Wireless implementation of high sensitivity radiofrequency probes for the dielectric characterization of biological tissues

G. Masilamany, P.-Y. Joubert
Université Paris Sud
IEF – CNRS UMR 8622
Bât. 220, 91 400 Orsay, France

S. Serfaty, B. Roucaries, P. Griesmar
Université Cergy-Pontoise
SATIE, CNRS UMR 8029, ENS Cachan
61 avenue du Président Wilson, 94 230 Cachan, France

Abstract — The paper reports on an original wireless inductive radiofrequency probe dedicated to the distant sensing of the dielectric properties of biological tissues, with non-invasive, non-contact, and easy-to-implement diagnosis of tissues in view. The sensing technique lies on the distant monitoring of a high-quality factor radiofrequency resonator electromagnetically coupled to the investigated tissue. The sensing if carried out through the measurement of the modifications of the resonator impedance with the changes of the tissue properties. Firstly, a method dedicated to the wireless measurement of the resonator impedance is proposed and discussed. Secondly, the method is implemented for the sensing of dielectric liquids standing for tissue phantoms. Preliminary sensing results show that the electrical conductivity and the permittivity of the considered phantoms can be distantly and separately sensed. These results open the way to new easy-to-implement tissue characterization techniques.

Keywords—Radiofrequency sensors, dielectric characterization, electromagnetic resonators, inductive coupling, wireless sensing, diagnosis of biological tissue, non-invasive and non-contact sensing.

I. INTRODUCTION

The accurate knowledge of the dielectric properties of organic medium such as biological tissues is a major issue for human health and well-being. Indeed, these properties are related to the physiological or pathological state of tissues, and may be used as relevant indicators, especially in the radiofrequencies (RF) and microwave bandwidths, either to detect or monitor pathologies, lesions or burns [1,2]. It has been shown that some tissue pathologies are characterized by variations of both the electrical conductivity \( \sigma \) and the dielectric permittivity \( \varepsilon \) [3]. Considering the values of these parameters in the RF bandwidth (typically \( \sigma \approx 0.2 \) S/m and \( \varepsilon \approx 40 \) at 100 MHz for human skin [1]), RF techniques appear to be particularly relevant for sensing both parameters with equal sensitivity, conversely to microwaves techniques which are more sensitive to permittivity changes [4].

Measurement techniques using open-ended coaxial probes have been proposed and implemented in the RF bandwidth to that purpose [2,4]. However due to contact issues [4], they are not likely to be easily used for in-vivo, for non-invasive or even for wearable implementations. Conversely inductive contactless RF resonators combine high sensitivity and possible non contact or distant implementations. It has been successfully used in various NMR / IRM in-vivo or ex-vivo tissue characterizations [6-8]. These accurate characterization techniques imply heavy instrumental set-ups and implementations.

In this paper, the authors aim at assessing the accuracy of dielectric characterization through the impedance measurement of a high-Q RF resonator, with contactless and easy-to-implement dielectric characterizations of biological tissues in view. To that purpose, a probe featuring a wireless flat RF resonator inductively coupled to a monitoring bobbin coil is proposed. The probe constitutes a radiating transmit and receive inductive sensor electromagnetically interacting with its direct environment (e.g. featuring a biological tissue). The dielectric properties of this environment may be sensed through the impedance changes of the resonator loaded by the tissue. In this scheme, the resonator is remotely addressed by means of a coupled RF monitoring coil, which is itself connected to an RF impedance analyzer.

The proposed measurement method is presented in section II. We firstly describe the used resonator. Then, a lumped electrical elements model of the interactions between the resonator and the monitoring coil is proposed. The estimation of the resonator impedance is carried out starting from experimental measurement data, and using the coupling model. The obtained results are used to evaluate the measurement accuracy and its optimization. In section III, the measurement technique is implemented for the sensing of liquids featuring various dielectric properties, used as biological tissue phantoms. Section IV is on conclusions.

II. WIRELESS MEASUREMENT METHOD

A. RF resonator

The resonator used in this study is a planar, 19 mm large multi-turn split-conductor transmission-line resonator (MTLR). A detailed study of such MTLR may be found in [6]. Here it is constituted of two rolled-up 1 mm width
transmission lines constituted of 35 µm thick copper tracks deposited by photolithography on each side of a 250 µm thick low loss dielectric substrate (CuFlon), as depicted in Fig.1. The obtained concentric rings are split in an unbalanced way, so as to constitute a resonator with high transmitting ability and high sensitivity to magnetic field changes. When operating in the air at an operating frequency close to the resonance frequency, the unloaded MTLR can be modeled using electrical lumped elements denoted \( L_1, R_1, \) and \( C_1 \) (Fig. 2.).

Inductance \( L_1 \) can be analytically modeled and estimated starting from the geometrical and material properties of the resonator [6]. Besides, resistance \( R_1 \) gathers radiating losses in the environment and in the dielectric substrate, as well as conductive losses within the copper. These losses may be evaluated based on the study of losses in transmission lines [9-12]. Here, since losses in the copper override other losses, \( R_1 \) may be approached using [10]:

\[
R_1 \approx \frac{4\pi l_0 K/\sqrt{\varepsilon}}{w}
\]

(2)

![Fig. 1. High-Q RF probe using multi-turn transmission line resonator (MTLR) and monitoring bobbin coil. MTLR and coil features are: \( d_{in}=19 \) mm, \( d_{out}=15 \) mm, \( t=0.25 \) mm, \( w=1 \) mm, \( d_p=8 \) mm, \( h_p=11 \) mm.](image)

where \( l \) is the total length of the turns, \( R_s \) is the sheet resistance of the strip due to skin effect, \( w \) is the width of the strip-lines, and \( K_i \) and \( K_e \) are correction factors relative to the current distribution and to the ruggedness of the copper layer, respectively.

Furthermore, the MTLR resonance condition is [6]:

\[
\frac{L_1}{\omega_0} \tan \left( \frac{\omega_0 h_p}{2c} \right) = 1
\]

(3)

where \( \omega_0 \) is the resonance angular frequency, \( L_m \) is the average length of the turns, \( c \) is the wave celerity in the vacuum, \( \varepsilon_{rd} \) is the effective relative permittivity of the strip-line constituting the turns. Finally, the capacitance \( C_1 \) may be estimated starting from the resonance angular frequency (4).

\[
C_1 = \frac{1}{\omega_0^2 l_0^2}
\]

(4)

According to the geometrical and material properties of the used MTLR and using (1) to (4), one finally gets to the estimation of the MTLR electrical lumped elements gathered in Table 1 (\( \omega_0 = 2\pi f_0 \)).

![Fig. 2. Lumped element electrical model of MTLR coupled to monitoring coil.](image)

### Table 1. Modeled and experimentally estimated parameters of the resonator.

<table>
<thead>
<tr>
<th>At ( d = 16 ) mm</th>
<th>( R_1 ) (mΩ)</th>
<th>( L_1 ) (nH)</th>
<th>( C_1 ) (pF)</th>
<th>( f_0 ) (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Model</td>
<td>703</td>
<td>109.2</td>
<td>2.48</td>
<td>306</td>
</tr>
<tr>
<td>Experimental</td>
<td>710</td>
<td>108.8</td>
<td>2.55</td>
<td>302.6</td>
</tr>
<tr>
<td>±30</td>
<td>±0.44</td>
<td>±0.06</td>
<td>±0.60</td>
<td></td>
</tr>
<tr>
<td>Deviation</td>
<td>1.0 %</td>
<td>0.4%</td>
<td>2.8%</td>
<td>1.1 %</td>
</tr>
</tbody>
</table>

**B. Wireless measurement method**

Feeding the monitoring coil with a sinusoidal “transmit” voltage \( V_T \) of angular frequency \( \omega \), a local magnetic field distribution appears within the vicinity of the coil, according to Lenz’s law. By reciprocity, this magnetic field distribution can be considered as resulting from the currents induced in the RF resonator, and as generating a “received” voltage \( V_R \) at the ends of the coil. These interactions may be represented by the electrical circuit of Fig. 2 which features the lumped elements of the resonator and of the coil [13], and the electromagnetic coupling. As a result, the impedance of the monitoring coil reads:

\[
Z_{mes} = Z_c + \frac{(\omega M)^2}{Z_1}
\]

(5)

where \( M \) is the mutual inductance between the monitoring coil and the resonator, \( Z_c \) and \( Z_l \) are the equivalent impedance of the coil and of the MTLR, respectively. The proposed measurement method aims at accurately estimating the impedance of the resonator. Indeed, this impedance is expected to change in the presence of a dielectric lossy medium such as a biological tissue (see section III). Here,
considering the probe in the air, and assuming that 
\[ Z_c = R_c + j\omega L_c \]
is the impedance of the monitoring coil, \( Z_{mes} \) actually reads:
\[
Z_{mes} = Z_c + \frac{k^2 L_1 L_3 (\omega)^2}{\frac{1}{j\omega L_3} + R_1}
\]  
where \( k \) is the coupling coefficient between the probe and the resonator so that \( M = k\sqrt{L_1 L_3} \). Knowing the impedance \( Z_c \) of the monitoring coil (\( R_c = 50 \Omega \) and \( L_c = 0.22 \mu H \)), and assuming that the coupling between the coil and the MTLR is sufficiently weak, the lumped elements of \( Z_1 \) can be estimated from the measured frequency response of \( Z_{mes} \). In practice, this frequency response is measured in a 10 MHz bandwidth centred on the MTLR resonance frequency. It is used to fit the complex polynomial frequency response of \( Z_{mes} - Z_c \) deriving from (6). The fit is carried out by means of a damped Gauss-Newton iterative minimization algorithm [14] allowing \( R_1, L_1 \) and \( C_1 \) to be estimated.

C. Experimental results and discussion

In order to evaluate the influence of the coil/MTLR coupling on the measurement accuracy, an experimental set-up was implemented in absence of lossy medium. The MTLR and the monitoring probe described in Fig. 1 are mounted on a device allowing the probe/resonator distance \( d \) to be adjusted (Fig. 3). The whole set-up is placed in a thermoregulated enclosure, and the coil is connected to an RF network analyzer (HP 4195A) operated as a reflectometric impedance analyzer.

In practice, a “\( Z_c \)” compensation procedure is carried out. It consists in measuring the frequency response of the uncoupled monitoring coil (i.e. \( Z_c \) alone) and recording it within the analyzer. Then, an internally compensation procedure \( Z_{mes} - Z_c \) is carried out so that the equivalent impedance of the resonator may be directly provided by the analyzer, for the given monitoring coil. In order to make sure that the “\( Z_c \)” compensation procedure is effective (i.e. \( Z_c \) remains unchanged in presence and in absence of the resonator) it is necessary that the coupling is weak enough so that the resulting mutual inductance is negligible compared to the impedance of the monitoring coil.

In practice, frequency responses are measured for \( d \) ranging from 0 to 16 mm. Each measurement point is carried out at least 6 times. As an illustration, the experimental and fitted
frequency responses of \(Z_{\text{mes}}-Z_i\) which are obtained for \(d = 16\) mm are plotted in a complex impedance plane (Fig. 4).

The estimation results obtained for \(R_i\) and \(L_i\) as a function of \(d\) are presented in Fig. 5, and Fig 6, respectively. One can note that fair agreements between estimated values and expected values are obtained for large distances resulting in weak couplings between the coil and the resonator. Indeed observed deviations between expected and estimated values are less than 1% for both \(R_i\) and \(L_i\) for \(d \geq 14\) mm. Estimation results for \(d = 16\) mm are gathered in Table 1. The quality factor of the MTLR is in the order of \(Q = 300\).

From these results, one may conclude that the larger the distance, the higher the measurement accuracy. However, the electromagnetic power generated within the coil environment rapidly decreases with the distance. In practice, a sufficiently high power should be transmitted from the resonator to the dielectric medium to be characterized, so as to enhance the sensitivity to its dielectric properties.

As a result, the distance \(d\) should be adjusted so as to result from a trade-off between accuracy and sensitivity to the properties of the investigated medium. In this study, distance \(d\) was empirically fixed to 8 mm, i.e. such that \(d_{\text{out}}/d = 0.5\).

III. IMPLEMENTATION OF THE RF PROBE FOR THE SENSING OF DIELECTRIC SOLUTIONS

A. Experimental set-up

In this section, the RF probe and the wireless measurement method are implemented for the sensing of dielectric lossy medium used as tissue phantoms. The experimental set-up is the same as the one described in section II with \(d = 8\) mm, except for additional glass pill-bottles containing dielectric liquids that are placed above the MTLR in the experimental device (Fig. 3). Since the pill-bottles are placed directly above the resonator, the distance between the resonator and the liquid in these experiments is fixed. It is equal to the thickness of the bottle base (3 mm). These pill-bottles are cylindrical and feature a 35 mm height and a 30 mm diameter. They contain various liquids exhibiting electrical conductivities ranging from 0 to 4 S/m and relative permittivity ranging from 18 to 80, as presented in Table 2.

B. Implementation of the wireless measurement method for loaded resonator

Placed next to the transmitting MTLR, liquid samples undergo the induction of eddy currents related to the electrical conductivity \(\sigma\) of the medium, and displacement currents related to the permittivity \(\varepsilon\) of the medium [16]. Therefore, from the viewpoint of the monitoring coil, the loaded MTLR features new dielectric losses, which are actually due to the currents induced in the lossy medium coupled with the MTLR. Also, the energy transmitted and stored in the dielectric medium may be modeled by a lumped inductance. As a result, these new features may be modeled by additive lumped element \(R_i\) and \(L_i\) in the electric model of the loaded MTLR, as depicted in Fig. 7.

![Fig. 7. Lumped element electrical model of monitoring coil coupled to MTLR loaded with tissue phantom.](image)

As established in [17] for circular loop-antennas interacting with infinite or semi-infinite dielectric media, resistance \(R_i\) stands for the additional losses due to the circulation of eddy currents induced within the lossy medium. It is therefore expected to be related to the conductivity \(\sigma\) of the medium. Furthermore, inductance \(L_i\) stands for the energy leakage due to the induced displacement currents. It is related to the permittivity \(\varepsilon\) of the medium [18].

Taking these additional lumped elements into consideration, the measured impedance now reads:

\[
Z_{\text{mes}} = Z_c + \frac{\varepsilon_k}{1 + \varepsilon_k} \frac{L_3 (L_4 + L_3) (1 + L_4)}{(1 + L_4)(1 + L_3)(1 + L_4)} (\omega)^3
\]

In order to sense the dielectric properties (\(\sigma\) and \(\varepsilon\)) of the liquid samples, it is necessary to estimate the additional lumped elements \(R_i\) and \(L_i\) of the equivalent impedance of the MTLR. To do so, frequency responses of \(Z_{\text{mes}}\) have been measured while loading the MTLR with each of the dielectric liquids presented in Table 2. Operating conditions are as described in subsection II.C. In particular, distance \(d\) is

### Table 2. Dielectric Parameters of Used Liquids

<table>
<thead>
<tr>
<th>Liquid</th>
<th>Measured electrical conductivity (S/m)</th>
<th>Relative permittivity</th>
</tr>
</thead>
<tbody>
<tr>
<td>KCl solutions</td>
<td>0.14 to 3.99 (13 values)</td>
<td>(\approx 80)</td>
</tr>
<tr>
<td>Isopropanol</td>
<td>(&lt;10^{-3})</td>
<td>17.9</td>
</tr>
<tr>
<td>Acetone</td>
<td>(&lt;10^{-3})</td>
<td>20.7</td>
</tr>
<tr>
<td>Ethanol</td>
<td>(&lt;10^{-3})</td>
<td>24.5</td>
</tr>
<tr>
<td>Methanol</td>
<td>(&lt;10^{-3})</td>
<td>30</td>
</tr>
<tr>
<td>Acetonitrile</td>
<td>(&lt;10^{-3})</td>
<td>37.5</td>
</tr>
<tr>
<td>Dimethyl Sulfoxide</td>
<td>(&lt;10^{-3})</td>
<td>46.7</td>
</tr>
</tbody>
</table>

* using a DC conductimeter
adjusted to 8 mm, each measurement point is repeated at least 6 times, and the used acquisition bandwidth is 10MHz around the resonance frequency of the loaded MTLR. The implemented compensation procedure enables $Z_c$ to be internally compensated. The coil/resonator coupling is assumed to remain unchanged, as well as $R_i$, $C_i$ and $L_i$.

The obtained frequency responses are used to fit the impedance model deriving from (7). The fitting is carried out using the same estimation algorithm as used in subsection II.B. It allows $R_i$ and $L_i$ to be estimated knowing $R_i$, $C_i$ and $L_i$.

C. Measurement and estimation results

Estimation results of $R_i$ obtained for the conductive liquids are presented in Fig. 8, and the estimation values of $L_i$ obtained for liquids featuring various permittivity values are presented in Fig. 9. One can note that a fair linear dependence between $R_i$ and the electrical conductivity $\sigma$ on one hand, and between $L_i$ and the permittivity $\varepsilon$ on the other hand, is obtained.

These results validate the ability of the proposed RF probe associated to the wireless measurement method, to sense both dielectric parameters of the considered medium.

IV. CONCLUSION

In this study, a wireless RF probe dedicated to the non contact characterization of dielectric media was presented. The probe is constituted of a high-Q MTLR distantly coupled to a monitoring coil. A distant measurement method was proposed so as to accurately evaluate the lumped elements of the MTLR electrical model. Furthermore the measurement accuracy was studied as a function of the monitoring distance.

The measurement method was then implemented in order to sense the dielectric properties of liquids used as tissue phantoms. The measurement is carried out through the changes of the complex impedance of the MTLR loaded with the different tissue phantoms. Obtained measurement results show the ability of the proposed RF probe to distantly and separately sense both the electrical conductivity and the permittivity of biological tissues.

Further works will focus on the solving of the inverse problem, which consists in estimating the dielectric properties of the investigated media starting from the experimental estimation of $R_i$ and $L_i$. For now, the obtained preliminary results open the way to the development of easy-to-implement dielectric characterization techniques of organic media. Such techniques are likely to find many applications in the field of medical applications for non-invasive characterizations (wearable sensors) or even in-vivo implementations, providing that biocompatible MTLR such as developed in [8] are considered.


