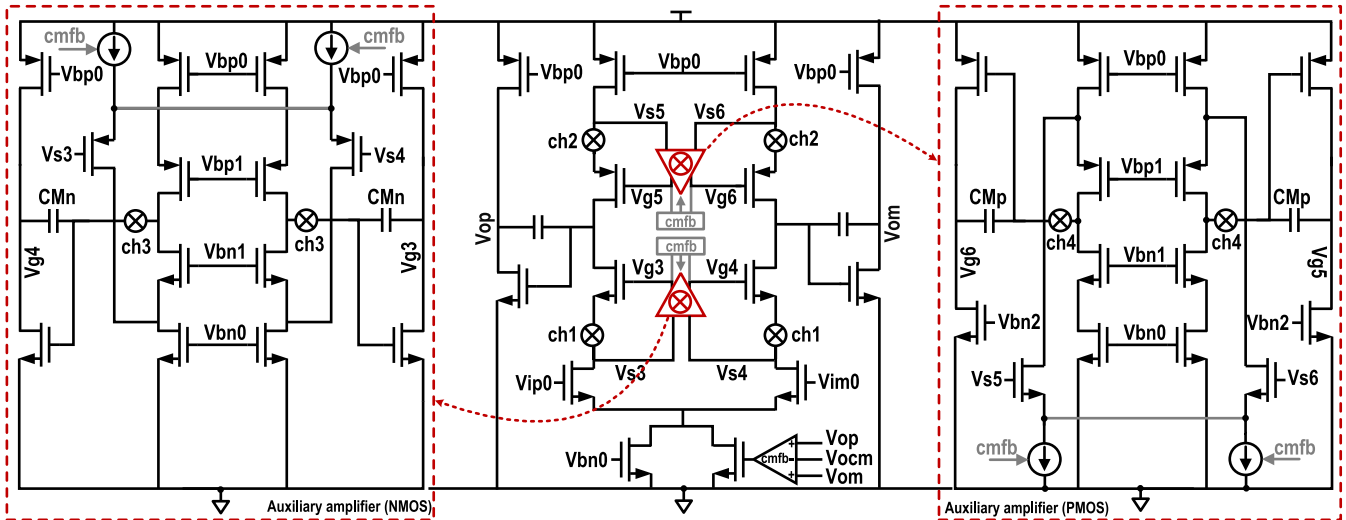


Fig. 7. Schematic of the amplifier  $A_v$  used to realize TIAs. Two-stage chopped auxiliary amplifiers are used for gain boosting.

using the chopping technique in this paper. As shown in Fig. 7, both the amplifier  $A_v$  and auxiliary amplifiers are chopped to mitigate their offset and low frequency noise. Choppers are always placed at low impedance nodes to reduce their residual offsets, caused by self-mixing. PSS simulations show that the amplifier achieves a 113-dB dc gain over a 3-dB bandwidth  $\approx 4$  kHz. The amplifier is not unity gain compensated.

### C. Quadrature Clock Generation

This section illustrates the implementation of the quadrature clock comparator needed for the generation of the clock which has a  $\pi/2$  rad phase lead compared to the excitation oscillator outputs. The design of a comparator, that is fast enough, is very important because any excess phase difference over  $\pi/2$  rad attenuates the effective input signal to the TIA and degrades the input SNR to the TIA. This attenuation is proportional to  $\cos(\Delta\phi)$ , where  $\Delta\phi$  is the phase difference between the ideal quadrature clock phase and the phase of the comparator output. As an example, at  $f_{exc} = 200$  MHz, a 500-ps excess delay translates to  $\Delta\phi = 36^\circ$  causing the signal to be attenuated by 20%. For this reason, a fast continuous-time comparator has been designed.

Fig. 8 shows the schematic of the continuous-time comparator designed for the interface. The comparator consists of three stages: 1) quadrature generation stage; 2) self-biased preamplifier; and 3) regenerative latch. The quadrature generation stage provides a low input impedance so that capacitors  $C_{comp}$  ( $= 200$  fF) generate currents proportional to input voltages  $V_{sen}$  and  $V_{ref}$ . These currents, in quadrature to the excitation oscillator voltages, are converted into voltages  $V_{op1}$  and  $V_{om1}$  at low impedance nodes, resulting in smaller delay at the expense of a low voltage gain. The self-biased preamplifier stage [33] amplifies the output of the first stage and also sets the optimum amplifier-to-latch transfer point which desensitizes the effect of process-voltage-temperature variations on the comparator delay [34]. This is important for minimizing the delay due

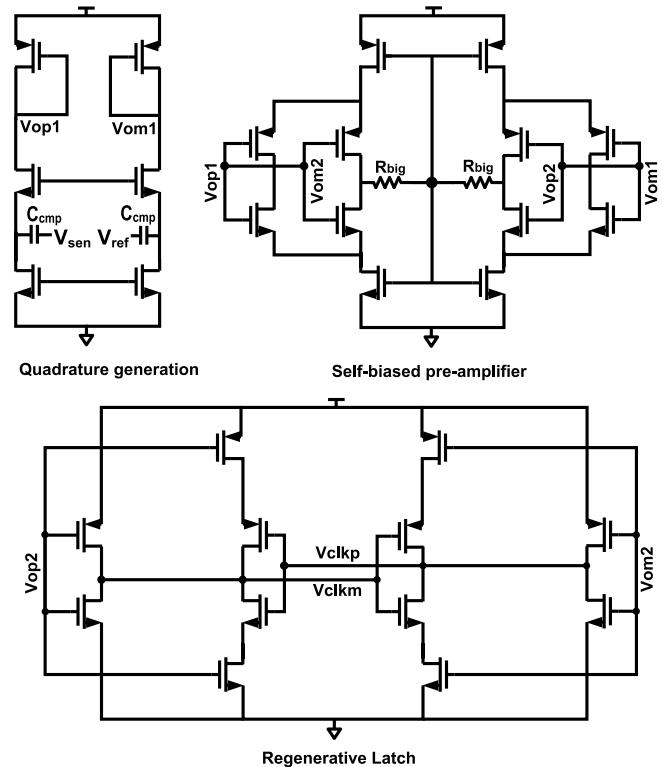


Fig. 8. Continuous-time comparator to generate clock in quadrature to excitation oscillator's outputs.

to the regeneration stage. At  $f_{exc} = 126$  MHz,  $\approx 0.4$ -ns delay of the comparator causes  $\approx 5\%$  signal attenuation.

## IV. EXPERIMENTAL RESULTS

To evaluate the usefulness of proposed concepts in precision displacement sensing, a microchip was fabricated. As shown in Fig. 9, the ECS interface occupies only  $1.18$  mm<sup>2</sup> in TSMC  $0.18$ - $\mu$ m CMOS technology. The interface consumes  $11$  mA from a  $1.8$ -V supply, which includes  $2.2$  mA consumed by two on-chip unity gain output buffers used to drive an



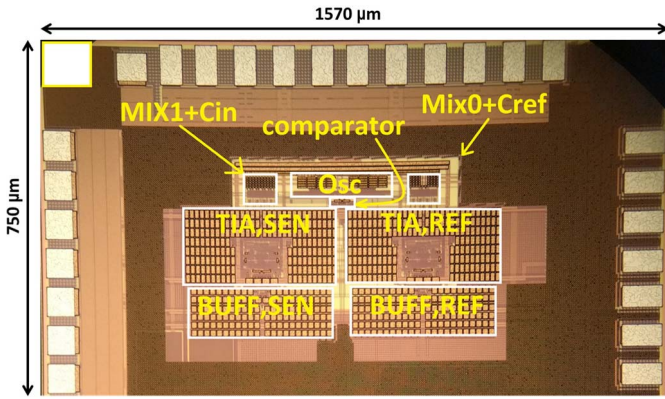


Fig. 9. Chip micrograph of the ECS interface.

off-chip analog-to-digital converter (ADC). The analog output of the chip is digitized using an evaluation board of the 24-bit  $\Sigma\Delta$  ADC AD7779.

### A. PCB-Based ECS

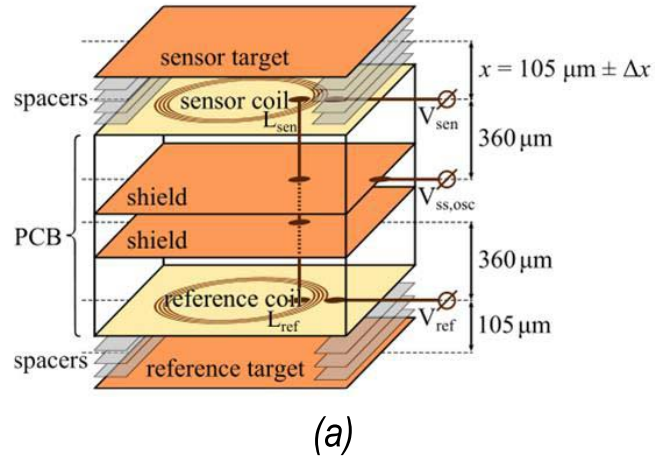
A printed circuit board (PCB)-based sensor was used to evaluate the performance of the microchip and its usefulness in sensing displacement. Fig. 10(a) depicts the structure of the PCB-based sensor. It consists of two flat coils, of identical geometry, implemented in the top and bottom layer of a custom four-layer PCB. Each coil consists of four,  $35\ \mu\text{m}$  thick, turns with a  $200\ \mu\text{m}$  pitch and an 8 mm outer diameter. The coils are shielded from each other by two ground planes to prevent mutual inductance. Their distance to the moving sensor target and fixed reference target is established by a number of precision  $10\text{-}\mu\text{m}$  stainless-steel spacers. Spacers are positioned 3 mm away from the coils so as not to affect the sensitivity of coils.

Fig. 10(b) illustrates the custom PCB designed for the characterization of the ECS interface. Nylon nuts and bolts are used to position sensor and reference targets with the help of stainless steel spacers with high stiffness. With a copper target and nominal standoff  $x_{so} = 105\ \mu\text{m}$  (14 spacers minus coil thickness of  $35\ \mu\text{m}$ ), the effective standoff inductance  $L_{so}$  is found to be  $\approx 100\ \text{nH}$  (including  $L_{ic} \approx 20\ \text{nH}$ ) using an impedance analyzer. The coil used for the characterization of the ECS interface is similar to coil 3, annotated in Fig. 10(b) (but underneath the copper target in the figure).

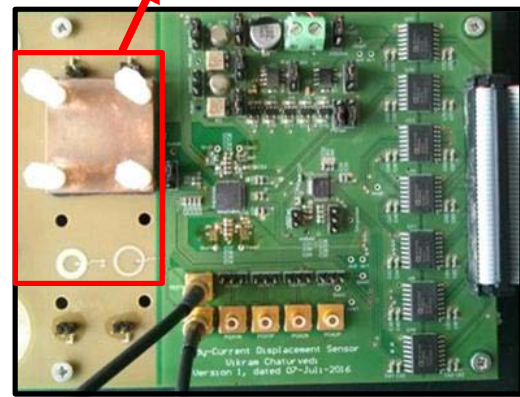
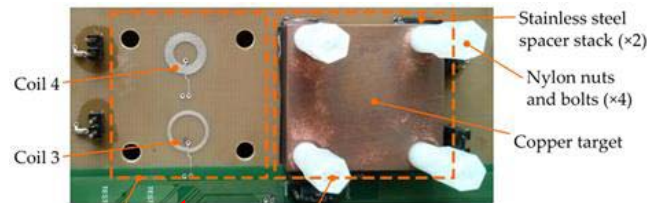
### B. Transfer Characteristic

Fig. 11 shows the measured transfer characteristic obtained using the ECS interface and the PCB-based sensor, where  $D_{out} = V_{o,sen}/V_{o,ref} = \Delta x/\Delta x_{max} \approx \Delta L(x)/(\beta * L_{ref})$  (1). To introduce a displacement  $\Delta x$ , the number of spacers between the sensor target and the sensor coil is varied. This introduces an effective displacement of  $10\ \mu\text{m}$  with each spacer, around nominal standoff  $x_{so} = 105\ \mu\text{m}$  (Figs. 5 and 10). The distance between the reference coil and reference target is kept fixed at  $105\ \mu\text{m}$ .

The transfer characteristics is evaluated at three different values of  $I_{ss,osc}$ , 1.2 mA, 1.52 mA, and 1.84 mA, to evaluate the effectiveness of the ratiometric principle in suppressing the



(a)



(b)

Fig. 10. PCB-based ECS. (a) Conceptual sensor structure. (b) Customized PCB with copper target.

effect of the oscillator bias current ( $I_{ss,osc}$ ). Output  $V_{o,sen}$  has a higher slope at higher  $I_{ss,osc}$  due to higher gain. Three  $V_{o,sen}$  curves meet at a single point where  $L_{sen} = L_{ref}$ . The variation in  $V_{o,ref}$  is dominated by the improvement in the quality factor of the sensor coil, in spite of a decrease in  $f_{exc}$ , with higher  $x_{so}$ . The ratiometric readout  $D_{out}$  (1) is well matched for three values of  $I_{ss,osc}$ .

The presence of the null in  $D_{out}$  around the nominal  $x_{so} = 105\ \mu\text{m}$  shows that the sensor offset is indeed compensated by the ECS interface. The deviation of zero crossing from  $x_{so} = 105\ \mu\text{m}$  is due to a small mismatch ( $\approx 3\ \text{nH}$ ) between  $L_{sen}$  and  $L_{ref}$ , caused primarily by the difference in parasitic inductances. The maximum displacement range  $\Delta x_{max}$  was found to be  $10\ \mu\text{m}$  which matches closely to  $\beta * x_{so}$ , with  $\beta = 0.1$  and  $x_{so} = 105\ \mu\text{m}$ . The limited output voltage range of the TIA manifests itself as the saturation of  $D_{out}$  at larger displacements. The ratiometric readout  $D_{out}$  for higher ( $I_{ss,osc}$ ) saturates at smaller standoffs due to larger signal swing from the excitation oscillator.

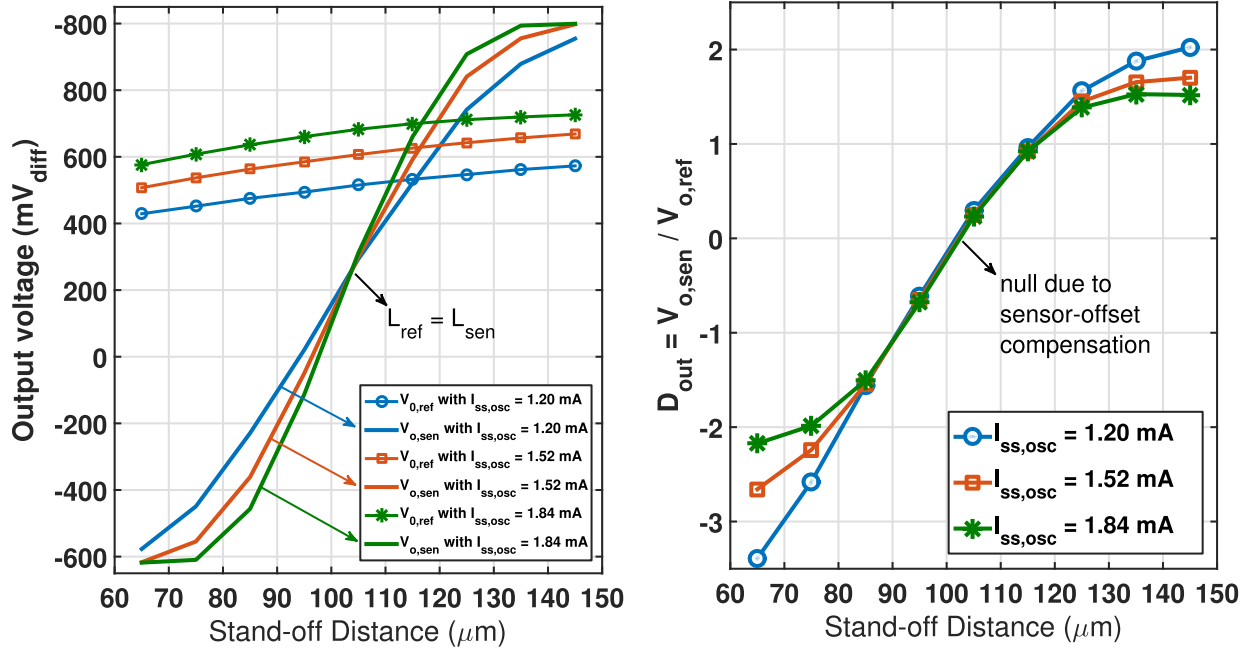


Fig. 11. Measured transfer characteristics at different values of  $I_{ss,osc}$ . The effective displacement is standoff distance minus nominal standoff ( $=105 \mu\text{m}$ ).

Due to pseudo-differential structure of the sensor, the excitation frequency  $f_{exc}$  also undergoes change with the displacement. An excitation frequency deviation of  $\Delta f_{exc} = 1.3 \text{ MHz}$ , around  $f_{exc} = 126 \text{ MHz}$ , for a displacement change of  $\Delta x_{max} = 10 \mu\text{m}$  was noticed. This confirms that the frequency output in the ECS is indeed less sensitive when compared to the output amplitude ( $\Delta f_{exc}/f_{exc} \approx 1\%$ ).

### C. Noise Characterization

To evaluate the noise performance of the ECS interface, fast Fourier transform (FFT) was used. Measured  $2^{17}$  point FFTs of two output voltages and  $D_{out}$  are shown in Fig. 12, with  $\Delta x_{max} = 10 \mu\text{m}$  (standoff distance  $=115 \mu\text{m}$ ). The noise spectrum in two output voltages was found to be dominated by the oscillator amplitude noise. The suppression of the low-frequency amplitude noise of the excitation oscillator, due to ratiometric readout  $D_{out}$ , is clearly noticeable. The spectral shaping is primarily due to the decimation filter of the off-chip  $\Sigma\Delta$  ADC. Low frequency noise was found to be dominated by residual flicker noise from the tail current source of the excitation oscillator. In a 2-kHz noise bandwidth, the calculated SNR was found to be 86.6 dB, which corresponds to a resolution of 14.1 bits over  $\Delta x_{max}$  ( $=10 \mu\text{m}$ ). This corresponds to a displacement resolution of 0.6 nm and effective noise floor of  $13.4 \text{ pm}/\sqrt{\text{Hz}}$ . For coil inductance  $=100 \text{ nH}$ , this corresponds to an inductance resolution of 0.58 pH. The signal bandwidth could not be directly measured because of practical difficulty of introducing  $\mu\text{m}$  range single-tone vibration into the sensor target. However, the signal bandwidth in the ECS interface is limited by the bandwidth of the amplifier  $A_b$  used in the TIA. Parasitic extracted PSS simulations reveal that the amplifier  $A_b$  has 3-dB bandwidth of  $\approx 4 \text{ kHz}$ .

Another experiment was also performed to evaluate the suppression efficacy of the ratiometric measurement. A 2-kHz

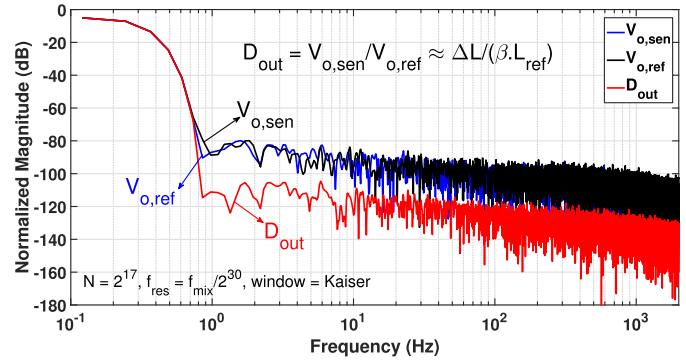


Fig. 12. Measured spectrum of the output voltages and  $D_{out}$ .

frequency component was introduced in the oscillator bias current by switching ON and OFF  $80 \mu\text{A}$  in parallel to a fixed current of  $1.12 \text{ mA}$ . The output spectrum of output voltages and  $D_{out}$  is shown in Fig. 13. It can be noted that the signal tone at  $2 \text{ kHz}$  is suppressed by  $34 \text{ dB}$  by the ratiometric measurement. This shows that the ratiometric measurement suppresses variation in  $I_{ss,osc}$ , which is instrumental in improving the noise and drift performance of the ECS interface. A  $34\text{-dB}$  suppression also endorses the high linearity of the demodulators, when compared to [4] which achieved  $\approx 24 \text{ dB}$  of ratiometric suppression. A higher noise floor is also noticed in the spectrum which we feel is due to the inter-modulation effect of the switched-current source.

Using a low noise amplifier (SR560) and a low-noise spectrum analyzer (HP4395A), the output noise floor of the reference channel is found to be  $136 \text{ nV}/\sqrt{\text{Hz}}$ . This would translate into a  $92.9\text{-dB}$  SNR, if all of oscillator's amplitude noise was suppressed by the ratiometric readout. In practice, it is slightly less because of the finite input impedance of the demodulator which allows some of oscillator's amplitude

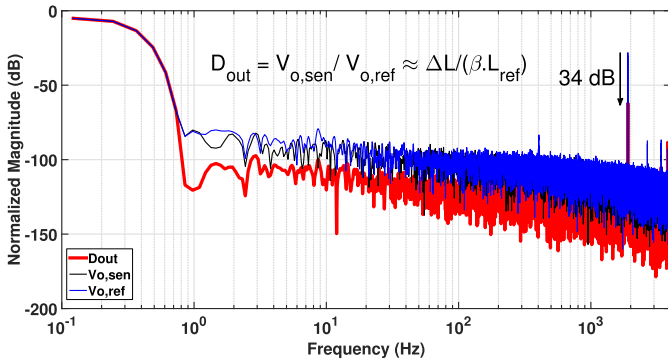


Fig. 13. Suppression of a 2-kHz tone in oscillator bias current by ratiometric measurement.

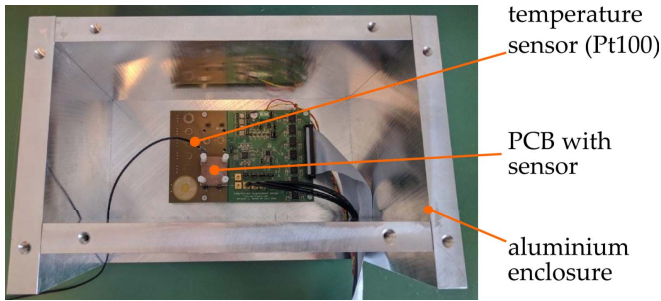


Fig. 14. Test setup for stability characterization of the ECS interface.

noise to escape from the  $LC$  tank. This illustrates an important tradeoff when using capacitors as V2I instead of a  $g_m$  block.

#### D. Stability Characterization

Stability tests were performed on the ECS interface using the climate chamber ACS DY110C. The temperature and humidity were controlled using the software LabVIEW. As shown in Fig. 14, the PCB-based ECS and the readout interface were placed inside a thick Aluminum box to ensure a highly stable environment. During the experiment, a high precision temperature sensor Pt100 was used to find the exact temperature of the PCB-based ECS inside the Aluminum box (Fig. 10). The Aluminum box, with the ECS and interface inside, was placed in the climate chamber, and precautions were taken to isolate mechanical vibrations of the climate chamber from coupling to the sensor.

For thermal stability characterization, humidity was kept fixed at 40% and temperature was stepped from 20 °C to 40 °C, in steps of 5 °C, and then back to understand the hysteresis effects. Fig. 15 shows the drift in  $D_{out}$  caused by the change in temperature. A temperature change of 17.14 °C caused an effective displacement error of 124 nm ( $\Delta x = \Delta D_{out} * \Delta x_{max}$ ), which corresponds to a thermal drift of 7.24 nm/°C. This amount of thermal drift is very low because various applications requiring such a high displacement resolution have very well-controlled environments ( $\Delta T < 10\text{--}20$  mK). The difference in plots for the two directions of the temperature change is attributed to the long-term drift (the experiment took  $\approx 22$  h to complete).

For long-term stability, temperature, and humidity of the climate chamber were kept fixed at 25 °C and 40%,

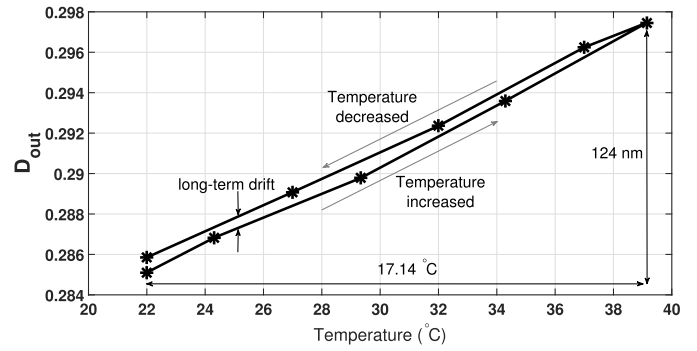


Fig. 15. Thermal stability characterization of the ECS interface.

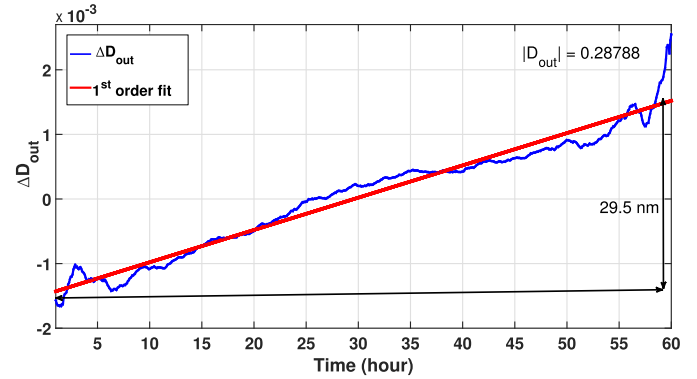


Fig. 16. Long-term stability characterization of the ECS interface.

respectively. The experiment was carried out for 60 h, during which data was captured once every 5 min for a duration of 1 s. Fig. 16 shows the effect of long-term drift on  $D_{out}$ . After the first-order fit, the effective drift was calculated as  $\approx 29.5$  nm over 60 h. The deviation from the linear behavior is attributed to mechanical vibrations of the climate chamber which were not attenuated enough by the vibration isolator. High precision applications such as wafer scanners are frequently calibrated between each run ( $< 2$  min). Hence, a long-term drift of 29.5 nm over 60 h is quite benevolent.

#### E. Performance Comparison

Table I summarizes the performance of the ECS interface and compares it to the state of the art from both academia and industry. The presented ECS interface achieves 0.6 nm in a 2-kHz noise bandwidth. This is at least  $30\times$  better than [9]. Despite a higher  $f_{exc}$ , better resolution and implementation in a sub-micrometer CMOS process, the interface dissipates similar power [4], [7]. Thermal drift is found to at least one order better than the previous state of the art [4]. Furthermore, it achieves an inductance resolution of 0.58 pH which outperforms that of precision LCR meters.

To compare the power efficiency of ECS interfaces, a figure-of-merit ( $FoM_{ECS}$ ) was introduced in [4]

$$FoM_{ECS} = \frac{P}{BW \cdot 2^{ENOB}} \cdot \frac{1}{f_{exc}} \quad (6)$$

where  $P$  is the power consumption,  $BW$  is the measurement bandwidth,  $ENOB$  is the effective resolution of the interface (in bits), and  $f_{exc}$  is the excitation

TABLE I  
PERFORMANCE SUMMARY AND COMPARISON WITH THE STATE OF THE ART

	This work	[4]	[8] <sup>#</sup>	[7]	[9] <sup>#</sup>	[35]
Standoff Distance	105 $\mu\text{m}$	3 mm	1 mm	3 mm	2.286 mm	3 mm
Excitation Frequency (MHz)	126	20	-	15	-	0.3125
Resolution (nm)	0.6	65	60	135	21	2930
Noise Bandwidth (kHz)	2	1	0.52	1	1	10
Inductance Resolution (pH)	0.58	1.5	-	3.8	-	-
Thermal Stability (nm/ $^{\circ}\text{C}$ )	8	90	few 1000s	-	few 1000s	-
Power Consumption (mW) <sup>§</sup>	19.8	18	$\approx 750$	18 <sup>*</sup>	$\approx 700$ <sup>*</sup>	7.3
Technology	0.18 $\mu\text{m}$ CMOS	0.35 $\mu\text{m}$ BICMOS	-	0.35 $\mu\text{m}$ BICMOS	-	0.6 $\mu\text{m}$ CMOS
$FoM_{ECS}$ (aJ/Hz)	4.73	19.4	-	55.5	-	220

\* digital output from the ECS interface.

§ power consumption of an output ADC ( $\ll 1$  mW) is negligible compared to that of ECS interfaces.

# high precision ECS products available in the market.

frequency of the ECS interface. The presented ECS interface is at least four times better than the state of the art, when it comes to power efficiency.

## V. CONCLUSION

An ECS interface for sub-nm displacement sensing is presented. The degradation of the stability due to the skin effect is mitigated using a high excitation frequency ( $> 100$  MHz). A small standoff distance, along with integrated read-out circuitry in close proximity to the ECS coils, is employed to improve the sensitivity of the system. Ratiometric measurement is found to be instrumental in suppressing the noise and drift of the excitation oscillator. To assist this ratiometric suppression, a highly linear amplitude demodulation scheme is realized using a passive (capacitive) V2I conversion approach. Sensor offset compensation is found to be helpful in obviating the detrimental effect of sensor offset on the linearity of the interface, particularly demodulator blocks.

At a nominal standoff of 105  $\mu\text{m}$ , the ECS interface achieved a displacement resolution of 0.6 nm (in 2-kHz noise bandwidth) within a 10- $\mu\text{m}$  displacement range. The thermal drift was found to be  $< 7.3$  nm/ $^{\circ}\text{C}$ , in the range of 20  $^{\circ}\text{C}$  to 40  $^{\circ}\text{C}$ . A 60-h long-term stability experiment showed a drift of only 29.5 nm. What is more, all of this is achieved by consuming only 11 mA from a 1.8-V supply. By increasing  $f_{exc}$ , reducing the standoff distance and compensating sensor-offset, this is the only ECS interface to achieve the sub-nm displacement resolution.

## ACKNOWLEDGMENT

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