Modeling Inductive Coupling for Wireless Power Transfer to Integrated Circuits

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***Abstract*—With the trend of portable electronics to go battery-less, Inductive Coupling Wireless Power Transfer (ICW Power Transfer) is becoming commonplace. Inductive Coupling has shown to be a good technique for proximity and wireless power transfer in general, because it permits the easy use of impedance matching and resonance circuits. An important potential application is powering Integrated Circuits (ICs) from a PCB (Printed Circuit Board) without using conductive pins. While using ferromagnetic core materials improves coupling, conductive non-ferromagnetic ones like most IC's substrates, decrease coupling and ICW Power Transfer performance. In this paper we developed a simple theoretical model for ICW power transfer and compared it with 3D Electromagnetic simulations of an inductive link system to ICs. The results show that enough power can be supplied to very low power consumption ICs. As this technique can be also used to perform wireless communications, it opens the possibility to design ICs without pins at all.**

***Keywords*: PCB - *Printed Circuit Board*, IC - *Integrated Circuit*, ICW Power Transfer - *Inductive Coupling Wireless Power Transfer*, Near Field, EM - *Electromagnetic Fields*, Loop Wire, Range...**

I. Introduction

Inductive coupling transformer is being used for a long time mainly to transfer power with galvanic isolation. Nowadays, it is also used in energy harvesting applications such as inductive cellphone chargers and medical implants [1]. It should be noted that this near field scheme used to transfer energy is safer to humans than a far field solution, because it is more confined in space and the usage of lower frequencies (when compared with radiating solutions). [2] points out the possibility of powering CMOS ICs, but it refers to an old high resistivity substrate process that is no longer effective (due to latch-up and yield problems) and in other cases [3] special and expensive technologies are used.

If a similar technique can be used to power todays conventional CMOS ICs, that would potentially reduce their price as no pins and interconnections were needed, would easy board assembly as ICs were no more soldered to PCBs (allowing a manufacture process with only low temperatures), would provide galvanic isolation by default, among many other foreseeable advantages.

The practical application is represented in Fig. 1, where we could see the IC glued on top of a printed circuit board with an internal receiver inductor. On the bottom of the Printed Circuit Board (PCB) there is a transmitter inductor while inside the chip will lay the receiving counterpart (probably the PCB inductor can be made large enough so only one was needed per board). However, transferring Energy to ICs by Inductive Coupling is a quite dificult task (compared to a non-integrated application), because only a small amount of space (small chip area) can be used and the conductive substrate of traditional CMOS technologies acts as a lossy material which degrades the performance of the receiver inductor. This effect is well documented in reducing the quality factor of inductors in RF applications [4]. Similarly in the application described here causes a reduction of the received voltage. In order to verify how much these constraints can or cannot hinder our goal, we developed an inductive coupling model (presented in Section II), we checked it against 3D electromagnetic simulations, including a realistic scenario similar to Fig. 1 (Section III) and finally the paper ends in Section IV drawing some conclusions.

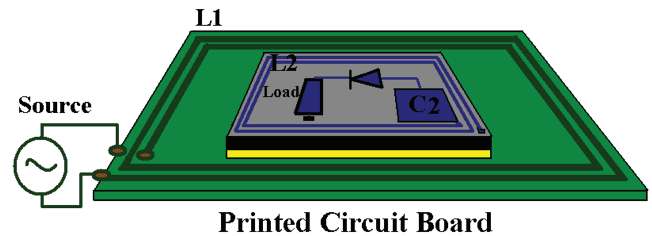

Fig. 1: ICW Power Transfer from PCB to ICs illustration.

II. Modeling Inductive Coupling

Inductive Coupling link could be seen as an transmitter inductor and a receive inductor transferring energy by common variable magnetic field lines. By Lenz Law ($\varepsilon = -N_e \cdot \frac{d\Phi_B}{dt}$), a variable magnetic flux produced by the transmitter inductor will cause an electromagnetic force in the second inductor. The mutual inductance (M) is unique and is related to the self inductances and the coupling factor (that is the ratio of inductors magnetic flux common to both inductors by the total magnetic flux K_Φ [5]).

$$M = K_\Phi \cdot \sqrt{L_1 \cdot L_2} \tag{1}$$

In the case of an ideal transformer, as the coupling factor is one (the total magnetic flux is the same in both inductors), the Lenz law implies that the voltage at the receive inductor will be related to the voltage at transmitter inductor by a constant factor (transformation ratio - N). When we take a current from the receiver inductor, a related current will circulate in the transmitter inductor (making an ideal power transfer link). Again, the unity coupling factor implies that when we take a current from the receiver inductor, the change in magnetic flux will be compensated in the transmitter inductor by the input current, so that the common magnetic flux remains the same.

But in our scenario the coupling factor will be far from unity, therefore we should start from the mutual inductance generic circuit of the Fig. 2, where the secondary (or receiver) inductor is made the series of two inductors in a way that the central equivalent circuit is an ideal transformer (satisfy $M = \sqrt{L_1 \cdot L_{2tf}} = K_\Phi \cdot \sqrt{L_1 \cdot L_2}$). L_{2tf} is the equivalent ideal transformer secondary inductor, and $Lout$ is the equivalent output inductance added to ideal transformer model to make the real one (low coupling factor). In this case the following relations apply:

$$L_2 = {K_\Phi}^2 \cdot L_2 + (1 - {K_\Phi}^2) \cdot L_2 \tag{2}$$

$$L_{2tf} = {K_\Phi}^2 \cdot L_2 \tag{3}$$

$$L_{out} = (1 - {K_\Phi}^2) \cdot L_2 \tag{4}$$

$$N = \frac{M}{L_1} = \frac{{K_\Phi}^2 \cdot L_2}{M} = K_\Phi \cdot \sqrt{\frac{L_2}{L_1}} \tag{5}$$

To obtain a simple first order model, some simplifications will be made: the source inductor (magnetic dipole) is perfectly aligned with the receiver inductor (as well their geometric centers) and the distance is large compared to the dimensions of the inductors (see Fig. 3). This last is not what what is depicted in Fig. 1 but simplifies the problem and gives accurate results for the case where the receiver coil is much smaller than the transmitter one. This comes out from the simplification made in magnetic induction mathematical integration (as the reader could see later in (9), we consider only the central magnetic field value for our model aproximation).

From (5) becomes clear that most of the work is calculating K_Φ. Note that for low coupling factors $L_{out} \simeq L_2$. Therefore taking Fig. 3 in consideration:

$$M^2 = K_\Phi^2 \cdot (L_1 \cdot L_2) \tag{6}$$

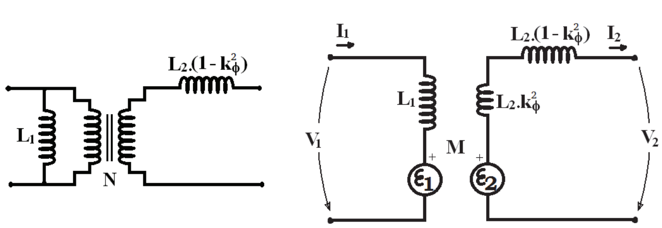

Fig. 2: Generic transformer model.

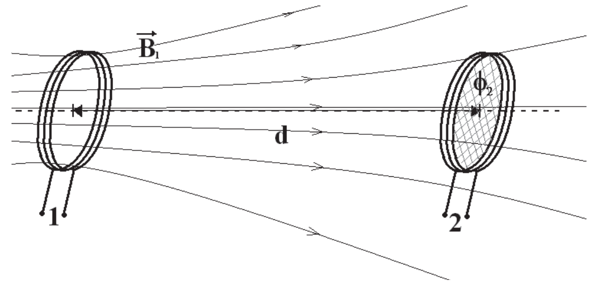

Fig. 3: Perfectly Aligned Inductive Coupling Link.

$$K_\Phi = \frac{\Phi_2}{\Phi_1} \tag{7}$$

where Φ_1 and Φ_2 are the total magnetic flux of the transmitter inductor (inductor 1) and the part of magnetic flux common with receiver inductor (inductor 2), respectively.

$$N_1 \cdot \Phi_1 = L_1 \cdot I_1 \tag{8}$$

$$\Phi_2 = A_2 \cdot B_2 \tag{9}$$

with B_2 being the magnetic field produced by transmitter sensed by receiver, $N1$ the number of turns of the transmitter inductor and $A2$ the area of the receiver inductor. For closer distances, B_2 varies substantially along the receiver inductor area and the magnetic flux should be performed by mathematical integration.

The magnetic field produced by transmitter in the receiver is [6]:

$$B_2 = \frac{\mu_0}{2} \cdot \frac{r_1^2}{(d^2 + {r_1}^2)^{\frac{3}{2}}} \cdot N_1 I_1 \tag{10}$$

with μ_0 being the vacuum magnetic permeability, r_1, N_1, I_1 being the transmitter inductor radius, number of wire loops, transmitter inductor current and d the distance between the transmitter and receiver inductors. The coupling factor is then:

$$K_\Phi = \frac{\mu_0}{2} \cdot \frac{N_1^2}{L_1} \cdot \frac{r_1^2 \cdot A_2}{(d^2 + {r_1}^2)^{\frac{3}{2}}} \tag{11}$$

The situation will also be slightly more complicated, because planar inductors have a different inner and an outer radius - therefore we should use the effective radius and effective inductor area values. Also, as already said, if a conductive material is added somewhere near the coils it will reduce the magnetic field in receiver (as well the coupling factor). On the other hand, if a ferromagnetic material is used instead, we expect to have concentration of magnetic field lines and to increase the coupling factor. We will introduce K_{core} into the model to take into account the effect of the core of inductors, but it should be noted that the coupling factor is restricted to one!

$$K_\Phi = \frac{\mu_0}{2} \cdot \frac{N_1^2}{L_1} \cdot \frac{r_1^2 \cdot A_2}{(d^2 + {r_1}^2)^{\frac{3}{2}}} \cdot K_{core}(f, d) \tag{12}$$

The equivelent transformer ratio (or voltage gain) is:

$$N = \sqrt{\frac{L_2}{L_1}} \cdot \frac{\mu_0}{2} \cdot \frac{N_1^2}{L_1} \cdot \frac{r_1^2 \cdot A_2}{(d^2 + {r_1}^2)^{\frac{3}{2}}} \cdot K_{core}(f, d) \qquad (13)$$

As the receiver's electronics concerns, it is optimum to have a voltage level as high as the IC technology allows. This tends to favor higher relation $\frac{L_2}{L_1}$ and a higher area at the receiver side, which is unfortunately bounded by silicon costs (changing the receiver's side does not change the factor $\frac{N_1^2 \cdot r_1^2}{L_1}$ of (13)). This happens because magnetic induction in conductive (non-ferromagnetic) core will reduce more and more the surounding magnetic field with frequency, K_{core} will decrease with the frequency, and lower frequencies will correspond to greater turns ratio.

Although it is possible to convert ac/dc with more or less sophisticated circuits, all inductive harvesting circuits have an equivalent model to that of Fig. 4:

$$Vo = Vi \cdot N \qquad (14)$$

$$L_o = (1 - {K_\Phi}^2) \cdot L_2 \qquad (15)$$

$$R_{po} = 2\pi \cdot L_o \cdot f_r \cdot Q_{L_o} \qquad (16)$$

$$C_o = \frac{1}{(2\pi f_r)^2 \cdot L_o} \qquad (17)$$

Maximum transfer power happens at resonance (f_r) and when $R_L = R_{po}$. We have to adjust the variables to reach a minimum voltage that we arbitrary set to $1.4V_p$. We can change the distance a little (d), the inductor parameters (N_1, r_1, N_2, A_2, Q_1, Q_2) and change or put some material (for example ferromagnetic glue to bond the IC to the PCB) between the primary and secondary (increase K_{core}). It's interesting to note, that for the same N and Q_{L_o} (the same output voltage), more power could be transfered with lower frequencies (lower equivalent R_{po})!

III. Electromagnetic 3-D Simulations

Ansoft HFSS [7] was used to make 3D electromagnetic simulations. Lumped Ports were used and the important values to the circuit model of Fig. 4 were extracted from the Z-matrix. The first set of simulations were done with the large setup

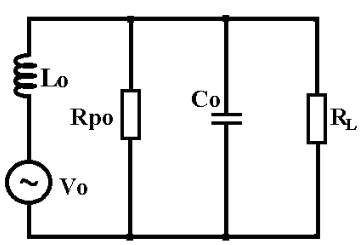

Fig. 4: Inductive-based harvest equivalent circuit.

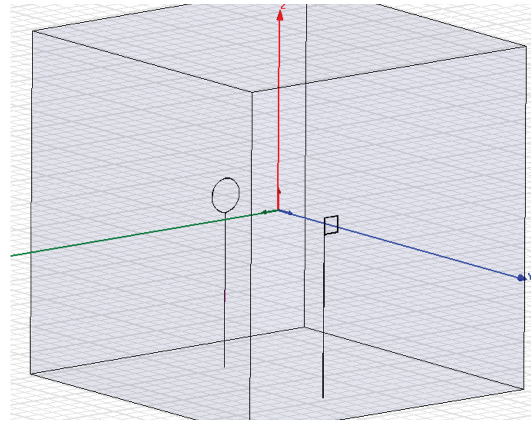

Fig. 5: Simulation layout for testing the model accuracy.

depicted in Fig. 5 (1MHz frequency, a round 1-turn L_1 with a diameter of 20cm and a square 3-turns L_2 with a 10cm of diameter) to check the accuracy of theoretical model. As we could see from Fig. 6, the simulated and predicted transformer ratio are consistent and agree very well. As the coupling factor is low, Fig. 7 presents correctly a L_{out} approximately constant and equal to receiver's inductance.

The main goal of this work is to show that transferring energy to ICs by inductive coupling is feasible. Fig. 8 shows our setup: a $1mm$ thickness FR4 material PCB, with an inductor L_1 on the bottom side and a receiver chip glued with a bonding material on the top side. The parameters used are based upon the AMS 0.35μ IC CMOS technology [8].

From Fig. 9 it is possible to verify that once more, the model and the simulation results are within 7dB for the 50

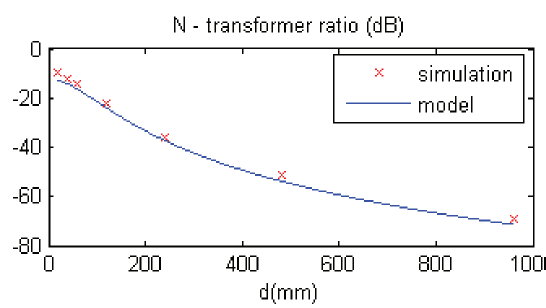

Fig. 6: Simulation results: L_1 to L_2 equivalent N.

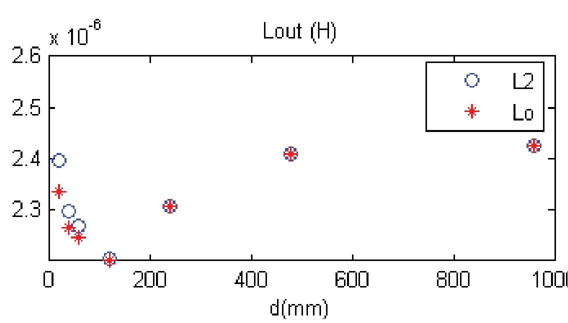

Fig. 7: Simulation results: L_1 to L_2 equivalent L_o.

and 250 MHz cases. For the 550MHz, as expected, radiation, capacitive coupling, high frequency losses in materials and other phenomena become relevant, and the results diverge (actually we expected that to happen already at 250MHz and even at lower frequencies). We checked the L1 value against the simplified formulas given in [9] and a difference less than 3% is verified. Extracting the parameters of the model of Fig.4 and considering $V_i = 14Vp$ we obtained: $f_r = 50MHz$; $N = -40dB$; $Q_{L_o} = 20$; $L_o = 100nH$. For a $C_o = 100pF$ and $R_{po} = 628\Omega$. It is possible to extract $1.5mW$ of power (with $1.4V_p$ voltage level), which is a promising result.

IV. Conclusions

We proposed an Inductive Coupling Wireless Power transfer to power CMOS ICs. We developed a simple model for calculating the equivalent turns ratio (N) of two coils weakly coupled. The model predicts with an error less than 7dB (having the simulation results as reference) the equivalent turns ratio (N).

For a realistic distance of $3mm$ between the inductors and $50MHz$ frequency it was possible to harvest energy with a $1.5mW$ of power. But, by making fine adjustments in receiver's coil we could reach the goal of transferring more power (tens of milliwatts) which is enough to many low-power ICs. Improvements can be achieved by selectively put some ferromagnetic material in the bonding process to the PCB. Anyhow the performance already pointed by model/simulations justify further studies and the fabrication of a test chip, so this technique along with wireless data transmission can lead to the practical design and manufacture of integrated circuits without any pins!

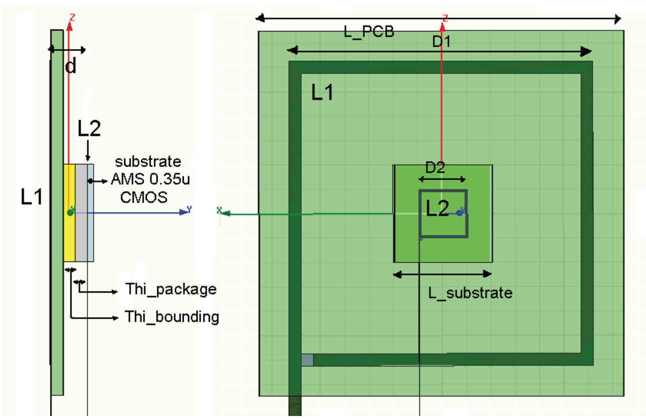

Fig. 8: ICW power transfer from PCB to IC: simulation setup

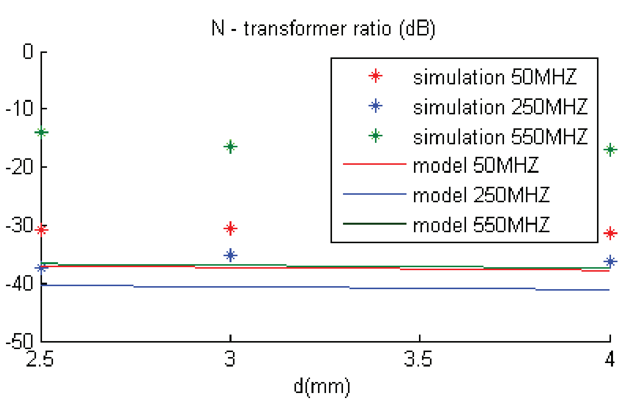

Fig. 9: ICW power transfer from PCB to IC: simulated N.

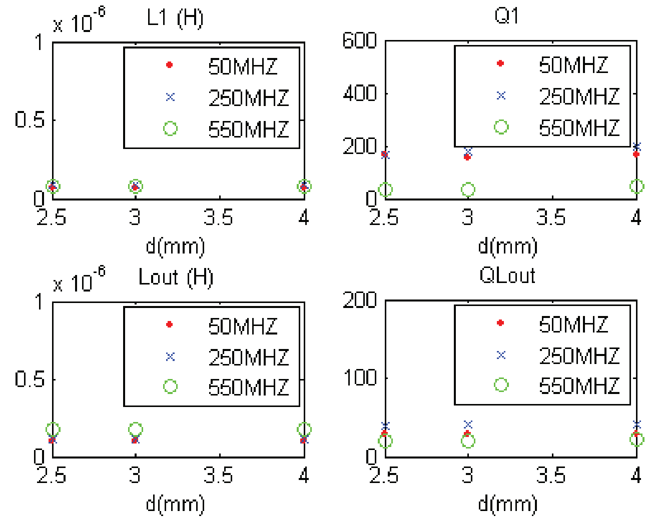

Fig. 10: ICW power transfer from PCB to IC: more simulation results.

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