Accurate and Efficient Analytical Electrical Model of Antenna for NFC Applications

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***Abstract*— An accurate, compact, efficient analytical model at the electrical level of antennas dedicated to NFC (Near Field Communication) applications is presented in this paper. The model takes into account the skin effect, which is usually neglected in existing electrical models while it constitutes a major issue in the NFC context. The proposed model is validated with respect to finite element simulation. Comparison with most significant state-of-the-art models proves the proposed analytical model to be more accurate and efficient for antenna modeling in the NFC context.**

***Keywords*— Antenna; Near-Field Communication (NFC); analytical electrical model; skin effect**

I. Introduction

Planar antennas are widely used in many different applications: mobile phones, Radio Frequency Identification (RFID), static conversion of power electronics, NFC (Near Field Communication)… The electromagnetic phenomena induced by a current flowing in a conductor have been studied for a very long time [Ros1907], but none of the proposed models turns out to be well suited for NFC applications, which are booming with the democratization of smart phones.

In the context of NFC applications, specific physical parameters of the antenna have to be considered to develop a proper model. First of all, the conductor cross-section is rectangular. Secondly, the antenna is generally made of N rectangular-shaped turns, as illustrated in fig. 1. Thirdly, operating parameters must be carefully taken into account. In particular, NFC antennas operate at 13.56 MHz. Although the skin effect is often neglected at this frequency, simulations as well as experimental measurements show that the impact of the skin effect on NFC antenna performances at 13.56 MHz is significant enough to make such models inaccurate.

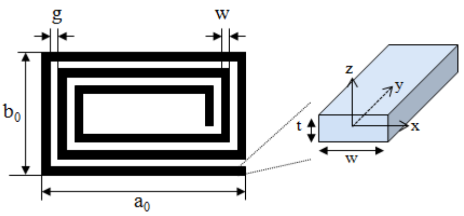

Figure 1. Top view (left) and cross-section (right) of a NFC antenna.

In addition, the context in which an analytical model is needed for NFC antennas is very specific. Indeed, the manufacturer of the NFC Integrated Circuit (IC) is usually not the same as the physical antenna manufacturer. The NFC IC manufacturer must design a tuning block with respect to the application antenna without knowing the final environment of the antenna (typically the mobile phone device). Moreover, the integrator generally uses the same NFC IC for different devices, which means that a dedicated tuning is required for each application. In such a frame of intelligent control of the tuning, a model for the NFC antenna is needed that enables fast and controlled calculations. Simulation of physical antennas with finite element models offers the required accuracy but is too much time consuming to be performed in each particular situation. As a consequence, the NFC IC manufacturer needs an efficient, accurate and compact analytical electrical model in order to perform suited tuning.

Nevertheless, in order to be efficient, the model does not have to take into account the application environment of the antenna. Indeed, when the NFC IC manufacturer designs the tuning, the final application environment is often unknown, either because the integrator prefers to keep it secret or because it is not yet finalized. In addition, the effective application environment does not depend only on the system characteristics (for instance the phone shell) but also on the environment around the system when used (typically the second device involved in the NFC communication, external magnetic perturbations…) that cannot be forecasted at the system development phase. The aim of the analytical model is to promptly provide the manufacturer with a reasonable estimation of the necessary tuning. Rather than an inaccurate environment model, an ideal environment model is more valuable as a reference. In practice, the tuning done without consideration of the application environment turns out to be quite close to the final one. Moreover, the tuning achieved without environmental perturbations gives a valuable idea of the best possible tuning in an ideal context.

In this paper, we present an analytical electrical model of antenna dedicated to NFC applications that takes into consideration the physical antenna characteristics. The paper is organized as follows. In section II, the reference model is presented. Section III summarizes a state of the art of the existing antenna models. The proposed model is developed in section IV and validated in section V. Finally, section VI concludes the paper.

II. Reference Model

Spiral planar antennas are inductors. As such, they can be represented by an equivalent lumped RLC electrical circuit in which *La* is the equivalent inductance of the antenna, *Rs* the ohmic loss resistance in series with *La* and *Cs* the parasitic series feed-forward capacitance in parallel with *La* and *Rs*. The aim of an electrical model of the NFC antenna is to derive the electrical parameters *La*, *Rs* and *Cs*, which are frequency-dependent, from the structural antenna characteristics. The electrical parameters can be estimated either by the computation of an analytical expression or by extraction from a finite element simulation.

Our objective is to propose an accurate analytical model for the estimation of the electrical parameters of the lumped equivalent circuit. The results are confronted to finite element simulation computed using EM Pro simulator. The advantages of such simulation-extracted parameters are twofold: firstly, they are perfectly reproducible; secondly, they are completely independent of environmental perturbations. On the contrary, experimental measurements on real antennas cannot be trusted as reliable references given that they are not reproducible and are significantly influenced by the environmental conditions. As a consequence, we will use only Finite Element Method (FEM) simulation results from EM Pro as a reference for validation.

III. STATE OF THE ART

In the literature, several lumped electrical models of antennas can be found. In most cases, the available models deal with antennas that exhibit circular section and circular turns. Fewer publications refer to rectangular section conductors and/or rectangular antenna turns. Existing analytical models are either not accurate enough or too complex in terms of calculation with respect to our objectives and context. We briefly review in this section the main contributions related to analytical electrical models and remind the basic principles that will be used in the proposed model.

The antenna parameters are frequency-dependent, especially the series resistance. This is due to two main and complementary effects: the skin and the proximity effects. Usually, these effects are not taken into account or are roughly evaluated in NFC systems because of the relatively low operating frequency. It is true that for antenna with large inter-spire like in our application, the proximity effect is negligible, but skin effect has to be estimated to be able to define a viable analytical model of NFC antenna at 13.56 MHz.

The skin effect is due to internal feedback mechanism inside a conductor when the electrical current varies. When an alternating current flows in a conductor, the magnetic field is induced in the conductor and modifies the current distribution within the conductor. As a result, the current tends to flow near the surface and the current density decreases from the surface to the center of the conductor. Generally, we define a skin effect depth δ by the following equation:

$$\delta = \frac{1}{\sqrt{\pi . f \cdot \sigma \cdot \mu}} \tag{1}$$

where σ, μ are respectively the conductivity and the permeability of the conductor and f the frequency of the signal. In the NFC context, the skin depth δ is of the same order of magnitude as half the thickness of the conductor *t/2*. The skin effect has been widely studied for conductor with circular cross-section, especially for high current in power electronics [Nan04] or for high frequency applications [Mer10].

For planar antenna, the section being rectangular, the above references for the impact of the skin effect loss on the resistance are not applicable. In the case of rectangular cross-section conductor, a few studies of the skin effect loss are available in the literature. One of the classical approaches for skin effect estimation considers that the conductor is close to a conducting carrier (masse plan or silicon substrate) [Zol07]. In this context, the skin effect appears only on one side of the conductor. Unfortunately, when the two sides are likely to be affected by the skin effect, an accurate estimation of the skin effect is much more complex. In [Yao13 and Eo93], an analytical model based on a set of experimental measurements is proposed. Unfortunately, the accuracy of this empirical model is very sensitive to the environmental conditions. This model is not accurate in the case of our application as we show in section V.

Some solutions are proposed in the case of UHF RFID (Ultra High Frequency RFID). For these kinds of applications, the skin effect is taken into account; unfortunately to ease its estimation the proposed solutions split the problem into two parts: *t/2*<<δ and *t/2*>>δ. Different models are given for both cases and no valid model exists for the in-between range, which is precisely the range of concern in the NFC context as mentioned before. Sometimes the assumption on the ratio between skin depth and conductor thickness is not explicit in the paper, like in [Mer10], where the resistance due to skin effect on the two opposite sides of the conductor is looked upon as two parallel resistances, but it is only true if the current density is close to zero in the middle of the conductor (i.e. if *t/2*>>δ).

To conclude, there is a need of an accurate analytical electrical model of antenna dedicated to NFC applications.

IV. ANALYTICAL MODEL DEVELOPMENT

In this section, we develop the proposed analytical electrical lumped model of NFC antenna: firstly the series resistance *Rs*, secondly the inductance *La* and thirdly the parasitic capacitance *Cs*.

A. Series Resistance Rs

Our objective is to estimate the exact contribution of the skin effect on the equivalent impedance of the conductor. The real part of this impedance leads to the series resistance *Rs*.

In using the Maxwell's equations and the Ohm's law we can obtain the following equation for the diffusion of the current density:

$$\Delta\vec{J} - \mu . \sigma . \frac{\delta\vec{J}}{\delta t} = \vec{0} \tag{2}$$

We assume that the current density only depends on z (fig. 1). This means that the current density is uniform in the x-y plan. Along the *z*-axis, the diffusion equation of *J* is:

$$\frac{\partial^2 J(t,z)}{\partial z^2} - \sigma\mu \frac{\partial J(t,z)}{\partial t} = 0 \tag{3}$$

Using the time domain Laplace transformation, the general solution of (3) can be written as:

$$J(p,z) = J1(p) \cdot e^{-kz} + J2(p) \cdot e^{kz} \tag{4}$$

With $k^2 = \sigma \cdot \mu \cdot p$.

Considering the boundary conditions, we find:

$$J\left(p, \frac{t}{2}\right) = J\left(p, -\frac{t}{2}\right) = Jo(p) \tag{5}$$

This implies:

$$J1(p) = J2(p) = \frac{Jo(p)}{\left(e^{-k\frac{t}{2}}+e^{k\frac{t}{2}}\right)} \quad (6)$$

Finally, the current density variation in the rectangular conductor due to skin effect is given by (7):

$$J(p,z) = Jo(p)\frac{e^{-\sqrt{\sigma\mu p}z}+e^{\sqrt{\sigma\mu p}z}}{\left(e^{-\sqrt{\sigma\mu p}\frac{t}{2}}+e^{\sqrt{\sigma\mu p}\frac{t}{2}}\right)} = Jo(p)\frac{cosh(z\cdot\sqrt{\sigma\cdot\mu\cdot p})}{cosh\left(\frac{t}{2}\cdot\sqrt{\sigma\cdot\mu\cdot p}\right)} \quad (7)$$

Because the current density depends only on z, it can be expressed as:

$$I(p) = w\int_{-\frac{t}{2}}^{\frac{t}{2}} J(p,z)\cdot dz = \frac{2\cdot w\cdot Jo(p)}{\sqrt{\sigma\cdot\mu\cdot p}}tanh\left(\frac{t}{2}\cdot\sqrt{\sigma\cdot\mu\cdot p}\right) \quad (8)$$

Antenna voltage can be computed by integration of the electric field on the length of the conductor *l*.

$$V(p) = \int_0^l E\left(p,\frac{t}{2}\right)\cdot dy = l\cdot E\left(p,\frac{t}{2}\right) = \frac{Jo(p)}{\sigma}\cdot l \quad (9)$$

Finally, we take the real part of the internal impedance to obtain the series resistance:

$$Rs(f) = \mathrm{Re}\left\{\frac{l\cdot\sqrt{\sigma\cdot\mu\cdot p}}{2\cdot w\cdot\sigma\cdot\tanh\left(\frac{t}{2}\cdot\sqrt{\sigma\cdot\mu\cdot p}\right)}\right\} = \frac{l}{2\cdot w\cdot\sigma\cdot\delta}\frac{\sinh\left(\frac{t}{\delta}\right)+\sin\left(\frac{t}{\delta}\right)}{\cosh\left(\frac{t}{\delta}\right)-\cos\left(\frac{t}{\delta}\right)} \quad (10)$$

B. Inductance La

Our objective is to define an accurate analytical expression of the inductance of the multi-turn rectangular antenna. In [You03], a complete development is presented to estimate this inductance. This solution is accurate but difficult to apply because the entire complex computations have to be done for each new antenna layout. Conversely, [Zol07] provides a fast analytical expression of *La*, but this expression is not accurate enough as far as it is based on an extended expression of a multi-turn circular antenna.

In our context, we want to define a model that is accurate but that involves a simple analytical expression. The key idea consists in converting the N-turn rectangular antenna into an equivalent single-turn rectangular antenna. This allows us to preserve the specific features of a rectangular antenna while simplifying the computation. Given that the cross-section shape has a negligible influence on the inductance, the rectangular cross-section conductor is approximated by a circular cross-section of equal area $\left(d = \sqrt{4.w.t/\pi}\right)$. This single-turn antenna has the following new geometrical properties:

$$a_{avg} = a_0 - N.w - (N-1).g \quad (11)$$

$$b_{avg} = b_0 - N.w - (N-1).g \quad (12)$$

The total inductance is the sum of internal and external inductances.

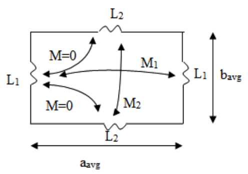

Figure 2. Partial and partial mutual inductances of a single-turn antenna

The external inductance consists of the contribution of partial inductances *L1* and *L2* and partial mutual inductances *M1* and *M2* between all parallel conductor parts due to external magnetic field as illustrated in figure 2.

$$L_{external} = 2(L1 - M1) + 2(L2 - M2) \quad (13)$$

The above calculation of the external based on self and mutual inductances is not new in the spirit, but the detailed calculation proposed here is original. We obtain the following expressions:

$$L1 = \frac{\mu o.}{2\pi}b_{avg}\left[\ln\left(\frac{2b_{avg}}{d} + \sqrt{\left(\frac{2b_{avg}}{d}\right)^2 + 1}\right) - \sqrt{\left(\frac{d}{2b_{avg}}\right)^2 + 1} + \left(\frac{d}{2b_{avg}}\right)^2\right]$$

$$\approx \frac{\mu o.}{2\pi}b_{avg}\left(\ln\left(\frac{4b_{avg}}{d}\right) - 1\right) \quad (14)$$

In practice, the length of the antenna segment b_{avg} is larger than the approximated wire radius *d*. Therefore, we have the following approximations:

$$M1 = \frac{\mu o.}{2\pi}b_{avg}\left[\ln\left(\frac{2b_{avg}}{2a_{acg} + d} + \sqrt{\left(\frac{2b_{avg}}{2a_{avg} + d}\right)^2 + 1}\right) - \sqrt{\left(\frac{2a_{avg} + d}{2b_{avg}}\right)^2 + 1} + \left(\frac{2a_{avg} + d}{2b_{avg}}\right)^2\right]$$

$$\approx \frac{\mu o.}{2\pi}b_{avg}\left[\ln\left(\frac{b_{avg}}{a_{avg}} + \sqrt{\left(\frac{b_{avg}}{a_{avg}}\right)^2 + 1}\right) - \sqrt{\left(\frac{a_{avg}}{b_{avg}}\right)^2 + 1} + \left(\frac{a_{avg}}{b_{avg}}\right)^2\right] \quad (15)$$

L2 and *M2* have similar expressions by swapping b_{avg} with a_{avg} in the equations of *L1* and *M1*.

The internal inductance is due to magnetic variations in the conductor. It strongly depends on the frequency due to skin effect. This inductance is directly derived from the imaginary part of the internal antenna impedance obtained from equation (8) and equation (9):

$$L_{internal} = \frac{1}{\omega}.Im\left\{\frac{l\cdot\sqrt{\sigma\cdot\mu\cdot p}}{2\cdot w\cdot\sigma\cdot tanh\left(\frac{t}{2}\sqrt{\sigma\cdot\mu\cdot p}\right)}\right\} = \frac{l}{2\cdot w\cdot\sigma\cdot\delta}\frac{sinh\left(\frac{t}{\delta}\right)-sin\left(\frac{t}{\delta}\right)}{cosh\left(\frac{t}{\delta}\right)-cos\left(\frac{t}{\delta}\right)} \quad (16)$$

Based on this estimation of the total inductance of a single-turn antenna, we take into account the effective number of turns *N* by multiplying the sum of the external and internal inductances by a factor *N* to the power of *E*:

$$La = (L_{external} + L_{internal})\,N^E \quad (17)$$

The coefficient *E* is estimated empirically from a large set of simulations. In our application where the proximity effect can be neglected, this coefficient is constant and $E = 1.5$.

C. Parasitic Capacitor Cs

For multi-turn antenna, adjacent conductors create capacitors. The resulting parasitic capacitance *Cs* is expressed as [Zol07]:

$$C_s = \frac{\pi\cdot\varepsilon_0\,\varepsilon_r\cdot l_g}{ln\left(\frac{\pi\cdot g}{w+t}+1\right)} \quad (18)$$

where ε_0 and ε_r are respectively the permittivity of the vacuum and the relative permittivity of the inter-spire medium. The parameter l_g is the total length of the conductor facing an adjacent conductor. This parameter is slightly shorter than the length of the antenna and can be obtained from:

$$lg = (N-1)\cdot[2\cdot(a_0 + b_0) - 4N\cdot(w+g)] \quad (19)$$

V. VALIDATION

For the validation of the proposed model, we simulate different antenna topologies with parameters given in Table I. Three of these antennas have 3 turns and one has 5 turns. The inter-spire *g* is large enough to neglect the proximity effect.

TABLE I. GEOMETRICAL PARAMETERS OF CONSIDERED ANTENNAS

		Antenna A	Antenna B	Antenna C	Antenna D
Antenna geometry	a_0 x b_0 (mm)	20x30	20x30	72x42	72x42
	w (mm)	0.1	0.2	0.1	0.1
	g (cm)	0.35	0.33	0.65	0.55
	t (mm)	0.035	0.035	0.035	0.035
	N	3	3	3	5

Table II summarizes the values of the RLC model parameters obtained from the analytical expressions described in this paper (Proposal) in comparison with values extracted from the FEM simulation and with parameter estimation proposed in published papers. The parameter values are all estimated at the specific operating frequency of 13.56 MHz characteristic of NFC applications.

TABLE II. VALUES OF RLC MODEL PARAMETERS AT 13.56 MHZ COMPARISON WITH THE FEM SIMULATION AND PUBLISHED PAPERS

		Antenna A	Antenne B	Antenne C	Antenne D
Rs (Ω)	FEM	1.2087	0.7163	2.8175	3.9923
	[Yao13]	1.7554	1.2086	4.3231	5.7798
	[Zol07]	2.3699	1.4202	5.8362	7.8027
	[Mer10]	1.6343	0.9155	4.0287	5.3862
	Proposal	1.1258	0.6263	2.7768	3.7124
La (µH)	FEM	0.305	0.2584	0.8436	1.2772
	[You03]	0.3028	0.2565	0.8628	X
	Proposal	0.3039	0.2654	0.8736	1.3043
Cs (pF)	Proposal	0.8735	1.0159	1.9398	3.1577

Concerning the model inductance *La*, we compare the value obtained from our analytical expression with exact computation from [You03] and value extracted from FEM simulation. The exact computation is slightly more accurate than our estimation but it is much more complicated to implement: for 3-turn antennas, the computation requires to solve 12 equations; in the case of 5-turn antennas, this approach becomes prohibitive in terms of computational effort. Moreover, our direct estimation is very close to FEM extraction (estimation error < 4%). Actually, the impact of the skin effect, which creates the internal inductance, on the total inductance of the antenna is not significant. As a consequence, our analytical model can be simplified considering only the external inductance given by equation (12).

Regarding the estimation of ohmic loss represented by the series resistance *Rs*, our analytical model proposal is the only viable solution for an accurate estimation of *Rs* value at 13.56 MHz. Figure 3 gives the resistance value versus frequency in the case of antenna *A* for the five estimation approaches considered in Table II.

The computations from [Zol07] and [Yao13] significantly overestimate the value of the series resistance. For [Zol07], it is due to the approximation of the skin effect on only one side of the conductor. In [Mer10] the empirical model proposed is not adapted to NFC application. The estimation proposed in this paper converges toward our solution for high frequencies because their assumption that the skin effect loss on two sides of the conductor can be considered as two resistances in parallel becomes true when the skin depth is very small. But in our NFC application, the model must be accurate at the 13.56 MHz operating frequency. In this case, our model is the only one to provide an estimation error inferior to 4%.

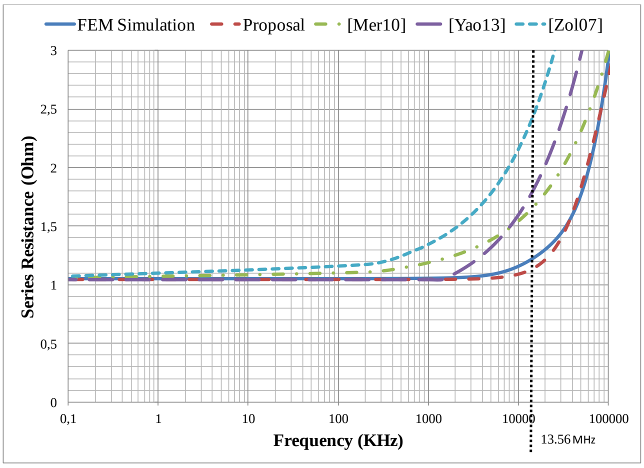

Figure 3. Series resistance *Rs* vs. frequency

VI. CONCLUSION

To respond to the established need from NFC IC designers of an accurate, compact and efficient model at the electrical level of antennas dedicated to NFC applications, we have developed the analytical lumped model presented in this paper. Comparison with finite element simulation has proven this model to be very accurate at the NFC operating frequency, taking properly into account the skin effect. Our model is more accurate for NFC applications than existing models, especially regarding the ohmic loss estimation. It is also more efficient in terms of computational efforts than exact computations concerning inductance estimation.

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