A Novel Smart Interface for Voltage-Generating Sensors

Xiujun Li, Gerard C. M. Meijer, *Member, IEEE*, and Erik J. Schnitger

***Abstract*—A novel low-cost smart interface for self-generating sensors, e.g., thermopiles and thermocouples, is presented. The proposed interface acts as an asynchronous converter for dc voltages employing a relaxation oscillator which output a period-modulated signal. A microcontroller is used to measure the output signal from the interface, to process data, and to output the measured sensor signal. The system provides an intrinsic A/D converter. Experimental results show that the interface is able to measure voltages in the range of 0 V to 5 V, with an accuracy and resolution of $\pm 64 \times 10^{-6}$ and 20×10^{-6}, respectively. The measurement time is less than 100 ms.**

***Index Terms*— Interface, relaxation oscillator, smart sensors, voltage measurement.**

I. INTRODUCTION

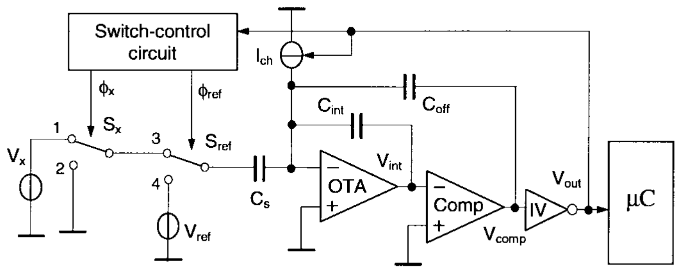
RECENTLY, some sensor interfaces have been presented in which the measurand is converted into a period-modulated output signal [1]–[5]. The advantage of this way of modulation is that powerful low-cost processing circuits, such as microcontrollers can directly process these signals without needing an extrinsic A/D converter. The combined features of such an interface and a microcontroller are very powerful. Using the memory and processing facilities of the microcontroller many nonidealities of the analog front end in the interface can be eliminated or significantly reduced [1], [2].

The interfaces which recently have been presented in [1]–[3], [5] have especially been designed to support passive sensing elements, such as capacitive, resistive, and resistance-bridge sensing elements.

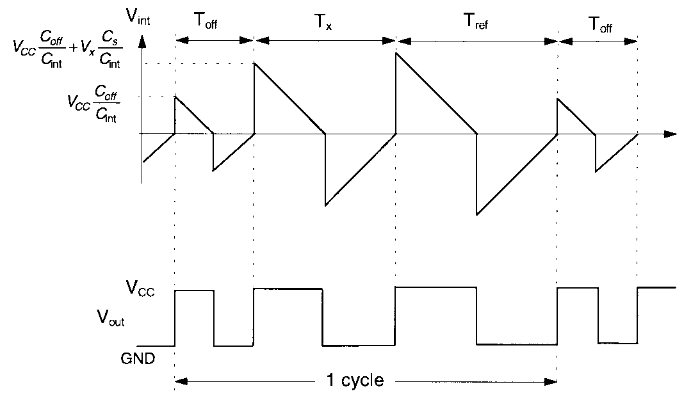
In this paper, we present a simple and low-cost smart interface which is suited to process signals of voltage-generating sensors, such as thermocouples. The interface can be implemented using low-cost CMOS IC technology.

II. MEASUREMENT CONCEPT

Fig. 1(a) shows a schematic diagram of the interface for the measurement of voltage signals. In this circuit, V_x denotes an unknown voltage (measurand), V_{ref} is a known reference voltage. The interface consists mainly of a modified Martin relaxation oscillator [6], sampling switches and capacitor, and a switch-control circuit. The modified Martin oscillator is a simple but very useful relaxation oscillator which is



(a)



(b)

Fig. 1. (a) The schematic diagram of the interface and (b) output signals V_{int} and $V_{\rm out}$ of the interface.

implemented with an operational transconductance amplifier (OTA), a comparator, an inverter, the capacitances $C_{\rm off}$, C_{int}, and a controlled current source [4], [6], [7]. The capacitor C_s and switches S_x, S_{ref} sample the voltage signals V_x and V_{ref} alternatively.

The three-signal technique [7] is applied to ensure the accuracy and reliability of the circuit by eliminating the gain and offset parameters of the interface system. To perform this three-signal technique, a reference signal (V_{ref}) and a constant part (including offset voltages) are measured in exactly the same way as the sensor signal (V_x) during two additional phases.

In the following, we explain the measurement principle.

The operational transconductance amplifier (OTA) and the capacitor C_{int} form an integrator with a low input impedance for the switched current I_{ch}. The sign of this current depends on the output state of the inverter IV. The output state of the comparator changes as soon as the output voltage of the integrator crosses the threshold level (in this case 0 V).

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Fig. 1(b) shows that the waveform of the voltage V_{int} at the integrator output and the output signal V_{out} of the interface.

The charge on the capacitors C_{off} is periodically dumped into and undumped from the capacitor C_{int}. The charge/discharge states of the switched current source I_{ch} are synchronized with the output signal of the oscillator. The states of the sampling switches are controlled by the switch control circuit which is synchronized with the output signal of the oscillator. During the offset measurement the logic control signals ϕ_{ref} and ϕ_x both are low, and the switches S_{ref} and S_x are in the fixed positions "3" and "2," respectively. The period T_{off} of the square-wave signal at the output of the oscillator amounts to

$$T_{\text{off}} = 2\frac{V_{cc}}{I_{ch}}C_{\text{off}} \tag{1}$$

where V_{CC} is the power supply voltage, I_{ch} is the amplitude of the charge/discharge current.

Now assuming, for example, the unknown voltage V_x is measured, the output of logic control ϕ_{ref} is low and ϕ_x changes its state to high. Then the switch S_{ref} will be in position "3," and the switch S_x and capacitor C_s will sample voltage signal V_x and convert it into a charge. The change Δq_x in the charge of the capacitor C_s causes an equal change of charge in C_{int}, where it holds that:

$$\Delta q_x = V_x C_s. \tag{2}$$

A periodic control signal ϕ_x causes this charge to be dumped into and undumped from the integrating capacitor C_{int} and the voltage V_{int} increases or decreases by the value

$$\Delta V_{\text{int}} = V_x \frac{C_s}{C_{\text{int}}}. \tag{3}$$

Now the integrating capacitor C_{int} is charged or discharged during a time T_x until the next zero crossing [see Fig. 1(b)]. Then, the period of the output signal of the oscillator T_x amounts to

$$T_x = 2\frac{V_x}{I_{ch}}C_s + T_{\text{off}}. \tag{4}$$

Similarly, when the voltage signal V_{ref} is measured, the switch S_x will be in a fixed position "2," and the switch S_{ref} and capacitor C_s will periodically sample voltage signal V_{ref}. Then, we obtain a period T_{ref} which it holds that

$$T_{ref} = 2\frac{V_{ref}}{I_{ch}}C_s + T_{\text{off}}. \tag{5}$$

Fig. 1(b) shows the output signal of the interface, which is very simple and has a discrete amplitude (V_{CC}), so that it is compatible with microcontrollers. With (1), (4), and (5), the measurand V_x is found from

$$V_x = \frac{T_x - T_{\text{off}}}{T_{ref} - T_{\text{off}}}V_{ref}. \tag{6}$$

Formula (6) shows that the values and inaccuracies of the capacitances C_s and C_{int}, and the charge/discharge current I_{ch} do not affect the measured result because of the applied three-signal technique. Moreover, in [7] it is shown that the interface is immune to most of the nonidealities of the OTA and the comparator, such as slewing, limitations of bandwidth and gain, offset voltages, and input bias currents. These nonidealities only cause additive or multiplicative errors which are eliminated by the three-signal technique.

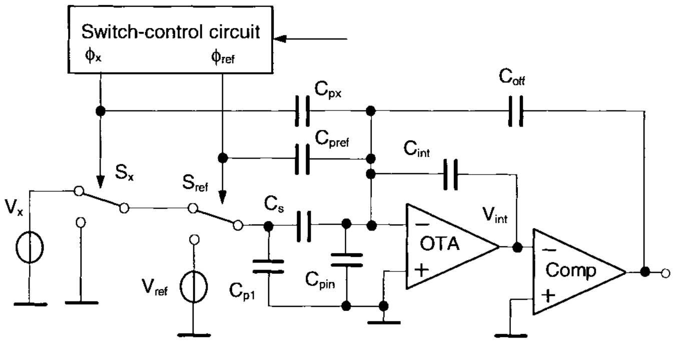

Fig. 2. The parasitic capacitances of the interface.

III. Nonidealities

Many nonidealities of the interface are eliminated by using the three-signal technique. However, some effects cannot eliminated by this technique, as discussed below.

A. The Effect of the Parasitic Capacitors

Fig. 2 shows some critical parasitic capacitances C_{px}, C_{pref}, C_{p1}, and C_{pin} of the interface.

1) The Parasitic Capacitance C_{pin} and Limited OTA Transconductance: The parasitic capacitance C_{pin} and the limited OTA transconductance g_{m0} limit the high-frequency (HF) bandwidth B of the integrator which equals

$$B = \frac{g_{m0}}{2\pi}\frac{1}{C_{pin} + C_{\text{off}} + C_s}. \tag{7}$$

This finite HF bandwidth will cause a nonlinearity of the conversion of the voltage signal to the output period. The corresponding relative nonlinearity ε_{NL} amounts to [2]

$$\varepsilon_{NL} \cong \left(\frac{1}{2} + \frac{I_{ch}}{V_{cc}C_{\text{off}}}\frac{C_{pin} + C_{\text{off}} + C_s}{g_{m0}}\right) \cdot \exp\left(-\frac{T_p}{2}\frac{g_{m0}}{C_{pin} + C_{\text{off}} + C_s}\right) \tag{8}$$

where T_p is the oscillator period.

A large transconductance of the OTA will reduce the effect of the parasitic capacitance on the nonlinearity, which is in agreement with general feedback theory.

The parasitic capacitance C_{pin} also results in an additive term for the period of the oscillator. This term T_{add} amounts to

$$T_{add} = \frac{2(C_{pin} + C_{\text{off}} + C_s)}{g_{m0}}. \tag{9}$$

For a constant parasitic capacitance, this additive term will be eliminated, by using the three-signal technique.

2) The Parasitic Capacitances C_{px} and C_{pref}: The influence of the parasitic capacitances C_{px} and C_{pref} on the results can be expressed by the following equation:

$$\frac{T_x - T_{\rm off}}{T_{ref} - T_{\rm off}} \cong \frac{V_x}{V_{ref}}\left(1 - \frac{V_{cc}}{V_{ref}} \cdot \frac{C_{pref}}{C_s}\right) + \frac{V_{cc}}{V_{ref}} \cdot \frac{C_{px}}{C_s}. \quad (10)$$

Equation (10) shows that even after applying the three-signal technique the parasitic capacitors C_{pref} and C_{px} cause a gain error and an offset on the result. These systematic errors can be eliminated after calibration while using a compensation algorithm in the microcontroller. For such a compensation, it is necessary that the parasitics do not change after calibration.

B. Output Impedances R_{oref}, R_{ox} of Reference Voltage Source and V_x

The output impedance R_{ox} or R_{oref} of the reference source or measurand V_x and capacitances associated with resistors R_{ox} and R_{oref} form an RC circuit with a time constant of

$$\tau_i \cong (2R_{\rm ON} + R_{oi})(C_s + C_{p1}) \quad (11)$$

where R_{oi} represents the output impedances R_{ox} or R_{oref}. Further, $R_{\rm ON}$ is the switch ON resistance of switches S_x and S_{ref}. This effect results in a relative error for the result

$$\varepsilon_{NL} \cong \exp\left(-\frac{T_p}{2}\frac{1}{t_i}\right). \quad (12)$$

Example: for a relative error of, for instance smaller than 10^{-5}, with $R_{\rm ON} = 200\ \Omega$, $T_p = 10\ \mu$s and $C_s + C_{p1} = 10$ pF, R_{oi} must be smaller than 43 kΩ.

C. Switch Clock Feedthrough

It is common practice to reduce the effect of switch-clock feedthrough by compensation. Because both switches S_x and S_{ref} consist two sub-switches which are controlled by opposing clock signals, the effect of the clock feedthrough is partly canceled. Moreover, it is possible to obtain a further reduction of the clock feedthrough by using CMOS switches. In our measurement setup no significant residual of the clock feedthrough has been observed.

D. Noise

In the voltage measurement, by using the interface shown in Fig. 1(a), the noise originates mainly from two parts: the oscillator noise and the quantization noise caused by sampling in the microcontroller. The standard deviations of the relative errors caused by these two noise sources can be expressed by, respectively, [8]

$$\varepsilon_{no} = \sqrt{\frac{C_{pin}}{C_s^2}\frac{2f_T C_{int} v_{eq}^2}{V_{cc}^2 N}} \quad (13)$$

$$\varepsilon_{nqt} = \frac{1}{\sqrt{3}}\frac{1}{NT_p f_c} \quad (14)$$

where f_T is the bandwidth of the OTA, v_{eq} is the rms value of the equivalent input noise (V/$\sqrt{\rm Hz}$) of the OTA, f_c is the sampling frequency of the microcontroller and N is the number of measured periods.

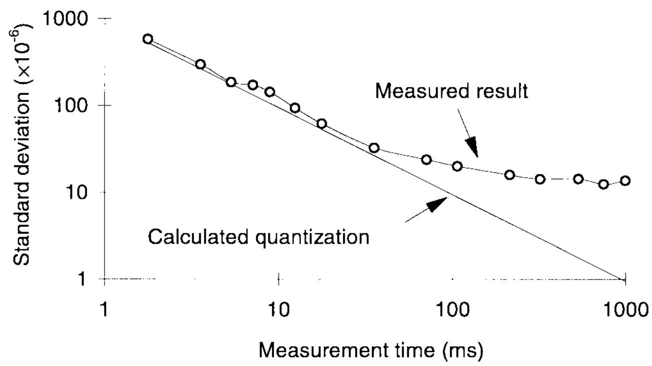

Fig. 3. Measured standard deviation versus the total measurement time.

TABLE I
MEASURED RESOLUTION AND NONLINEARITY FOR DIFFERENT MEASUREMENT RANGES

Meas. range	Resolution	Nonlinearity	C_s (pF)
0 V to 5 V	20×10^{-6}	$\pm64\times10^{-6}$	2
0 V to 1 V	19×10^{-6}	$\pm47\times10^{-6}$	10
-2 mV to 60 mV	28×10^{-6}	$\pm100\times10^{-6}$	150

IV. MEASUREMENT RESULTS

A prototype has been built based on the circuit shown in Fig. 1(a). The sampling switches are implemented with a simple quad bilateral switch (CD4066). The oscillator is fabricated as an ASIC which has been described in [2]. The frequency of the oscillator varies from 90 kHz to 45 kHz, depending on the sensor signals. A microcontroller of the type INTEL D87C51AF, which has a 3 MHz counting frequency, is employed to measure the output period of the interface, to process the data, and to communicate with the outside digital world. The system is powered by a single 5 V supply voltage.

The standard deviation has been determined as a function of the measurement time and is shown in Fig. 3 for the case that $V_x = V_{ref} = 1$ V and $C_s = 10$ pF. The measurement time T_{meas} is $N(T_{\rm off} + T_x + T_{ref})$, which is changed by changing the number N of periods.

Fig. 3 also depicts the quantization noise as calculated using (14). For short measurement intervals, the quantization noise, which originates from the sampling by the microcontroller, is dominant. For these short intervals, the resolution is inversely proportional to the measurement time.

For the different measurement ranges the nonlinearity and resolution of the interface are also measured, for a measurement interval of 100 ms (see Table I).

It is shown that the interface can measure an appropriate negative signal in a single positive power supply configuration. This is very useful for the application of thermocouple measurement.

V. CONCLUSIONS

In this paper a low-cost and accurate interface for voltage-generating sensors has been presented. This interface is based on the use of a modified Martin relaxation oscillator. The three-signal technique applied results in complete elimination of additive and multiplicative errors. This enables the realization of very accurate systems even when using low-cost inaccurate components. A microcontroller is used to process the signal from the interface and to output the sensor signal measured, eliminating the need for an external A/D converter. A prototype of the measurement system has been built and tested. The resolution and accuracy of the interface amount to 20×10^{-6} and $\pm 64 \times 10^{-6}$, respectively, over a range of 0 V to 5 V for a measurement time of 100 ms. The interface also can measure an appropriate negative signal with a single positive power supply. Such an interface is very easy to be implemented using low-cost integrated CMOS technology.

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