A High-Accuracy High-Speed Signal Processing Circuit of Differential-Capacitance Transducers

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***Abstract*—For high-accuracy signal processing of differential-capacitance transducers, an interface circuitry is developed based on a relaxation oscillator. The interface consists of an integrator, a differentiator, and a comparator, and it uses two capacitors of the transducer—one for the integration and the other for the differentiation. This configuration allows the ratiometric operation in the amplitude domain and provides a square wave whose amplitude is proportional to the ratio of the capacitance difference between the two transducer capacitors to their sum. A circuit analysis shows that the interface can detect the capacitance change as small as 0.1% of the total capacitance in 10 μs. Experimental results are also given to confirm the analysis.**

***Index Terms*—Analog circuit, capacitive transducer, intelligent transducer, pressure measurement, relaxation oscillator, signal processing.**

I. Introduction

DIFFERENTIAL capacitance transducers consisting of ganged two capacitors are widely used for detecting pressure difference, linear displacement, acceleration, and rotational angle [1]–[3]. To extract such measurands from the complementary capacitance changes of the two capacitors, the ratiometric operation, which divides the capacitance difference between the two capacitors by their sum, is required for the interface circuit. Several techniques have so far been proposed for such a ratiometric signal processing, including switched-capacitor (SC) analog-to-digital (A/D) [4], capacitance-to-frequency [5], capacitance-to-phase [6], and capacitance-to-voltage conversion [7], [8]. Of these, the relaxation oscillation and SC techniques are best suited for the interface because they allow a high-accuracy measurement with the simple configurations, but take a long time to perform the ratiometric operation [9], [10].

Applications of differential capacitance transducers to accelerometers require the ratiometric operation performed in a few milliseconds. Such a speed requirement can be easily met by the SC sample and hold (S/H) circuit [11], but the accuracy in ratiometric operation when the total capacitance of a transducer is a few pico Farad (pF) is limited to 1% by clock feedthrough. Though slower in speed than the SC S/H circuit, the relaxation-oscillator based interface recently proposed also meets the speed requirement and allows much higher accuracy [12]. Using the dual-slope integration technique, the interface converts the capacitance ratio into the duty ratio. Given the capacitance ratio not in the duty ratio but in a voltage form, one can eliminate the post processing of counting and division and thereby expect much higher speed ratiometric signal processing.

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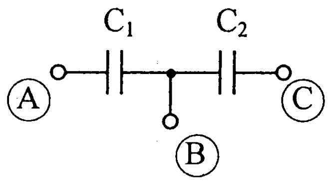

Fig. 1. An equivalent circuit of a differential capacitance transducer.

Based on this idea, an interface circuitry of differential capacitance transducers is developed. In the followings, its circuit configuration, accuracy, and speed estimates, and performances of a prototype interface will be described.

II. Ratiometric Signal Processing

A differential capacitance transducer can be represented by two capacitors with a common electrode, as shown in Fig. 1 [8].

In a linear displacement encoder, areas of the capacitor electrodes change linearly with the displacement x. The capacitances C_1 and C_2 can then be expressed as follows:

$$C_{1,2} = \frac{C_0}{2}\,(1 \pm x) \tag{1}$$

where C_0 is the total capacitance of the transducer. In a differential pressure transducer, on the other hand, the spacings between the electrodes change linearly with the pressure difference x. Therefore, C_1 and C_2 are expressed as

$$C_{1,2} = \frac{C_0}{2}\,\frac{1}{1 \mp x}. \tag{2}$$

The capacitance change of a differential capacitance accelerometer is also given by (2).

Whether the capacitance change is linear or hyperbolic, the measurand x can be detected independently of the total capacitance C_0 by the following ratiometric operation:

$$x = \frac{C_1 - C_2}{C_1 + C_2}. \tag{3}$$

Besides a linear detection of a measurand x for easy calibration, the ratiometric operation given by (3) has another distinct

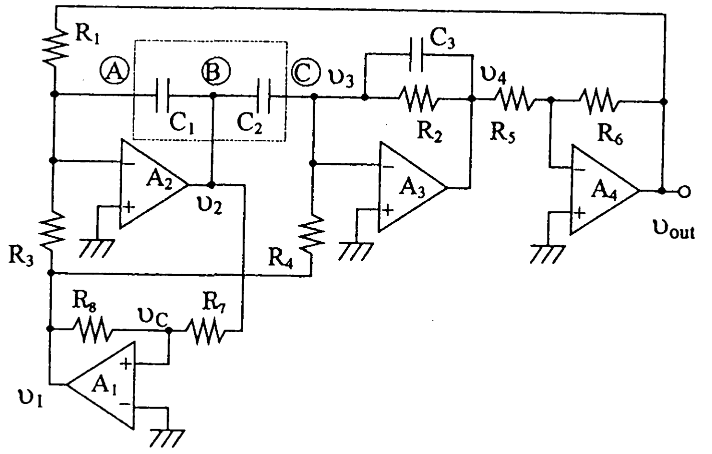

Fig. 2. The circuit diagram of the interface.

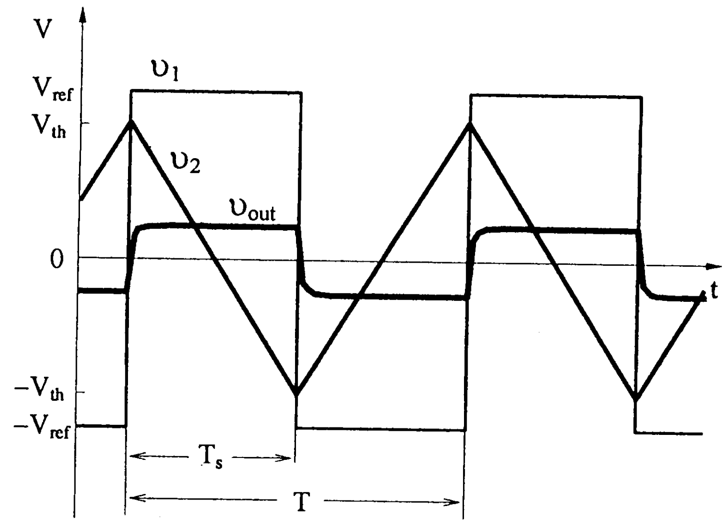

Fig. 3. Voltage waveforms in the interface.

feature that the capacitance changes due to temperature, which would otherwise be a major error source are cancelled. The other error sources are stray capacitances associated with terminals Ⓐ, Ⓑ, and Ⓒ. A processing circuit should be configured such that these stray capacitances have no effect on the ratiometric operation.

III. Circuit Description

An interface circuit of a differential capacitance transducer is shown in Fig. 2. The core is the relaxation oscillator composed of the integrator A_2 and the comparator A_1. The differentiator A_3 and the inverting amplifier A_4 are added to the core for the ratiometric operation. Two capacitors, one C_1 for integration and the other C_2 for differentiation, represent a differential capacitance transducer. Assume that the comparator A_1 provides $+V_{\rm ref}$ when the weighted sum of v_1 and v_2 is negative and $-V_{\rm ref}$ when it is positive

$$v_1(t) = V_{\rm ref}\,\mathrm{sgn}[v_c(t)] \tag{4}$$

$$v_c(t) = \frac{R_7}{R_7+R_8}\,v_1(t) + \frac{R_8}{R_7+R_8}\,v_2(t). \tag{5}$$

The output of the integrator A_2 then takes a triangular waveform and the current flowing into the differentiator takes a square waveform, as depicted in Fig. 3. The response of a differentiator to square wave excitation would be oscillatory. To avoid such an oscillatory response, C_3 is connected in parallel with R_2.

Assuming ideal op-amps and neglecting C_3, one can derive $v_{\rm out}$ as follows:

$$v_c(t) = \frac{C_1\,\dfrac{R_2}{R_4} - C_2\,\dfrac{R_2}{R_3}}{C_1\,\dfrac{R_5}{R_6} + C_2\,\dfrac{R_2}{R_1}}\,V_{\rm ref}\,\mathrm{sgn}[v_c(t)]. \tag{6}$$

If $R_3 = R_4$ and $R_1R_5 = R_2R_6$, then (6) reduces to

$$v_{\rm out} = V_o\,\mathrm{sgn}[v_c(t)] \tag{7}$$

where

$$V_o = \frac{R_1}{R_3}\,\frac{C_1 - C_2}{C_1 + C_2}\,V_{\rm ref} = kxV_{\rm ref}. \tag{8}$$

Equation (8) indicates explicitly that the interface circuit performs the ratiometric operation in the amplitude domain. The time T_s required for the ratiometric operation is a half period of the oscillation

$$T_s = T/2 = (C_1 + C_2)R_3. \tag{9}$$

It is noted that the oscillation frequency is independent of a measurand x and measures the total capacitance C_0 of a transducer.

IV. Performance Estimates

The accuracy and speed of the ratiometric operation achievable with the interface will be estimated in this section.

The factors which affect the operational accuracy are resistance mismatches and nonideal performances of op-amps. Let $\delta_{i,j}$ be the resistance mismatch between R_i and R_j

$$\frac{R_i}{R_j} = 1 + \delta_{i,j}. \tag{10}$$

Substituting (10) into (6), we have to first order

$$V_o = kx(1+\varepsilon_R)V_{\rm ref} + \Delta V_R \tag{11}$$

where

$$\varepsilon_R = \frac{\delta_{3,4} + (\delta_{1,2} + \delta_{5,6})(1+x)}{2} \tag{12}$$

$$\Delta V_R = \frac{\delta_{3,4}kV_{\rm ref}}{2} \tag{13}$$

are the nonlinear and offset errors due to resistance mismatches, respectively.

The finite open-loop gain A and the offset voltage $V_{\rm os}$ of an op-amp are two main error sources that affect the operational accuracy. Their effects are also described in terms of the nonlinear and offset errors as follows:

$$V_o = kx(1+\varepsilon_A+\varepsilon_V)V_{\rm ref} + \Delta V_A + \Delta V_V \tag{14}$$

where ε_A and ΔV_A are nonlinear and offset errors due to the finite gains and ε_V and ΔV are those due to the offset voltages

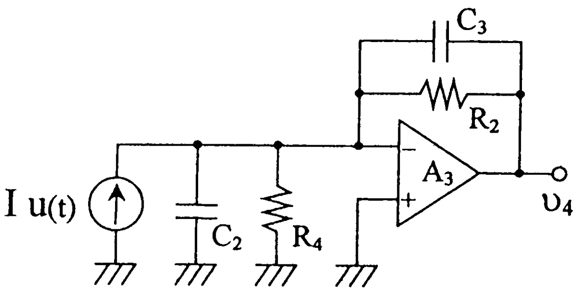

Fig. 4. The phase-compensated differentiator under step excitation.

of op-amps, respectively, given by

$$\varepsilon_A = \frac{2}{A}(1+x) \tag{15}$$

$$\varepsilon_V = \frac{R_3}{R_2}\frac{V_{\rm os}}{V_{\rm ref}} \tag{16}$$

$$\Delta V_A = V_{\rm ref}/A \tag{17}$$

$$\Delta V_V = \left\{1 + R_1\left(\frac{1}{R_2}+\frac{1}{R_3}\right)\right\} V_{\rm os} \tag{18}$$

In deriving the first-order expression (14), $R_2 = R_4, R_5 = R_6, \omega C_2 R_2 \ll 1$, and $\omega C_1(R + 1//R_3) \cong 1$ are assumed.

Assuming typical values $\delta = 0.05\%, A = 100$ dB, $V_{\rm os} = 0.5$ mV, $V_{\rm ref} = 10$ V, $R_1 = R_2 = R_3$, and -0.5×0.5, one can evaluate each error term as follows: $\varepsilon_R = 0.1\%, \varepsilon_A = 3 \times 10^{-5}, \varepsilon_V = 0.5 \times 10^{-4}, \Delta V_R = 2.5$ mV, $\Delta V_A = 0.1$ mV, $\Delta V_V = 1.5$ mV. The offset error can be easily nullified by an offset adjustment and thus the ultimate accuracy is determined by the nonlinear error. From the above evaluation, it is concluded that resistance matching is crucial for a high resolution.

The operational speed is limited by the differentiator because the phase compensation using C_3 is accompanied by the time delay. As described previously, a step current is applied to the inverting input terminal of op-amp A_3. The equivalent presentation of the differentiator is shown in Fig. 4. The response $V_4(s)$ to the step excitation is then given by

$$V_4(s) = \frac{-R_2 I}{s} \cdot \frac{\omega_3}{s+\omega_3} \cdot \frac{\omega_2\omega_u}{s^2 + \omega_2 s + \omega_2\omega_u} \tag{19}$$

where ω_u is the unity-gain bandwidth of A_3 and

$$\omega_2 = \frac{1}{C_4 R_4} \tag{20}$$

$$\omega_3 = \frac{1}{C_3 R_2}. \tag{21}$$

For the phase compensation to be effective, placing ω_3 at one tenth of ω_u is preferable. Let $\omega_3 = \omega_2/2 = 0.1\omega_u$. Then, the step response $v_4(t)$ becomes

$$\begin{aligned} v_4(t) = &-R_2 I\{1 - (1.0 - 0.05\ \cos 0.44\omega_u t \\ &+ 0.23\ \sin 0.44\omega_u t)e^{-\omega_3 t}u(t). \end{aligned} \tag{22}$$

The amplitudes of oscillatory terms are small owing to the phase compensation and the unit response (22) can be approximated by a simple exponential function. The 0.1% settling time τ_s is then given by

$$\tau_s = 7\omega_3^{-1}. \tag{23}$$

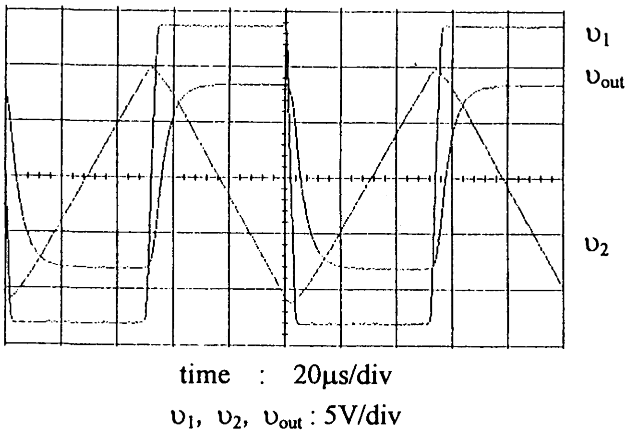

Fig. 5. Experimentally observed waveforms.

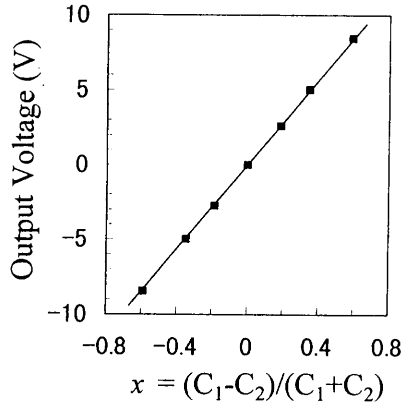

Fig. 6. The output amplitude V_o versus x.

This is the minimum time required for the ratiometric operation of 0.1% accuracy if A_3 slews much faster. Specifically, if $\omega_3 = 2 \times 10^5$ rad/s, then $\tau_s = 10\ \mu$s.

V. Prototype Interface

A prototype interface based on Fig. 2 was breadboarded using LF411 op-amps. The circuit parameters are $R_1 = R_2 = R_3 = R_4 = 1.2\ {\rm M}\Omega, R_5 = R_6 = 5\ {\rm k}\Omega, C_3 = 2$ pF, $V_{cc} = -V_{ss} = 15$ V, $V_{\rm ref} = 13$ V. Mica capacitors are used for C_1 and C_2 and their values are selected such that x range from -0.5 to 0.5.

Fig. 5 shows experimentally observed waveforms when $C_1 = 20$ pF and $C_2 = 5$ pF ($x = 0.6$). It can be seen that $v_{\rm out}$ assumes $+7.8$ and -7.8 V alternately and settles in 15 μs. These observed values agree quite well with theoretical values given by (8) and (23), respectively. The noise floor including the oscillatory amplitude is less than 2 mV. This also confirms the step response of the phase-compensated differentiator given by (22).

The amplitude V_o of the output signal measured over the wide range of capacitance change is plotted in Fig. 6. Fig. 7 shows the output amplitude plotted for the small capacitance change. In this measurement, a ganged capacitor composed of three parallel plates whose total capacitance is 6 pF was used in parallel with mica capacitors [7]. These preliminary results confirm the accuracy estimates in the previous section, indicating that a resolution as high as 0.1% is easily attainable with the interface.

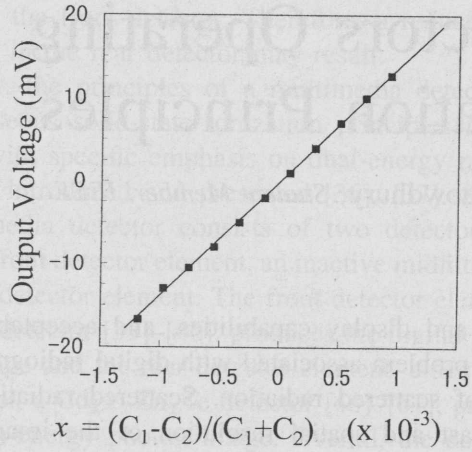

Fig. 7. The output amplitude for a small capacitance change.

VI. Conclusions

A relaxation-oscillator-based interface circuitry has been presented, which performs the ratiometric signal processing of a differential capacitance transducer in the amplitude domain. Circuit analyses have shown that a resolution higher than 0.1% and a sampling rate as high as 100 ksps are easily achievable with the simple configuration. A prototype interface built using off-the-shelf components has confirmed the circuit analyzes and demonstrated the validity in practical applications.

A one-chip implementation of the interface and its applications to practical transducers are future works.

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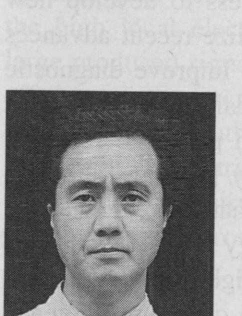

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