Novel Telemetric Technique for Passive Resistive Sensors Based on Impedance Phase Angle Measurement at Constant Frequency

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Abstract— Telemetric systems are a valid alternative to measurement devices exploiting batteries or cabled connections, which may be not suitable in some cases, such as in harsh or hermetic measurement environments. They permit either to supply a passive sensor or to carry out monitoring tasks without contact. In this paper, we propose a novel measurement technique for telemetric systems working with a resistive sensing element. Such technique is based on a reading of system impedance phase angle always at sensing inductor’s resonant frequency, while relative distance between system inductors is kept fixed. Analytical expressions have been derived to calculate sensor resistive output from phase angle value. They have been obtained from a proper circuit model of the telemetric device. The proposed method has been applied to the wireless measurement of temperature inside an oven. Experimental tests have been carried out, by putting a resistive sensor inside the oven and increasing its internal temperature from about 28 °C to about 134 °C. Preliminary results show that calculated temperature values present an average deviation from reference equal to 1.7 °C. Furthermore, uncertainty related to such values is 1.2 °C. Therefore, they prove the feasibility of implementing the proposed technique in all the applications that require the use of telemetric devices with passive resistive sensors.

Keywords— Temperature measurement; Telemetric device; Resistive sensor; Impedance phase angle; Constant frequency.

I. INTRODUCTION

Telemetric devices are employed for the wireless measurement of multiple physical and chemical quantities. Examples touch pressure [1]-[2], temperature [3]-[4], humidity [5]-[6], strain [7]-[8], pH [9], and the presence of gases [10]. Furthermore, they are realized through different manufacturing technologies, according to the operation requirements of the specific application for which they are designed. For instance, passive sensing circuit can be fabricated with materials that keep their physical and chemical properties stable even in harsh conditions, like Low Temperature Co-fired Ceramics (LTCC) [1], [10]. In other cases, passive elements are printed directly on a flexible substrate [2]-[3]. In this way, they fit surfaces with different shapes. Finally, sensor can be designed as a microelectromechanical system (MEMS) [1], [3]-[4] with reduced dimensions, in order to occupy small spaces.

Literature reports telemetric systems equipped with capacitive [1], [4]-[7], [9]-[10], inductive [2], [8], or resistive [3] sensors. Thus, passive sensing circuits have a LC (in the first and second cases) or LR (in the third one) configuration. The former is the most considered solution, whereas the latter is less investigated. Anyway, the corresponding measurement techniques are based on system frequency analyses to detect circuit resonances, which shift according to sensor output change. However, such methods imply performing continuous frequency sweeps, in order to identify the resonances. This results in an increase of measurement time and complexity of the electronics that composes the readout unit.

On the contrary, this paper proposes an alternative technique for telemetric devices operating with a high-value resistive sensor, i.e., whose output is in the order of the kilohm. The proposed method relies on a mathematical formula that expresses the relationship between sensor output, the other device elements, and a reading of system electrical impedance, executed always at the same frequency. The first part of the work will present the analysis carried out to derive our solution from a proper circuit model of the system. Then, the technique has been applied for wireless temperature measurement inside a hermetic environment. Therefore, the second part will describe the experimental tests carried out on a real apparatus and the achieved results.
II. MEASUREMENT TECHNIQUE

A. Telemetric system model

A new circuit model is proposed to describe a telemetric system with a high-value resistive sensor. It is a lumped-parameter model that includes the parasitic effects and magnetic flux exchange between readout inductor and sensing inductor, which are connected to readout unit and passive sensor, respectively. This model is illustrated in Fig. 1, where each element is identified with the corresponding physical principle. Elements have the following definition [11]. Inductances \( L_R \) and \( L_S \) symbolize leakage fluxes from readout and sensing inductors, respectively. On the contrary, \( L_M \) represents the flux exchanged between them. These parameters depend on the distance at which inductors are coupled [12]. Parasitic resistance \( R_S \) takes into account the losses in sensing inductor’s windings, because of Joule effect. On the other side, parasitic capacitance \( C_2 \) represents the electric field generated between the windings of sensing inductor and between windings and its substrate [11]. Then, \( C_1 \) is a fixed capacitor that can be attached in parallel to sensing inductor, in order to set its resonant frequency. Resistance \( R_s \) models sensor output, which varies as a function of the measured quantity. It can assume values from about a kiloohm up to some tens of kiloohm. Finally, coefficient \( n \) is the windings ratio.

Combining the elements of the model of Fig. 1 permits to obtain the expression of impedance at readout inductor’s terminals (which is indicated with “Z” in Fig. 1). It is reported in (1) as a function of frequency \( f \). Parameter \( C_p \) is the sum between \( C_2 \) and \( C_1 \). \( L_1 \) and \( L_2 \) are the equivalent inductances for readout and sensing inductors, respectively. They are linked to model elements through the following relations [11]:

\[
L_1 = L_M + L_R \tag{2}
\]

\[
L_2 = L_S + L_M / n^2 \tag{3}
\]

\( L_1 \) and \( L_2 \) are constant parameters. In particular, they are not affected by the relative position between the inductors, since leakage flux growth when inductors get far is perfectly compensated by coupling flux decrease, and vice versa.

Finally, variable \( \beta \) depends on \( L_M \), as defined by (4). Thus, it changes if relative position between the inductors varies.

\[
\beta = \frac{L_M^2}{L_S n^2} \tag{4}
\]

Used notation with \( L_R, L_S, L_M, \) and \( \beta \) is equivalent at all, from a mathematical point of view, to the more common representation employing coupling factor to model the interaction between the inductors [1], [3], [7], [9]-[10].

B. Study on impedance phase angle

Equation (1) provides some information on how \( Z \) varies with system parameters and, therefore, with \( R_s \). Conversely, we started from such expression to find an equation that allowed us calculating sensor output from an impedance measurement performed at readout inductor’s terminals. For this purpose, we evaluated which impedance point was the most suitable to detect. In a previous work [13], we studied the response of a telemetric device when the value of its resistive sensing element changed. In particular, we represented its impedance as magnitude and phase angle. Fig. 2 reproduces the results related to the latter. Reported qualitative curves show how it behaves when \( R_s \) augments. We found that phase angle decreased when sensor output increased. In particular, its minimum was a convenient point, since its variation with \( R_s \) presented a linear trend, for sensor output range in the order of some kiloohms. In addition, the corresponding frequency \( f_{\text{min}} \) presented a change with respect to its starting value equal to 0.12% at most. Reference [10, 13] reports similar findings.

\[
Z(j2\pi f) = \frac{-4\pi^2(L_1L_2 - L_3\beta + R_s C_p L_S) f^2 + j2\pi L_1(R_s + R_x) f - 8\pi^2 R_s C_p L_S(L_1 - \beta) f^3}{(R_s + R_x) - 4\pi^2 R_s C_p L_S f^2 + j2\pi(L_1 + R_s C_p R_x) f} \tag{1}
\]
Therefore, \( f_{\text{min}} \) can be considered constant. If coupling distance between sensing inductors is sufficient, then \( f_{\text{min}} \) can be identified with sensing inductor’s resonant frequency \( f_{\text{res}} \) [1], with further addition of capacitance \( C_s \). This frequency is defined as:

\[
f_{\text{res}} = \frac{1}{2\pi \sqrt{L_s C_p}}
\]  

Therefore, we based our measurement technique on a phase angle reading performed always at \( f_{\text{res}} \). Working at a fixed frequency leads to some advantages related to the aspect that a complete sweep for the identification of the frequencies of interest is not necessary. Thus, both measurement time and readout electronics complexity can be reduced.

C. Mathematical analysis

Equation (6) provides phase angle mathematical function of frequency \( f \). It is derived from (1).

\[
\angle Z(f) = \frac{\pi}{2} - a \tan 2\left[ \frac{2\pi(L_s + R_s C_p R_s)f}{R_s + R_j - 4\pi^2 R_s C_p L_s f^2} \right] + a \tan 2\left[ \frac{2\pi(L_s L_j - L_j \beta + R_s C_p L_s R_j)f}{L_s(R_s + R_j) - 4\pi^2 R_s C_p L_s^2(L_j - \beta)f^2} \right].
\]  

Since the proposed method is supposed to be applied for telemetric systems operating with resistive sensors in the order of the kilohm, then \( R_s \) is much greater than \( R_j \). In fact, such parasitic resistance is usually in the order of some ohms or tens of ohm. Therefore, this approximation can be introduced:

\[
R_s + R_j \approx R_s
\]  

Then, phase angle at frequency \( f_{\text{res}} \) is called \( \phi \) from now on:

\[
\phi = \angle Z(f = f_{\text{res}})
\]  

Equation (6) is used to calculate the expression of the tangent of \( \phi \), taking into account also (7) and (8):

\[
tan(\phi) = \frac{L_s(L_j - \beta) + L_s^2 R_j C_p R_s}{\beta R_j \sqrt{L_s^2 C_p}}
\]  

Working at \( f_{\text{res}} \), together with considering (7), permits to simplify (6) in a significant way, when obtaining (9). Finally, (9) is reordered to isolate sensor output:

\[
R_j = \frac{L_s(L_j - \beta)}{\beta \sqrt{L_s^2 C_p} tan(\phi) - L_s^2 C_p R_s}
\]  

III. APPLICATION TO TEMPERATURE MEASUREMENT

A. Experimental setup

We carried out an experimental analysis to apply the proposed technique to the wireless measurement of temperature in a hermetic environment. Fig. 3 reports a block scheme of the experimental setup realized for the analysis, whereas Fig. 4 shows an image with some of its components. We employed a telemetric configuration made by two planar inductors. They were fabricated through printed circuit board (PCB) technology, by patterning a square copper spiral on a rigid substrate of FR4 glass-reinforced epoxy. Table I lists their geometric and electrical characteristics. The latter are the elements of inductors’ equivalent circuits. They were measured with an HP4194A impedance analyzer. We derived the uncertainty associated to the electrical characteristics by taking into account analyzer’s accuracy in their estimation. It refers to a 95% confidence interval (CI). In Table I, the electrical characteristics related to sensing inductor comprise the addition of a 560 pF commercial capacitor, which works as \( C_s \). According to the measured values of Table I and Eq. (5), frequency \( f_{\text{res}} \) was equal to about 2.52 MHz.

A Class A (according to standard EN 60751) Pt1000 resistive temperature sensor was connected to sensing inductor and capacitor \( C_s \). These components form system’s sensing circuit, which was put in an oven, as shown in the inset of Fig. 4. The oven simulates a hermetic environment, where temperature is supposed to be measured.

We placed readout inductor near the closed oven, keeping it parallel and coaxial with respect to sensing inductor, at a
relative distance fixed to 20 mm, thanks to a mechanical support equipped with micrometric screws. Readout inductor’s terminals were then connected to HP4194A analyzer, which was interfaced to a laptop via a GPIB-USB high-speed module. A Virtual Instrument (VI) written in LabVIEW was executed on laptop for data acquisition from HP4194A analyzer and their elaboration. The ensemble made by these two instruments compose system’s readout unit. In addition, another resistive temperature sensor, a Class A Pt100, was inserted in the oven. It was exploited as a reference, to make a comparison with the results provided by the proposed method. It was connected to an Agilent 34401A digital multimeter, which was interfaced to the laptop and was driven by the same VI. Thus, a synchronized data acquisition from the two devices was achieved.

**B. Measurement protocol**

Process first step consisted in system single-point calibration. Room temperature (i.e., the initial state) was chosen as known condition. After a trigger command from laptop, HP4194A analyzer detected \( \phi_0 \). Then, the VI received the data from that reading and calculated \( \beta \), by applying (11), from system parameters and sensor output \( R_{x,0} \), which was found from its datasheet. Afterwards, the VI began to operate in continuous mode, and the real measurement procedure started. HP4194 analyzer identified \( \phi \). Subsequently, the VI calculated \( R_x \) by implementing (10), from system parameters and coefficient \( \beta \) obtained earlier. Then, it converted \( R_x \) into a temperature value, by exploiting sensor linear relationship, which is known from its datasheet. Finally, it compared calculated temperature with reference provided by Pt100 and digital multimeter. While system was working, we gave growing heat to the oven, increasing gradually its temperature from about 28°C to about 134°C corresponding to a variation from about 1111 Ω to about 1515 Ω.

**C. Preliminary results**

The experimental tests carried out so far permitted us to obtain preliminary results about method effectiveness in estimating resistive sensor output. Fig. 5 reports some of the read values of phase angle \( \phi \) as a function of temperature. The vertical bars show the uncertainty related to \( \phi \) measurement, which is affected by

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**TABLE I. SELECTED H-P 4194A ANALYZER**

<table>
<thead>
<tr>
<th>Property</th>
<th>Value (unit)</th>
<th>Value (unit)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Geometric characteristics</td>
<td></td>
<td></td>
</tr>
<tr>
<td>outer side</td>
<td>50 (mm)</td>
<td>27 (mm)</td>
</tr>
<tr>
<td>inner side</td>
<td>27 (mm)</td>
<td>4 (mm)</td>
</tr>
<tr>
<td>number of windings</td>
<td>28</td>
<td>23</td>
</tr>
<tr>
<td>windings width</td>
<td>150 (μm)</td>
<td>300 (μm)</td>
</tr>
<tr>
<td>windings spacing</td>
<td>250 (μm)</td>
<td>200 (μm)</td>
</tr>
<tr>
<td>Electrical characteristics</td>
<td></td>
<td></td>
</tr>
<tr>
<td>equivalent inductance</td>
<td>47.48 ± 0.09 (μH)</td>
<td>7.05 ± 0.01 (μH)</td>
</tr>
<tr>
<td>equivalent resistance</td>
<td>35.10 ± 0.02 (Ω)</td>
<td>1.899 ± 0.004 (Ω)</td>
</tr>
<tr>
<td>equivalent capacitance</td>
<td>3.062 ± 0.006 (pF)</td>
<td>566.6 ± 1.1 (pF)</td>
</tr>
</tbody>
</table>

(*) Parallel capacitance is included for electrical parameters estimation.

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Fig. 3. Block scheme of the used experimental apparatus.

Fig. 4. Picture showing the experimental apparatus.

Fig. 5. Phase angle \( \phi \) as a function of temperature.
two factors. These are analyzer’s accuracy and $f_{res}$ value calculation (which depends on the uncertainty on $L_2$ and $C_p$). On the other hand, the horizontal bars (which are barely visible in Fig. 5) derive from the uncertainty on reference temperature. Such uncertainty is affected by Pt100’s tolerance characteristics and digital multimeter’s measurement accuracy. In both cases, uncertainty refers to a 95% CI. Fig. 5 points out how $\phi$ decreases when temperature grows, since resistive sensor output increases as well. This confirms the trend shown in Fig. 2. Phase angle behavior is almost linear within the temperature range that we took into account. Indeed, a linear regression fits the points with a coefficient of determination $R^2$ close to 0.998. We found that $\phi$ changed from $81.70 \pm 0.16$ degrees to $79.33 \pm 0.15$ degrees for a temperature increase from $28.5 \pm 0.2$ °C to $134.4 \pm 0.5$ °C. This means that $\phi$ sensitivity to temperature is equal to 0.022 degrees/°C, on average.

Fig. 6 shows the values of temperature calculated by implementing the proposed method from phase angle points reported in Fig. 5. In addition, bars referring to temperature uncertainty for a 95% CI are included. Furthermore, Fig. 6 compares calculated values with reference. Fig. 6 points out how the two series are close to each other. Indeed, their difference is equal to 1.7 °C on average. In general, it is between 0.2 °C and 3.7 °C along all the studied temperature range. Furthermore, uncertainty on calculated temperature results to be equal to 1.2 °C, on average. It was estimated by applying the rules of the “Guide to the expression of Uncertainty in Measurement” (GUM) to the mathematical equations illustrated in Section II, starting from the uncertainty related to $\phi$ measurement and system parameters. We took into account Pt1000’s tolerance characteristics at the same time. Therefore, final uncertainty is affected by all these factors, which were previously described. Achieved results are satisfying, considering also the examined range, which is quite wide. They prove that calculating the output of a resistive sensor through the proposed solution is a valid approach.

**IV. CONCLUSIONS**

This paper has presented a novel measurement technique for telemetric systems that work with passive high-value resistive sensors. It is based on two features. The first one is a reading of impedance phase angle always at the frequency identifying sensing inductor’s resonance, keeping system inductors at a fixed relative position. The second one is the implementation of equations that permit to calculate sensor output remotely from such reading and system parameters. We described how we derived these mathematical formulas from a proper circuit model of the system. Furthermore, the proposed solution has been applied to the wireless measurement of temperature inside an oven. For this purpose, we used a real telemetric device made of PCB planar inductors and a Pt1000 resistive sensor. Performed experimental analysis has shown that method provides a temperature estimation that deviates of 1.7 °C from reference, on average. Furthermore, derived uncertainty on calculated temperature is 1.2 °C. Such results point out the feasibility of working at a constant frequency. This has the advantage of avoiding continuous sweeps to identify the frequencies of interest. In general, results demonstrate the validity of our approach, which could represent a useful alternative, especially when a direct estimation of the output of a resistive sensor is not possible to obtain because of the characteristics of measurement environment.

Future studies will investigate the possibility to find a way to compensate the influence of a change in inductors’ relative position (as achieved in [11] for devices that operate with capacitive sensing elements). In fact, this could overcome the limitation due to the necessity of always maintaining the inductors at a fixed relative position during the entire measurement process.

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**REFERENCES**


