Contactless Transmission of Measurement Information Between Sensor and Conditioning Electronics

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***Abstract*—In many cases, contactless transmission can be a solution to applications where the measurement environment has characteristics that are unsuitable for the classical cable transmission. In this paper, the contactless transmission problem is analyzed when the distance between the readout and the sensor circuits changes. A model of the system, which considers the effect of the parasitic capacitance and the change of coupled and leakage fluxes, is proposed. A simulation of the theoretical analysis on the developed model shows that one frequency is more sensitive to the transducer capacitance and has no dependence on stray capacitance. A parameter that has very little sensitivity to sensing capacitance and high sensitivity to distance is identified. The measurement method has been tested in the laboratory, and the relative experimental results are reported.**

***Index Terms*—Biological measurements, contactless measurements, high-temperature measurements, inductive telemetry, vacuum measurements.**

I. Introduction

IN MANY applications, the sensitive element and the conditioning electronics cannot be directly connected by a wire, because the measurement environment can have unsuitable characteristics for the correct operation of the electronics. Some application examples are listed as follows:

1) measurement with high temperatures;
2) measurement in inaccessible environments like hermetic boxes;
3) measurement inside the human body;
4) measurement on rotating systems.

For some of the previous examples, traditional wireless techniques can offer aid, but they need a supply system that periodically requires maintenance or substitution. In this paper, we propose a different approach that consists of two systems: One is the sensitive element, which is passive and can have characteristics compliant with the extreme conditioning of the measurement environment, and the second one is the processing electronics hosted in an appropriate environment. The two parts are connected to each other by a magnetic field.

Recent research reports some telemetry techniques and describes prototypes for pressure measurements; these systems are used in many fields, ranging from industrial control processes [1]–[3] to biomedical systems [4]–[6]. Intraocular monitoring [6] is an example in which the sensitive element is an inductor–capacitor resonant circuit that is made by a capacitive pressure sensor in parallel with an internal planar coil. The readout circuit is positioned on a particular spectacle frame. In the field of industrial applications, wireless tire strain monitoring can be quoted; it is based on the capacitance variation due to deformation of steel wire belts of a reinforced tire [4]. These papers describe telemetry systems that consist of a planar spiral in the sensing circuit and coil inductance in the readout circuit, whose diameter is much larger than the planar one. The physical quantity is measured as a resonance frequency variation. In most cases, the configuration of the measurement system is modeled as an ideal transformer, but when the capacitance of the transducer is comparable with one of the readout circuits, the presence of the parasitic capacitances cannot be neglected, and a more accurate model has to be taken into consideration. Moreover, the distance between the two inductors (sensing and readout) can change due to practical use, inducing a variation of the resonant frequency; in this case, a complete analysis should include the effect of the coupled and leakage fluxes between the two inductors.

To give a contribution to the previous problems, this paper proposes a study of a telemetry system that consists of two planar coupled inductors. A capacitive transducer is connected to the first one, whereas an impedance analyzer is connected to the second one; the distance between the two coils can change. On the readout coil, the impedance is measured, and its plot shows three resonance frequencies. The relation of these frequencies with the system parameters has been studied, and some experimental tests have confirmed the simulation results.

In the following paragraph, we propose a model and identify each element with its physical nature. The telemetric system is modeled by two coupled inductors, including the effect of the fluxes that are related to the distance between the readout and the sensing circuit and the parasitic capacitances that influence the resonant frequencies of the system. The model has been simulated with PSpice using values of the model parameters that were deduced from a real system. Preliminary experimental results validate the theoretical considerations.

II. Model and Measurement Method

In previous papers [1]–[6], the inductive telemetric system consists of an inductance in the sensing circuit and a coil inductance in the readout circuit, with a diameter that is much

Manuscript received October 2, 2006; revised September 20, 2007.

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Digital Object Identifier 10.1109/TIM.2007.909414

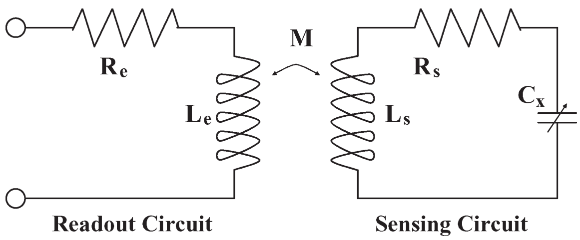

Fig. 1. Model that is commonly used to analyze telemetric systems.

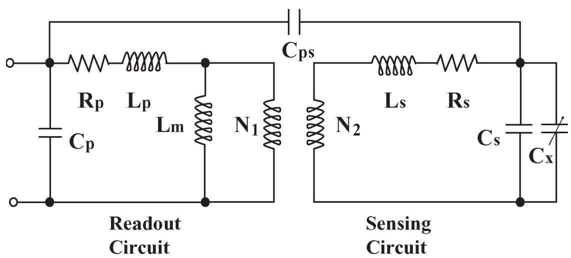

Fig. 2. Physical model of a system composed of two coupled planar inductances.

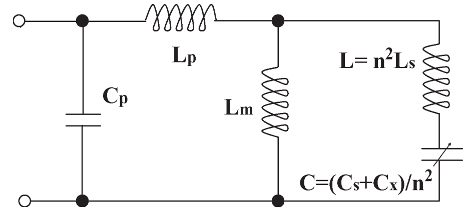

Fig. 3. Simplified model for mathematical study, with $n = N_1/N_2$.

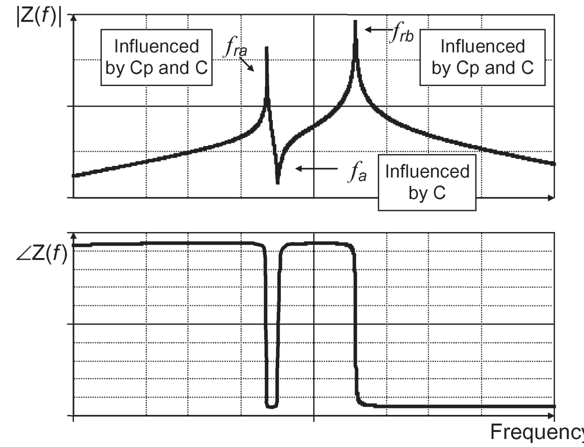

Fig. 4. Qualitative frequency response for typical parameters value of a telemetric system.

larger than the planar one. The equivalent circuit that describes the two inductances is reported in Fig. 1. The inductance is modeled with an inductor L_s, a series parasitic resistance R_s, and a variable capacitor C_x, representing the capacitance sensor. The readout circuit is modeled by an inductor L_e and a parasitic resistance R_e. When the two circuits are close, there is a mutual inductance coupling M between the inductor L_s and the inductor L_e. The impedance at the readout circuit terminals is measured, and the frequency, whose phase in a short frequency interval is at its minimum value, corresponds to the resonant frequency of the sensing circuit [1].

In some applications, this model shows some limits, e.g., to measure capacitive sensors whose values are low compared with those of the parasitic capacitances of the inductances of the sensing circuits.

A more accurate model of the system is proposed, as shown in Fig. 2: R_p and R_s are the equivalent resistances of the readout and sensing circuits; C_p and C_s are the parasitic capacitances of the readout and sensing circuits; L_p and L_s are the leakage inductances of the readout and sensing circuits; C_x is the sensing element; C_{ps} is a coupled capacitance of the sensing circuit with the readout circuit; L_m is referred to as the coupled flux; and N_1 and N_2 are the numbers of the equivalent inductor windings.

By increasing the distance between the readout and sensing circuits, the coupling capacitance C_{ps} decreases and can be neglected because it has a lower value than the other capacitors. Since the aim is the evaluation of the resonant frequencies, R_p and R_s, which are in the operating frequency range, can also be neglected. The elimination of the resistances modifies the module of the impedance close to the resonant frequencies, but the values of these frequencies remain significantly the same, allowing a simpler mathematical calculation. The major difference is in the amplitude of the module. The dissipative components tend to reduce the amplitude in the presence of high values, and then, some of the resonant frequencies could be hidden.

Neglecting C_{ps}, R_p, and R_s and reporting the secondary on the primary side, the model of Fig. 2 could be simplified into the one reported in Fig. 3, where L and C are, respectively, equal to $n^2 L_s$ and $(C_s + C_x)/n^2$, with $n = N_1/N_2$.

As seen from the terminal of the circuit of Fig. 3, the impedance is given in (1), shown at the bottom of the page, which has a qualitative frequency response shown in Fig. 4, where the presence of three resonances is evident.

In Fig. 4, the two resonances $f_{r,a}$ and $f_{r,b}$ depend both on C_p and C, whereas f_a is influenced only by C, as reported in (2) and (3), shown at the bottom of the next page. As previously reported, C_p is the parasitic capacitance of the readout circuit, whereas C is the sum of the parasitic capacitance C_s of the sensing circuit and the capacitance under measurement C_x.

Moreover, a sensitivity analysis to the sensing capacitance has been conducted by simulating, with the aid of PSpice, the proposed model whose parameters have values that are measured or estimated by two real planar inductors that constitute a real inductive telemetric system.

$$Z(s) = \frac{s^3 \left(L_m LC + L_p C(L_m + L)\right) + s(L_m + L_p)}{s^4 C_p \left(L_m LC + L_p C(L_m + L)\right) + s^2 \left(C_p(L_m + L_p) + C(L_m + L)\right) + 1} \quad (1)$$

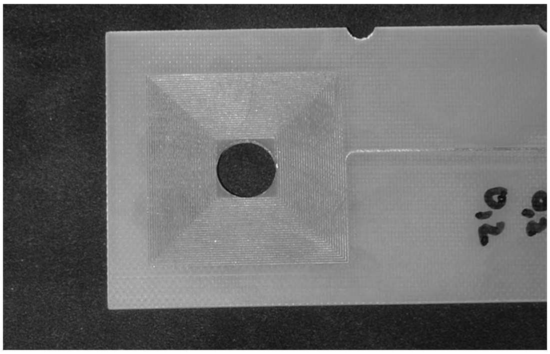

Fig. 5. Realized planar inductor.

TABLE I
VALUES OF EQUIVALENT CIRCUIT PARAMETERS FOR ONE INDUCTOR

L_{ser}	R_{ser}	C_{par}
19.9 µH	7.28 Ω	1.94 pF

III. TELEMETRIC SYSTEM AND EXPERIMENTAL APPARATUS

We have built up two equal planar inductors: one functioning as a readout circuit and the other as a sensing circuit. Each planar inductor consists of 27 square windings; the interwindings distance is 0.2 mm, and the thickness of the trace is 0.2 mm. The external area is about 35×35 mm^2, with a central hole that is 10 mm in diameter. Fig 5 shows a photograph of a single planar inductor.

A single planar inductor has been characterized. The inductor has been considered with its equivalent physical model (an inductance L_{ser}, a series resistor R_{ser}, and a parallel capacitance C_{par}) and the values of the three parameters have been measured (Table I) by an impedance analyzer (HP4194A). At the operating range of the frequency (more than 10 MHz), the impedance of R_{ser} is negligible with respect to the impedance of L_{ser}.

The next operation is to estimate the parameters of the physical model of the telemetric system reported in Fig. 2. In particular, C_{ps}, which is calculated at a distance of 10 mm between the readout and sensing circuits, can be roughly estimated to be 0.4 pF. The calculus has been made by using the formula of the ideal capacitor in which the area corresponds only to the copper surface.

TABLE II
VALUES OF EQUIVALENT CIRCUIT PARAMETERS FOR THE TWO INDUCTORS

L_m	6 µH	C_p	1.9 pF
L_p	14 µH	C_s	1.9 pF
L_s	14 µH	R_p	7.3 Ω
N	1	R_s	7.3 Ω

Table II reports the value of the parameters of the model of Fig. 2 for a distance of 5 mm between the readout and sensing circuits. The values of L_m, L_p, and L_s depend on the distance between the two inductances. Due to the difficulty of measuring such quantities, they have been estimated with the aid of an electromagnetic simulator. Each spiral has been simulated by 27 concentric circles. The ratios between the flux generated by the readout and the flux, coupled with the sensing at different distances, are reported in Table III.

The two inductors are fixed in an L-shaped support; these inductors are horizontally placed, facing each other with the two central axes being coincident. The inductor clamp is made of plastic, and the handling systems are aluminum micrometer screws. Fig. 6 shows a photograph of the supporting system. It is possible to note the presence of two micrometer screws. The vertical one regulates the distance between the two inductors from 0 to 25 mm, whereas the horizontal one is used for alignment purposes.

IV. SIMULATIONS

A sensitivity analysis has been conducted on the equivalent system of the telemetric circuit with the simulation software PSpice using the circuit proposed whose parameters are shown in Tables I and II.

Fig. 7 shows the simulation obtained from PSpice, changing C_x from 1.9 up to 3.5 pF and showing that the changes of C_x influences all the resonances but with different sensitivities. The results give 6, 4.4, and 1.8 MHz/pF for the second, first, and third resonances, respectively.

A second analysis has been done, changing the parasitic capacitance C_p. Fig. 8 shows the impedance module for C_p from 1.5 up to 2.4 pF; f_a is not influenced. Results show that the second resonant frequency is independent of C_p variation.

$$f_a = \frac{1}{2\pi\sqrt{C\left(L + \frac{L_m Lp}{L_m + Lp}\right)}} \tag{2}$$

$$\begin{aligned}(2\pi f_{r_{a,b}})^2 = {} & \left[2C\left(L + \frac{L_p L_m}{L_p + L_m}\right)\right]^{-1} + \left[2C_p\left(L_p + \frac{L L_m}{L + L_m}\right)\right]^{-1} \\ & \pm \sqrt{\left[4C^2\left(L + \frac{L_p L_m}{L_p + L_m}\right)^2\right]^{-1} - \left[4{C_p}^2\left(L_p + \frac{L L_m}{L + L_m}\right)^2\right]^{-1} - \frac{{L_m}^2 - L_p L_m - L L_p - L L_m}{2CC_p(L L_p + L L_m + L_p L_m)^2}}\end{aligned} \tag{3}$$

TABLE III
ESTIMATED FLUX RATIOS AT DIFFERENT DISTANCES BETWEEN THE TWO INDUCTANCES

Distance	*5 mm*	*10 mm*	*15 mm*	*20 mm*	*25 mm*
Flux Ratio	48%	26%	15%	9%	6%

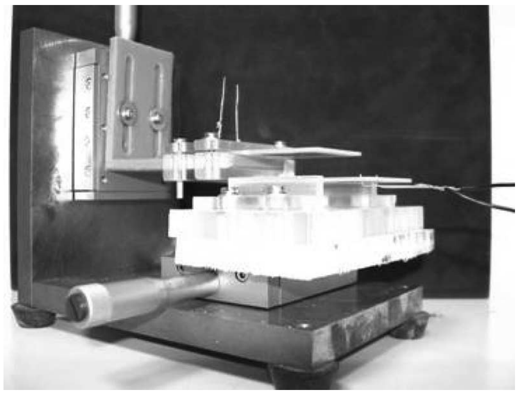

Fig. 6. Experimental system structure to obtain a measurable distance between the readout and sensing circuits.

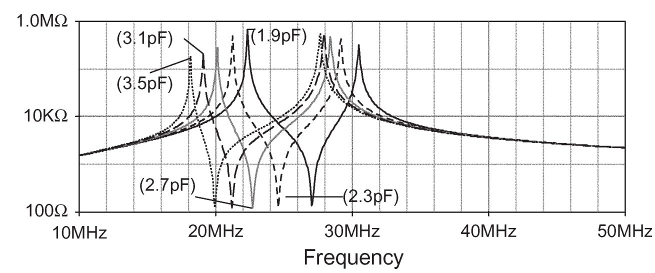

Fig. 7. Impedance module of the circuit of Fig. 3 with different values of C.

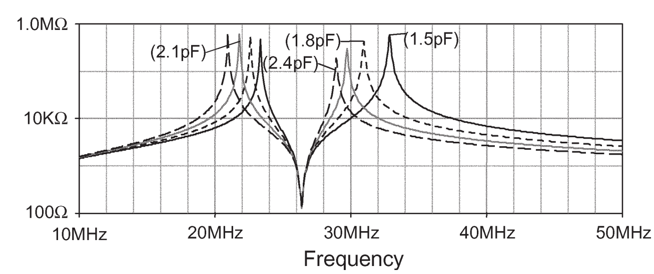

Fig. 8. Impedance module of the circuit of Fig. 3 with different values of C_p.

In a contactless system, the distance between the sensitive and readout circuits can change, as well as the magnetization and leakage fluxes. Consequently, L_p, L_m, and L_s change, as well as the previous resonance frequencies. Fig 9 reports the impedance module as seen from the readout circuit when the ratio of coupled and total fluxes change from 10% up to 50%. This range approximately corresponds to the telemetric system under analysis to a distance shift from about 5 up to 20 mm. Approximately, in this range, $f_{r,a}$ and $f_{r,b}$ can change to 400 kHz/mm.

The possibility of finding a parameter that depends only on distance has been studied. Combining the formula of the three resonances, it is possible to define an expression, called F, as shown in the following:

$$F = (2\pi f_{r\ a})^2 + (2\pi f_{r\ b})^2 - (2\pi f_a)^2. \tag{4}$$

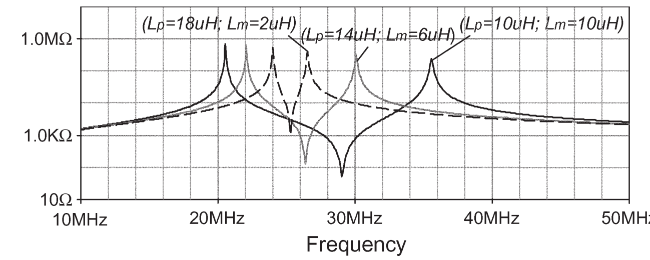

Fig. 9. Impedance module of the circuit of Fig. 3 with different values of L_m, L_p, and L_s.

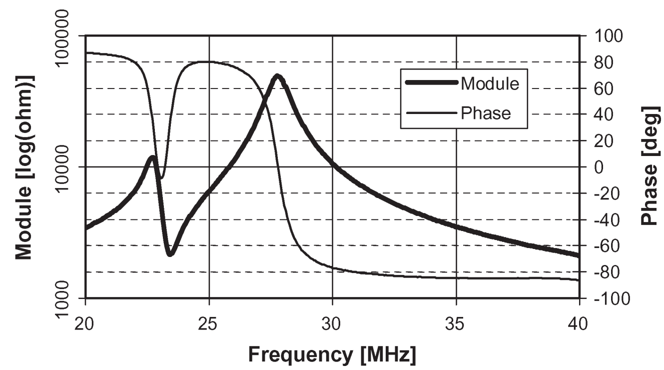

Fig. 10. Diagram of the impedance of the system measured on the readout with $C_x = 2$ pF at 25 mm.

Substituting (2) and (3) into (4), we have

$$F = \left[C_p \left(L_p + \frac{L_m L}{L_m + L}\right)\right]^{-1}. \tag{5}$$

F does not depend on C_s and C_x but only on C_p, L_p, L_m, and L. If C_p can be considered fixed, F depends only on magnetization and leakage fluxes, which change according to the geometry of the system and, in particular, to the distance between the primary and secondary coils. A method to compensate the effect induced by the distance variation is proposed and experimentally proven in [8].

V. EXPERIMENTAL RESULTS

The experimental test to validate the simulation results has been conducted on the previously described telemetric system. Using an impedance analyzer (HP4194A), the module and phase of the impedance, as seen from the readout circuit, have been measured. The three frequencies have been measured, changing the value of C_x and the distance between the two coupled planar inductances. In Fig. 10, we show the module and the phase of the impedance measured when the sensor capacitance is 2 pF and the distance from the readout circuit is 25 mm. It is possible to verify the presence of the three resonances in the frequency range predicted by the simulator.

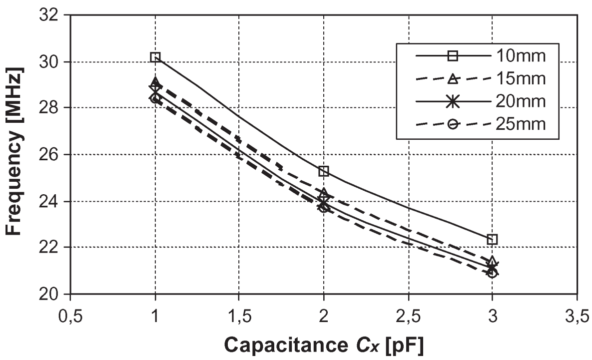

Fig. 11. Diagrams of f_a for different distances and different values of C_x.

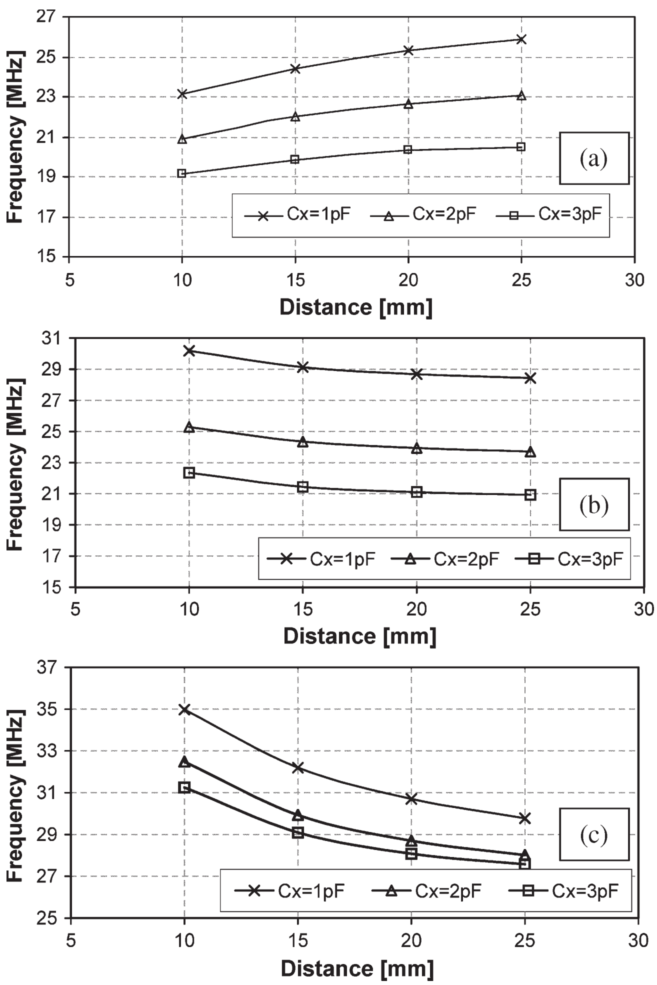

Fig. 12. Resonant frequencies (a) $f_{r,a}$ and (c) $f_{r,b}$ and (b) the frequency f_a for different values of sensing capacitance and different distances.

In Fig. 11, we can see that f_a decreases when capacitance C_x increases. However, it is also evident that the distance also changes the value of the frequency, as the results from the simulation test forecasted.

Fig. 12 shows three different graphs: one for each previously identified frequency. Notice that all of them change when the distance between the readout and sensing circuits changes. Three different values of sensor capacitance for each distance are also considered. As expected, the increase of the sensor capacitance shifts the value curves to the low frequencies.

TABLE IV
FREQUENCY VARIATION FOR DIFFERENT DISTANCES MEASURED, CHANGING C_x BY 2 pF

ΔC_x [pF]	dist. [mm]	Δf_{ra} [MHz]	Δf_a [MHz]	Δf_{rb} [MHz]
2	10	3,975	7,825	3,725
2	15	4,525	7,675	3,1
2	20	4,975	7,563	2,613
2	25	5,363	7,493	2,205

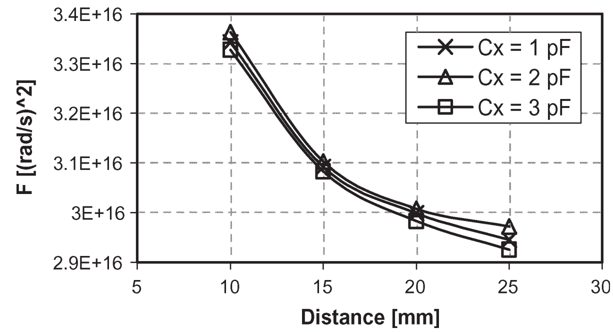

Fig. 13. Values of F measured with different C_x's and distances.

Table IV reports the three frequencies when C_x changes by 2 pF (i.e., ΔC_x). The second frequency f_a has the highest sensitivity. This is in accordance with Fig. 7. The sensitivity values obtained with the simulation also have a good correspondence with these experimental results.

Parameter F has also been calculated. For each distance value, the three frequencies have been measured, so it is possible to calculate the F values according to (4). These experimental tests are carried out with different values of C_x. Fig. 13 shows the values of parameter F as a function of distance. The results demonstrate that F mainly depends on distance, showing very little sensitivity to C_x.

VI. CONCLUSION

In this paper, a model of an inductive telemetric system considering the parasitic capacitances and the leakage fluxes has been proposed. The experimental response frequency of a real telemetric system agrees with the response frequency of the model. Three resonance frequencies $f_{r,a}$, f_a, and $f_{r,b}$ have been identified as possible measurement parameters. Comparing the frequencies, f_a is more sensitive to the measurand and is not dependent on the parasitic capacitance of the readout circuit. Furthermore, the experimental results show that f_a is more sensitive than the other two frequencies.

An analysis on the effect of the distance variation between the readout and sensing inductances highlights that all the previous three frequencies change in a sensible way. A parameter that is obtained from appropriate combinations of the three resonant frequencies measures the distance and can be used to measure the distance or to compensate the change in f_a due to the distance variation [8].

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Emilio Sardini was born in Commessaggio, Mantova, Italy, in 1958. He received the M.Sc. degree in electronic engineering from the Politecnico di Milano, Milan, Italy, in 1983.

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