A microcontroller-based interface circuit for non-linear resistive sensors

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Abstract

This article proposes a circuit based on a microcontroller unit (MCU) for the direct measurement and linearization of non-linear resistive sensors, such as thermistors. The measurement relies on an embedded digital timer and does not require (either embedded or external) operational amplifiers or an analog-to-digital converter, thus resulting in a low-cost, low-power design solution. The circuit includes a known resistor with a twofold function: it is a reference for circuit auto-calibration purposes, and it is in parallel with the non-linear resistive sensor for linearization purposes. A prototype is implemented with an 8-bit MCU (ATtiny2313) applied to a commercial thermistor, and the results show a non-linearity error smaller than 1% full-scale span.

Keywords: Embedded system, microcontroller, resistive sensor, sensor interface electronics, thermistor

1. Introduction

Direct sensor-to-MCU interface circuits have been extensively analyzed in the last fifteen years for low-cost, low-power sensor applications. In these circuits, the sensor is connected to the MCU with a minimum number of external components, and without requiring (either internal or external) an OpAmp-based circuit or an analog-to-digital converter (ADC). Its operating principle generally relies on measuring, through an embedded digital timer, the charging or discharging time of an RC or RL circuit that includes the sensor. This technique has been applied to measure resistive [1, 2], capacitive [3, 4], inductive [5, 6], and voltage-output [7] sensors.

As for resistive sensors, new ideas complementing the topology proposed in [1] have been recently suggested. For instance, circuits for remote resistive sensors [8], for low-value resistive sensors [9], and with a reduced measuring time [10]. However, direct interface circuits for the measurement and hardware linearization of non-linear resistive sensors, such as thermistors, have not been tackled in the literature so far. In [11], a negative temperature coefficient (NTC) thermistor was directly measured by a MCU but without applying any hardware linearization technique. Likewise, in [12], an integrated circuit (IC) so-called USTI, which relies on an operating principle similar to that proposed in [1], was applied to measure an NTC but the linearization was performed in software.

On the semiconductor market, there are ICs specifically designed for the measurement of thermistors, such as MAX6682, MAX6691, UTI, among others. In these ICs, the thermistor and an external known resistor form a voltage divider whose output is read by an embedded ADC in MAX6682, a voltage-to-PWM converter in MAX6691, or a voltage-to-period converter in UTI. Other circuits based on more general-purpose ICs, such as the NE 566 [13], have also been proposed for the linearization of thermistors. However, in all these previous design solutions, a MCU is still necessary to read the output signal of these ICs. So, the challenge is how the MCU can directly read and linearize a non-linear resistive sensor, such as a thermistor, without using any active IC in between.

This article aims to prove that the circuit proposed in [1], but with a different configuration of the digital inputs/outputs of the MCU, can also be applied for the measurement and linearization of non-linear resistive sensors. This is carried out by placing a resistor in parallel with the sensor, which is a common practice, but this resistor is the same employed for the circuit auto-calibration and, hence, no additional components are required. The proposed method is applied to the measurement and linearization of a commercial NTC thermistor.

2. Operating principle

Figure 1 shows the circuit proposed to measure and linearize an NTC thermistor (R_x). The circuit includes an external known resistor (R_{CL}) with a twofold function: 1) circuit auto-calibration so that the output becomes independent of several variables and components of the circuit; and 2) sensor linearization so that the output changes linearly with the measurand.

The circuit in figure 1 relies on sequentially measuring the discharging time of three RC circuits and then applying the 3-signal auto-calibration technique [1]. First, pin 1 provides a digital "1" (with an analog output voltage V_1), whereas the other pins are set in high-impedance (HZ) state. Consequently, the capacitor C_{d} is charged towards V_1 through the external resistor R_{i}, which improves the rejection of power supply noise/interference [14], for a time longer than $5R_{\mathrm{i}}C_{\mathrm{d}}$. Once C_{d} is charged to V_1, pin 1 is set in HZ, and pins 2, 3 and 4 are set in one of the three configurations indicated in Table I. At this instant, the embedded timer starts, and C_{d} is discharged towards ground through pin 2 in the offset measurement, pin 4 in the reference measurement, and both pins 3 and 4 (to have R_x in parallel with R_{CL}) in the sensor measurement. The equivalent discharging resistance for the three measurements is specified in Table I, where R_0 is a resistor that limits the current to the maximum current sunk by a pin, and R_{n} is the internal resistance of a pin, which is assumed equal for the three discharging pins. Then, when the discharging exponential signal reaches the lower threshold voltage (V_{TL}) of the Schmitt-trigger buffer embedded into pin 1, the timer stops. The time interval needed to discharge C_{d} from V_1 to V_{TL} is proportional to the equivalent resistance, as indicated in Table I, where $k = C_{\mathrm{d}} \ln(V_1/V_{\mathrm{TL}})$.

Once we have the three discharging times (T_{off}, T_{ref} and T_{s}) related to the offset, reference and sensor measurements, respectively, the following time-based equation is applied to estimate the linearized sensor resistance:

$$R_{x\mathrm{L}} = \frac{T_{\mathrm{s}} - T_{\mathrm{off}}}{T_{\mathrm{ref}} - T_{\mathrm{off}}} R_{\mathrm{CL}} \tag{1}$$

According to (1), $R_{x\mathrm{L}}$ does not depend on k and, therefore, the tolerance and low-frequency variability (due, for instance, to thermal and time drifts) of C_{d}, V_1 and V_{TL} are not critical.

The circuit in figure 1 (excluding the sensor) is highly linear when reading resistances in the kiloohms range, with a non-linearity error of 0.01% full-scale span [1]. Therefore, the selection of the value of R_{CL} is not critical in terms of circuit auto-calibration. However, R_{CL} must be appropriately selected for sensor linearization purposes. Following the criterion of having an inflection point in the response of $R_x||R_{\mathrm{CL}}$ versus temperature at the central point (T_{c}) of the measurement range [15], the appropriate value of R_{CL} should be:

$$R_{\mathrm{CL}} = \frac{\beta - 2T_{\mathrm{c}}}{\beta + 2T_{\mathrm{c}}} R_{x,\mathrm{c}} \tag{2}$$

where β is the characteristic temperature of the NTC thermistor, and $R_{x,\mathrm{c}}$ is the value of R_x at T_{c}; both β and T_{c} are expressed in kelvins in (2).

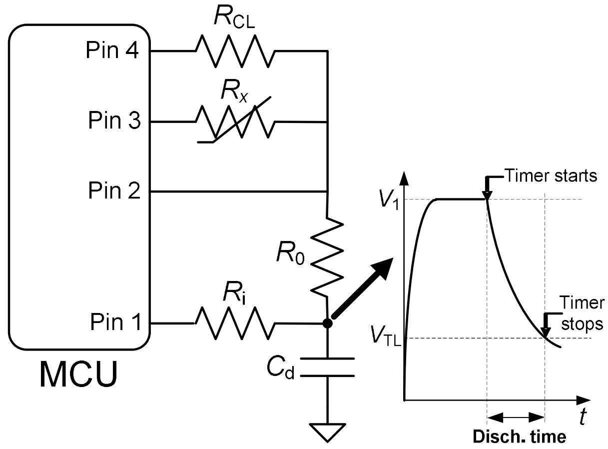

Figure 1. MCU-based circuit for the direct measurement of an NTC thermistor.

Table 1. State of pins 2, 3 and 4 in figure 1, equivalent resistance (R_{eq}), and resulting discharging time for each of the three measurements.

	P2	P3	P4	Eq. resistance	Disch. time
Offset	"0"	HZ	HZ	$R_{\mathrm{eq1}} = R_0 + R_{\mathrm{n}}$	$T_{\mathrm{off}} = k\, R_{\mathrm{eq1}}$
Ref.	HZ	HZ	"0"	$R_{\mathrm{eq2}} = R_{\mathrm{eq1}} + R_{\mathrm{CL}}$	$T_{\mathrm{ref}} = k\, R_{\mathrm{eq2}}$
Sens.	HZ	"0"	"0"	R_{eq3}[a] $\approx R_{\mathrm{eq1}} + (R_{\mathrm{CL}}\|\|R_x)$	$T_{\mathrm{s}} = k\, R_{\mathrm{eq3}}$

[a] This approximation is demonstrated in the appendix.

The main energy consumption of the circuit in Fig. 1 is due to [16]: 1) the fact of charging (three times) C_{d} to V_1, and 2) the measurement of the three discharging times with the embedded timer running at high frequency to reduce quantization effects. Note that the current consumption of the MCU itself while controlling the charging stage is in principle negligible since this task can easily be controlled by a timer running at low frequency (e.g., 32 kHz), which involves a very low current consumption. Accordingly, the energy required to carry out a complete measurement can be expressed as:

$$E \approx V_1 C_d \left[3V_1 + I_{timer} \left(R_{eq1} + R_{eq2} + R_{eq3} \right) \ln\left(\frac{V_1}{V_{TL}} \right) \right] \quad (3)$$

where I_{timer} is the current consumption of the MCU when the timer is running (and the CPU is in sleep mode) to measure the discharging time. If the timer operates at a higher frequency, the value of I_{timer} increases, but a lower value of C_d can be employed to have the same (relative) quantization effects. Therefore, there is a compensation in terms of energy, as inferred from (3).

3. Results

A prototype of the circuit in figure 1 was designed using an 8-bit MCU (ATtiny2313 from Microchip) running at 20 MHz with a software written in C language but at a register-level programming. The supply voltage was 5.0 V, thus resulting in $V_1 = 5.0$ V and $V_{TL} = 2.2$ V. The function of pins 1, 2, 3 and 4 was carried out by pins PD6/ICP, PB3, PB5 and PB7, respectively. The discharging time was measured via a 16-bit timer with a time base of 50 ns and applying the methodology explained in [17]. To be precise, a capture module (associated to pin PD6/ICP) was employed so that the uncertainty due to the interrupt-response delay is avoided and, in addition, the CPU was in sleep mode while waiting for the crossing of V_{TL} in order to avoid the effects of any software-related noise. The value of R_n for such a MCU is less than 20 Ω [1].

The circuit in figure 1 was applied to measure a commercial NTC thermistor (B57164K type from TDK) with a nominal resistance (at 25ºC) of 10 kΩ and $\beta = 4300$ K. This NTC was placed inside a small thermal chamber (9102 Hart Scientific) and subjected to temperatures from 5ºC to 45ºC, in steps of 5ºC; the actual temperature was known via a platinum thermal sensor. The other components selected were: $R_{CL} = 7500$ Ω [applying (2)], $R_0 = 470$ Ω, $R_i = 100$ Ω, and $C_d = 100$ nF. The board with the MCU and the components (excluding the NTC) was at room temperature during the tests.

At each step of temperature under test, the resistance of the NTC thermistor was first measured using a digital multimeter (Keysight 34410A, with an error of ±4 Ω in the worst case), as shown in red dashed line in figure 2. As expected, the resistance shows an exponential dependence versus temperature. Next, the same NTC thermistor was measured and linearized using the circuit in figure 1 and applying (1), thus resulting in the response shown in blue continuous line in figure 2. As can be seen, the output of the proposed circuit offers a remarkably linear response, with a non-linearity error smaller than 1% full-scale span. However, as usually occurs when applying hardware linearization techniques [15], there is a linearity-sensitivity trade-off. The improvement of linearity in Fig. 2 is at the expense of a lower sensitivity in the resistance detected.

A complete measurement involving three charging/discharging stages lasted around 1.3 ms in the worst-case scenario, which is short enough for many sensor applications. The value of I_{timer} was 4 mA, thus resulting in $E \approx 30$ μJ in the worst case. If the MCU operating conditions changed from 5 V – 20 MHz to 3 V – 4 MHz, then $I_{timer} = 430$ μA and $E \approx 4$ μJ. However, this is assuming the same value of C_d, so the energy required becomes smaller but quantization effects are higher (to be precise, five times greater) and, hence, the measurement resolution worsens. Therefore, there is another trade-off between resolution and energy consumption. For comparison purposes, it is worthy to mention that the energy required by the ADC-based MAX6682 IC is 46 μJ at 3.3 V. This energy, which is clearly higher than the numbers provided before, takes into account only the ADC conversion, but not the serial data transmission and the reading carried out by the MCU.

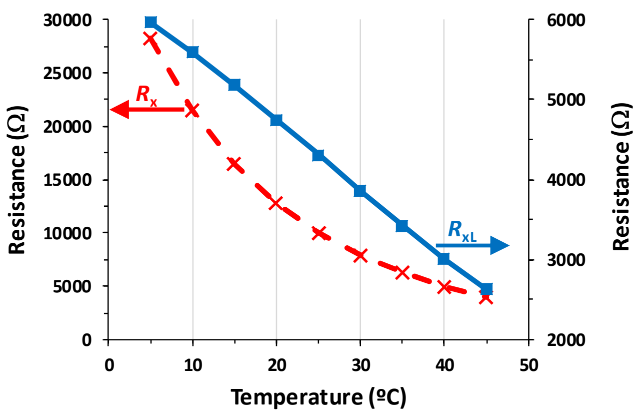

Figure 2. Resistance of the NTC thermistor versus temperature in red dashed line, and linearized sensor resistance estimated by (1) in blue continuous line.

4. Conclusions

In the context of direct sensor-to-MCU interfaces, this paper has demonstrated that non-linear resistive sensors, such as thermistors, can also be measured and linearized. This is achieved using the same circuit proposed so far for linear resistive sensors, but with a different configuration of the digital inputs/outputs of the MCU. The key of the design is the selection of a reference resistor, in principle only though for circuit auto-calibration purposes, that is placed in parallel with the non-linear sensor to overcome its non-linearity.

Appendix

For the sensor measurement, the equivalent resistance equals:

$$R_{eq3} = \left[\left(R_n + R_{CL} \right) \| \left(R_n + R_x \right) \right] + R_0 \quad (A.1)$$

which can be expressed as

$$R_{\mathrm{eq3}} = \frac{(R_{\mathrm{CL}} \parallel R_x) + R_{\mathrm{n}} + \dfrac{R_{\mathrm{n}}^2}{R_{\mathrm{CL}} + R_x}}{1 + \dfrac{2R_{\mathrm{n}}}{R_{\mathrm{CL}} + R_x}} + R_0 \qquad \text{(A.2)}$$

Assuming that ($R_{\mathrm{CL}} + R_x$) >> $2R_{\mathrm{n}}$, (A.2) can be approximated to

$$R_{\mathrm{eq3}} \approx \left[(R_{\mathrm{CL}} \parallel R_x) + R_{\mathrm{n}} + A\right] B + R_0 \qquad \text{(A.3)}$$

where

$$A = \frac{R_{\mathrm{n}}^2}{R_{\mathrm{CL}} + R_x} \qquad \text{(A.4)}$$

$$B = \left(1 - \frac{2R_{\mathrm{n}}}{R_{\mathrm{CL}} + R_x}\right) \qquad \text{(A.5)}$$

According to (A.4) and (A.5) and the value of the components employed in the set-up, we have $A = 0.03\ \Omega$ and $B = 0.997$ in the worst case. Therefore, in a first approximation, (A.3) can be simplified to

$$R_{\mathrm{eq3}} \approx (R_{\mathrm{CL}} \parallel R_x) + R_{\mathrm{n}} + R_0 \qquad \text{(A.6)}$$

which agrees with that indicated in Table 1.

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References

[1] Reverter F, Jordana J, Gasulla M and Pallàs-Areny R 2005 Accuracy and resolution of direct resistive sensor-to-microcontroller interfaces *Sens. Actuators A Phys* **121** 78-87

[2] Anandanatarajan R, Mangalanathan U and Gandhi U 2020 Enhanced microcontroller interface of resistive sensors through resistance-to-time converter *IEEE Trans. Instrum. Meas.* **69** (6) 2698-2706

[3] Reverter F and Casas O 2010 A microcontroller-based interface circuit for lossy capacitive sensors *Meas. Sci. Technol.* **21** 065203

[4] Czaja Z 2020 A measurement method for capacitive sensors based on a versatile direct sensor-to-microcontroller interface circuit *Measurement* **155** 107547

[5] Kokolanski Z, Jordana J, Gasulla M, Dimcev V and Reverter F 2015 Direct inductive sensor-to-microcontroller interface circuit *Sens. Actuators A Phys.* **224** 185-191

[6] Asif A, Ali A and Abdin M Z U 2019 Resolution enhancement in directly interfaced system for inductive sensors *IEEE Trans. Instrum. Meas.* **68** (10) 4104-4111

[7] Bengtsson L 2012 Direct analog-to-microcontroller interfacing *Sens. Actuators A Phys.* **179** 105-113

[8] Nagarajan P R, George B and Kumar V J 2017 Improved single-element resistive sensor-to-microcontroller interface *IEEE Trans. Instrum. Meas.* **66** (10) 2736-2744

[9] Hidalgo-López J A, Botín-Córdoba J A, Sánchez-Durán J A, Tejero-Calado J C and Oballe-Peinado O 2020 Improved calibration method for resistive sensors using direct interface circuits *IEEE Trans. Instrum. Meas.* **69** (8) 5693-5701

[10] Hidalgo-López J A, Sánchez-Durán J A and Oballe-Peinado O 2020 Reducing measurement time in direct interface circuits for resistive sensor readout *Sensors* **20** 2596 (13 p)

[11] Demori M, Baù M, Dalola S, Ferrari M and Ferrari V 2019 Low-frequency RFID signal and power transfer circuitry for capacitive and resistive mixed sensor array *Electronics* **8** (6) 675 (15 p)

[12] Yurish S Y and Cañete J 2013 Simple interface circuit for high-resolution, multichannel, smart temperature sensing based on NTC thermistors *Proc. 4th Intern. Conf. Sensor Device Technologies and Applications* 46-52

[13] Narayana K V L and Bhujanga R A 2012 A novel method of linearizing thermistor characteristic using voltage controlled oscillator *Sensors & Transducers* **136** (1) 1-11

[14] Reverter F, Gasulla M and Pallàs-Areny R 2007 Analysis of power-supply interference effects on direct sensor-to-microcontroller interfaces *IEEE Trans. Instrum. Meas.* **56** (1) 171-177

[15] Pallàs-Areny R and Webster J G 2001 *Sensors and signal conditioning* (New York: John Wiley & Sons)

[16] Sifuentes E, Gonzalez-Landaeta R, Cota-Ruiz J and Reverter F 2019 Seat occupancy detection based on a low-power microcontroller and a single FSR *Sensors* **19** (3) 699 (11pp)

[17] Reverter F and Pallàs-Areny R 2006 Uncertainty reduction techniques in microcontroller-based time measurements *Sens. Actuators A Phys.* **127** 74-79