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Direct inductive sensor-to-microcontroller interface circuit☆

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ABSTRACT

This paper proposes and analyses a microcontroller-based interface circuit for inductive sensors with a variable self-inductance. Besides the microcontroller (μC) and the sensor, the circuit just requires an external resistor and a reference inductor so that two RL circuits are formed. The μC appropriately excites such RL circuits in order to measure the discharging time of the voltage across each inductor (i.e. sensing and reference) and then it uses such discharging times to estimate the sensor inductance. Experimental tests using different commercial μCs at different clock frequencies show the limitations (especially, due to parasitic resistances and quantisation) and the performance of the proposed circuit when measuring inductances in the millihenry range. A non-linearity error lower than 0.3% full-scale span (FSS) and a resolution of 10 bits are achieved, which are remarkable values considering the simplicity of the circuit.

1. Introduction

Inductive sensors are widely used in industry electronic instrumentation because they are robust and compact and, in comparison to capacitive sensors, they are less affected by environmental factors such as humidity and dust [1]. Within the group of inductive sensors, those based on a variable self-inductance (due to changes of either the magnetic reluctance or the number of turns of the coil winding) are quite common to measure displacement (e.g. of ±1 mm [2], ±5 mm [3] and ±6 mm [4]), position (e.g. of a piston inside a power cylinder [5]) and proximity. Other physical quantities that indirectly cause a displacement are also measured through these sensors, for instance: inductive pressure sensors based on either a Bourdon tube [6] or a vertical coil embedded into an integrated circuit (IC) package [7]. Inductive sensors have also been proposed to measure temperature by using cores with a low Curie temperature [8]. In the previous applications [3,4,6,7], the sensor inductance is in the range of units or tens of millihenry, but lower values (e.g. units or tens of microhenry) can also be found.

Two main types of interface circuits have been proposed for inductive sensors with a variable self-inductance and both usually operate at low-medium frequencies. The first type is a relaxation oscillator (based, for instance, on a 555 IC timer [7]) providing a time-modulated signal that can be read by a digital system (e.g. a μC with an embedded timer) without using an analog-to-digital converter (ADC). The second type is an AC-excited bridge (such as the Maxwell bridge [6] or the Maxwell-Wien bridge [3]) providing an amplitude-modulated signal that needs to be demodulated and digitised before being read by the digital system. The reference inductor usually required in bridge circuits is proposed to be emulated by a generalised impedance converter in [9]. Another interface circuit suggested for inductive sensors that does not belong to the previous two groups is the dual slope inductance-to-digital converter whose output can be read by a digital system without an ADC, as in the oscillator circuits, but it needs operational amplifiers (OpAmp) and analogue comparators in the signal path [4]. There are also a few commercial ICs (e.g. LDC1000 from Texas Instruments) that perform an inductance-to-digital conversion, but these are mainly intended for inductive sensors based on eddy current operating at higher frequencies.

With the aim of reducing the cost and power consumption of sensor electronic interfaces, the concept of “direct interface circuit” has been widely proposed, analysed and tested for resistive [10–12] and capacitive [13–15] sensors. In these circuits, the sensor resistance (or capacitance) together with a capacitor (or resistor) form an RC circuit whose charging or discharging time is directly measured by a μC through an embedded digital timer and without using any intermediate active circuit (such as comparators, OpAmps, timers and/or ADC). The performance of such circuits is quite remarkable taking into account their simplicity, for instance:

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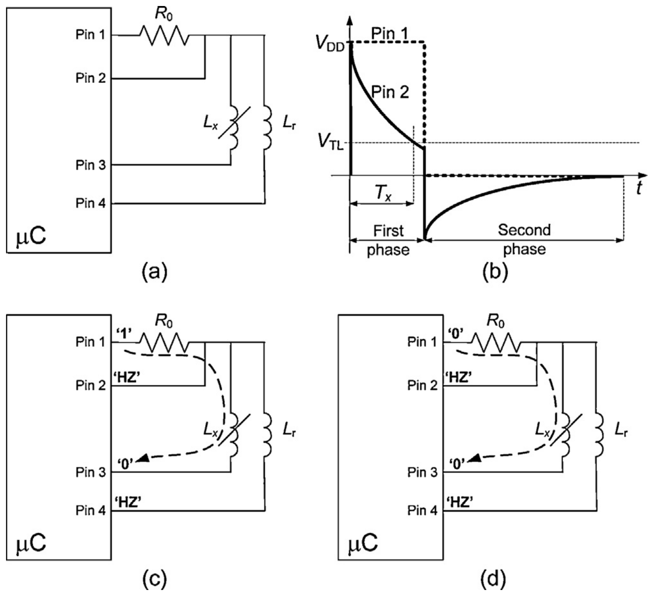

Fig. 1. (a) Proposed direct interface circuit for an inductive sensor (L_x). (b) Transient response of the voltage at pins 1 and 2 when measuring the RL circuit that includes L_x. (c) First phase for the measurement of L_x. (d) Second phase for the measurement of L_x.

a non-linearity error (NLE) of 0.01% full-scale span (FSS) and an effective resolution of 13 bits when measuring resistive sensors in the kiloohm range [10,11], and 0.1% FSS and 9 bits when measuring capacitive sensors in the picofarad range [13]. Although the same operating principle could be applied to measure inductive sensors by employing an RL circuit, instead of an RC circuit, formed by the sensor inductance and a resistor, no attempts to do so have been reported so far. Just in [16,17] we can find very preliminary circuit proposals but these have not been either analysed or tested.

As a continuation of the work presented in [18,19], this paper proposes, theoretically analyses and experimentally evaluates a direct interface circuit for inductive sensors with a variable self-inductance. In the proposed circuit, the inductive sensor is excited by a single step pulse and the result of the measurement is the inductance value at low frequencies; this is assuming that the frequency dependence of the inductance (due to the frequency dependence of the permeability) starts decreasing at high enough frequencies. For this reason, the proposed circuit is not suitable for those inductive sensors whose operating principle involves the measurement of the inductance at medium–high frequencies; this is the case, for instance, of eddy-current sensors that are generally excited by an AC signal of high frequency (say, units or tens of MHz) so as to have an appropriate penetration depth in the metallic target to be detected [20,21]. Furthermore, the sensor is expected to have an inductance of some units or tens of millihenry [3,4,6,7]; lower values of inductance would require a very high speed reference oscillator that is not feasible nowadays in common low-cost 8-bit microcontrollers.

The paper is organised as follows: Section 2 describes the operating principle of the circuit, Section 3 analyses the error sources, Section 4 shows experimental results, and Section 5 provides the main conclusions.

2. Operating principle

The proposed direct interface circuit for inductive sensors is shown in Fig. 1a. Besides the μC and the sensor (L_x), this electronic interface just needs a reference inductor (L_r) that is used for a single-point calibration, and an external resistor (R_0) that limits the current provided by the μC. With these components, two RL circuits are formed: R_0 together with L_x, and R_0 together with L_r. Then, each RL circuit is appropriately excited by the μC so as to measure the discharging time (T_x and T_r) of the voltage across each inductor (L_x and L_r, respectively). As for the μC, only digital peripherals (specifically, a timer) and digital input/output ports are required, thus resulting in a fully digital sensor interface circuit.

The measurement of the discharging time of each RL circuit requires two phases. Fig. 1c and d show, respectively, the state of the digital ports of the μC during the first and second phase when measuring the RL circuit that includes L_x. In the first phase (Fig. 1c), pin 1 generates a step pulse (i.e. from a digital '0' to '1', or from 0 V to the supply voltage, V_{DD}) that is synchronised with the start of the timer, pin 3 provides a digital '0' (i.e. 0 V), and pins 2 and 4 are in a high-impedance (HZ) state. This configuration results in a discharging voltage across L_x, as shown in Fig. 1b, that is monitored by pin 2. When such a discharging voltage reaches the low threshold voltage (V_{TL}) of the digital Schmitt-trigger (ST) buffer embedded into pin 2, the timer stops and a digital number with information about the length of T_x is registered. Under ideal conditions, T_x is equal to

$$T_x = \frac{L_x}{R_0} \ln\left(\frac{V_{DD}}{V_{TL}}\right) \quad (1)$$

In the second phase (Fig. 1d), pin 1 provides a digital '0', whereas the other pins do not change their state. With this configuration, the inductor current is discharged towards zero; this phase must be long enough (at least five times the discharging time constant) so as to be sure that the energy stored before in the inductor is removed. Afterwards, the circuit operates similarly for the measurement of the RL circuit that includes L_r, but pin 3 is in HZ state and pin 4 provides a digital '0'. In that case, the result is a digital number with information about the length of T_r, which is ideally equal to

$$T_r = \frac{L_r}{R_0} \ln\left(\frac{V_{DD}}{V_{TL}}\right) \quad (2)$$

After measuring T_x and T_r, the following single-point calibration technique is proposed to be applied

$$L_x^* = \frac{T_x}{T_r} L_r \quad (3)$$

where L_x^* is the estimated value of L_x. Replacing (1) and (2) in (3) yields $L_x^* = L_x$ and, hence, the estimated value has no error under ideal conditions. Furthermore, changes of temperature affecting the circuit are cancelled out by (3) whenever (i) temperature remains constant during the measurement of T_x and T_r, and (ii) the reference inductor has a low temperature coefficient. Note that changes of temperature can affect the values of R_0, V_{DD} and V_{TL} in (1) and (2), but if they do in the same way in both measurements, then such thermal effects are compensated through (3). Time drifts affecting R_0, V_{DD} and V_{TL} are also auto-calibrated by (3). In order to compensate for the temperature dependence and time drifts of L_x, the circuit would require an L_r with the same dependence. The application of a three-signal calibration technique [10,13] seems in principle unnecessary since the offset parasitic inductance (of some units of nanohenry) introduced by the circuit itself (for instance, due to the interconnections on the printed circuit board or to the bonding pad of the μC chip) is much lower than the sensor inductance (of some units or tens of millihenry).

The current consumption of the proposed circuit can be reduced by following the guidelines suggested in [22]. In the first phase, the discharging time should be measured by the timer running at high frequency (e.g. units or tens of MHz) so as to have a good timing resolution, but the CPU (Central Processing Unit) should be off whenever this does not stop the interrupt system and the timer. In the second phase, just the CPU should be on but