A 0.18-µm CMOS time-domain capacitive-sensor interface for sub-1mG MEMS accelerometers

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Abstract: A high-resolution capacitive-sensor interface for sub-1mG MEMS accelerometers is presented herein. A time-domain capacitive-sensor interface based on a relaxation oscillator with noise reduction is proposed to achieve a high resolution. A prototype interface is fabricated using a 0.18-µm CMOS process. The prototype is linked with a sub-1mG MEMS accelerometer, and its performance is investigated experimentally. The results confirm that the proposed interface is able to detect sub-1mG acceleration with a signal-to-noise ratio of 90.3 dB (an acceleration noise-floor of 9.0 µG/√Hz with a bandwidth of 12 Hz).

Keywords: time domain, capacitive-sensor interface, relaxation oscillator, sub-1mG, microelectromechanical systems (MEMS), accelerometer

Classification: Micro- or nano-electromechanical systems

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1 Introduction

The development of microelectromechanical systems (MEMS) has allowed the production of miniaturized capacitive accelerometers, which are now used in applications including inertial navigation [1], platform stabilization in space [2], earthquake prediction [3], and early diagnosis of diseases by monitoring of human activity [4]. The small capacitive accelerometer comprises a MEMS accelerometer and a capacitive-sensor interface [5]. To extend the range of applications, the MEMS
accelerometer and interface should achieve high resolution within a small device. In the pursuit of a low-noise MEMS accelerometer, capacitive MEMS accelerometers with gold proof masses were fabricated using multi-layer metal technology [6, 7]. Capacitive-sensor interfaces with low noise and a simple structure are required to counterpart the capacitance values of a MEMS accelerometer.

Several types of capacitive-sensor interfaces are available, including switched-capacitor charge amplifiers [8] and voltage synchronous modulator/demodulators [9]. Although they achieve a high-resolution performance, these interfaces exhibit three main drawbacks: (i) large chip size, (ii) the inability to operate at low voltage, and (iii) the requirement of an analog-to-digital converter (ADC) if digital output is required [10].

To replace voltage domain capacitive-sensor interfaces, research has focused on the development of time/frequency-domain capacitive-sensor interfaces [11, 12, 13]. These interfaces possess two key features: (i) they can output a digital signal directly using a simple counter without ADC, and as the output signals are period modulated, they can be directly connected to the controller and (ii) they offer high-resolution even with a simple structure. However, the resolution of such interfaces is limited by the noise at both the switches and the current sources to charge measured capacitors. Recently, a highly sensitive oscillating accelerometer that uses MEMS resonance with PLL circuits has been reported [13]. However, its interface requires the addition of complex control circuits.

Therefore, time-domain capacitive-sensor interfaces for a sub-1mG MEMS accelerometer with a simple structure were investigated. Time-domain interfaces can achieve a better signal-to-noise ratio (SNR) than voltage domain interfaces because the voltage noise can be converted to a marginal jitter (time-domain noise).

This paper is structured as follows. In Section 2, our approach to achieving high resolution using a time-domain capacitive-sensor interface is introduced. Next, the proposed noise-reduction techniques for the time-domain capacitive-sensor interface are discussed. Finally, the results of experiments that were conducted to test the validity of our approach are reported.

2 Approach to achieve high resolution

MEMS accelerometers detect changes in acceleration as variations in capacitance. To process the sensing data, a capacitance-to-digital (CTD) interface is required. Fig. 1 shows two types of CTD interfaces. The voltage domain interface shown in Fig. 1(a) converts capacitance variations into voltage variations via an amplifier. These variations are then converted to a digital signal via the ADC. ΔΣ-ADC is often used to output the digital signal [14, 15, 16, 17]. When ΔΣ-ADC exhibits sufficiently high resolution, the SNR that limits the resolution of the interface is given by Eq. (1).

$$\text{SNR} = \frac{V_s}{v_n},$$  

where $V_s$ and $v_n$ are the signal and noise of the voltage. $v_n$ is then given by
where \( S_v(f) \) and \( f_w \) are the voltage-noise spectral density and signal bandwidth, respectively. To maximize the SNR, \( V_s \) should be as large as possible and \( v_n \) as small as possible. However, the maximum achievable value of \( V_s \) is determined by the supply voltage, and supply voltages have been steadily lowered to reduce power consumption of digital circuits [18]. In practice, therefore, the only way of achieving a high SNR is to reduce the value of \( v_n \).

In contrast, the time-domain interface shown in Fig. 1(b) has the potential for high-resolution sensing of capacitance variation. As the figure shows, the circuit converts capacitance variations into variations in the oscillation period, which are then converted to digital signals by the counter. If rise and fall times are sufficiently short, the jitter is minimal. When applied to a time-domain interface, the SNR is then given by Eq. (3).

\[
\text{SNR} = \frac{T}{\sigma},
\]

where \( T \) and \( \sigma \) are the signal period and jitter of a square voltage, respectively, and \( \sigma \) is the root-mean-square value defined as follows:

\[
\sigma = \sqrt[2]{\int_0^{f_w} S_{\sigma(f)}^2 df}.
\]

Here, \( S_{\sigma(f)} \) is the jitter spectral density. A large signal period value minimizes the jitter and yields high SNR.

The SNR can be further improved by measuring the multi-signal periods instead of enlarging the single-signal period. This produces the averaging effect shown in Fig. 2. The total jitter \( \sigma_t \) of the total period \( T_t \) is given as follows:

\[
\sigma_t = \sqrt{\sigma_1^2 + \sigma_2^2 + \cdots + \sigma_n^2},
\]

where \( \sigma_n \) is jitter in the n-th period. In this case, the averaged SNR (SNRavern) is derived as

![Fig. 1. Block diagram and sensing signal of (a) voltage-domain capacitive-sensor interface and (b) time-domain capacitive-sensor interface.](image-url)
\[ \text{SNR}_{\text{avr}} = \frac{nT}{\sqrt{n\sigma}} = \sqrt{n}\text{SNR}. \quad (6) \]

SNR$_{\text{avr}}$ can be improved by $\sqrt{n}$. By measuring repeated signals over an extended period, the rate of $\sigma_i$ in $T_i$ can be reduced and the SNR improved.

When the averaging technique is applied to the voltage domain interface, an extra memory circuit is required. An additional advantage of the proposed approach is that averaging can be easily performed by applying a counter circuit to the multi-signal period.

3 Time-domain capacitive-sensor interface

3.1 Proposed time-domain capacitive-sensor interface

A conventional time/frequency-domain capacitive-sensor interface based on a relaxation oscillator is shown in Fig. 3 [19, 20]. The interface outputs square waveforms, and the oscillation period is modulated by the change in capacitance $C_M$. The drawback of this interface is that noise from the switches (M$_H$ and M$_L$) and electromagnetic noise from the interconnection limit the sensing resolution. A reference capacitor $C_{\text{ref}}$ is also required, and it occupies a large area on the chip.

A second type of conventional interface uses current sources to charge the sensing capacitor [12]. However, the resolution is limited by noise from the current source.

A frequency-domain capacitive-sensor interface based on a phase-locked loop (PLL) has been recently reported [13]. High-resolution acceleration sensing was achieved by combining MEMS resonance with the PLL. However, the circuit was complex and required two sets of oscillators, the PLL itself, and a subtractor.
Fig. 4 shows the proposed time-domain capacitive-sensor interface based on a relaxation oscillator [21]. The time-domain interface detects variations in capacitance as period variations in the oscillation signals. The interface employs an operational amplifier (OPA) between switches \((M_H, M_L)\) and comparators (CMPs). The OPA is connected to a capacitive MEMS accelerometer \(C_M\) and functions as an integrator, as well as charges or discharges \(C_M\) using constant current sources comprising the resistors \(R\) and switches \((M_H, M_L)\). As shown in Fig. 5, triangle waveforms are produced at the OPA output by controlling the switches. The CMPs switch states of charging or discharging \(C_M\) by the triangle waveform using reference voltages of \(V_{\text{high}}\) and \(V_{\text{low}}\); these states are fed back to the switches. Thus, an oscillation is produced, and the period of the squared-output waveform corresponds to the value of \(C_M\).

The interface has the following features: (i) The feedback-loop of the OPA is tolerant of electromagnetic noise at the interconnection between the oscillator and the MEMS accelerometer \(C_M\). (ii) The effect of noise from \(M_H, M_L\), and the OPA is reduced. (iii) The reference capacitors often used in capacitive-sensor interfaces [11, 12, 13, 14, 15, 16] are no longer required. (iv) A single electrode of the MEMS accelerometer can be used. (v) Supply-voltage noise can be canceled. (vii) Time averaging can be performed by merely using a simple counter circuit.
3.2 Cancellation of supply-voltage noise

The proposed interface can cancel supply-voltage noise. When \( M_L \) is in the ON state and \( M_H \) is OFF, the voltage of the OPA rises, as shown in Fig. 5. The constant current \( I_{LC} \), given by Eq. (7), then flows through a resistor \( R \).

\[
I_{LC} = \frac{V_{com} - V_{low}}{R},
\]

(7)

where \( V_{com} \) is the reference voltage and is defined as follows:

\[
V_{com} = \frac{V_{high} + V_{low}}{2}.
\]

(8)

The input voltage of the OPA is set by \( V_{com} \) as the output voltage of the OPA \( V_{OPA} \) rises. When \( V_{OPA} \) reaches \( V_{high} \), the CMPs switch the states of \( M_H \) and \( M_L \). The rising time \( T_{rise} \) is given by Eq. (9).

\[
T_{rise} = \frac{C_M I_{LC}}{V_{high} - V_{low}} = 2RC_M.
\]

(9)

The falling time \( T_{fall} \) is equal to \( T_{rise} \). The output period \( T_{out} \) is divided by a period \( T_{OPA} = T_{rise} + T_{fall} \) and derived as follows:

\[
T_{out} = 4T_{rise} = 8RC_M.
\]

(10)

\( T_{out} \) is determined by the time constant \( RC_M \); thus, \( T_{rise} \) does not depend on either \( V_{high} \), \( V_{low} \), or \( V_{com} \). This allows the supply-voltage variation and noise at \( V_{high} \), \( V_{low} \), and \( V_{com} \) to be canceled. The interface is therefore robust against supply-voltage variation and noise.

3.3 Cancellation of OPA noise

The proposed interface can also reduce the noise of the OPA. In low-frequency sensing operation, \( 1/f \) noise (flicker noise) is dominant; this flicker noise limits the resolution. When the equivalent input noise is \( v_{noise} \), as shown in Fig. 6(a), \( V_{com} \) becomes \( V_{com}' \) and is given by

![Diagram](image-url)

Fig. 6. Cancellation of OPA noise. (a) OPA model, (b) voltages in OPA, and (c) output wave form of the proposed frequency-domain capacitive-sensor interface.
\[ V_{\text{com}} = \frac{V_{\text{high}} + V_{\text{low}}}{2} + v_{\text{noise}}, \]  

(11)

as shown in Fig. 6(b). \( v_{\text{noise}} \) changes the rising time \( T_{\text{rise}} \) of \( C_M \), which is expressed by Eq. (12).

\[ T_{\text{rise}} = \frac{2RC_M}{V_{\text{high}} - V_{\text{low}} + 2v_{\text{noise}}} \times (V_{\text{high}} - V_{\text{low}}). \]

(12)

In the same way, the falling time \( T_{\text{fall}} \) is expressed by

\[ T_{\text{fall}} = \frac{2RC_M}{V_{\text{high}} - V_{\text{low}} - 2v_{\text{noise}}} \times (V_{\text{high}} - V_{\text{low}}). \]

(13)

The one signal period \( T_{\text{OPA}} \), as shown in Fig. 6(c), is given by Eq. (14).

\[ T_{\text{OPA}} = T_{\text{rise}} + T_{\text{fall}} = \frac{4RC_M(V_{\text{high}} - V_{\text{low}})^2}{(V_{\text{high}} - V_{\text{low}})^2 - 4v_{\text{noise}}^2} \approx 4RC_M. \]

(14)

When the value of \( V_{\text{high}} - V_{\text{low}} \) is much larger than \( v_{\text{noise}} \), \( T_{\text{OPA}} \) is mainly determined by the time constant of \( RC_M \) and is not affected by \( v_{\text{noise}} \). The effect of OPA noise in the proposed interface is therefore reduced. It should be noted that the value of \( v_{\text{noise}} \) is assumed to be similar in the rising and falling operations. In a real interface, noise error may appear.

### 4 Design and fabrication

A prototype interface was fabricated using a 0.18-\( \mu \)m CMOS process. Fig. 7 shows a micrograph of the fabricated chip. The core area was 700 \( \mu \)m².

Fig. 8 schematizes the sub-1mG MEMS accelerometer [6, 7]. The proof-mass is shifted by an acceleration input, thereby changing the capacitance between the

![Fig. 7. Chip micrograph of the proposed frequency-domain capacitive-sensor interface.](image)

![Fig. 8. Cross section of the sub-1mG MEMS accelerometer.](image)
proof-mass and the fixed electrode. The MEMS accelerometer was fabricated using a post-CMOS gold electroplating process, as shown in Fig. 9. Using a high density of gold for the proof-mass can reduce the generated mechanical noise, so-called Brownian noise $B_n$.

To evaluate the sensitivity of the MEMS accelerometer, the capacitance change was measured as a function of the input acceleration. The MEMS accelerometer was placed on a vibration exciter (WaveMaker05, Asahi Seisakusho), and the change in capacitance was measured using a semiconductor device analyzer (B1500A, Agilent Tech., Inc.). The MEMS accelerometer showed a sensitivity of 3.91 pF/G and an initial capacitance of 6.45 pF. Next, the capacitance-frequency characteristics were investigated using an LCR meter (IM3533-01, HIOKI E.E. Corp.), and the $B_n$ was evaluated from an analytical model [6]. The values of resonant frequency $f_{res}$ and the quality factor $Q$ were 324 Hz and 2.15, respectively. The value of $B_n$ was 0.42 µG/√Hz.

The fabricated chip and the MEMS accelerometer were next mounted on a test board, as shown in Fig. 10, for evaluating the MEMS accelerometer performance.

5 Experimental results

5.1 Evaluation of the proposed capacitive-sensor interface

The output period as a function of test capacitances $C_c$ (ceramic capacitors) was measured using an oscilloscope (DSOS404A, Keysight Tech., Inc.), as shown in Fig. 11. The periods were demonstrated to be proportional to the capacitance value, and the sensitivity was 0.555 µs/pF. The interface consumed 122 mW at a supply voltage of 3.3 V.
5.2 Detection of sub-1mG acceleration

The sensitivity to acceleration was investigated by applying vertical acceleration to the test board using the vibration exciter. The experimental setup is shown in Fig. 12(a). Delay detection was used to derive the input acceleration from the output signal of the test board.

Fig. 12(b) shows the measured period variation as a function of the input sinusoidal acceleration. Acceleration with an amplitude of 10 mG and frequency of 10 Hz was applied using the vibration exciter. The period was set as 1,000 output waveforms. The period variation yielded the variation in input sinusoidal acceleration.

Fig. 12(c) shows the measured period change $\Delta T_{out}$ as a function of the input acceleration, which was increased from $-30$ to 30 mG at 10 Hz. A linear response was obtained across this acceleration range. The sensitivity was 1.37 ms/G.

5.3 Averaging effect against noise

The output period was averaged using an oscilloscope. Fig. 13 shows SNR$_{avr}$ as a function of the averaging number $n_{avr}$. The proposed interface was connected to an internal reference capacitance $C_{ref}$ 4 pF prepared in the interface for self-testing. Measured SNR$_{avr}$ was 82.5 dB without averaging, and 110 dB (capacitive noise-floor of 2.4 aF/$\sqrt{\text{Hz}}$) with $10^4$ times averaging (sampling time of 21 ms). When the proposed interface was connected to the sub-1mG MEMS accelerometer $C_M$, the measured SNR$_{avr}$ was 63.6 dB without averaging and 90.3 dB (acceleration noise-floor of $9.0 \mu\text{G}/\sqrt{\text{Hz}}$) with $10^4$ times averaging (sampling time of 43 ms). These results confirmed that SNR$_{avr}$ improved by averaging.

However, the measured slope of SNR$_{avr}$ was approximately 6 dB/dec, as shown in Fig. 13; the measured value was below the theoretical value of 10 dB/dec. The decrease in the slope was attributed to the correlation jitter between the periodic signals. If random noise is dominant, SNR$_{avr}$ approaches the theoretical slope value.
The measured acceleration noise-floor was 9.0 µG/√Hz with a bandwidth of 12 Hz. The experimental results confirmed that the proposed CMOS interface with the MEMS accelerometer was able to detect sub-1mG acceleration.

Fig. 12. Vibration test results. (a) experimental setup, (b) measured period as a function of acceleration with amplitude 10 mG, and (c) measured period change as a function of acceleration with amplitudes from -30 mG to 30 mG at a frequency of 10 Hz.
6 Conclusions

A time-domain capacitive-sensor interface based on a relaxation oscillator with noise reduction is proposed herein, and its ability to detect sub-1mG MEMS acceleration has been investigated. A prototype interface was fabricated using a 0.18-μm CMOS process. The interface was then linked to the sub-1mG MEMS accelerometer and evaluated experimentally. The proposed interface was demonstrated to achieve enhanced resolution with an SNR of 90.3 dB. The acceleration noise-floor was 9.0 μG/√Hz with a bandwidth of 12 Hz.

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