SC INTERFACE WITH EXTENDED LINEAR RANGE

Ventzeslav Petrov Iordanov, Gerard Meijer, Stoyan Nihtianov Nihtianov

Delft University of Technology, Electronic Instrumentation Lab., DIMES, Mekelweg 4, 2628 CD Delft, The Netherlands, Phone +31 15 278 3342, Fax +31 15 278 5755, E-mail: v.p.iordanov@its.tudelft.nl

Abstract. *In this paper a novel design for a switched-capacitor (SC) interface with extended linear range will be presented. The circuit is based on a simple relaxation oscillator. The improvement in the dynamic range is achieved by controlling the charge transfer of the sampling (measured) capacitor to the input integrator, using a charge transfer control loop.*

INTRODUCTION

Capacitive sensors are widely applied in cascade-sensor systems, in which a physical, chemical or mechanical quantity is converted into a capacitance value and further processed by an electronic circuit: the modifier. A simplified circuit of a typical front-end for such a modifier is shown in Figure 1.

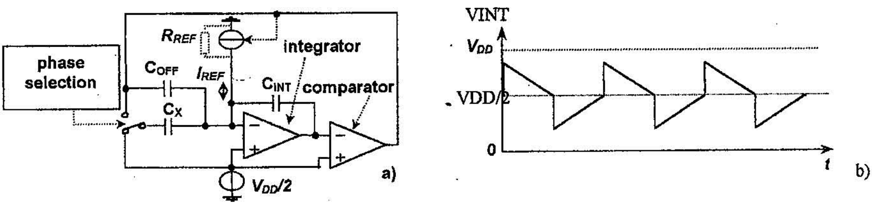


Figure 1: Voltage controlled oscillator a) and the output voltage of the integrator b).

In a sensor circuit the measurand can modulate the capacitor C_X or the voltage V_S. In both cases the measurand is transferred to a charge. At the moment t_0, this charge Q_X is transferred to the integrator. Next, integration of the (constant) reference current I_{REF} removes this charge. A comparator is monitoring the integrator output signal and actuates the switch control. When, at the moment $t_0 + T$, the integrator output voltage crosses the threshold level it holds that:

$$Q_X + I_{REF}T = 0\,, \tag{1}$$

where T is the time needed to remove the charge Q_X. By repeating this process periodically a charge-to-period converter is obtained. According to this simple principle many high-performance modifiers for sensor signals are made [1]- [4].

To reduce the effect of the comparator noise, the integrator output voltage should be large enough, which is achieved by selecting the capacitor value C_{INT} small enough. We have selected C_{INT} =10 pF. In integrated smart sensors such a small capacitor value offers the additional advantage to occupy only a small chip area. However, with such a small value of C_{INT}, a problem arises when this front-end circuit is used with a large capacitor C_X or a large voltage V_S: When a large charge Q_X is transferred to the integrator at once, this will cause an overload of the applied amplifier. If the output is clipped, the open loop gain of the integrator OpAmp (Operational Amplifier) decreases strongly. Then, the input signal of this amplifier is no longer negligible small, and the reference current I_{REF} changes, as caused by the source resistor R_{INT} not being infinite. As can be concluded from Equation (1), this causes non-linearity in the charge-to-period conversion. An even worse non-linearity can occur when input (protection) diodes are present. In case of a large input voltage of the OpAmp, these diodes can start to conduct. This will cause signal-charge loss at the integrator input. To extend the linear region, of the charge-to-period converter, without increasing the value the capacitor C_{INT}, a new front-end circuit has been designed.

SC FRONT-END WITH CHARGE-TRANSFER CONTROL

The clipping of the integrator's output voltage occurs because the charge of the measured capacitor C_X is dumped in a very short time into C_{INT}. If the charge transfer is spread out over an available time then the amplitude of the output voltage will not be clipped by the supply [5].

In Figure 2(a) circuit clipping of the amplifier output is avoided by using a negative feedback circuit, which controls the discharge of the capacitor C_X. The action of this feedback circuit can be explained by considering the integrator output voltage over a complete period (Fig. 2(b)).

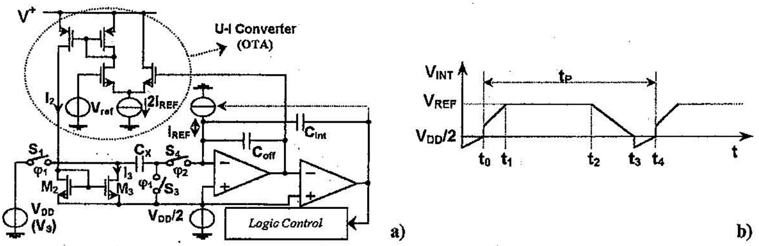


Figure 2: SC front-end with charge-transfer control: (a) diagram, (b) the integrator output voltage versus the time.

During the time interval just in advance of t_0 the switches S_1 and S_2 are closed and the voltage source V_{DD} charges the capacitor C_X. Next, S_1 and S_2 are opened and S_3 is closed. Now the charge transfer to the integrator starts. During the time interval t_0^+ to t_1 the integrator output voltage is small. Consequently, it holds that $(I_2 = I_1) > I_{REF}$. The differential current $(I_2 - I_{REF})$ charges the integrator capacitor C_{INT}, which causes an increase of V_{INT}. During the time interval t_1 to t_2 the output voltage V_{INT} is approximately equal to V_{REF}. The input stage of the *U-I* converter is in balance. This causes that $I_1 = I_2 \cong I_{REF}$. Because $(I_2 - I_{REF})$ is approximately zero, the integrator output voltage remains constant. However, the capacitor C_X is still discharged by the current I_2. This situation continues till the drain-source voltage of the MOS transistor is dropped so far that M_3 starts to be biased in the triode region. During the time interval t_2 to t_3 M_3 is biased in the triode region. Consequently, the left-hand side of the capacitor C_X is connected to ground via the ON resistance R_{ON} of M_3. The small remaining charge in C_X is transferred to C_{INT}. In practical circumstances the time constant R_{ON} C_X appears to be negligible small. Next, the reference current IREF linearly discharges the capacitor C_{INT}. The time interval t_3 to t_4 is the sampling interval in which the capacitor C_X is connected to the voltage source V_{DD}. At the start of this interval, the comparator switches, thus causing pumping of the charge of C_{OFF} into C_{INT}. Next, the comparator reverses the direction of I_{REF} to discharge C_{INT} and to generate a periodic signal, which starts for the next period at t_0 with a small charge-pumping action of C_{OFF} into C_{INT}. The requirements to the U-I converter are not very critical. The transfer function is allowed to be non-linear; thus a simple OTA can be used (Figure 2). Although the output voltage of the integrator has a strange shape, the Operational Amplifier operates in linear region. The period of the oscillator depends linearly on the capacitance to be measured. No charge is lost at the input of the integrator. Taking into account the effect of the offset capacitor C_{OFF}, for the period time t_P it can be found that:

$$t_P = \frac{V_{DD}C_X}{I_{REF}} + 2\frac{V_{DD}C_{OFF}}{I_{REF}} \tag{2}$$

The first term in the right-hand side of Equation (2) is proportional to the measurand (C_X), while the second term introduces a certain offset. In the next section is shown how auto-calibration technique compensates for the offset part.

EXPERIMENTAL RESULTS

Resolution and non-linearity have been tested with the measurement setup shown in Figure 3. In the first step, the offset measurement, the MUX connects both capacitors C_{REF} and C_X to $V_{DD}/2$. In the next two steps sequentially one of the capacitors C_{REF} and C_X is connected via the MUX to V_{DD}, while the other one remains connected to $V_{DD}/2$. The offset problems are expected due to a parasitic capacitance between the terminals of the transmitting and receiving electrodes (Figure 4 (a)). The parallel parasitic capacitances lead for an additive error when they are equal.

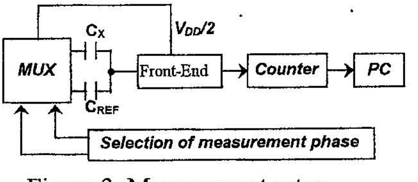
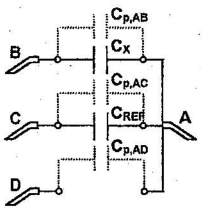


Figure 3. Measurement setup.

Applying the three-signal technique can eliminate their effect on the final measurement result C_X/C_{REF} in this case. Measuring pin D to which no capacitor has been connected performs the offset measurement.

a)



b)

$$T_{OFF} = T_0 + K_{\mathrm{mod},C}(C_{OFF} + C_{p,AD})$$
$$T_{REF} = T_0 + K_{\mathrm{mod},C}(C_{REF} + C_{OFF} + C_{p,AC})$$
$$T_X = T_0 + K_{\mathrm{mod},C}(C_X + C_{OFF} + C_{p,AB})$$

T_0 and $K_{\mathrm{mod},C}$ are constants of the modulator and C_{OFF} is the feedback capacitor in the VCO, ensuring the oscillations.

$$M = \frac{T_X - T_{OFF}}{T_{REF} - T_{OFF}} = \frac{C_X + C_{p,AB} - C_{p,AD}}{C_{REF} + C_{p,AC} - C_{p,AD}}$$

Figure 4. (a)The transmitting and receiving terminals and the parasitic capacitance between them; (b) Three-signal technique.

When the parallel parasitic capacitances are equal, their effect is completely eliminated by the three-signal technique. Any mismatch between these parallel parasitics leads to an error. Small differences in the offset can be obtained by applying an external multiplexer in combination with shielding or maintaining a sufficiently large distance between the transmitting and the receiving terminals.

The calculation of the non-linearity is based on four measurements. In addition to the measurement of the offset signal E_{OFF}, the signals $E_X^1+E_{OFF}$ and $E_X^2+E_{OFF}$ are measured separately, and, finally the sum $E_X^1+E_X^2+E_{OFF}$ is measured. This results in four measurement phases:

- $T_{OFF} = GE_{OFF}$
- $T_X^1 = G(E_X^1 + E_{OFF})$
- $T_X^2 = G(E_X^2 + E_{OFF})$
- $T_{X^1+X^2} = G(E_X^1 + E_X^2 + E_{OFF})$

The non-linearity is now defined by:

$$\lambda = \frac{T_X^1 + T_X^2 - 2T_{OFF}}{T_{X^1+X^2} - T_{OFF}} - 1$$

The non-linearity in the range 0 pF – 2000 pF has been measured with randomly chosen pairs of equal capacitors. In our test we found that $\lambda < 0.4\%$ (Table 1).

Table 1:

C_X, [pF]	C_{REF}, [pF]	σ, [pF]	λ, [%]
1,5	5,6	0,004	0,2
5,6	27	0,004	0,4
56	100	0,004	0,2
100	100	0,006	0,1
560	1000	0,08	0,08
1000	2000	0,07	0,05

Figures 5 and 6 show the noise performance of the oscillator. A capacitor of 100 pF was measured multiple times. It can be seen that without averaging the quantization noise is dominant. Averaging the signal leads to improved noise performance but increases the measurement time.

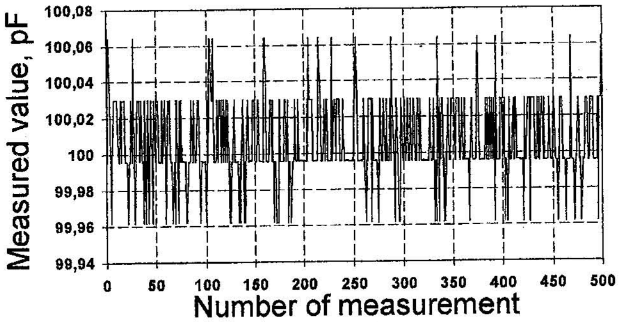


Figure 5: Set with 500 measurements of a capacitor $C_X = 100$ pF. Without any averaging the standard deviation amounts to 23,33 fF.

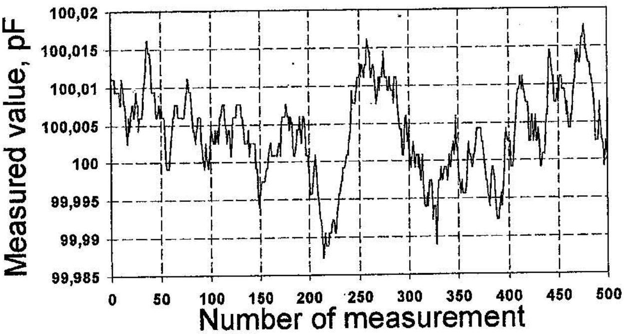


Figure 6: Set with 500 measurements of a capacitor $C_X = 100$ pF. Averaging every 20 measurements leads to standard deviation $\sigma = 5{,}83$ fF.

CONCLUSIONS

In this paper a novel design for a switched–capacitor (SC) interface for capacitive sensors with extended linear range is presented. The circuit is based on the use of a simple relaxation oscillator. The improvement in the dynamic range is achieved by controlling the charge transfer of the sensor capacitor to the input integrator, using a charge transfer control loop. In this way, the charge-transfer process is spread over the available time, which prevents clipping of the input amplifier. With the charge-transfer control loop the integrator operates in its linear region, although the output voltage of the integrator has a strange shape. The time period t_p varies linearly with the capacitance to be measured. Experimental results show a non–linearity of less then 0.4% over the range of 0 – 200 pF and resolution of 0.006 pF.

REFERENCES

[1] A.Cichocki and R. Unbehauen, "A switched-capacitor interface for capacitive sensors based on relaxatioin oscillators", *IEEE Trans. on Instr, and Measurement,* 1990, 39, (5), pp. 797-799.

[2] K. Mochizuki, K. Watanabe, T. Masuda and M. Katsura, "A relaxation-oscillator-based interface for high-accuracy ratiometric signal processing of differential-capacitance transducers", *IEEE Trans. on Instr, and Measurement,* 1998, 47, (1), pp. 11-15.

[3] F.N. Toth, G.C.M. Meijer and H.M.M. Kerkvliet, A. Very Accurate Measurement System for Multielectrode Capacitive Sensors, *IEEE Trans. on Instr, and Measurement,* 1996, 45, (2), pp. 531-535.

[4] F.M.L. van der Goes, Low-Cost smart sensor interfacing, *PhD. Thesis, Delft University of Technology,* April 1996.

[5]. Meijer G. and Iordanov V., "A novel switched-capacitor front end for capacitive sensors with wide dynamic range", *Proc. of ELECTRONICS'99,* Book 1, Sozopol, Bulgaria, September 1999, pp. 24-30.