Analytical considerations for an ISO/IEC14443 compliant SmartCard transponder

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***Abstract*— We derive and use an analytical formula to describe some aspects in the RF behaviour of Smartcard transponders at the contactless air interface. The aspects we consider contain minimum operating H-field, thermal loading, card loading and maximum load modulation depending on transponder equivalent circuit parameters. We not only consider results, but also carefully discuss the approximations and simplifications we make, which should give an understanding of the principle and thereby allow for adaptations to selected fabrication technologies among the variety of existing options.**

I. Introduction

Network analysis allows to consider problems for integrated semiconductor transponder properties based on simple analytical formulas based on equivalent circuits for antenna and chip. But is it correct to apply this principle in the context of Radio Frequency Identification (RFID) and secure, batteryless, contactless Smartcards? The electrical part of a contactless Smartcard is called a *transponder*, consisting of antenna, assembly and controller chip. So, apparently not, as microprocessors undergo pulse-shaped changes in the current consumption (e.g. for read and write memory access, or due to parasitic capacitance for state-switches in synchronous logic). An analogue simulation of the complete integrated semiconductor chip on transistor or even on layout level would be more appropriate [16]. This covers all analogue and digital blocks combined, and an analysis of their main functions in time domain. An appropriate model for design should even include variations in the wafer production process, and variations due to environment in application (e.g. temperature dependency). However, on one hand such detailed informations are not provided by semiconductor manufacturers, and on the other hand, the complete circuit of a state-of-the-art Smartcard security controller is much too complex to allow a complete simulation on transistor level in reasonable time. Even for semiconductor manufacturers the trend is to use block level simulations. This means to specify in a more abstract way the behaviour of a complete functional block, and it allows to model the influence of one block to the rest of the circuit based on such assumptions. Analytic, quasi-static models can describe the voltage-limited, contactless transponder based on equivalent circuit parameters on the level of an RF coupling system, if certain preconditions can be met, like a continuous current consumption. In fact, for security Smartcard controllers it is essential that the current at the (accessible) antenna connection pads is smoothed and does not contain spikes, as these would be visible as non-intended modulation, and would allow to draw conclusions on chip operation. This anyhow is not allowed for security chips.

The influence of a variation of individual parameters can be studied very well with such analytical considerations, and parameter measurements can be related to each other. So it is possible to relate equivalent chip impedance, which is measured contact-based [8], to the RF system properties of the transponder resonance circuit, the *resonance frequency* and the *quality factor*. These are measured contactless. And this, basically, is the benefit and the practical use case of such analytical considerations.

II. An analytic formula for Hmin based on the transponder equivalent circuit

Requirements for the final product, the contactless Smartcard transponder, are defined by the base standard [1]. For proximity cards, values for the behavior in the contactless system are defined at the Air Interface, and a test standard [2] defines an antenna assembly as test environment, and the methods for measurement. This coaxial loop antenna assembly shown in fig. 1 is specified in such way, that the vector of the emitted *H*-field is perpendicular to the transponder loop antenna, and the signal strength of the alternating *H*-field is homogenous in radial direction in distance of the device under test (DUT) position, over the card antenna area. The *H*-field strength is measured as induced voltage in the Calibration Coil, defined in the test standard. This coil is placed coaxial, on the opposite side of the DUT transponder, but in equal distance to the circular emitting loop antenna.

These special conditions in most cases do not apply to a transponder in arbitrary position in a reader emitting *H*-field. But the transponder will show equal behavior, if we take the equivalent average (spatially averaged over the transponder antenna area) homogenous *H*-field strength of the vector perpendicular to the transponder antenna.
This *H*-field strength is one base requirement for the transponder, as it defines the power which is available for the chip operation. So we will deduce the chip (AC) voltage as function of this *H*-field strength.

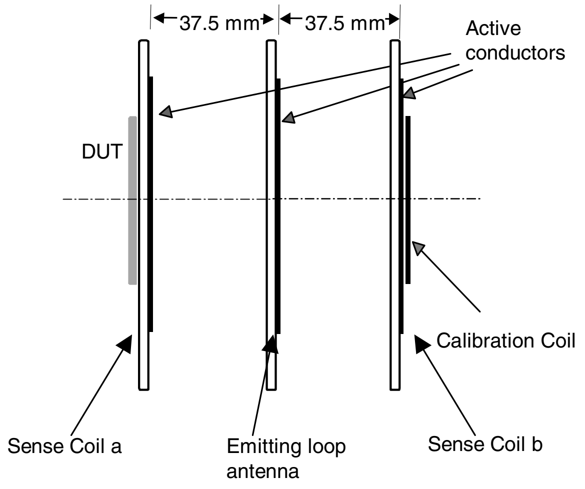

Figure 1. Coax. antenna assembly for contactless transponder test [2].

For a non-resonant 1-turn coil like the Calibration Coil, the induced voltage $u_i(t)$ in the conductor loop is the derivative of the magnetic flux $\Phi(t)$.

$$u_i(t) = -\frac{d\Phi(t)}{dt} \quad (1)$$

For a harmonic sine-wave oscillation at the angular frequency ω, we can use root mean square (*RMS*) values, neglect a 90 ° phase shift and further expand the equation to

$$U_I = \Phi \cdot \omega_C = [B \cdot A_A] \cdot \omega_C = [(\mu_0 \cdot H) \cdot (N \cdot A)] \cdot \omega_C . \quad (2)$$

For the size of the Calibration Coil 1 as given in the test standard [2], for $N = 1$ turn, an antenna area of $A \cong 3000\, mm^2$, for the magnetic field constant of $\mu_0 = 4\pi 10^{-7} Vs/Am$ and for the angular carrier frequency $\omega_C = 2\pi 13.56\, MHz$ this results a relation of about 0.32 A/m(rms) per 1 V(rms) induced voltage.

To consider the antenna voltage for a contactless Smartcard transponder, we have an *LCR* resonant circuit. Maybe the simplest approach is to use a parallel equivalent circuit, as given in fig. 2. The equivalent circuit is split up into antenna inductance L_A and voltage source, and in equivalent parallel transponder capacitance C_T and resistance R_T, which are usually dominated by the chip input impedance. We need to note that mainly R_T, but also C_T are voltage dependent, due to this reason.

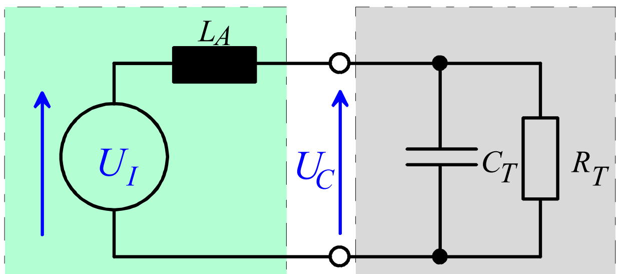

Figure 2. Transponder equivalent circuit for power considerations.

For harmonic sinewave considerations, network calculation allows us in a few steps to calculate the chip voltage U_C at the connection pads as function of induced voltage U_I and of resonant circuit equivalent elements.

$$U_C = U_I \cdot \frac{\dfrac{R_T}{sR_TC_T+1}}{sL_A + \dfrac{R_T}{sR_TC_T+1}} \quad (3)$$

We can give this relation as a function of equivalent network elements, as in (4)

$$U_C = U_I \cdot \frac{1}{1 + j\omega\dfrac{L_A}{R_T} - \omega^2 L_A C_C}, \quad (4)$$

or as function of the more general RF system properties of a resonant circuit, the angular resonance frequency ω_R and the quality factor Q (5):

$$U_C = U_I \cdot \frac{1}{\left| j\dfrac{\omega}{\omega_R} \cdot \dfrac{1}{Q} + \left(1 - \dfrac{\omega^2}{\omega_R{}^2}\right) \right|} \quad (5)$$

Two fundamental equations relate equivalent network elements and resonant circuit system properties: One is the relation for the *natural (angular) self-resonance frequency* ω_R, also known as Thomson equation

$$\omega_R = \frac{1}{\sqrt{L_A C_T}}, \quad f_R = \frac{1}{2\pi\sqrt{L_A C_T}}, \quad (6)$$

the other one is the relation for the quality factor, which we indicate with a T as *transponder system Q-factor* Q_T.

$$Q_T = \frac{R_T}{\omega_R L_A} = \omega_R C_T R_T . \quad (7)$$

If we use the absolute value for the relation of U_C as function of U_I and substitute U_I with (2), we finally get an equation which relates the *H*-field required for chip operation to the equivalent transponder circuit properties, and to the chip voltage.

$$H = \frac{\sqrt{\left[\left(1 - \left(\dfrac{\omega}{\omega_R}\right)^2\right)^2 + \left(\dfrac{\omega L_A}{R_T}\right)^2\right]}}{\omega \mu_0 N A} \cdot U_C \quad (8)$$

This analytical formula is a very useful expression, which we will use for further considerations in this paper.

III. Parameter Variation for Typical Smartcards

In fact, the equivalent circuit for a transponder most commonly used, is a little bit more complex in structure than the circuit given in fig. 2. It consists of the antenna equivalent circuit (L_A, R_A, C_A), a representation for the assembly (R_{AS}, C_{AS}), and the simplest equivalent circuit for the chip (R_C, C_C).

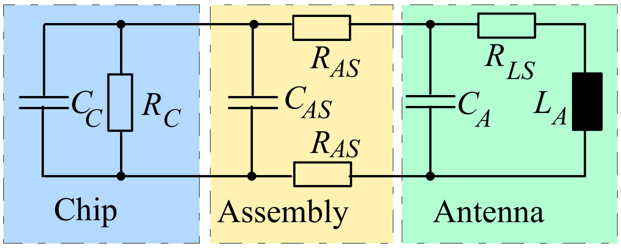

Figure 3. Transponder equivalent circuit for chip, assembly, antenna.

The antenna resistance R_{LS} is the serial conductor resistance of the coil. It is frequency independent, if we neglect the skin-effect and other possible losses (e.g. in the dielectric card material). We can calculate an equivalent parallel antenna resistance R_A using the relation for the antenna Q-factor and the (angular) transponder resonance frequency ω_R, by

$$R_A = \frac{(\omega_R L_A)^2}{R_{LS}} \tag{9}$$

For the chip on the other hand, the equivalent parallel resistance R_C can be considered frequency independent, but is voltage dependent (due to the voltage limiter). Provided that we can neglect the assembly resistance ($R_{AS} \equiv 0$), we can summarize all parallel resistances to R_T and all parallel capacitances to C_T, to end up at the more simple equivalent parallel resonant circuit of fig. 2.

$$\frac{1}{R_T} = \frac{1}{R_C} + \frac{1}{R_A} \tag{10}$$

$$C_T = C_C + C_{AS} + C_A \,. \tag{11}$$

A set of typical parameters for this transponder equivalent circuit is given in table 1.

TABLE 1
TRANSPONDER EQUIVALENT CIRCUIT TYPICAL PARAMETER SET

Parameter	*Value*
L_A	1.8 µH
C_A	2 pF
R_{LS}	1.5 Ω
N	3.8
A_A	0.0014 m²
U_C	2.4 V(rms)
C_C	70 pF
R_C	1.5 k

As can be seen, also the voltage divider derived from the simple equivalent circuit in fig. 2 is more complex according to fig. 3. For typical values nevertheless the transponder resistance R_T is dominated by the chip resistance R_C, and the transponder capacitance C_T is dominated by chip capacitance C_C. So our analytical formula (8) is a valid *approximation*.

As a next step, we can use this analytical formula to consider parameter variations. We assume the minimum chip voltage for operation U_{CMIN} as our operating point, to define typical contactless Smartcard parameter values according to table 1. Unless specified otherwise, these values are used in fig. 4 – 9.

Fig. 5, 7 and 9 show the influence of a variation of the parameters U_{CMIN}, R_T and L_A on the required minimum alternating *H*-field for transponder operation, for the energy-optimum case of resonance equal to the 13.56 MHz carrier frequency. Expressed in equivalent circuit elements, the required minimum *H*-field strength for transponder operation, H_{TMIN}, in principle decreases with decreasing start-of-operation chip voltage U_{CMIN}, with increasing R_T and with decreasing L_A. Expressed in resonant circuit system parameters, this means the required minimum *H*-field decreases with increasing transponder *Q*-factor, for resonance at carrier frequency.

However, this ideal consideration can be misleading in practice, as there are on one hand production tolerances in transponder capacitance and inductance, and on the other hand there is detuning due to coupling of the transponder to the reader in the inductive near field. Both root causes require to consider traces over a resonance frequency range rather, as shown in fig. 4, 6 and 8. Here we can see in fig. 4, a lower U_{CMIN} results in a reduction of H_{TMIN} at carrier frequency f_C and a much wider resonance frequency range (due to a flat trace), while fig. 6 shows that an increase of R_T (which is equal to a reduction of current) decreases H_{TMIN} at f_C, but does not increase much the useful resonance frequency range below the minimum *H*-field for operation required by the standard, H_{SMIN}. A decrease of inductance (and an adequate increase of capacitance) shows an improvement for H_{TMIN} at f_C, but on the expense of a reduced resonance frequency range, as we can see in fig. 8. This rises the question for optimization, especially for the relation of *L* and *C*.

A. Optimum chip input capacitance

One of the important conclusions of chapter II is that the minimum *H*-field for transponder operation, H_{TMIN}, depends on the resonance frequency and on Q_T of the transponder. There are different options in antenna design to vary this parameter, mainly to overcome the limited voltage condition, e.g. using a a matching network [3], an auto-transformer or a transformer [6]. In our context we can also use a variation of inductance and capacitance (to still meet the right resonance frequency) of the transponder [4]. Re-arranging (6) and (7) results

$$Q_T = R_T \sqrt{\frac{C_T}{L_A}} \tag{12}$$

for the quality factor of a parallel resonant circuit, like the equivalent transponder circuit in fig. 2. Assuming the parallel resistance remains unchanged (this means, negligible losses in capacitors), Q_T will increase with the square root of the relation of capacitance to inductance. This increases "Card Loading" to the reader. On the other hand our consideration is only true for the optimum tuning to a resonance frequency equal to carrier frequency in the coupling system. In practice, there are production tolerances for the integrated capacitance (e.g. +/- 10 %), for antenna inductance, and other parameters.

So if nearly all samples out of production (e.g. +/- 6 σ) should meet the base standard requirement of H_{SMIN}, there will be a trade-off. To increase Q_T also means to increase the variance in the behaviour of all parts out of production

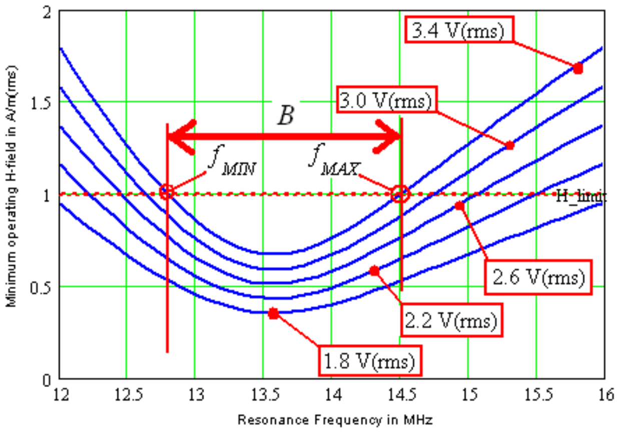

Figure 4. (Required) minimum *H*-field for start of chip operation over resonance frequency, minimum chip voltage U_{CMIN} as parameter.

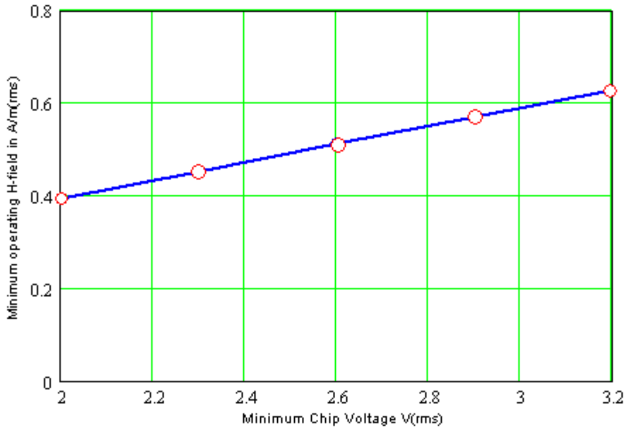

Figure 5. (Required) minimum *H*-field for start of chip operation as function of minimum chip voltage, for resonance at carrier frequency.

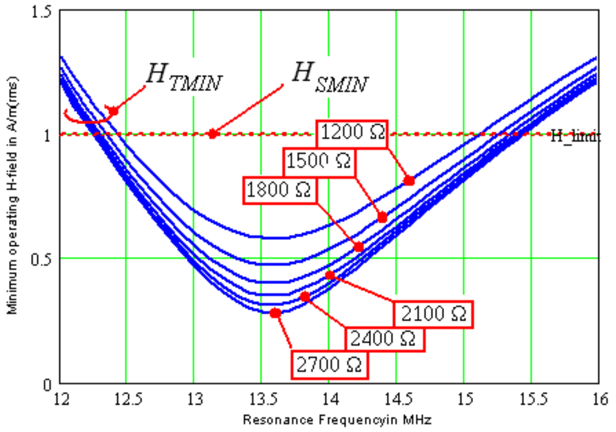

Figure 6. Minimum *H*-field for start of chip operation as function of resonance frequency, equivalent parallel transponder resistance R_T as parameter.

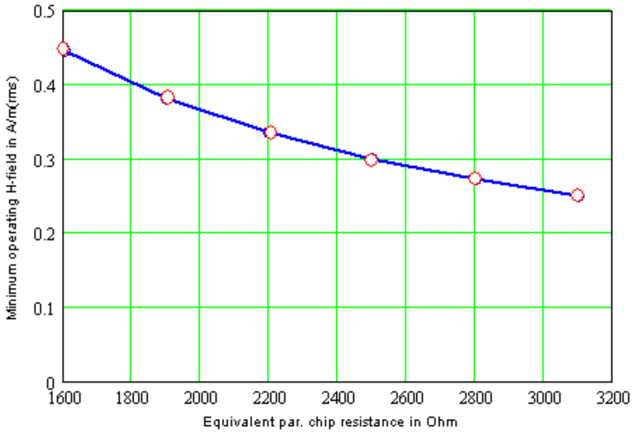

Figure 7. Minimum *H*-field for start of chip operation as function of equivalent parallel transponder resistance R_T, for resonance at carrier frequency.

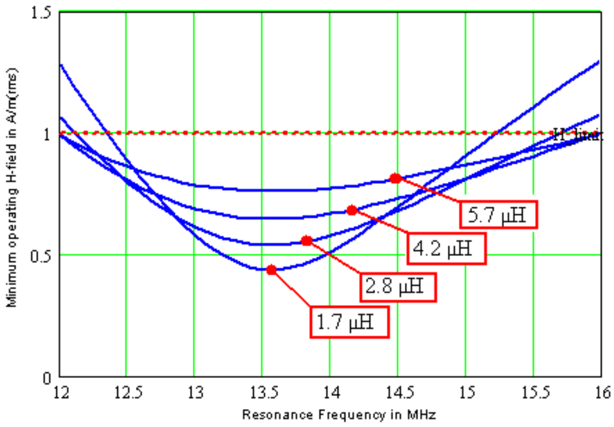

Figure 8. Minimum *H*-field for start of chip operation as function of resonance frequency, antenna inductance L_A as parameter.

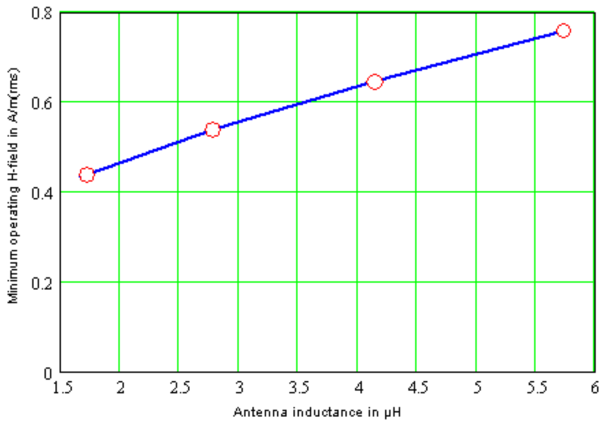

Figure 9. Minimum *H*-field for start of chip operation as function of antenna inductance L_A, for resonance at carrier frequency.

in practice. Re-arranging (8) allows to determine the allowable upper and lower resonance frequency limit f_{LIM}.

$$f_{LIM} = \frac{f_C}{\sqrt{1 \pm \sqrt{\left(\frac{H_{SMIN} 2\pi f_C \mu_0 NA}{U_{CMIN}}\right)^2 - \left(\frac{2\pi f_C L_A}{R_T}\right)^2}}} \quad (13)$$

f_C is the 13.56 MHz carrier frequency, and U_{CMIN} is the minimum chip voltage (rms) for operation. These frequency points are indicated as red vertical lines in fig. 4 (here for an H_{SMIN} limit of 1 A/m).

Another approach to consider an optimum capacitance may be to allow only a certain range for a contactless system property among all produced samples, e.g. an H_{TMIN} increase of a factor $K = \sqrt{2}$ compared to the ideal value, the lowest achievable H_{TMIN} for energy optimum resonance frequency. This approach allows to relate the allowable tolerance to the absolute capacitance value, as shown in fig. 10.

So if we consider the optimum H_{TMIN} at resonance equal to carrier frequency, (8) simplifies to

$$H_{TMIN}(@\,\omega_R \equiv \omega_C) = \frac{L_A U_{CMIN}}{\mu_0 NAR_T (@\,U_C \equiv U_{CMIN})}. \quad (14)$$

Assuming R_T is frequency independent, we can use (13) to calculate a resonance frequency range, or bandwidth B.

$$B = f_{MAX} - f_{MIN} = \frac{f_c}{\sqrt{1 - \sqrt{2} \cdot \left(\frac{f_C \pi L_A}{R_T}\right)}} - \frac{f_c}{\sqrt{1 + \sqrt{2} \cdot \left(\frac{f_C \pi L_A}{R_T}\right)}} \quad (15)$$

Our formula (15) is expressed in equivalent network elements, but of course, it can also be expressed by the contactless system property Q_T. The substitution is given by

$$\frac{1}{Q_T^{\,2}} = \left(\frac{f_C \pi L_A}{R_T}\right)^2. \quad (16)$$

Fig. 10 shows the relation between absolute inductance value and allowable tolerances, for our above mentioned assumptions. The corresponding capacitance to meet resonance at carrier frequency can be calculated from (6) and is given as parameter for antennas used in fig. 8.

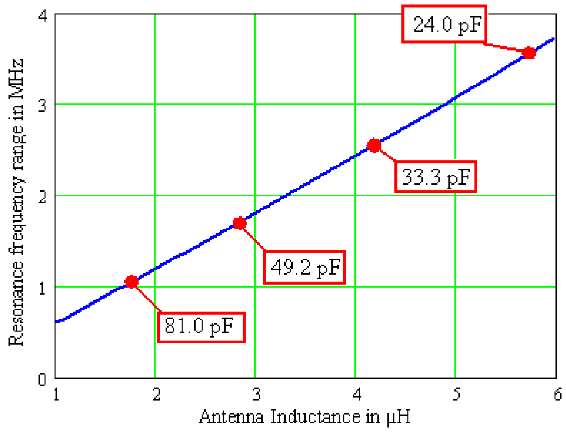

Figure 10. Relation between antenna inductance and allowable resonance frequency tolerance range.

The trade-off for optimum capacitance will depend on achievable production tolerances then.

B. *Antenna size dependency*

The parameters in our analytical formula (8) cannot be varied independently, as we have seen for the set of resonance frequency, transponder capacitance and antenna inductance, in the previous section. For the antenna, inductance, number of turns and area are related. So we can use an analytical formula to calculate inductance as a function of the other parameters, e.g. for a circular planar loop antenna. We will neglect parasitic capacitance and resistance changes in this context. The inductance of a cirular planar loop according to Maxwell [6] is given by

$$L_A = L_{1TURN} \cdot N^2 = \mu_0 \cdot a \cdot \left[\ln\left(\frac{8a}{r_W}\right) - 2\right] \cdot N^2 \quad (17)$$

where N is the number of turns, a is the average loop radius, and r_W is the conductor wire diameter. So it is the inductance of one single turn, times the number of turns to the power of two, or slightly less depending on the wire distance. For the average loop radius as function of average area A

$$A = a^2 \pi \rightarrow a = \sqrt{\frac{A}{\pi}}, \quad (18)$$

we could give (8) for a fixed area as function of number of turns. But in the typcial application case the chip capacitance is fixed (e.g. to 70 pF), and the antenna size should be adapted e.g. to an object. To achieve a constant, energy-optimum resonance frequency equal to the system carrier frequency, we can relate N and A. For this purpose we first solve (17) as a function of the number of turns, N, and then substitute L_A using (6). As result we find (19).

$$N = \frac{1}{\omega_C \sqrt{C_T \cdot \mu_0 \sqrt{A/\pi} \left[\ln\left(\frac{8\sqrt{A/\pi}}{r_W}\right) - 2\right]}} \quad (19)$$

This allows to compare the minimum required H-field for transponders H_{TMIN} as function of antenna area, using (14)

$$H_{TMIN} = \frac{\sqrt{\sqrt{\frac{A}{\pi}} \left[\ln\left(\frac{8\sqrt{\frac{A}{\pi}}}{r_W}\right) - 2\right]}}{\sqrt{\mu_0 C_T \omega_C R_T A}}. \quad (20)$$

So under all these pre-assumptions and if we neglect Card Loading to the reader, H_{TMIN} is nearly inversely proportional to the average antenna area [7].

IV. Considerations for higher H-field strength

A. Available chip current out of the loop antenna

So far we have considered Q_T at the minimum chip voltage for start of chip operation, U_{CMIN}. If the *H*-field increases, also the chip voltage will increase. To provide a constant supply to the digital part in the chip, a shunt regulator usually limits the chip voltage to an U_{CLIM}, which may be higher than U_{CMIN}. For an *H*-field higher as required to generate U_{CLIM}, the antenna basically acts as a current source, and it is then possible to give a simple conversion ratio between the *H*-field at transponder, and the (active) current provided to the chip. In the analytic formula (8) this is reflected accordingly by a change of R_T, which is voltage dependent. Again, we can re-arrange (8), this time to find an expression for R_T out of the system frame conditions.

$$R_T = \frac{\omega_C L_A}{\sqrt{\left(\frac{\omega_C \mu_0 N A H}{U_{CLIM}}\right)^2 - \left[1 - \left(\frac{\omega_C}{\omega_R}\right)^2\right]^2}} \quad (21)$$

$$I_A = \frac{U_{CLIM}}{R_T(@\,\omega_R \equiv \omega_C)} \quad (22)$$

So the above mentioned conversion ratio (in mA per A/m) for the real part of the antenna current I_A is depending on antenna area *A*, number of turns *N*, resonance condition, magnetic field constant, limited chip voltage U_{CLIM} at antenna connection, and the *H*-field$(H \geq H_{TMIN})$. Such a conversion ratio depending on antenna area was already mentioned in a previous work for typical conditions [5].

B. Thermal consideration – maximum power dissipation

The above mentioned consideration can be useful for an estimation, how much current will be available for a chip out of a certain loop antenna area, if it should be compliant to a requirement related to standard and production tolerances. But it can also be used to estimate the upper current limit, and the upper power limit, a chip has to withstand in operation (maximum power dissipation). In this context Q_T will be very low, and there will be nearly no resonance frequency dependency. So the dissipation power P_{TH} can be calculated by

$$P_{TH} = U_{CLIM} \cdot I_A(@\,H \equiv H_{MA}) \quad (23)$$

The maximum alternating *H*-field H_{MA} (for the transponder to withstand without permanent damage) is defined in part 1 of the base standard [1], as long-time average it is 4/3, and as peak value it is 8/5 of H_{SMAX} for each antenna size class, e.g. 10 A/m for class 1 antennas (the operational *H*-field range range, H_{SMIN} and H_{SMAX} are defined in part 2 of the base standard).

We can ask also for this context, if a higher or a lower Q_T brings an advantage. In fact, the chip must be capable to conduct a higher *reactive* current for higher Q_T at U_{CMIN} (start of chip operation) at comparatively low *H*-field. The value can be calculated according to

$$I_C = U_{CMIN}/Z_C = U_{CMIN} \cdot j\omega C \rightarrow |I_C|_{mA} = 0.0852 \cdot U_{CMIN} \cdot C|_{pF} \quad (24)$$

But for high *H*-field, the total current, mainly active current then, exceeds this value. So for thermal considerations at high *H*-field, less antenna turns are an advantage, according to (21), (22). In other words: Q_T can decrease more then.

V. Card Loading and Load Modulation

A. Maximum Card loading

In addition to considerations on parameter variations in the previous section, also requirements for standard compliance must be considered. The test standard [2] specifies a *maximum card loading* to the reader. This requirement is specified as a test case for a card, relative to a so-called "Reference PICC" (a piece of hardware, which emulates the RF system properties of a transponder card) for the appropriate antenna class size. For this purpose e.g. the class 1 Reference PICC is adjusted to a resonance frequency of 13.56 MHz and 6VDC, to be measured at the rectifier. In general RF system properties, this means a specification for the resonance frequency, and the maximum transponder Q-factor at a defined *H*-field strength. Basically the card loading increases with increasing transponder Q-factor, and for readers the sensitivity to Card loading also increases with increasing reader antenna Q-factor. Out of this consideration, it seems appropriate to compare the operational Q-factor for Reference PICC and transponder Card, if we assume an equal loop antenna size. If the transponder Q-factor of a card is below the value for the Reference PICC at H_{SMIN} of the indivicual transponder antenna size class as specified in the base standard, the Card Loading test will be passed by this transponder.

For resonance at carrier frequency (21) simplifies to

$$R_T(@\,\omega_R \equiv \omega_C) = \frac{U_C L_A}{\mu_0 N A H}, \quad (25)$$

which means for the maximum limit of the transponder system Q-factor at this *H*-field strength

$$Q_{TLIM} = \frac{U_{CR}}{2\pi f_C \mu_0 N A H_{SMIN}}. \quad (26)$$

We can calculate Q_{TLIM} for the Reference PICC, for frame conditions as specified in the base standard [1]. As the Reference PICC voltage is specified as DC voltage at a full wave rectifier, and the voltage drop is not known, one critical point is to derive the Ref. PICC antenna coil voltage U_{CR}, which is needed for the calculation. The voltage drop will depend on the input current. For our consideration, we assume

$$U_{CR} = \frac{U_{DC} - 0.9V}{\sqrt{2}} \quad (27)$$

TABLE 2
REFERENCE PICC MEASUREMENT FOR TRANSPONDER SIZE CLASSES

*Class**	*PCD**	H_{SMIN}*	L_A**	U_{DC}*	R_2typ**	ΔH**	Q_{TLIM}***
		A/m(rms)	µH	V_{DC}	Ω	%	
1	1	*1.5*	*2.29*	*6.0*	*975*	*6.9*	*3.0*
2	1	*1.5*	*2.38*	*4.5*	*1191*	*3.1*	*3.3*
3	1	*1.5*	*2.38*	*4.5*	*1308*	*3.2*	*3.6*
4	2	*2.0*	*2.36*	*4.5*	*1074*	*7.2*	*3.0*
5	2	*2.5*	*2.36*	*4.5*	*1092*	*4.7*	*3.1*
6	2	*4.5*	*2.25*	*4.5*	*839*	*2.1*	*2.5*

*) defined, **) measured, ***) calculated from measurement

Our assumptions for Q_{LIM}, the maximum allowable Q_T at H_{SMIN} for each antenna size class as defined in the base standard is given in table 2. These values give a good indication for design, as the Reference PICC antenna size can be considered typical for each class.

B. Maximizing Load Modulation

Load modulation for the reader means a change in the complex antenna voltage (phasor), initiated by a change of the magnetic flux of the transponder, which passes the reader loop antenna. It can be considered as an external modulation, for which no modulation index can be defined. Furthermore, it can be pure amplitude modulation of the RF carrier, or pure phase modulation, or anything inbetween, depending on the coupling conditions and the two resonant antenna circuits.

For the typical case of a larger reader loop antenna and a smaller transponder loop antenna, the magnetic momentum of the transponder is a good parameter for a general consideration, independent of the antenna shape. It consists of the number of antenna turns *N*, the current in one turn *I*, and the average loop antenna area *A*.

$$M = NIA \quad (27)$$

If we consider fixed coupling conditions and antenna size, the transponder anatenna current is the modulated parameter. This current consists of active and reactive current – the transponder Q-factor determines this relation. Q_T is highest for the transponder being quiet, at the lowest *H*-field at which it is functional and can perform load modulation. For "resistive" load modulation, a low-ohmic shunt is switched in parallel to the transponder, in effect decreasing the transponder Q-factor to a low value.

So on one hand, to increase load modulation means to increase this delta *Q*. The maximum value is limited by the minimum chip current consumption, and antenna losses. The minimum value is depending on the shunt implementation – if it is on the DC side of the rectifier, some voltage drop will remain and increase Q_{TMIN} over 0.

On the other hand, there is a requirement for the switching time constant from the base standard [1], which requires a subcarrier frequency of 847.5 kHz.

This trade-off allows from RF system perspective, to find an optimum (= maximum) transponder Q-factor, allowing the maximum load modulation for a passive transponder.

The impulse response of a parallel resonant circuit (like our equivalent transponder circuit) in time domain is given by

$$y_n(t) = Ae^{-(\zeta\omega_n)t} \cdot \cos(\omega_d t + \Phi) \quad (28)$$

where ω_d is the damped angular self resonance frequency, *A* is the initial amplitude envelope, ζ is the attenuation coefficient and ω_R is the natural resonance frequency of the circuit. Q-factor and attenuation coefficient are related by

$$Q = \frac{1}{2\zeta} = \frac{\tau\,\omega_R}{2} = \omega_R R_T C_A = \frac{R_T}{\omega_R L_A} = R_T\sqrt{\frac{C_T}{L_A}}\,. \quad (29)$$

The time constant is given by

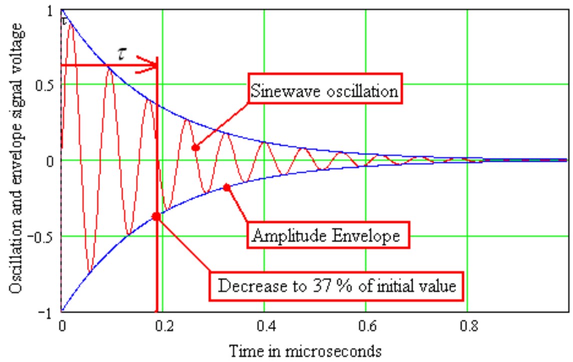

Figure 11. Attenuated harmonic sine wave oscillation as impulse response of a parallel resonant circuit of Q-factor 8.

$$\tau = \frac{2Q}{\omega_R}\,, \quad (30)$$

the damped angular self resonance frequency in time domain is given by

$$\omega_D = \omega_R\sqrt{1-\zeta^2} = \sqrt{\frac{1}{L_A C_T} - \left(\frac{1}{2R_T C_T}\right)^2} \quad (31)$$

and finally the amplitude envelope *e(t)* is given by

$$e(t) = Ae^{-(\zeta\omega_n)t} = Ae^{-t/\tau} = Ae^{-t\cdot\frac{\omega_n}{2Q}}\,. \quad (32)$$

We should remark in this context, we consider Q for the general resonant circuit, in this case for a transponder. In load matching, e.g. for the emitting loop antenna in the test assembly, the operational Q has half this value [15].

Considering the rising amplitude of the signal after a switch from zero to maximum *Q*, we find

$$e_R(t) = 1 - e^{-t\cdot\frac{\omega_n}{2Q}}\,. \quad (33)$$

Integrating over this curve will be a measure for the load modulation phasor. This results in

$$\int_t \left(1 - e^{-t\frac{\omega}{2Q}}\right)dt \;=\; t + \frac{2Q}{\omega}e^{-t\frac{\omega}{2Q}} \quad (34)$$

To find the maximum load modulation, we can determine the extremum by building the derivation of (34) for *Q*. This yields

$$\frac{d}{dQ}\left(t+\frac{2Q}{\omega}e^{-t\frac{\omega}{2Q}}\right)=\frac{2}{\omega}e^{-t\frac{\omega}{2Q}}+\frac{t}{Q}e^{-t\frac{\omega}{2Q}}, \quad (35)$$

which is a maximum in this case.

The specifications of the base standard [1] determine for the data transmission direction from card to reader, data is transmitted in frames, using manchester coding as channel coding. This data modulates a subcarrier frequency by on-off-keying. The subcarrier period is defined to be 16 carrier periods, so we typically have a rectangular wave-shape signal with 8 carrier periods in high, and 8 carrier periods in low state. This signal is used for load modulation, which means a switch of the transponder system Q-factor in our consideration.

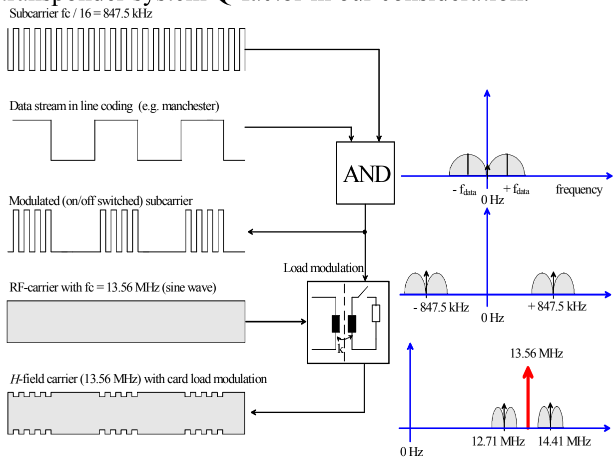

Figure 12. Loadmodulation principle according to [13], modified.

Considering a switch of Q_T between *0* and a maximum value for the non load modulated state, we need to let *t* be half a subcarrier period, and we find an optimum value

$$Q_{TOPT}=\left|-t\cdot\frac{\omega_C}{2}\right|=\frac{8}{f_C}\frac{2\pi f_C}{2}=8\pi\approx 25.13. \quad (36)$$

So under these frame conditions (e.g. neglecting any coupling effects, resonance at carrier frequency), the optimum transponder Q-factor is 25. A lower Q-factor means less reactive current and consequently less (change in) magnetic momentum of the transponder. A higher Q-factor will increase the time constants for the transition, and also decrease the load modulation.

VI. CONCLUSION

We have considered in an analytical way properties of contactless proximity transponders at the communication air interface, on one hand described by general RF system properties, and on the other hand using equivalent circuit elements. We have shown the impact of chip parameters like voltage, resistance and integrated capacitance on the minimum operating *H*-field and the resonance frequency range. In contrast to the ideal case this is essential for practice, as consequence of fabrication tolerances as well as due to (intended) de-tuning due to coupling. We have also considered the loop antenna size and in addition to the minimum required power for chip operation we have also considered maximum current and power conditions, which a product thermally has to withstand. Finally, we have considered the Q-factor for maximum allowed loading, and we have derived an optimum Q-factor to achieve the maximum load modulation for a battery-less transponder independent of antenna size.

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