Improving contactless technology by increase of transponder load modulation with serial capacitor

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***Abstract*—Proven chip technology based on International Standards can be used to implement person-related, secure contactless technology also into objects defining smaller antenna sizes than cards. Provided that sufficient power for chip operation is available, low transponder load modulation is the most limiting factor for the communication distance. A simple capacitance matching network allows to balance between power requirements and load modulation amplitude, so that compliance to the requirements of the Standard and to existing terminal infrastructure can be achieved.**

I. INTRODUCTION

Contactless technology for person-related applications is most advanced in contactless cards today. Billions of transponder cards are already in the field, mainly for applications in Access Control, Public Transport Ticketing, e-Passports, or Credit Cards with contactless interface. There was also a huge investment in terminal infrastructure during the past decade and up to date over 300 million devices of different manufacturers are deployed. Most of these components are compliant to the ISO/IEC14443 Proximity Standard [10] or to FeliCa, all these are inductively coupled systems operating in the Near Field in the 13.56 MHz world-wide harmonized frequency band for Short Range Devices.

As the reliability of this kind of RFID technology is already proven by its history, it is desirable to enable the integration of the same technology also into other objects, most prominent may be watches, but also key-fobs, cell-phones, memory sticks or even SIM-modules. The product standard describes system properties of terminal reader and transponder at the air interface, offering a Type A interface and a Type B interface for ISO/IEC14443 and a third interface option for FeliCa. The operating principle of all these interfaces is intentional de-tuning of the reader antenna circuit by the transponder antenna circuit in proximity of each other. Properties of this de-tuning are important for the air interface. Up to now the evaluation of standard conformance was related to a Class 1 transponder antenna size which fits into the ID-1 card format specified in ISO/IEC7810, as this is the main application case. But other objects can require different antenna sizes, typically smaller ones. To be able to use the same chip platforms, smart antenna design is required to generate as good as possible similar transponder characteristics at the air interface as larger contactless cards have, to fit into existing infrastructure. For the analogue performance of transponders 3 main aspects have to be considered, which are: (1) energy transmission from reader to the batteryless transponder, (2) reader command demodulation and decoding in the transponder, and (3) transponder load modulation as well as reader sensitivity to this de-tuning, to allow transmission of information from transponder to reader.

A survey through literature shows, inductance as the most important electrical antenna parameter already was calculated by Maxwell [1] as analytical expressions for special cases. Many useful approximations and informations in the practical RFID context are given in [2], and descriptions of state of the art production technologies for transponder antennas are described in the books of Finkenzeller [3] and Paret [4]. A more detailed treatment of the electrical antenna properties is given in the PhD thesis of Cichos [5]. However, for the contactless transponder, powered by the H-field, it is not as straightforeward as in other communication technologies, to combine a good antenna with a good receiver to have a good system. In fact, the contactless transponder consisting of chip and antenna must be considered as one system and must be treated in context, application specific, to be able to understand the system behaviour. One reason is the chip impedance, which varies over a wide range. To get optimum energy transfer, the transponder resonance frequency should be close to the 13.56 MHz carrier frequency, this is well understood. A method to describe the resonance frequency detuning is presented by Witschnig et al. [6].

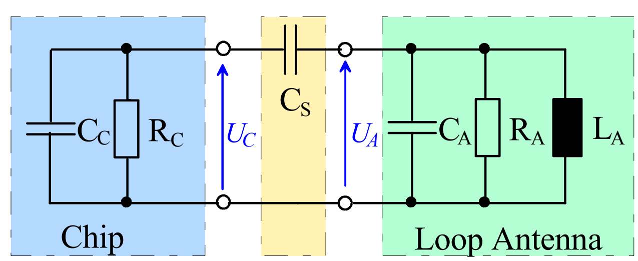

Figure 1. Transponder equivalent circuit.

An option to increase load modulation (on the expense of energy consumption) was presented by Klapf et al. [7] as the concept of inverse load modulation. One drawback is that this would require changes in chip architecture. A different option for the same goal is to use smart transponder antenna design. This paper describes, how a capacitor in series between chip and antenna can be used to boost the load modulation of transponders with small antennas to values similar to larger ones. For objects like watches or memory sticks it is an option to integrate discrete passive components in addition to wire loop antennas, and for some antenna technologies it is also possible to integrate capacitance in the antenna itself. For specific product applications, we have recommended the principle since some time, in this paper we give the theoretical explanation combined with measurements of minimum H-field required for operation and of load modulation, for a practical application case. Considerations on LCR network analysis which are very useful in this context also can be found in the book of Scott [8].

II. CONSIDERATION OF THE ANTENNA PARALLEL EQUIVALENT CIRCUIT PROPERTIES

As a starting point, practical values were measured for a planar, circular loop wire antenna and an NXP Smart Card chip. These values are given in table 1:

TABLE I. SAMPLE ANTENNA AND CHIP PARAMETERS

Parameter	Value	Unit
Antenna diameter	131	mm
Copper wire diameter	150	µm
Number of turns	10	
Antenna Inductance L_A	7.73	µH
Antenna Capacitance C_A	7.9	pF
Antenna Resistance R_A	46.4	kOhm
Chip Capacitance C_C	17	pF
Chip Resistance R_C	2000	Ohm

We can use network analysis, to consider properties of the circuit. As a starting point, let us first only consider the antenna parallel equivalent circuit, consisting of inductance L_A, resistance R_A and capacitance C_A as function of the complex frequency $\sigma = s + j\omega$. The admittance Y_A of this network can be calculated according to

$$Y_A = \frac{1}{sL_A} + \frac{1}{R_A} + sC_A = \frac{s^2 R_A L_A C_A + sL_A + R_A}{sR_A L_A} \tag{1}$$

So the network function for the antenna impedance Z_A is

$$Z_A = \frac{sR_A L_A}{s^2 R_A L_A C_A + sL_A + R_A} \tag{2}$$

A parallel resonance is given by the zero of the denominator (high impedance) while a serial resonance would be given by the zero of the counter (low impedance) of Z. If we set $s \to j\omega$ we get

$$R_A + j\omega L_A - \omega^2 R_A L_A C_A \equiv 0 \tag{3}$$

Considering only the real part, we get the well-known solution (Thomson equation) at

$$\omega^2 R_A L_A C_A = R_A \left| \frac{1}{R_A L_A C_A} \right. \tag{4}$$

$$\omega^2 = \frac{1}{L_A C_A} \quad \to \quad f_{RES\_A} = \frac{1}{2\pi\sqrt{L_A C_A}}$$

which gives us the undamped, natural self-resonance frequency f_{RES_A} of the antenna. The complete solution is given by the eigenvalues of the characteristic equation, by

$$s_{1,2} = -\frac{1}{2R_A C_A} \pm \sqrt{\left(\frac{1}{2R_A C_A}\right)^2 - \frac{1}{L_A C_A}} \tag{5}$$

A comparison with the harmonic oscillation as given in (6) allows to compare for (radian) self-resonance ω_0 and damping coefficient ζ

$$s = -\zeta\omega_0 \pm \omega_0\sqrt{1-\zeta^2} \tag{6}$$

so the damping coefficient is given in electrical parameters by

$$\zeta = \frac{1}{2R_A C_A \omega_0} = \frac{1}{2R_A}\sqrt{\frac{L_A}{C_A}} \tag{7}$$

and alternatively the antenna Q-factor at self-resonance frequency is given by

$$Q_A = \frac{1}{2\zeta} = \frac{1}{4R_A C_A \omega_0} = \omega_0 R_A C_A = \frac{R_A}{\omega_0 L_A} \tag{8}$$

where the bandwidth B is given by

$$B = \frac{\omega_0}{Q_A} = 2\xi\omega_0 \tag{9}$$

In the complex s-plane, the system parameters are represented as shown in fig. 2.

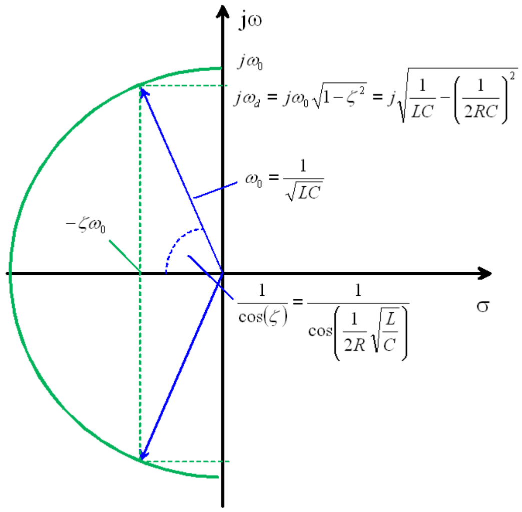

Figure 2. System parameters of the parallel resonant circuit.

For practical conditions we will find the weakly attenuated case for an antenna (critically damped case and over-damped case can be neglected), with Q-values around 70 (as for antenna values in table 1). Important for us is the (un-damped) self-resonance radian frequency ω_0, although also the damped natural radian frequency ω_d at which the antenna would oscillate after excitation will be very close.

III. CONSIDERATION OF THE COMPLETE NETWORK PROPERTIES

For the next step, we will consider the whole network of fig. 1 but neglect the chip resistance ($R_C \to \infty$). This allows a simple, but useful approximation, as we can find an equivalent capacitance and use the equations for the parallel resonant circuit. The two capacitances in series, C_S and C_C, can be replaced by one capacitance C_{SE} according to

$$\frac{1}{sC_{SE}} = \frac{1}{sC_S} + \frac{1}{sC_C} = \frac{C_S + C_C}{sC_S C_C} \qquad (10)$$

$$\rightarrow \quad C_{SE} = \frac{C_C C_S}{C_C + C_s}$$

We can add this one equivalent capacitance C_{SE} in parallel to the antenna capacitance C_A and get an estimation for the resonance frequency of the whole network according to

$$f_{RES\_E} = \frac{1}{2\pi\sqrt{L_A(C_A + C_{SE})}} \qquad (11)$$

Now, let us consider the complete network with all components, and calculate the network function seen from the antenna side. The chip admittance Y_C is given by R_C and C_C in parallel, so

$$Y_C = \frac{1}{R_C} + sC_C = \frac{sR_C C_C + 1}{R_C} = \frac{1}{Z_C} \qquad (12)$$

We can add the serial capacitance to the chip impedance and get the intermediate impedance result

$$Z = \frac{R_C}{sR_C C_C + 1} + \frac{1}{sC_S} = \frac{sR_C(C_S + C_C) + 1}{s^2 R_C C_C C_S + sC_S} \qquad (13)$$

Finally, we can add this admittance to the antenna admittance to get the complete network admittance function

$$Y_N = \frac{s^2 R_C C_C C_S + sC_S}{sR_C(C_S + C_C) + 1} + \frac{s^2 R_A L_A C_A + sL_A + R_A}{sR_A L_A} \qquad (14)$$

We can consider the corresponding network impedance as function of serial capacitance and of frequency, shown in figure 3 and 4.

The red dotted line gives the absolute value of the impedance, and the blue straight line gives the real part. The diagrams are evaluated with parameters given in table 1. For fig. 3, the frequency is the carrier at 13.56 MHz, and for fig. 4, the serial capacitance is 12 picofarad (pF).

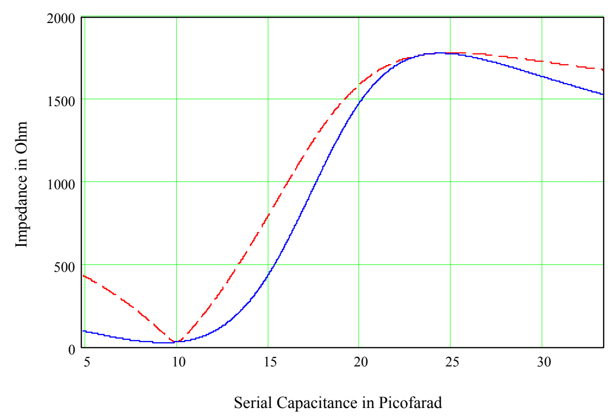

Figure 3. Network impedance for varied C_S.

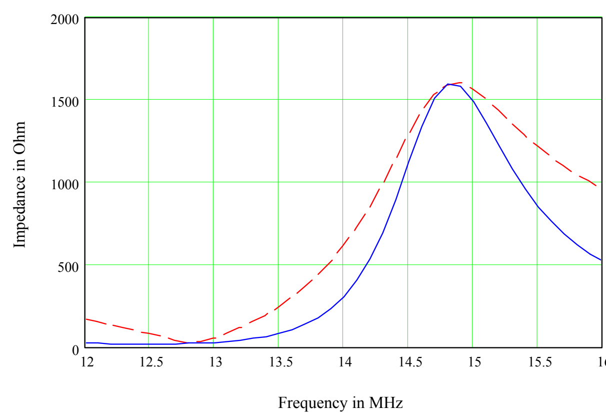

Figure 4. Network impedance over frequency.

The denomiator of Z_N results in a number of terms, which we present in (15).

$$s^3 R_A R_C L_A (C_A C_S + C_A C_C + C_C C_S) + \\ s^2 L_A (R_A C_A + R_A C_S + R_C C_C + R_C C_S) + \\ s(R_A R_C C_S + R_A R_C C_C + L_A) + \\ s^0 R_A \tag{15}$$

With $s \rightarrow j\omega$ the relevant parallel resonance frequency f_{RES_P} is given by

$$f_{RES\_P} = \frac{\omega}{2\pi} = \\ = \frac{\sqrt{\frac{R_A R_C (C_S + C_C) + L_A}{R_A R_C L_A (C_A C_S + C_A C_C + C_C C_S)}}}{2\pi} \tag{16}$$

as this solution gives very similar values to our approximation in (11).

IV. DETERMINATION OF THE VALUE FOR THE SERIAL CAPACITANCE

The most efficient energy transmission to a transponder is achieved, if the transponder resonance frequency is set very close to the carrier frequency. So as one option, we can re-arrange (16) to determine a value for the serial capacitance depending on the radian carrier frequency ω

$$C_S = \frac{R_A R_C C_C + L_A - \omega^2 R_A R_C L_A C_A C_C}{\omega^2 R_A R_C L_A (C_A + C_C) - R_A R_C} \tag{17}$$

As can be seen, not for all antenna parameters a positive capacitance can be found to satisfy the condition, mainly the inductance has to be high enough.

However, the main reason for our matching network was to achieve an increase in transponder load modulation by an increase of the antenna voltage. This is related to the transponder system Q-factor, which is modulated by the transponder in the up-link communication phase, from a maximum value (determined by antenna losses R_A and the chip current consumption, represented by the shunt resistor R_C) to a lower value (R_C // R_{SHUNT}). So let us consider this Q-factor for a moment. In general, the Q-factor of a resonant network is defined as the relation between stored energy W in one carrier period, and loss power P in the same time

$$Q = \frac{\omega W}{P} = \frac{\omega C U_{C,RMS}^{\;2}}{P} = \frac{\omega L I_{L,RMS}^{\;2}}{P} \tag{18}$$

For state of the art CMOS chip integration, the voltage for chip operation is limited to low voltages (e.g. 2.5 V(rms)) typically by a shunt regulator. To allow further cost reduction, the reduction of silicon area and consequently integration at smaller structure scales (below 100 nm) will continue to be a trend, so it is essential to consider the limited chip voltage as a frame condition. As the voltage of our transponder in operation is limited to a fixed value, we can use the relation (18) based on voltage and capacitance to calculate the system Q-factor of a typical transponder and further also the antenna current for known resonance frequency and inductance. This means for the antenna current

$$I_A = n \cdot I_{L,RMS} = n \cdot \frac{U_{C,RMS}}{\omega L} \tag{19}$$

where n is the number of turns of the loop antenna and $L \sim n^E$ with E close to 2. So less antenna turns will result in more transponder antenna current for voltage limited transponders considered for the same resonance frequency. This shows an alternative option to increase transponder load modulation, if the antenna design is executed in such a way, that parallel capacitance is increased and antenna inductance is reduced, to reach the same resonance frequency. The question how a change in the relation between L and C can increase the load modulation was also considered by Marechal et al. [9]. and by the authors. [10].

The aim of our matching network is different; we aim to increase the transponder system-Q-factor by an increase of the antenna voltage (by a factor) over the limited chip voltage. To consider this voltage transformation ratio, we can calculate a voltage divider between our antenna impedance Z_A and the serial capacitance.

$$\frac{U_A}{U_C} = \frac{Z_A}{Z_A + \frac{1}{sC_S}} = \frac{sZ_A C_S}{sZ_A C_S + 1} \tag{20}$$

We can solve the equation and as we do not care for the phase relation but only for the amplitude, we can get a rational function for the antenna voltage U_A

$$U_A = U_C \cdot \frac{\omega^2 R_A L_A C_S}{\sqrt{\left(\omega^2 R_A L_A (C_A + C_S) - R_A\right)^2 + (\omega L_A)^2}} \tag{21}$$

This means, the voltage at the antenna can be much higher than the limited chip voltage, which also allows to increase the transponder system Q-factor, and so to increase load modulation. On the other hand, the required minimum H-field for chip operation also will increase. This means in the given scenario, C_S can be dimensioned to balance between increased load modulation (due to increased antenna voltage) and the minimum required H-field for operation. For the given antenna and chip parameters of table 1, (18) gives a voltage transformation factor as shown in fig. 5.

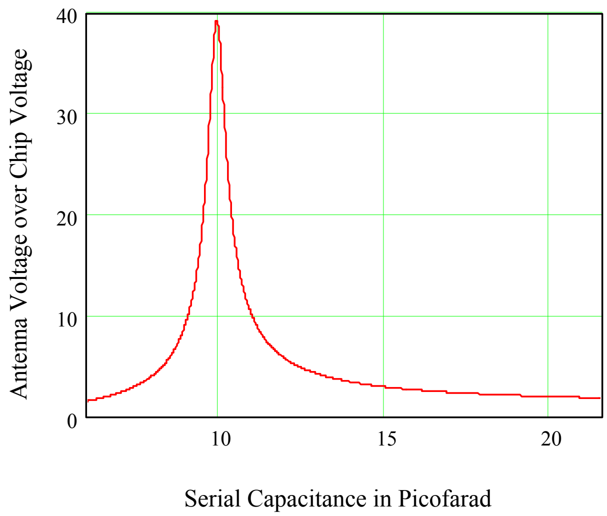

Figure 5. Voltage transformation as function of C_S.

Still in practice there are two more points to consider, not to increase the voltage too much:

- Tolerances in antenna and chip parameters are critical, if the system is operated very close to a maximum voltage transformation, and
- the system Q-factor also has influence on the received modulation signal of the reader command, so the receive path of the transponder is affected.

In practice, calculation of tolerances and measurements to also take into account other effects (e.g. metal parts in the objects) will allow to find optimum practical values. Of course, the network also can be extended, e.g. one more capacitor in parallel to the antenna parasitic capacitance will allow to make to the system more robust against tolerances in the antenna parasitics. Such practical measurements are presented in the next section.

V. PRACTICAL MEASUREMENTS

Using the antenna and chip parameter values presented in table 1, we can verify our considerations with a practical measurement series of load modulation using an ISO/IEC10373-6 test bench and varied values for the serial capacitor C_S. The voltage transformation in this series ranges from 1.8 to 5.7 at minimum H-field, according to the newtork calculation. For interpretation it should be noted, that the chip impedance varies in operation depending on the H-field strength: The equivalent parallel chip resistance varies due to the voltage limiter, and also the equivalent parallel input capacitance has a dependency of about 2.5 pF over the complete voltage range. As a consequence, transponder resonance frequency and transponder system Q-factor vary over the applied H-field range, which is one part of the explanation for the difference between upper and lower subcarrier level in the measurement series of fig. 6 – 9.

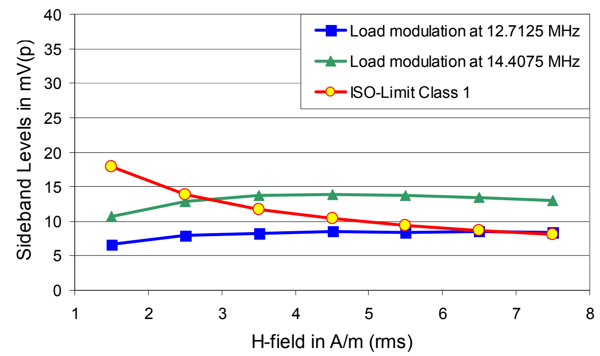

Figure 6. Load Modulation measured for Cs ~ 22 pF .

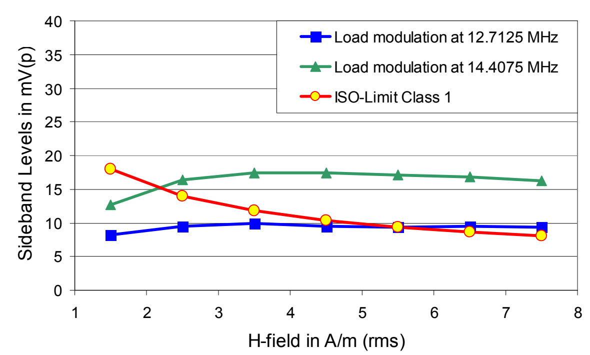

Figure 7. Load Modulation measured for Cs ~ 18 pF .

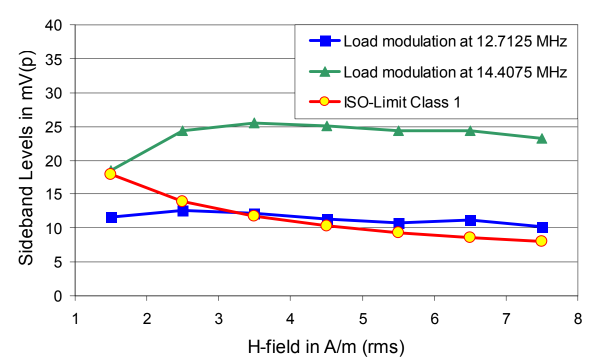

Figure 8. Load Modulation measured for Cs ~ 15 pF .

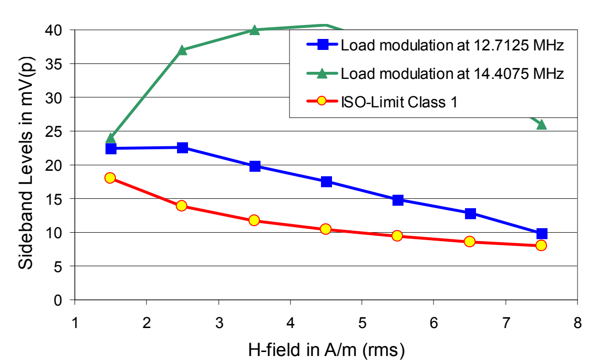

Figure 9. Load Modulation measured for Cs ~ 12 pF .

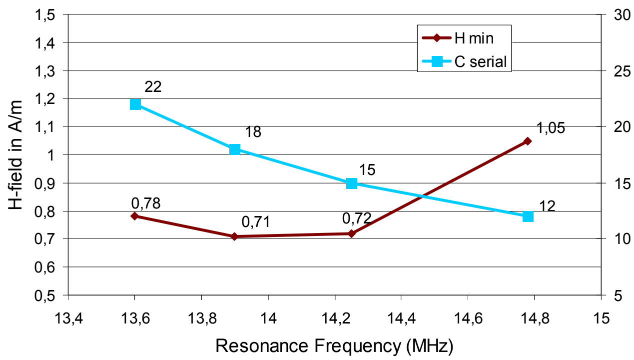

Figure 10. Minimum H-field for operation and serial capacitance values for previous figures.

As can be seen in fig. 6 – 9, the load modulation measured according to the ISO/IEC10373-6 Standard shows an increase of about 200 % or about a factor 3 (and even more compared to a standard transponder circuit). According to the red line indicating the ISO/IEC14443 minimum limit for load modulation this means the difference from a clear fail to a full pass in compliance to the product Standard requirements in this regard. On the other hand, the required H-field strength increases less than 50 %, as can be seen in fig. 10. As the Standard allows up to 1.5 A/m minimum H-field for transponder operation, this can be tolerated.

VI. CONCLUSION

A matching network of at least one capacitor in series between transponder chip and antenna allows a voltage transformation, which can increase load modulation for voltage-limited transponders in modern architecture for small structure scales on the expense of increased H-field required for operation. This can allow Standard conformance and longer communication distances for small transponder antenna form factors. As the loop antenna inductance should be higher than usual for such chips, this type of network is preferred for copper wire antennas of high antenna Q-factors, because the number of turns can easily be increased.

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