

RFID and wireless sensing

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Abstract - This paper presents RFID issues, concerning the implementation of wireless sensing systems in networked infrastructures. The powering and communications, multisensors management, wireless sensing extension and hardware implementations are discussed in an integrated and unified form.

Keywords - Wireless sensors network, RFID, protocols.

I. INTRODUCTION

Today, the demand of markets to use active transponders, is reaching a saturation point. The main reason is because these ones get its power from batteries, which must be periodically replaced or recharged. This increases the final costs (maintenance + management), but contrarily to the passive transponders, theirs advantage is to no require remote powering, since an adequate lifetime and tight current margins be previewed in advance. Thus, the use of passive transponders are an interesting option, since it is not required batteries and the lifetime is virtually infinite. However the low-power consumptions constraints are huge and these type of elements must work with small duty-cycles. Moreover, passive transponders has low ranges, which are imposed by power/data nature of the link, and whose signal amplitude is far great than the norm in radio receivers and may be equal to or even greater than the supply voltage. The radio link aspects imposes two transmission paths, e.g., one to communicate with and energise the transponder and another to communicate with the base-station. In most applications, idle periods are used to energise the transponder and even without the powering field, the transponder must have power yet and synchronisation problems will arise. The passive transponders use two major types of powering and communication schemes: the near field type, where inductive coupling can be used, as it happens in the majority of RFID systems, and also the electrostatic and magnetostatic coupling can be used. However, these two types of couplings haven't commercial interest. The far field type is also an additional option, where the electromagnetic coupling helps to achieve major ranges compared with the inductive one, and it can use of some RFID, transceiver, or both concepts.

II. NEAR FIELD USAGE

A tank circuit is used to "catch" the information and power, and have high losses. Thus, the field path-loss of $p(d)=k \cdot d^{-3}$ imposes very low ranges, which means that this coupling is suitable for very low range, d [m], applications (tens of cm) and is imposed by the minimum working voltage of the electronics. Figure 1 helps to illustrate what happens with the

transmissions. The Figure 1(a) represents the general RFID transmission system, where the transformer is of coreless type, with $\mu_r=1$. The quantities L_1 , L_2 are the inductances of the primary and secondary coil, R_1 and R_2 are the ohmic and losses in the coils of the transformer, C_1 and C_2 are the capacitances in the receiver and at the base-station, and the tuning/coupling of the carrier frequency is made at the value $\omega=(L_1C_1)^{-1/2}=(L_2C_2)^{-1/2}$, and R_{LOAD} is the effect of the switch in the backscattered field. As seen in Figure 1(b), the variable DC voltage in the primary can be transformed in a fixed value, by putting a variable resistor, R_s , in the primary. This resistor represents the effect of the bitstream transmission in the uplink. The Figure 1(c) is the reference to the primary, to help to get the transfer function of the whole transmission circuit.

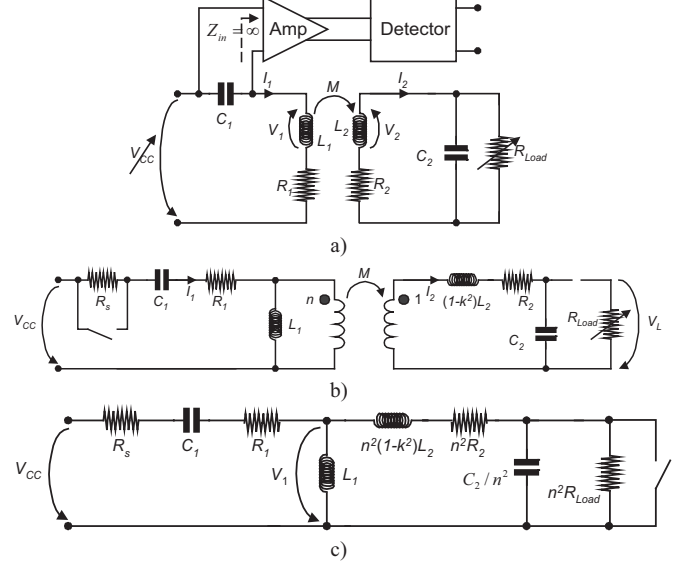


Figure 1: A general RFID data transmission system and the equivalent circuits for the downlink and uplink.

This system is rolled by the following equations:

$$V_1 = i\omega L_1 I_1 + i\omega M I_2 \quad (1)$$

$$V_2 = i\omega M I_1 + i\omega L_2 I_2 \quad (2)$$

where M is the mutual coupling inductance of the transformer, I_1 and I_2 are the currents flowing in the primary and in the secondary of the transformer and $i=(-1)^{1/2}$.

For this circuit, the coupling factor, k , which is distance dependent and the mutual inductance, M , are respectively:

$$M = k\sqrt{L_1 L_2} \quad (3)$$

$$k(d) = R_1 R_2 \cos(\theta) / [\sqrt{R_1 R_2} \times (\sqrt{d^2 + R_1^2})^3] \quad (4)$$

thus, the transform relation between the coils is

$$n = \sqrt{L_1 / (k^2 L_2)} \quad (5)$$

From Figure 1(c), the transfer function is:

$$T(s) = V_L / V_{CC} = bd / [a(b+c+d) + b(c+d)] \quad (6)$$

where $a=R_1+R_2$, $b=sL_1$, $c=sn^2(1-k^2)L_2+n^2R_2$ and $d=n^2R_1/(sR_{LOAD}C_2+1)$. The detection of the downlink transmission is made by measuring the voltage across the primary coil, whose voltage is:

$$V_1 = (R_s + a)b / [(R_s + a + b) + b(R_s + a)/(c + d)] \times V_{cc} \quad (7)$$

where $d=n^2R_{LOAD}/(sR_{LOAD}C_2+1)$, $a=R_1$, $b=sL_1$ and $c=sn^2(1-k^2)L_2$. The former term, d , is dependent from the load resistor and will affect the voltage detected in the primary of the transformer, when a downlink transmission is made. In this case, the current and voltage drop at C_1 are respectively:

$$I_1 = (b + c + d) / [a(b + c + d) + b(c + d)] \times V_{CC} \quad (8)$$

$$V_1 = (b + c + d) / [sC_1a(b + c + d) + b(c + d)] \times V_{CC} \quad (9)$$

The voltage V_1 is sensed by an amplifier ($Z_{in} \rightarrow \infty$) placed before the demodulator, and two output voltages are possible to obtain under normal conditions: $V_1(R_{LOAD}^0)$ and $V_1(R_{LOAD}^1)$. Normally, these two values are very close between them, which makes the modulation index to be very small:

$$m = |V_1(R_{load}^1) - V_1(R_{load}^0)| / |V_1(R_{load}^1) + V_1(R_{load}^0)| \quad (10)$$

thus, very efforts concerning the filtering and amplification, must be made to a perfectly detection on a weak envelope. Moreover, a special care must be considered, due to the non-linearity nature of the serial, R_s , and load, R_{LOAD} , resistors. Thus, using a moderated baudrates, the system can be classified as a wide sense stationary system (WSSS) [1].

For this coupling, the inductances must be calculated and this task can be done using one of the close formulas in [2]. Sometimes, the losses of the inductances can't be neglected because these ones has a resistance with material and skin effect components.

Taken the Figure 2(a) as reference and with respect to the power handling, the voltage at the receiver is:

$$v_2(t) = -\frac{d\phi}{dt}, \text{ with } \phi = \iint_{S_2} \mathbf{B}_{21} \cdot d\mathbf{s} \quad (11)$$

and the magnetic field is obtained from the Biot-Savart law:

$$\mathbf{B} = \mu_0 i_1 / (4\pi) \oint_{(S_1)} \mathbf{r} / r^3 \times d\mathbf{l}, \text{ with } \mathbf{B}_{21} = \mathbf{B} | \mathbf{u}_z. \quad (12)$$

Normally, the total field must be calculated numerically, for inductances with no circular geometries or exhibiting some kind of misalignment. Thus, the voltage is given by:

$$v_2(t) = -\frac{\mu_0}{4\pi} \frac{di_1}{dt} \iint_{(L_2)} \left(\oint_{(S_1)} \left(\frac{\mathbf{r}}{r^3} \times d\mathbf{l} \right) \cdot d\mathbf{s} \right) | \mathbf{u}_z \quad (13)$$

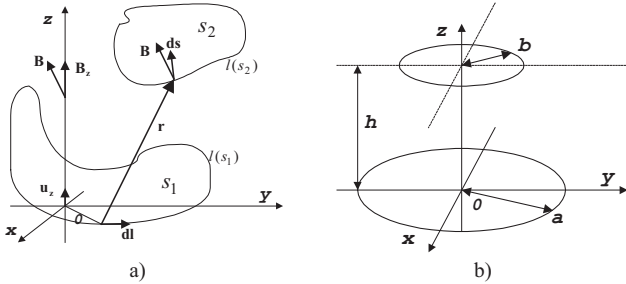


Figure 2: a) Power coupling in a near field usage, and b) power coupling for the particular case of two circular inductances.

Figure 2(b) show a particular case of two circular inductances with N_1 and N_2 turns at the base-station and at the receiver, respectively. The voltage in the secondary is:

$$v_2(t) = -[\mu_0 \pi N_1 N_2 (ab)^2] \cdot [2(a^2 + h^2)]^{-\frac{3}{2}} \times \frac{di}{dt} \quad (14)$$

The trade-off between RMS voltage at the inductance of the receiver and the current×turns at the inductance of the base-station is solved by obtaining the field \mathbf{B} that produces this voltage, then calculating the current necessary to produce that voltage and dividing it by the number of turns, until the current take a suitable value for all the cases.

III. FAR FIELD USAGE

This type of coupling uses the electromagnetic field to powering the transponders and to communicate with it. The information and the power are "caught" by an antenna, which resonates at the carrier frequency. The spherical propagation of the electromagnetic field helps to achieve highest ranges compared with those obtained with the inductive coupling. The field path-loss of $p(l(d))=k \cdot d^{-1}$ imposes high propagation ranges, which means that this coupling is limited only by the minimum voltage of the electronics to work. Figure 3(a) helps to illustrate what happens with the far field coupling, where a transmitter sends a modulated or unmodulated carrier towards the receiver. Then, if the received power is enough to powering the electronics of the receiver, this one process the received data (if is the case) and sends a bitstream towards the base-station for processing. This last transmission can be a vary from a simple receiver identification (RFID) to data acquired from a some sort of remote sensors. Figure 3(b) helps to know who this type of coupling is made in the receiver. First, it is well known that the voltage at the receiver antenna is field strength dependent and for deduction purposes, let's consider a wire antenna of length l . Thus and taking in account that for high distances, the field can be approximated by a plane wave then, the voltage at the terminals of the antenna is:

$$v_{ant} = l | \mathbf{E} = l \cdot \mathbf{u}_{ant} | \mathbf{E} = l \cdot \text{proj}(\mathbf{E}, \mathbf{u}_{ant}) \quad (15)$$

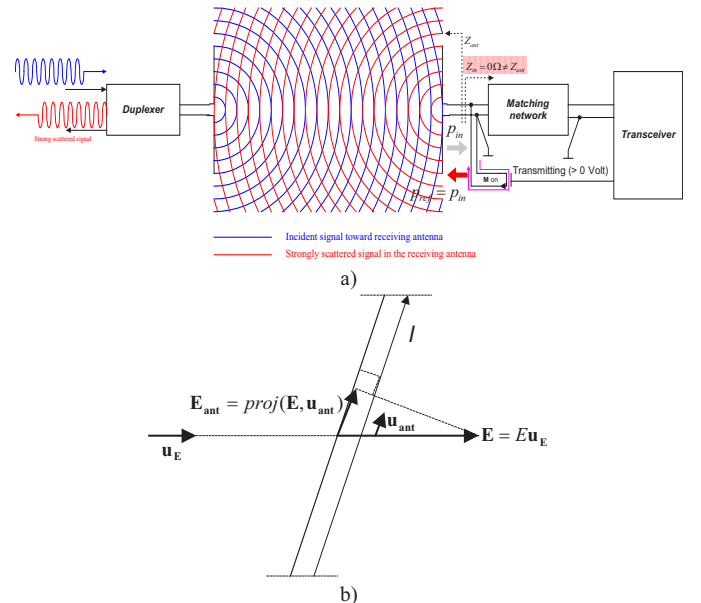


Figure 3: a) A far field system, based in an electromagnetic (EM) coupling and b) the EM field being "caught" in an wire antenna of length l .

The voltage, V_{RMS} [V], at the receiver is strongly dependent of the feeding power, P_{FEED} [W], which is a part of the received power, p_r [W], in the antenna, e.g., $V_{RMS}=f(P_{FEED})$. Thus, the received power, P_r [dB] is

$$P_r = P_{FEED} + \sum Gains + \sum Losses \text{ [dB]} \quad (16)$$

In the situation with ideal elements, the received power is:

$$\begin{aligned} P_r &= P_t + G_{ant}^t + G_{ant}^r + 20 \log_{10}[\lambda / (4\pi d)] = \\ &= P_t + \underbrace{G_{ant}^t + G_{ant}^r}_{\Sigma Gains} - \underbrace{20 \log_{10}(fd)}_{\Sigma Losses} + 20 \log_{10}[c / (4\pi)] \end{aligned} \quad (17)$$

Thus, as higher is the distance, or the frequency then higher is the loss, and in order to have the received power unchanged, the transmitted power, p_t [W], at the base-station must be increased. However, this is not always possible, thus the range is then limited and it can't be increased as desired. Moreover, the increase of the transmitted power is legally subjected to constraints imposed by the regulations, in order to guaranty the safety of people. Normally, these values are imposed with a safety factor of ten. These limits are calculated according the specific absorption rate (SAR) to the human beings [3]:

$$SAR = \sigma E^2 / \rho \text{ [W / kg]} \quad (18)$$

σ [S/m] is the tissue conductivity, ρ [kg/m³] is the tissue mass per volume and E [V/m] is the RMS electric field strength.

The mismatch between the antennas and the power amplifiers and low-noise amplifiers give rise to an additional loss factor which wasn't present in the near field coupling, e.g., the transmission loss, given by $L_{TL}=10\log_{10}(1-|\Gamma|^2)$, with $\Gamma=(Z_1-Z_2)/(Z_1+Z_2)$ and $P_{trans}=P_{inc}+10\log_{10}(1-|\Gamma|^2)=P_{inc}+L_{TL}$ (see Figure 4). Thus, the knowledge of the antenna impedances is mandatory in order to reduce the return loss. But, this brings an additional loss associated with the matching network.

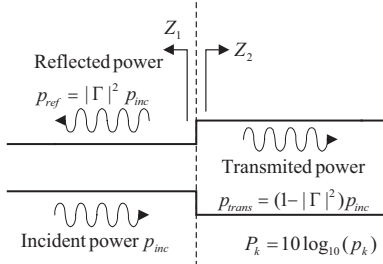


Figure 4: Illustration of losses due to mismatches in the antennas.

The backscattering transmission takes advantage of modification in the scattering characteristics of the antenna at the receiver. This changes the impedance match conditions by shorting the antenna with help of a switch. The backscatter properties can be modelled with the help of radar concepts [4]. The received power is obtained from the radar equation [5]:

$$p_r = p_t g_t^2 \lambda_0^2 / [(4\pi)^3 d^4] \cdot RCS \text{ [W]} \quad (19)$$

where RCS is the radar cross section of the target antenna. Thus, the received power from the sensor is:

$$p_r = p_t g_t^2 \lambda_0^2 / [(4\pi)^3 d^4] \cdot RCS_{sens} \quad (20)$$

It must be noted that the RCS of the sensor is dependent from the transmitted data. Thus, modulating the value of the RCS, the data arriving from the sensor can be successful recovered in the base-station. This don't invalidate the need of the

receiver to be sensible enough to detect tiny variations on tiny received signals. A variety of antennas can be used, since some major characteristics points be regarded, such a points includes the dimensions, the directivity and the impedance of the antennas. It can't be forgot that low frequencies implies prohibitive dimensions for the antennas. Independently the type of antenna chosen, the impedance matching is important in the RF circuits design, in order to achieve a maximal power transfer from the antenna to the receiver and to the powering subsystems.

IV. WIRELESS SENSING EXTENSION

The power management and the communication subsystems can be separated between them. The main tasks of the power management subsystem includes the rectification, which can be made with diodes, bridges constituted by MOSFETs or voltage multipliers. The voltage regulation is another task for this subsystem, and it can be made using a series of forward biased diodes connected between the two power rails, with reverse biased zener diodes, or using third party components (the MAX1672 commercial DC/DC step-up converter from Maxim IC is a good example).

The right chose of the encoding format to be used by the communication subsystem, constitutes an important task, because behind the data communication itself, timing information concerning clock and frame synchronisation must be provided at the same time. First, it is important that the data encoding occupy the minimum possible bandwidth. Moreover, data encoding brings other major aspects, such as the synchronisation capability and perhaps, the most important. Another aspect is the easiness of the implementation, e.g., the second most important. The cost of implementation is also important, but it depends from the former. Finally, it must be possible to provide mechanisms of signal resolving and collision detection.

For the demodulation, if ASK modulation is used then, a non-coherent detection (or envelope detection) can perform well this task. However, this technique has a drawback compared with the coherent detection, where an improvement of 3 dB is achieved, but with the cost to require the need of a in-phase carrier and a true four-quadrant multiplier circuit, followed by a low-pass filter [6]. All the trade-offs must be analysed in advance in order to know exactly the best choice. Moreover, if an envelope detector is chosen, this circuit can provide additional gain if necessary, as is the case of Figure 5 [7]. The gain increases the sensitivity of receivers.

Normally and taking the generic example of Figure 6, the transmitter subsystem in passive transponders, both using inductive and electromagnetic couplings is a transistor that is normally in the cut-off sate (for the bit '0' for example), but when a change in the bit occurs (changing to '1' for instance), the transistor enters in the conduction state and provoke respectively, disturbances in the magnetic field and scattering characteristics of antennas (mismatches). However, special care must be taken, in order to not discharge the retention

capacitor that supplies the transponder. When the inductive coupling is been used, a special care must be taken, e.g., ESD protections must be provided in order to avoid possible damages resulting from the switching in that transistor.

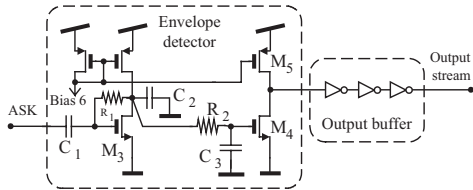


Figure 5: Envelope detector with gain.

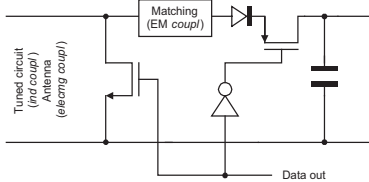


Figure 6: Illustration of a generic transmitter in a passive transponder.

Some improvements can be achieved in the communication link, e.g., in modulation and coding methods if transmitter blocks make use of active transmission, but low-power consumption enough to work properly and purely electromagnetic in the coupling - Figure 7(a). As depicted in Figure 7(b) and using active transmission yet, but still low-power consumption and combining inductive coupling to have more power available, then this new improvement will result in a suitable architecture for biometry purposes [8].

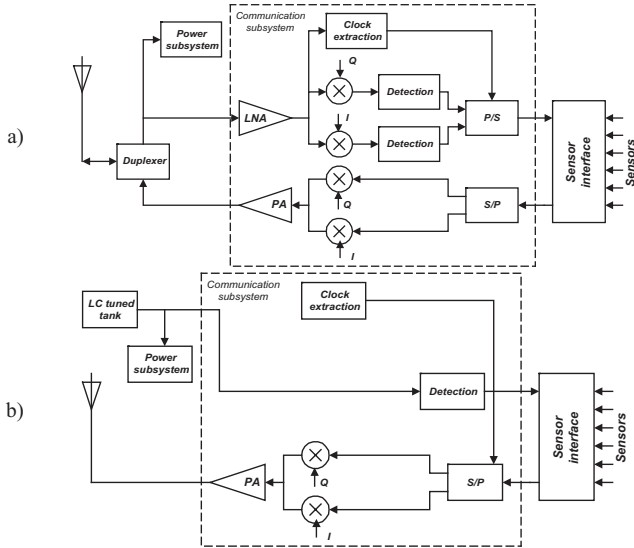


Figure 7: a) Communication subsystem improvements; and b) a new improvement in the communication subsystem.

V. UNIFICATION AND APPLICATIONS

A general implementation imply the identification of the application to be implemented, in order to decide the use of active or passive feeding. The cost is one the main aspects to be in account, as well as the reliability of the communication link. In relation to the type of coupling, the inductive or near-range is most suitable in applications which involves living beings. The electromagnetic coupling is normally reserved for far-ranges or to improve the capacity of the communication links. Thus, The appropriate architecture pf

the transceivers in the base-station and in the sensors (or transponders) must be chosen, as well as the selection of the suitable coding scheme to manage environments with multiple sensors and to be adapted to the previous chosen transceiver.

Systems based in RFID technology can be successfully applied to the telemetry and wireless sensing, where bi-directional communication and feeding of multiple sensors constitutes further steps to explore biometry principles in a multi-sensor fashion [9]. A first application uses an inductive links to feed and to communicate with the microchip's MCRF202, which is connected to a sensor. As shown in Figure 8, if the input SENSOR is '1', the MCRF202 uses a RF link to send a normal data stream, or else it sends an inverted information bitstream towards the base-station.

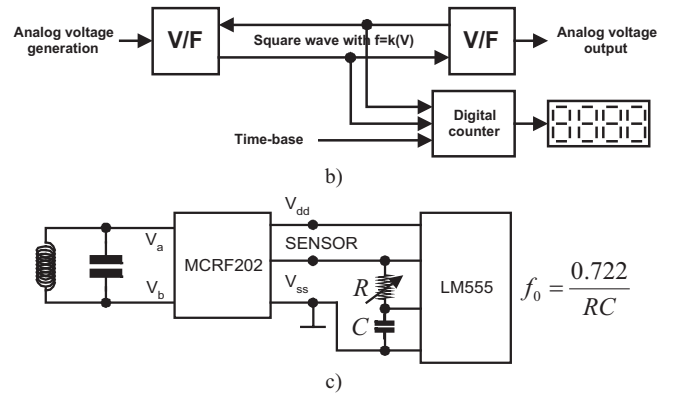


Figure 8: A sample applications of the RFID technology to the telemetry..

Figures 8(a) and (b) show a sample applications for the MCRF202, which behind provide power supply to external electronics, it also provides an interface for a voltage control oscillator (VCO or timer), respectively. The VCO interface is suitable for a frequency telemetry system, which is more robust to the presence of noise, when compared with a conventional amplitude (voltage or current) telemetry system.

Figure 9 shows another application for the MCRF202. It is the microcontroller interface, which uses the microchip's PIC16F84, whose internal time-base of 1 ms will allow to obtain the frequency f [kHz], and is directly coded in binary by the stored program and use the MCRF202 to transmit the data towards the base-station. To finish, the MCRF202 uses the FSK modulation with a carrier frequency of 13.56 MHz.

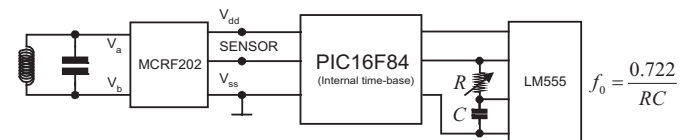


Figure 9: Microcontroller interface for wireless sensing.

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