Analysis of Power Transfer at UHF to RFID ICs by Miniaturized RFID Label Antennas

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INTRODUCTION

The manufacturing cost of RFID labels is an accumulation of IC design, IC manufacture, antenna manufacture, antenna and IC assembly and packaging costs. A significant barrier to reducing the cost is the lack of a streamlined process for attaching RFID label antennas to RFID ICs (about 0.25 mm^2 in size). A large portion, approximately 3 – 4 cents, of the label cost is allocated to the antenna manufacture, antenna and IC assembly and packaging. Hence there is a keen interest to produce a miniaturized on chip antennas to streamline the manufacturing process. This paper considers the feasibility of such an antenna for operation in the UHF ISM band 902 MHz – 928 MHz as defined by the FCC [1], based on analysing the power required to operate an RFID label using coupling volume theory.

The subject of small antennas has been considered in the past in notable publications [2, 3, and 4]. This analysis makes use of both coupling volume theory and radiating antenna theory while taking into account limitations expressed in [5]. Radiating antenna theory, commonly used in radar calculations, is appropriate in the context in which labels are placed in the far field, and when the label antenna size is large enough for the theoretically available source power from a lossless antenna to be actually extracted, or nearly so, within the constraints imposed by the facts that there is some loss, and complete extraction of the available source power is not possible and that the radiation resistance of the antenna is accompanied by a reactance which results in the fact that good transfer of power to an external load can only be accomplished over a limited bandwidth.

Coupling volume theory, first published in [6], is a powerful dimensionless analysis tool with a number of applications presented in [7 and 8]. Coupling volume theory was devised for situations in which labels are placed in the near field, i.e. energy storage field of a transmitter antenna, and also in the situation in which the radiation resistance of the label antenna is small in relation to the losses in that antenna. For operation in the HF ISM band centred at 13.56 MHz, both of these conditions are normally satisfied. For the situation when labels are placed in the far field of an interrogator antenna, but the labels are so small that their own losses are large in relation to the radiation resistance of the label antenna, it is appropriate to use a hybrid version of radiating antenna theory and coupling volume theory. Radiating antenna theory is used to calculate the energy density at the label position, and coupling volume theory is used to work out what useful power the label antenna can extract from the field.

PRACTICAL RANGE OF SMALL ANTENNAS

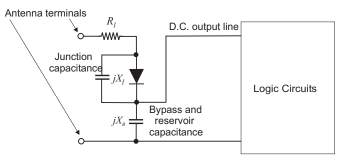

Fig. 1 RFID label schematic.

Considering practicable antennas for RFID applications restricts us to mostly planar structures that can be attached to items, cases and pallets. It is clear from Fig. 1 that the input impedance of an RFID chip is largely dictated by the junction capacitance of the rectification diode. As such the input impedance of an RFID chip at the threshold of operation is capacitive. Hence a planar inductive antenna is preferred over a capacitive antenna. Thus a magnetic dipole antenna whose output impedance is largely inductive will be considered in the following sections. It appears to be possible in many cases to arrange for resonance between these two elements to occur at the frequency of operation. It is well known in coupling volume theory that such resonance is required to make good use of the power transferred to the label antenna.

However before such actions can occur, it is necessary to produce some rectification. As discussed previously Schottky diode rectifiers have a junction capacitance which must be periodically charged and discharged to a sufficient potential for rectification to occur. That process involves an exchange of reactive power between the junction capacitance and some complimentary energy storage element. One of the benefits of magnetic dipole antennas is that the self inductance necessarily present in the antenna can be that energy storage element provided it has been suitably adjusted, and suitable adjustment is available by appropriately choosing the number of turns.

SMALL ANTENNAS

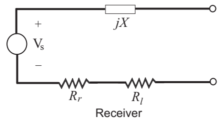

Fig. 2: Antenna equivalent circuit.

Fig. 2 shows an equivalent circuit for an electrically small antenna, operating in its receiving role. It shows that there is a radiation resistance R_r in series with an antenna reactance *jX*. When losses are to be taken into account the antenna will also have a loss resistance R_l. The value of R_r for a single turn loop of radius *a* where the propagation constant is β is given by (1).

$$R_r = 20\pi^2 (\beta a)^4 \quad \Omega . \tag{1}$$

The optimum load impedance for such an antenna is $R_r + R_l - jX$. When the antenna becomes very small, the radiation resistance of a loop antenna given by (1) reduces, however $R_r << R_l$. In this situation it is possible to apply coupling volume theory to determine the circuit behaviour. In this theory, the source voltage in the above antenna circuit might as well be calculated from Faraday's law, the radiation resistance neglected, the self inductance calculated from the magnetostatic formula, and the loss resistance determined taking into account that conduction will only occur within a skin depth of the metal surface.

MINIATURE ANTENNA PROPERTIES

A principal target of the analysis will be the determination of the feasibility of a miniaturised antenna providing adequate reactive power to the junction capacitance of a small rectifier diode. The analysis will be based on an operational frequency of 915 MHz. The antenna material will for a number of reasons be assumed to be copper with an electrical conductivity of 5.8×10^7 S/m. The antenna is assumed to be a square of side 2 mm. Where appropriate, the square shape may be replaced with a circle of equal area. Although the antenna coil will undoubtedly be realised as a planar structure with a flat strip conductor, the flat strip by may be replaced with an equivalent round wire of diameter half the width of the strip. The RFID interrogator will be assumed to radiate a power of 1 W through an antenna of gain 4 (6dBi), as is permitted in the FCC frequency hopping regulations [1]. An interrogation distance of 1 m will be assumed as adequate, which is well into the far field, but is not at such a distance as to represent an unfair challenge to the feasibility of the system under investigation. The skin depth δ, for copper at a frequency of 915 MHz is 2.18 μm. It should be noted here that it is feasible to deposit material for an antennas to at least this depth however depositing more metal will not produce an added benefit as the metal deeper than one skin depth will not contribute to the conduction. It follows that all resistance calculations are based on the assumption that inductors of all sizes are made from strip material of the above thickness.

$$R_s = \frac{1}{\delta\sigma} = \sqrt{\frac{\omega\mu}{2\sigma}} \tag{2}$$

Calculation of the coil resistance involves the determination of the surface resistivity R_s for the selected material and frequency. The surface resistivity is evaluated to be 7.909 mΩ per square using (2) where σ is the conductivity of copper, μ is the permeability of copper and ω is the angular frequency. Using the surface resistivity obtained above and a diameter ratio (diameter of the circle to the diameter of the wire) of 5, a coil loss resistance of R_l = 39.55 mΩ can be caluculated. In this scenario it is reasonable to use the coupling volume theory, in which radiation resistance is assumed to be negligible. This can be confirmed using (1). Provided the number of turns are not changed, an antenna in the form of a single circular coil of round wire in which the diameter ratio is a constant will have a constant resistance, independent of size.

$$L = \frac{\mu_o D}{2}\left[\mathbf{log}_e\left(\frac{8D}{d}\right) - 2\right] \tag{3}$$

The self-inductance of a single turn coil of diameter *D*, made from around wire of diameter *d* can be calculated by (3). Applying (3) to a single turn square coil of size 2 mm by 2 mm, with a diameter ratio *D/d* of 5 obtained by approximating it to a circular coil having the same area, the coil has a self-inductance L of 2.122 nH. At the operating frequency of 915 MHz, the reactance X of this self-inductance is 12.20 Ω. From the coil reactance and the loss resistance R_l we can calculate an inductor quality factor Q, of 308.

For a coil of N turns, both the self inductance and the coil resistance will both scale as the second power of N, and the series induced voltage will scale as the first power of N (provided the same amount of surface area is allocated to the provision of those turns as was allocated to the provision of a single turn). Most of the calculations below, such as the available source power, the short-circuit reactive power, the coupling volume, and the quality factor will be unaffected, and therefore are based on calculations for a single turn coil. This is one of the interesting properties of the theory.

REACTIVE POWER DENSITY PER UNIT VOLUME

$$\text{Power flow per unit area} = \frac{g_t P_t}{4\pi r^2} \text{ Wm}^{-2}. \tag{4}$$

Since the label is assumed to be in the far field, the reactive power density per unit volume can be calculated by the real power flow per unit area using the radiation antenna theory formula given in (4) where g_t is the antenna gain, P_t is the transmit power and r is the distance to a point in the field from the antenna. It can be shown that the reactive power density per unit volume can be obtained from the real power flow per unit area by multiplying by the propagation constant β [7] providing the formula below.

$$\text{Reactive power density per unit volume} = \frac{\beta g_t P_t}{4\pi r^2} \text{ Wm}^{-3} \tag{5}$$

The result, under the assumptions of interrogator power, antenna gain, and interrogator to label distance defined previously, is a reactive power density per unit volume of 6.1 VAm^{-3}. The power density given in (5) is an appropriate, power-like, measure of the field strength available for energising the RFID label.

LABEL COUPLING VOLUME

$$V_c = \frac{\text{Reactive power in the label indcutor when short circuit}}{\text{Reactive power density per unit volume at label position}} \tag{6}$$

The coupling volume V_c of a label can be defined as provided in (6). While the coupling volume for a planar coil of N turns area A and self inductance L is given by [6]

$$V_{c(Label)} = \frac{\mu_0 N^2 A^2}{L}. \tag{7}$$

For the coil antenna under consideration, with a flux collecting area of 1.6 mm by 1.6 mm, we find we have a coupling volume of $V_{c(Label)}$ = 3.881 x 10^{-9} m^3. Since the coupling volume does not change with the number of turns as discussed previously it is possible to alter the number of turns later to achieve a different inductance if that is desired to produce resonance with the junction capacitance of the rectifier.

REACTIVE POWER IN SHORT CIRCUIT LABEL

$$W_{S(Label)} = \left(\frac{\beta g_t P_t}{4\pi r^2}\right) V_{c(Label)} \text{ VA} \tag{8}$$

The reactive power $W_{S(Label)}$, which would flow within the label coil when it is short-circuited is given by the product of (5) and (7) in (8), resulting in a short circuit reactive power of 23.67 nVA. Since coupling volume scales as the third power of size (refer to (7)), obtaining a greater reactive power clearly involves increasing the label antenna size.

POWER DELIVERED TO A TUNED LABEL

The power $P_{R(Loss)}$, which will be delivered to the losses that exist within the label when it is tuned to resonance (thus we assume that the coil antenna is tuned to resonance by the capacitive load provided by the RFID IC) can be obtained using the standard result from coupling volume theory for coils coupling to the magnetic field given by (9)[8]. If we assume from our previous quality factor calculation a Q of 308, the a power of 7.29 µW can be delivered to the losses of the tuned circuit. There is a substantial quality factor reduction as a result of the introduction of a real load to which power must be delivered. If the substantial quality factor reduction has occurred through the introduction of a real load to which power must be delivered, then the calculated power is going substantially to that real load.

$$P_{R(Loss)} = W_{S(Label)} Q = \left(\frac{\beta g_t P_t}{4\pi r^2} \right) V_{c(Label)} Q \ \text{ VA} \tag{9}$$

REACTIVE POWER IN TUNED COIL

$$\mathrm{W}_{R(Coil)} = \left(\frac{\beta g_t P_t}{4\pi r^2} \right) V_{c(Label)} Q^2 \ \text{ VA} \tag{10}$$

The reactive power $W_{R(Coil)}$, which flows in the inductor, and also in the capacitance that it feeds at resonance can be calculated by (10) where (9) is multiplied by the quality factor of resonance [8]. For a quality factor of 308, a reactive power of 2.25 mVA can be calculated.

REACTIVE POWER NEEDED IN THE DEPLETION LAYER CAPACITANCE

Assuming that the rectifier requires a 1.5 V r.m.s. voltage across it for a useful output, and that the junction capacitance is approximately 0.1 pF, a reactive power of 1.294 mVA is needed in the depletion layer capacitance.

CONCLUSIONS

If the quality factor of 308 can be achieved then it is possible to generate a reactive power in excess of that needed to service the junction capacitance of the rectifier diode. However there is room for other sources of loss such as diode losses and losses in the rectification process. Nevertheless the results show that in the absence of any other losses substantial power transfer can take place, but only over extremely narrow bandwidths but when real losses in practically available materials are taken into account, or a greater operating bandwidth is desired, the picture becomes less rosy.

Estimates of the power likely to be required to run a backscatter label circuit is about 20 μW, which is in excess of quarter of the power calculate by (9) (the power available to a real impedance when the chip is a perfect match to the antenna). Thus with a high-quality factor inductance, it is probably possible to obtain an adequate rectified voltage from an unloaded rectifier, but obtaining adequate output power to run an RFID label is doubtful. It should be noted here that the power calculated as being able to be extracted from the antenna at the quality factors of 308 is still much less than the theoretically available source power which a completely lossless receiver antenna at this distance could provide by radiation antenna theory, which predicts an output power of 4.8 mW. However such an output power is unachievable in practice due to large ohmic losses in the small label antenna.

This analysis explains that it is not possible contemplate the use of tiny antennas for UHF RFID labels except for operation at very close range. Probably the only beneficial effect of such tiny antennas might be that they are easier to prevent from becoming environmentally detuned because their self fields extend only a small distance from the antenna. It should also be noted that the conclusions drawn here depend upon assumptions made about the feasible rectifier capacitance, achievable quality factor in coils, and label circuit current drain.

REFERENCES

[1] Federal Communication Authority (FCC), web page, http://www.fcc.gov.
[2] H. A. Wheeler, “Small Antennas”, *IEEE Trans. Antennas and Propagat.*, vol. AP-23, July 1975, pp. 462-469.
[3] H. A. Wheeler, “Fundamental Limitations of Small Antennas”, *Proc. IRE*, vol. 35, Dec. 1947, pp. 1479-1484.
[4] J. S. McLean, “A Re-Examination of the Fundamental Limits on the Radiation Q of Electrically Small Antennas”, *IEEE Trans. Antennas Propagat.*, vol. 44, no. 5, May 1996, pp.672-676.
[5] L. J. Chu, “Physical Limitations of Omni-Directional Antennas”, *J. Appl. Phys.*, vol. 19, 1948, pp. 1163-11175.
[6] K. Eshraghian, P. H. Cole, and A. K. Roy, “Electromagnetic coupling in subharmonic transponders,” *Journal of Electrical and Electronics Engineering*, No. 2, pp. 28–35, 1982.
[7] P. H. Cole, D. C. Ranasinghe and B. Jamali, “Coupling relations in RFID systems II: practical performance measurements”, *Auto-ID Center workshop*, Sept. 2003.
[8] P. H. Cole, “Coupling and Quality Factors in RFID”, *Proc. of SPIE Design, Characterisation and Packaging for MEMS and Microelectronics*, vol. 4593, 2001, pp. 1-11.