A NEW CONCEPT FOR IMPEDANCE SENSOR INTERFACING WITH FIRST ORDER OSCILLATOR

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Abstract

A concept for an accurate and low-cost impedance sensor interface circuit, based on first-order charge-balanced oscillator, is presented. With this circuit the resistive and the capacitive components of capacitance-dominated impedance can be measured. The principle of operation and some applications are discussed. Experimental results are presented.

INTRODUCTION

Nowadays a variety of different concepts and electronic circuits are used to interface capacitive, inductive or resistive sensors. They convert the capacitance, the inductance or the resistance of a sensing element into an electric signal (voltage, current, frequency, period, digital code, etc.)[1,2,3,4]. In fact, the above-mentioned sensor elements have a complex electrical behaviour, which is modelled by a multiple-component equivalent circuit. In this sense, we can call them "impedance sensors". The main objective of the existing interface circuits is to actually convert one of the impedance components and to be immune to the effects of the rest. The effects of the undesired components often limit the range of the measured physical quantity and sometimes bring non-linearity to the transfer characteristic.

If we can sense at least two independent components – resistive and reactive, module and phase, quality factor and resonance frequency, of the impedance sensor, a number of benefits can be encountered. First of all, we have two output signals and both of them could be informative – a kind of operational multisensor. We can extend the measurement range of both components and improve the accuracy.

The problem is even more interesting in the special case, when we have directly to measure the electrical properties of the object of interest – the so-called "electrical sensing principle". A lot of applications for such measurements can be found in medicine and biology. In this case the equivalent electric circuit of the object can be pretty complicated, so that at different frequencies of the applied signal different components may have dominant influence on the measured impedance. By just doing multiple-frequency impedance measurements, we can obtain information about many aspects of the object of interest.

Today accurate impedance measurements, especially in the frequency range above one megahertz, are carried out by expensive laboratory measurement equipment. It is not suitable for sensor interfacing. Our goal is to develop methods and to design electronic

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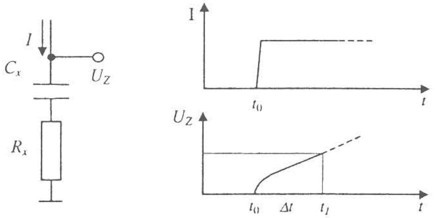
circuits to be used for accurate and low-cost impedance sensor interfaces. In the article we present a concept for such a circuit, based on a first order charge-balanced oscillator. With this circuit we can measure both the resistive and the capacitive component of impedance. The principles of operation and possible applications are discussed and some experimental results are presented.

MEASUREMENT STRATEGY

The measurement strategy is developed for an unknown impedance with serial R_x-C_x model. First, we can sense both the resistive and the capacitive components of the unknown impedance by applying a constant current to it (see Fig.1). The time Δt for which U_Z reaches a certain threshold level U_{ref} is a function of C_x and R_x, as it is shown in Fig.1.



$$\Delta t = \frac{U_{ref} C_x}{I} - R_x C_x$$

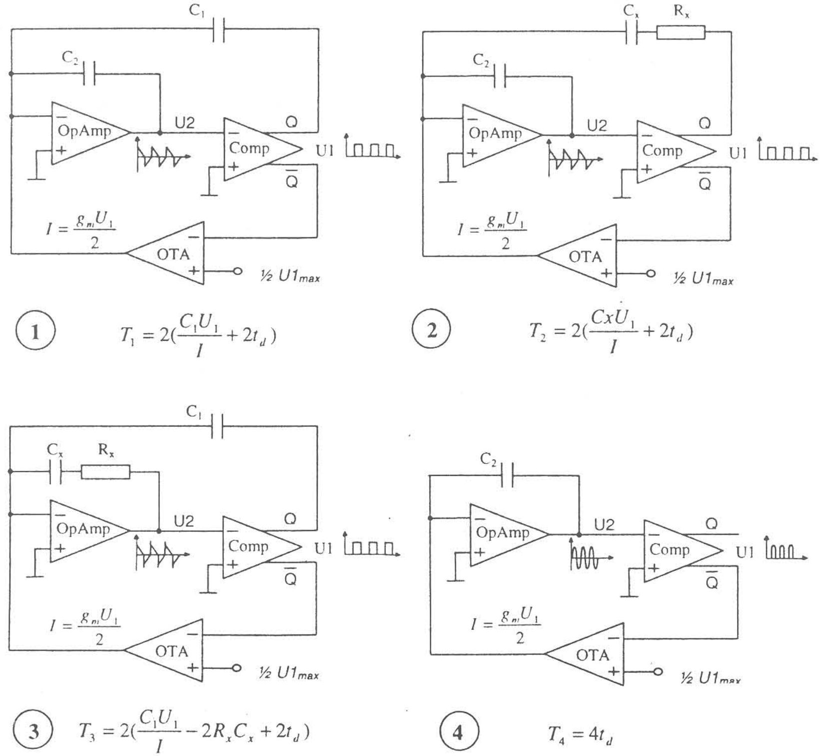
Fig. 1

Second, we can use one of the methods to sense the capacitive component of the impedance only. With the two results we can calculate the values of C_x and R_x.

CIRCUIT IMPLEMENTATION

To demonstrate the viability of the above-presented strategy we shall use the modified Martin oscillator, as described in [5,6]. The properties of this circuit as an accurate interface for capacitive sensors are already thoroughly explored and implemented in the available on the market Universal Transducer Interface (UTI) IC [7]. It ensures 15 bit of resolution for capacitors from 2 pF up to 300 pF. With currying out four measurements we can eliminate the undesired parasitics and the long-term drift of the oscillator frequency. Figure 3 shows the four steps for measuring the unknown impedance (R_x, C_x):

1) The value of a reference capacitor C_1 is converted into a period T_1;
2) The capacitive component C_x of the measured impedance is converted into period T_2;
3) To sense both C_x and R_x, the unknown impedance is placed in the negative feedback of the OpAmp, where runs a constant current $\pm I$. This leads to generation of pulses with period T_3;
4) Finally, the delay time $t_d = T_4/4$, due to the propagation delay in the comparator (Comp), the OpAmp and the Operational trance-conductance amplifier (OTA), is measured. It defines the highest frequency $f=1/4t_d$, which can be generated with this oscillator.



① $T_1 = 2(\frac{C_1 U_1}{I} + 2t_d)$

② $T_2 = 2(\frac{CxU_1}{I} + 2t_d)$

③ $T_3 = 2(\frac{C_1 U_1}{I} - 2R_x C_x + 2t_d)$

④ $T_4 = 4t_d$

Fig. 3

From the ratio $(T_2 - T_4)/(T_1 - T_4)$ we can calculate the value of C_x

$$C_x = C_1 \frac{T_2 - T_4}{T_1 - T_4} \qquad (1)$$

Next, from the difference $T_1 - T_3 = 2R_x C_x$ we can calculate the value of R_x

$$R_x = \frac{(T_1 - T_3)}{2} \frac{1}{Cx} = \frac{1}{2C_1} \frac{(T_1 - T_3)(T_1 - T_4)}{(T_2 - T_4)} \qquad (2)$$

To measure C_x (equation 1) we use the self-calibration three-signal method as it is done in [6]. It is important to point out that for the measurement of both impedance components - R_x and C_x, we need only one reference component: the capacitor C_1. The value of the other capacitor C_2 is not critical for the accuracy of this method. Nevertheless, to obtain the required linearity, it is important that both C_1 and C_2 are high-quality capacitors with values in the order of C_x.

There are some limitations for the measurement range of R_x. They are

$$R_x < \frac{C_1}{C_x}\frac{U_1}{2I} \qquad R_x C_x << T_2 \qquad (3)$$

If the first requirement in (3) is not met the circuit will start oscillating at its "natural" frequency $f \approx 1/T_4$ and will not be sensitive anymore to the values of R_x and C_x. If the second requirement is not met the generated period will be too short for all the charge in C_x to be transferred to C_2 in step 2 (see Fig.3) and this will lead to big error. To ensure high sensitivity for R_x high frequency oscillations are needed. That means that we have to use fast comparator, operational trance-conductance amplifier (OTA) and OpAmp. For good linearity the output resistance of the OTA and the input resistance of the integrating OpAmp have to be high. Otherwise the value of the current I will depend on the inevitable excursions of the voltage at the inverting input of the OpAmp during every transition period. Also part of the current will be lost during the transition periods, as it will go through the input resistance of the OpAmp.

PRACTICAL REALIZATION AND EXPERIMENTAL RESULTS

Figure 4 shows the circuit diagram of a charge-balanced modified Martin oscillator, with which the four measurement steps of Fig. 3 can be performed. Table 1 shows the state of the switches in each step.

The circuit was build with:

1. OpAmp LM7171 (NS) with: Slew Rate - 4100 V/µs; Open loop gain 85dB; Unity-gain bandwidth 200 MHz; Input resistance - 3.3 MΩ.
2. Comparator AD8561 (AD) with typical propagation delay of 7 nS;
3. Operational trance-conductance amplifier OPA660 (BB) with: BW 850 MHz; Slew rate 3000 V/µs ; Output impedance 25 kΩ||4.2 pF; Switching time < 1 nS.
4. Analogue switches MAX333A (MAXIM) with: R_{on}(typ.) < 17 Ω ; Input/output capacitance of 5 pF; Off leakage current < 1 nA.

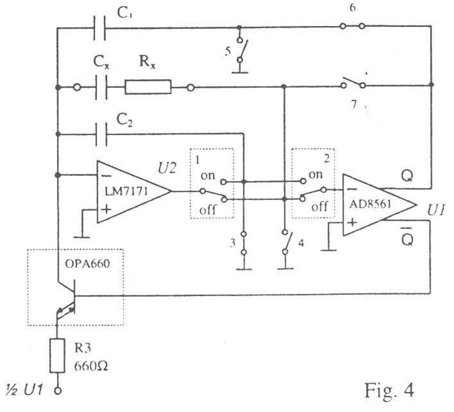


Fig. 4

Table 1

Step	1	2	3	4
Switch 1	on	on	off	on
Switch 2	on	on	off	on
Switch 3	off	on	on	off
Switch 4	on	off	off	on
Switch 5	off	on	off	on
Switch 6	on	off	on	off
Switch 7	off	on	off	off

All measurements were carried out by manually switching the switches 1÷7. We used $C_1 = C_2 = C_x = 220$ pF and $R_x = 50\ \Omega$. The integrating current was approximately 3 mA. The measured "natural" frequency of the oscillator $f = 1/T_4$ was higher than 30 MHz ($t_d \approx 8$ nS) and the frequency in steps 1, 2 and 3 was in the order of 2 MHz. The sensitivity for R_x was 1.2 nS/Ω. We could reliably sense 0.1 Ω change of R_x by averaging 1000 periods of the generated signal, which takes about 0.5 mS time.

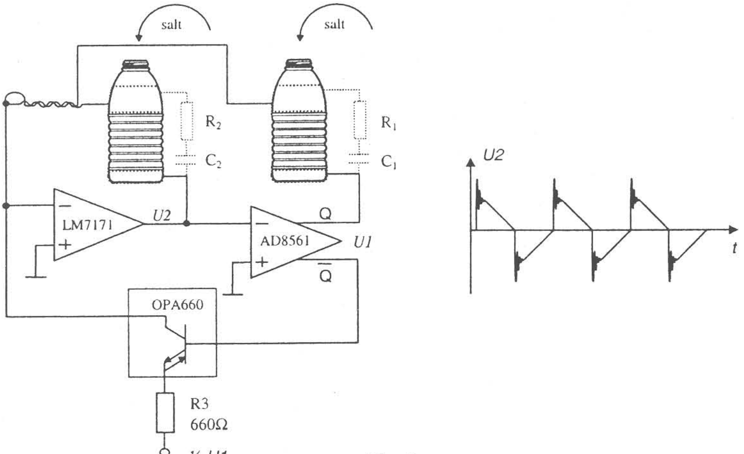


Fig. 5

To test the stability of this circuit when measuring the impedance of large objects we used two plastic 1-liter bottles with tap water in them, as it is shown in Fig.5. The initial values were: $C_1 \approx C_2 \approx 266$ pF, $R_1 = 258\ \Omega$. $R_2 = 262\ \Omega$. By adding salt first in Bottle 1 and then in Bottle 2 we decreased the resistance of R_1 and R_2. Figure 6 shows the period change of the generated pulses when the resistance R_2 was changed. The change of R_1 caused noticeable change of the generated period only when its value was well above 100Ω.

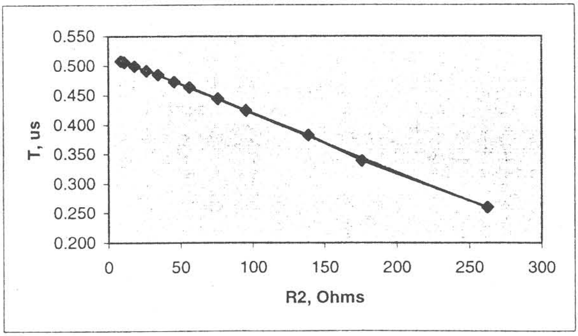


Fig. 6

We haven't yet tested the accuracy of this circuit and the effect of the main non-idealities, because to do so, we have to measure the periods T1÷T4 quickly one after the other by automatically switching the analogue switches 1÷7. In this way we shall avoid the long-term drift of the oscillator. The next step will be to build a micro-controller-based circuit, which will control the switches, will measure the generated time-periods and will calculate the components of the unknown impedance.

CONCLUSIONS

A new concept for an accurate and low-cost impedance sensor interface circuit is presented. The offered strategy has been discussed. A circuit implementation is also presented, based on a first order charge-balanced modified Martin oscillator. By measuring four times the generated period in four different configurations it is possible to calculate the resistive and the capacitive component of capacitance-dominated impedance, presented with a serial R-C model and to eliminate the effects of the inaccurate oscillator parameters. The principle of operation is discussed and experimental results are presented.

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