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Simulation Oriented Rectenna Design Methodology for Remote Powering of Wireless Sensor Systems

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***Abstract*— Performance of wireless sensor systems are evaluated on the basis of the quality of data communication and efficiency of power transmission. This work presents the design methodology of a miniaturized tag antenna for remote powering of wireless sensor systems at 2.45 GHz ISM band. Accurate simulations of the rectifier input impedance aimed at the maximization of transmitted power are presented. Simulation results combined with simple transmitting antenna scattering parameters and cable loss measurements provide straightforward characterization of the fabricated tag antenna. The experimental measurements on a 12x10mm miniaturized antenna designed with the proposed approach show a gain of -5.9 dB close by 0.8 dB to the simulation results.**

I. Introduction

Wireless remote powering of wireless sensor systems, such as implantable systems for wireless telemetry, is a key part of the design in terms of miniaturization of the system while eliminating the use of batteries as an energy source. By using wireless remote powering, the lifetime of the sensor system is no more dependent on a battery. Besides recharging an integral battery is no more necessary for continuous telemetry operation. Depending on the power consumption requirement of the sensor system, wireless remote powering can be performed with either near-field inductive coupling or far-field electromagnetic coupling. The choice of the remote powering frequency is based on the constraints of the application such as power consumption, device size, read range or proximity, transmission medium and data rate. While a high data rate and high read range makes it necessary to use high frequency communication, higher power delivery makes the use of near-field remote powering more preferable.

A method to design a miniaturized antenna for its target integrated rectifier for far-field electromagnetic energy harvesting is provided in [1]. However, unavoidable process variations causes different input impedances and efficiency performances depending on the process corner of the fabricated chip. The developed methodology can also be used with a rectifier built with well-characterized commercially available diode. This enables us to evaluate the performance of fabricated antenna more accurately. It is also possible to use the method to design and characterize antennas for remote powering of wireless implantable sensor systems for small animals [2] where the designed antenna's size in this work is small thanks to its high frequency of operation and geometry. In that case the designer has to consider the tissue absorption effects for the design of the tag antenna. In this work, we perform the design and measurements with free-space model to demonstrate the effectiveness of the given design methodology in a simple environment. In Section II, far-field remote powering and impedance matching concept between the chip and the antenna are explained. In Section III, the simulation of a discrete rectifier with ADS using Spice model is described. The design of a miniaturized antenna for the simulated rectifier is introduced in Section IV. Measurement results are given in Section V and results are discussed in Section VI.

II. Far-Field Remote Powering and Impedance Matching for Maximum Power Transfer

The power at the output of a rectifier connected to a receiving antenna in free space is given by modified Friis transmission equation as follows,

$$P_{OUT} = P_{TX} \cdot G_{TX} \cdot G_{RX(R)} \cdot \left[\frac{\lambda}{4\pi d}\right]^2 \cdot \eta \tag{1}$$

where P_{TX} is the delivered power to the transmitting antenna, G_{TX} is the gain of the transmitting antenna, λ is the wavelength of the electromagnetic wave and $G_{RX(R)}$ is the realized gain of the receiver antenna with d as the distance between the two antennas and η is the rectifier efficiency. The realized gain $G_{RX(R)}$ of the receiving antenna can be written as,

$$G_{RX(R)} = G_{RX} \cdot \tau \tag{2}$$

with G_{RX} is the far-field gain of the antenna, τ is the power transmission coefficient between receiving antenna and load which is given by [3],

$$\tau = \frac{4 \cdot R_{CHIP} \cdot R_{ANT}}{|Z_{CHIP} + Z_{ANT}|^2} \leq 1 \tag{3}$$

where Z_{CHIP} and Z_{ANT} are the complex impedances of the antenna and the chip in the form of $Z = R \pm jX$. Since passive RFIDs' have capacitive input impedance behaviour [4], the tag antenna type should be selected such that it has inductive reactive impedance behaviour in the frequency of interest.

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III. Determining Non-Linear Rectifier Impedance

The impedance matching between receiving antenna and the rectifier determines how much of the incident power on the antenna is delivered to the rectifier. In order to maximize impedance matching, the input impedance of the rectifier should be determined accurately. Experimental measurements are expensive and time consuming [5], [6]. On the contrary, large signal s-parameter (LSSP) simulations quickly obtain the input impedance of a rectifier. This work uses the Spice model of a commercially available diode used to build a bridge rectifier (Avago HSMS-282x) as shown in Fig. 1. Simulations account for the parasitic components of the package such as bondwires and lead frames. The test circuit includes a single tone source at 2.45 GHz with 6 dBm of output power when it is matched to its load.

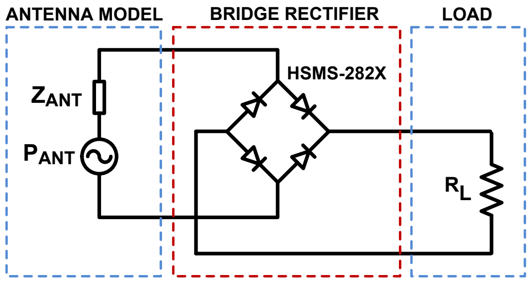

Fig. 1. Circuit schematic used to determine input impedance of the rectifier.

In order to ensure that the source and rectifier impedance are matched for obtaining correct input impedance of the rectifier we use the built-in optimizer of ADS. It minimizes the reflection coefficient between the source and the load which therefore maximizes τ by finding conjugate input impedance with the given input power. Fig. 2 shows simulation results of the input impedance of the rectifier versus frequency for 6 dBm of input power. Each point of the curve corresponds to a matched source impedance as determined by the optimizer.

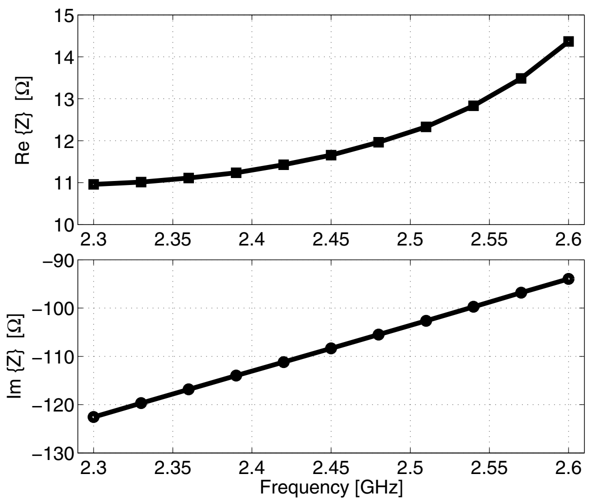

Fig. 2. Simulated input impedance of rectifier with matched source.

IV. Miniaturized Impedance-Matched Tag-Antenna Design

A. Antenna with Inductively Coupled Feed

Fig. 3 shows the used planar meandered antenna with inductively coupled loop feed. It ensures maximum impedance matching thanks to its geometry that enables various impedances where real and imaginary parts are independent of each other [7].
The real and imaginary parts of the antenna impedance $Z_{ANT} = R_{ANT} + jX_{ANT}$ are,

$$
\begin{aligned}
R_{ANT} =& \frac{(2\pi f M)^2}{R_{RB}} \\
X_{ANT} =& 2\pi f L_{loop}
\end{aligned}
\tag{4}
$$

where M is the mutual inductance between the radiating body of the antenna and the feed loop, L_{loop} is the inductance of the feed loop and R_{RB} is the resistance of the radiating body of the antenna at its resonance frequency. Notice that the input resistance of the antenna depends on mutual inductance M while input reactance is solely inductance of the feed loop L_{loop} itself.

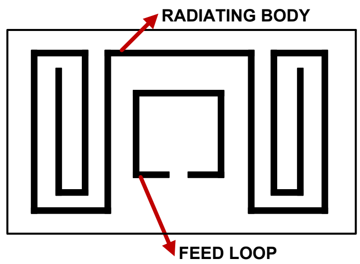

Fig. 3. Inductively coupled loop antenna layout.

B. Case Study

Suppose that the target input power is 6 dBm for the selected rectifier. Fig. 2 provides the corresponding rectifier impedance at 2.45 GHz: $Z_{RECT} = 11.7 - j108\ \Omega$. Maximum power transmission condition in (3) states that the rectifier and antenna have conjugately matched impedances. Therefore the target antenna impedance is determined as $Z_{ANT} = 11.7 + j108\ \Omega$. Using the above specifications we have designed and simulated the antenna with CST Microwave Studio. Simulated resistive and reactive input impedance of the designed antenna are depicted in Fig. 4. The simulated far-field radiation pattern of the antenna on XZ ($\Phi = 0°$) and YZ ($\Phi = 90°$) cut planes are depicted in Fig. 5 respectively.

The maximum simulated gain of the antenna is found as -5.1 dB and input impedance of the antenna is found as $Z_{ANT} = 11.7 + j108\ \Omega$. The antenna having dimensions of 12mm×10mm with dielectric thickness of 0.5 mm is fabricated on Rogers RO4003C dielectric with ($\epsilon_r = 3.55$). The fabricated antenna is depicted in Fig.6.

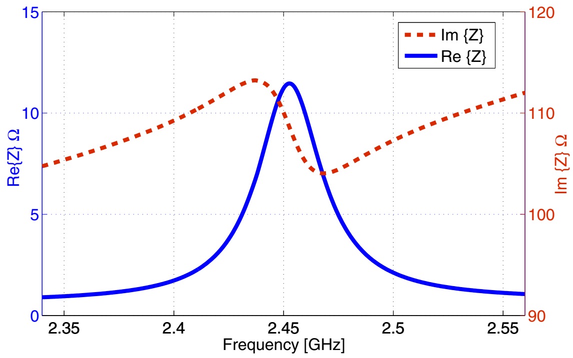

Fig. 4. Simulated real and imaginary impedance of the tag antenna.

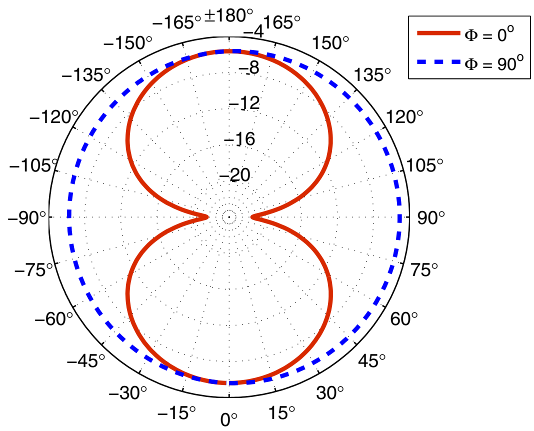

Fig. 5. Simulated far-field radiation pattern of the antenna.

V. Measurements

The fabricated antenna is aimed to be characterized by the output power of the rectifier. To do so, electromagnetic energy is radiated on the tag antenna by a coaxial-fed linearly polarized patch antenna as the transmitter antenna. The two antennas are placed facing each other with 12 cm separation. We need to ensure that the correct amount of power is delivered from transmitter antenna to the tag antenna. Therefore before measuring the tag antenna, transmitter antenna is characterized by using identical transmitter antennas facing each other as demonstrated in Fig. 7. This allows us to obtain transmitter antenna gain G_{TX} which is going to be used for tag antenna calculations.To characterize the transmitter antenna correctly, we have to determine different loss contributors. In RF communication systems, not only the free-space path loss but also the loss of other elements in the communication link effects the performance. Considering the demonstrated measurement setup for the transmitter antenna, the losses (in dB) can be accounted for with link budget equation given by,

$$P_{RX} = P_{TX} + G_{TX} - L_{TX} - L_{FS} + G_{RX} - L_{RX} \quad (5)$$

where P_{TX}, G_{TX} and G_{RX} are previously defined parame-

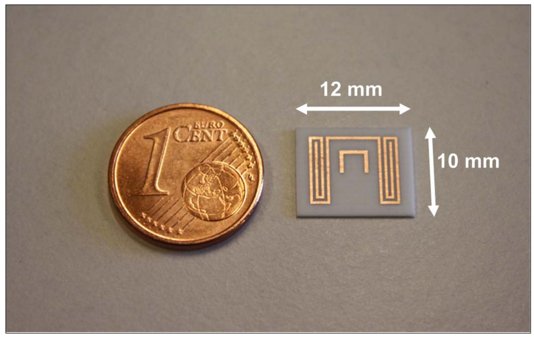

Fig. 6. Fabricated antenna.

ters, L_{FS} is free-space path loss defined by $(\lambda/4\pi d)^2$. The coaxial cable and mismatch losses of the transmitter and receiver antennas are denoted as L_{TX} and L_{RX} respectively.

A. Cable Loss

The use of a signal generator and a spectrum analyzer at both ends of a coaxial cable, which is then going to be used for feeding the transmitter antenna, determines the cable loss. The power of the signal which is fed to the cable should be in the range of the expected received signal power. The measured cable loss is 1.2 dB with an input signal power ranging between 0 dBm and 10 dBm.

B. Reflection Coefficient of TX Antenna

By a vector network analyzer (VNA) we measured the reflection coefficient (S_{11}) of the transmitting antenna. The resulting S_{11} in the frequency of interest is equal to -11.3 dB. It indicates that 7.4% of the power at the input of the transmitting antenna is reflected back. Thus, additional added power compensates for the power not delivered to the transmitter antenna due to the mismatch loss.

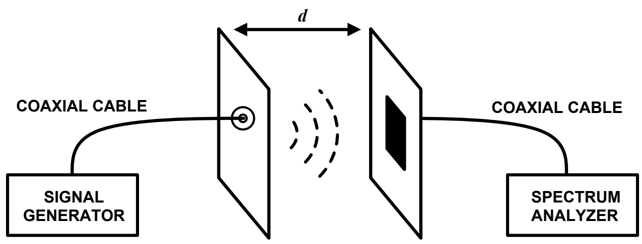

Fig. 7. Setup used to determine transmitter patch antenna gain.

C. Gain of TX Antenna G_{TX}

In order to verify the accuracy of simulations a simple gain measurement of transmitting antenna is performed. The experiment uses two identical transmitting antennas, then knowing the cable loss and reflection coefficients, the Friis transmission equation gives rise to the following relation,

$$P_{RX(T)} = P_{TX} \cdot G_{TX}^2 \cdot \left[\frac{\lambda}{4\pi d}\right]^2 \quad (6)$$

where $P_{RX(T)}$ is the received power by the receiving transmitter antenna. By using Eq. (6) including cable loss and reflection, the measured peak gain of transmitter antenna (G_{TX}) is 2.8dB. The simulated value is 3.6dB.

D. Measurement of the Power Link

We calculate the realized gain $G_{RX(R)}$ of the receiving antenna indirectly by simulating and measuring rest of parameters in (1) since the gain and the input impedance of the receiving antenna are not separately available. The measurement uses a distance between transmitter and tag antennas equal to 12 cm. Output power of the signal generator compensates for the cable and mismatch losses of TX antenna. The expected input power of the rectifier is 6 dBm (4 mW) and the simulated efficiency of the rectifier is 29.5% for 6 dBm of input power at 2.45 GHz. The use of (1) gives rise to the realized gain of the tag antenna $G_{RX(R)}$,

$$G_{RX(R)} = \frac{P_{OUT}}{P_{TX} \cdot G_{TX} \cdot \left[\frac{\lambda}{4\pi d}\right]^2 \cdot \eta} \tag{7}$$

According to the input power for simulation of the rectifier and separation distance $d = 12cm$ the equivalent isotropic radiated power $P_{EIRP} = P_{TX} \cdot G_{TX}$ is calculated as 2 W. Compensating for reflection at the input of the transmitter antenna, the radiated energy from the transmitter antenna results in 1.037 mW power at the output of the rectifier. Using (7) we calculate the realized gain $G_{RX(R)}$ of the tag antenna as -5.9 dB. On the other hand, by using the simulated antenna gain of -5.1 dB, the corresponding output power of the rectifier becomes 1.242 mW with the same transmitter and distance constraints but assuming ideal matching where $\tau = 1$. Based on these results, we find the difference between the simulations and the calculated realized gain $G_{RX(R)}$ as 0.8dB.

Fig. 8 depicts the realized gain $G_{RX(R)}$ of the tag antenna versus frequency from measured output power. It can be noticed that $G_{RX(R)}$ peaks with -5.9 dB at 2.42 GHz and the 3dB bandwidth is measured as 24 MHz. Also observe that the resonance frequency of the measured antenna is not exactly at 2.45 GHz, however the impedance variation of the rectifier at this frequency compared to the simulated impedance at 2.45 GHz is less than 1%. The parameters of the power link are summarized in Table-I.

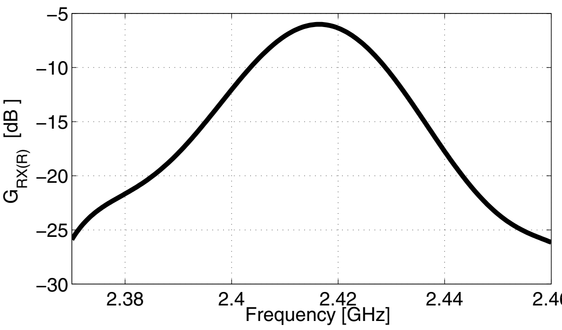

Fig. 8. Measured realized gain of the antenna versus frequency.

VI. Conclusion

A simulation oriented methodology for the co-design of a miniaturized tag antenna and a rectifier is presented. Performing simple measurements determines the parameters of the power link with the emphasis on the transmitter antenna in order to ensure that correct amount of power is transmitted. Using the Friis equation and the power at the output of the rectifier we calculate the realized gain $G_{RX(R)} = G \cdot \tau$ by combining measured and simulated parameters. Experimental results show that the realized gain of the tag antenna at 12 cm distance from the transmitter antenna is 0.8 dB less than the simulation results. Rather than performing time consuming and costly measurements the method uses full wave 3D simulation of tag antenna and LSSP simulation of rectifier to determine the input impedances quickly and accurately. Measurements verify that proposed design methodology exhibits good impedance matching performance for the tag antenna thanks to the simulated parameters of well characterized commercial diode.

TABLE I
Summary of Design Parameters and Outputs

Parameter	Value
Antenna Dimensions	12mm x 10mm x 0.5mm
Frequency	2.42 GHz
$P_{EIRP}(P_{TX} \cdot G_{TX})$	2 W
Distance	12 cm
P_{OUT} (Simulated / Measured)	1.242 mW / 1.037 mW
$G_{RX(R)}$ (Simulated / Measured)	-5.1 dB / -5.9 dB

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