AN 1629 UHF RFID Label Antenna Design

UHF Antenna Design

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Document information

Info	Content
Keywords	UHF, label antenna design, UCODE
Abstract	This document provides a general overview on basics of UHF wave propagation, as well as practical considerations of UHF label antenna design. The target is to guide the reader to a good understanding of UHF label antenna design in theory and in practice.

founded by Philips

Revision history

Rev	Date	Description
1.0	2008 / 09	First, initial release; Author : BR

1. Introduction

The aim of following UHF label antenna design guide is to provide the reader the necessary theoretical background on UHF as well as a practical insight on label antenna design.

In Chapter 2 the basics of UHF wave propagation are introduced. In order to be able to design a label antenna fitting the end application, the knowledge of the physical properties of an UHF wave is vital.

Chapter 3 deals with fundamental antenna parameters. The key characteristics of an antenna are described in the beginning. The second part of Chapter 3 shows schematic equivalent circuits of the components of the “system” label antenna.

Chapter 4 puts the theory of the preceding sections into practice. It solves the key question, of which steps to follow when designing an UHF label antenna.

Starting with the definition of parameters, it explains more details on a loop-dipole structured antenna. By means of an example (simulated in 3D EM simulator software), the antenna parameters are depicted and possibilities of antenna tuning are proposed.

2. UHF Wave Propagation

2.1 Freespace Transmission

A free space transmission system consists typically of a transmitting antenna, a transmission path, and a receiving antenna. Parameters of every component determine the functionality of the entire system.

The main criteria hereby are transmitted power, antenna parameters (gain, polarization and reflection coefficient), wavelength and distance between the transmitting and the receiving antenna. The mathematical relation between these components is described by the Friis Transmission Equation (equation 1):

$$P_R = P_T \left(\frac{\lambda}{4\pi R} \right)^2 G_T G_R \quad (1)$$

P_R: Transmitted Power

P_T: Received Power

R: Distance between transmitting and receiving antenna

λ: Wavelength

G_T: Gain of the transmitting antenna [dBi]

G_R: Gain of the receiving antenna [dBi]

G is measured relative to an isotropic antenna [dBi]

2.2 Absorption

Only vacuum is passed by electromagnetic energy without absorption. If an electromagnetic wave looses energy and this energy is converted into other forms, this process is called *Damping* or *Attenuation*.

The absorption is given by the imaginary part of the refractive index $\underline{n} = \sqrt{\underline{\varepsilon}_r}$.

The absolute value of the field strength decreases along the propagation path.

The electrical field strength at a certain position x depends on two parameters: Distance (x) from the RF source an the properties of the media along the transmission path

The propagation attenuation (or path loss) is calculated by referring to the ratio between the emitted electrical field strength at the source (antenna) and the electrical field strength at a certain distance x, after the transmission via the media. The propagation attenuation in dependency of the distance is given by equation 2.

$$A[dB] = 20 \lg \left| \frac{E(0)}{E(x)} \right| \quad (2)$$

The free space path loss at distance *d* is:

$$a_0[dB] = 20\lg\frac{4\pi d}{\lambda} \qquad (3)$$

Many dielectric materials, like dry paper or cardboard, dry wood, nonconductive plastics, most textiles, are substantially non absorbing and have modest refractive indices (ε_R 2–4) for 900-MHz radio waves.

Metals reflect essentially all the radiation that falls upon them. Water, with a dielectric constant of around 80, also reflects almost all of an incident wave and absorbs most of the rest.

The dielectric plays especially in the UHF RFID frequency range an important role, as the information transmission mainly takes place over the far field, where the electrical component (opposed to the magnetic component) is predominating.

The electrical flux density D is depending on the electric field E and the permittivity of the media ε. The permittivity can be considered as a scalar or a tensor, the D and the E field are parallel, see equation (4).

$$\mathrm{D} = \varepsilon\,\mathrm{E} = \varepsilon_0\varepsilon_\mathrm{R}\,\mathrm{E} \qquad (4)$$

ε_0 denotes the vacuum permittivity and ε_R the dielectric constant of the transmission media.

A label antenna applied on a carrier material of a high dielectric constant, has to be able to handle two effects, in order to keep a good performance:

Detuning: Detuning of the label resonance frequency will lead to mismatch and to performance loss. It needs a dedicated antenna design for high dielectric carriers.

Attenuation: Depending on the imaginary part of the dielectric constant of the media (carrier material), a certain amount of UHF waved will be absorbed and reduce as well the read range.

In order to be able to closely investigate the impact of different materials on label performance and to enable professional analysis of this behavior NXP composed a set of reference materials with various permittivity values (ε_R), as shown in Figure 1.

Table 1. Permittivity Values of the Reference Materials

PTFE	PMMA	PC	PET	PU/PUR	KITE	CARP	Glass
2,12 – 2,2	2,98 –3,02	3,19 –3,23	3,64–3,68	4,05 – 4,12	5,75 – 5,81	5,78 – 5,87	12,3 – 12,5

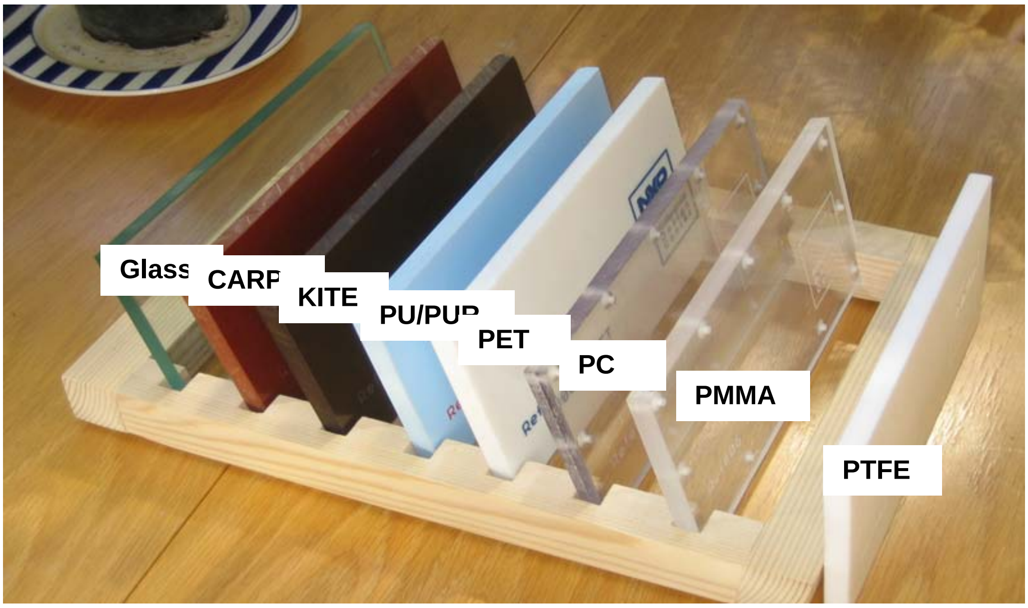

Fig 1. NXP Reference Materials

2.3 Reflection

A pure reflection of the traveling wave will conserve energy of the field. Wave-guides are an example of devices that uses these pure reflections for transmission. Multipath wave propagation leads to constructive or destructive interferences – one of the unwanted phenomenon's in the UHF area. There are two different kinds of reflection. The directional reflection occurs on plain surfaces and the diffuse reflection occurs on rough surfaces (Fig 2).

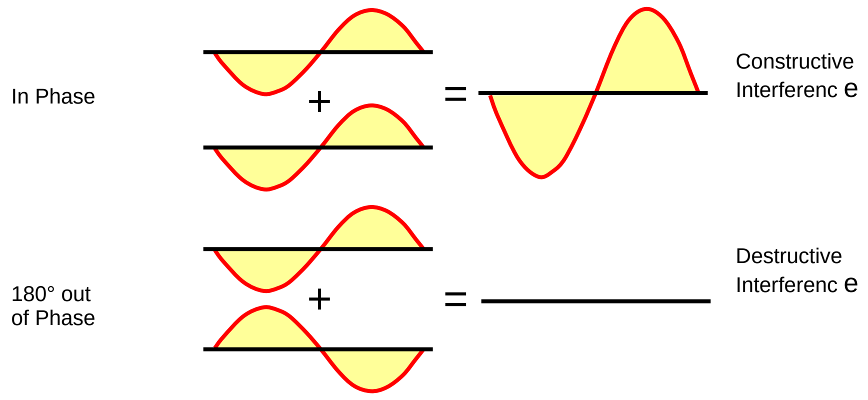

Fig 2. Constructive and destructive interferences

2.4 Diffraction and Refraction

When a wave passes from one medium into another medium that has a different velocity of propagation, a change in the direction of the wave will occur. This changing of direction as the wave enters the second medium is called refraction.

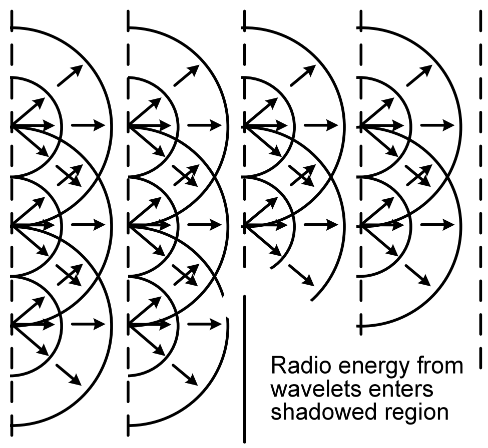

Fig 3. Diffraction

Diffraction is the bending of a wave around objects or the spreading after passing through a gap (Fig 3). It is due to any wave's ability to spread in circles or spheres in 2D or 3D space. Huygens' Principle is based on this process. In 1678 the great Dutch physicist Christian Huygens (1629-1695) wrote a treatise called 'Traite de la Lumiere' on the wave theory of light, and in this work he stated that the wave front of a propagating wave of light at any instant conforms to the envelope of spherical wavelets emanating from every point on the wave front at the prior instant (with the understanding that the wavelets have the same speed as the overall wave).

2.5 Polarization

The orientation of the electrical field component of an electromagnetic wave determines its polarization. There are three different polarizations possible. An *elliptical* polarization has an elliptical electrical field vector in the plane perpendicular to the propagation direction. Additionally there are the special forms of *circular*- and *linear* polarization.

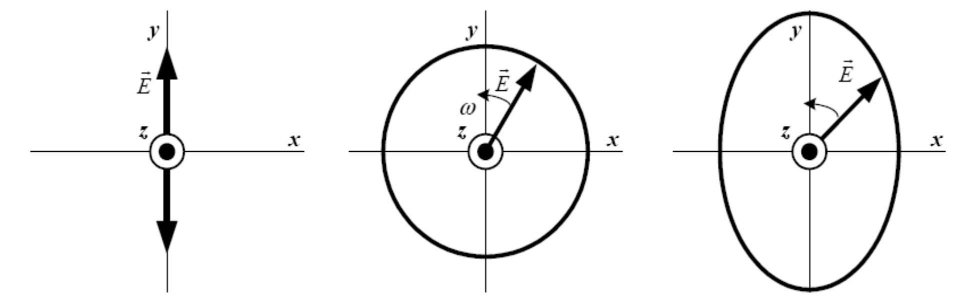

Fig 4. Linear-, circular-, elliptical polarization

Depending on the rotation direction of the circular polarization it is called *left*- or *right circular* polarized, where right means clockwise. In linear polarization, the electrical field vector has a relation to the earth's surface. Depending on this orientation one can differ between two forms of linear polarization. The *horizontal* polarization, where the electrical field vector is aligned parallel to the earth's surface, and the *vertical* polarization, where the electrical field vector is perpendicular to the earth's surface. In circular polarization, the wave front appears to rotate every 90 degrees between vertical and horizontal, making a complete 360-degree rotation once every period.

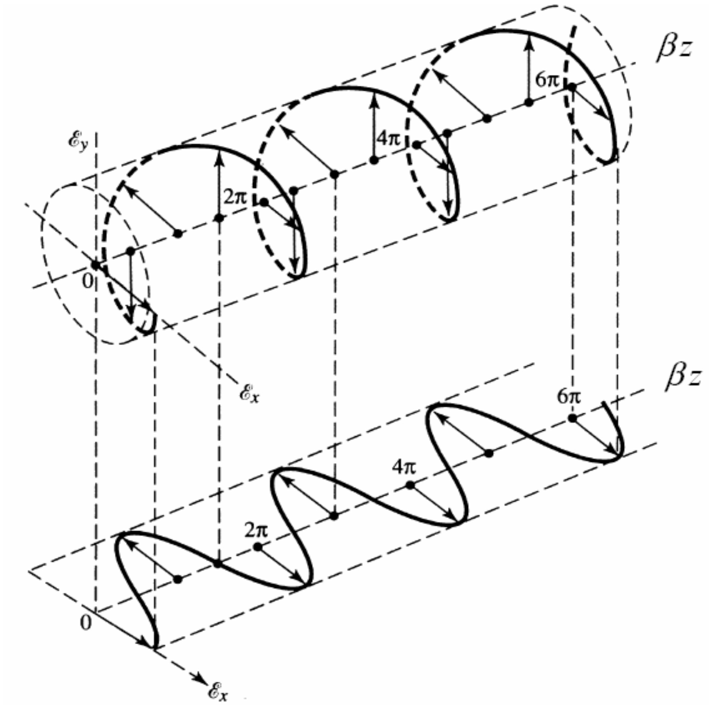

Fig 5. E-Field vector projection of a circular polarization

Label Orientation	Reader Antenna Polarisation: circular	vertical	horizontal
vertical	3dB	0dB	∞ dB
horizontal	3dB	∞ dB	0dB
slant	3dB	3dB	3dB
parallel to antennabeam	∞ dB	∞ dB	∞ dB

Fig 6. Polarization mismatch

The power transfer depends on the orientation of the receiving and transmitting antenna (Fig 6). Linear polarized antennas (horizontal or vertical) have 0dB loss if they are oriented in the same plane. If one antenna is linear and the other circular polarized the loss is always 3dB. However if a linear polarized label antenna (dipole) points direct longitudinal towards a reader antenna no power is transferred and the loss is infinite.

If the orientation of the two polarizations (label and reader) is considered in the vector plane, the efficiency of the power transfer (tag →reader / reader→tag) depends on how much the vector directions overlap. The non – overlapping components will result into losses.

3. Fundamental Antenna Parameters

3.1 Radiation Pattern

An antenna radiation pattern or antenna pattern is defined as a mathematical function or a graphical representation of the radiation properties of the antenna as a function of space coordinates.

The output characteristic of an antenna is described by the radiation pattern, which is the relative distribution of the radiated power as a function of the direction in space. Even if the radiation pattern is 3 dimensional very often only planar sections (cuts) are presented. The two most important views are those of the principal E-plane and H-plane patterns. The first one is a view of the radiation pattern obtained from a section containing the maximum value of the radiated field and in which the electrical field lays in the plane of the chosen sectional view. The later plane pattern is a sectional view in which the H field lies in the plane of the section, and this section is chosen to contain the maximum direction of radiation.

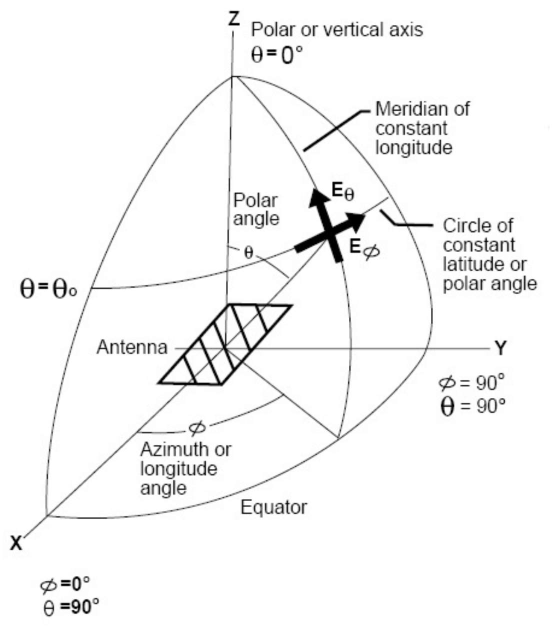

Fig 7. Radiation pattern

3.2 Field Regions

At high frequency measurements in the UHF region (and beyond) parasitic effects do play an increasing important role. In particular for radiating structures as e.g. antennas it is getting more difficult since every conducting- or non-conducting material (with a dielectric constant and/or a permeability constant greater than 1) located in the reactive near-field region causes changes of the boundary condition of the antenna (modifying the coupling between the antenna elements) and therefore changes its characteristic.

In general the space around a radiating element (e.g. an antenna) is divided into three different regions (Fig 8): The reactive near field region, the radiating near field (Fresnel) region, and the far field (Fraunhofer) region. Depending on the antenna design and geometry these regions can be defined as:

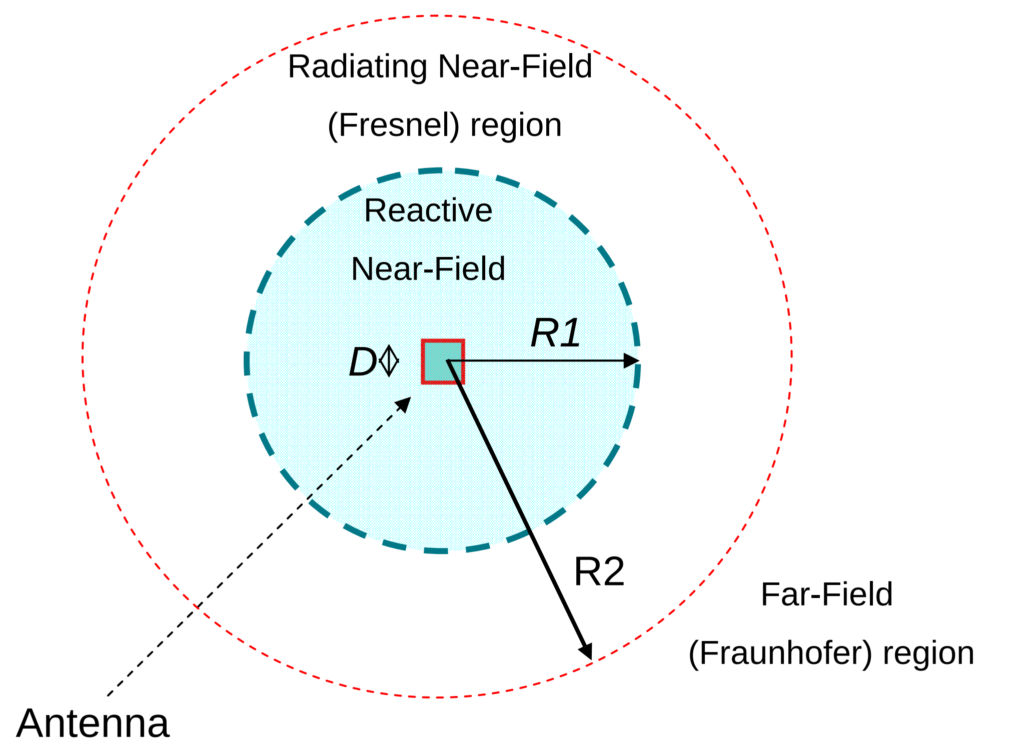

Fig 8. Field regions around an antenna

Whereas D is the largest dimension of the antenna, and R denotes the distance.

The three regions are discussed more in detail in the following sections.

3.2.1 Reactive Near Field Region

Reactive near field region is defined as that part of the near field region immediately surrounding the antenna wherein the reactive field predominates.

A commonly used formula for the boundaries of the reactive near field region is equation (5):

$$R_1 \leq 0.62\sqrt{\frac{D^3}{\lambda}} \tag{5}$$

The field decays rapidly in the order of $1 / R^4$.

Example: Assuming a typical UHF application, and substituting the parameters of equation 5 with real values (D= 25cm, f= 868 MHz, λ= 34.56 cm), using λ = c/f, we get to a reactive near field region with a radius of 13.1 cm.

3.2.2 Radiating Near Field (Fresnel-) Region

The radiating near field region is located between the reactive near field region and the far field region.

The boundaries are defined as follows in equation (6):

$$0.62\sqrt{\frac{D^3}{\lambda}} \leq R_2 \leq \frac{2D^2}{\lambda} \tag{6}$$

If an antenna has a maximum dimension that is not large compared to the wavelength, this region may not exist.

Example: Assuming a typical UHF application, and substituting the parameters of equation 6 with real values (D= 25cm, f= 868 MHz, λ= 34.56 cm), using λ = c/f, we get to a radiating near field region with a radius of 35.7 cm.

3.2.3 Far Field (Fraunhofer-) Region

The relative angular distribution does not vary with distance in the far-field region. The radiated power from an antenna decays in the far-field region according to the inverse square law ($1/R^2$) as a function of the distance. The Fraunhofer region is defined for distance as given in equation (7).

$$R \geq \frac{2D^2}{\lambda} \tag{7}$$

Example: Assuming a typical UHF application, and substituting the parameters of equation (7) with real values (D= 25cm, f= 868 MHz, λ= 34.56 cm), using λ = c/f, we get to the far field region if the distance to the antenna is greater than 35.7 cm.

3.3 Input Impedance

The input impedance is the parameter, which describes the antenna input behavior as a circuit element. As usual in electronic circuit design it is important to match this input impedance ($\underline{Z}_{antenna}$) to a given source impedance, which is in case of RFID applications usually the chip impedance of the transponder IC ($\underline{Z}_{chip}$). The maximum power delivered from the source to the antenna is given if the antenna input impedance is complex conjugate to the chip impedance.

Practical: The better the matching, the less power is reflected on the connection between IC bumps and label antenna pads. This will result in a maximum power transfer to the IC. An UHF RFID IC is an ultra low power design, therefore matching is essential in order to make the good sensitivity of the IC result in excellent read ranges in an end application.

Further investigation on this matter can be found in chapter 3.9.

3.4 Radiation Resistance

The radiation resistance of an antenna is equivalent to a resistance that would dissipate the same amount of power as the antenna radiates, when the current in this resistance is equal to the current at the antenna input terminals.

The total resistance of an antenna ($R_{antenna}$) can be separated into a series circuit of two

$$R_{antenna} = R_{rad} + R_{loss} \tag{8}$$

different resistors (equation 8).

Where R_{rad} is the radiation resistance and R_{loss} is the loss resistance representing the unwanted losses caused by the non-perfect conductors and (substrate) materials.

$$\mathrm{R}_{\mathrm{loss}} = \mathrm{R}_{\mathrm{loss,\,con}} + \mathrm{R}_{\mathrm{loss,\,sub}} \tag{9}$$

Accordingly one can further separate this loss resistance into those two contributions (equation 9).

3.5 Gain and Directivity

The ratio of the radiation intensity in any direction *d* to the intensity averaged over all directions is the directive gain of the antenna in that direction. The directive gain along the direction in which that quantity is maximized is known as the directivity of the antenna. The directivity of the antenna multiplied by the radiation efficiency is the gain of the antenna.

In the direction of maximum radiated power density, we get *G* times more power that we would have obtained from an isotropic antenna.

The gain of an antenna gives its ability to focus the radiated power in a particular direction relative to an isotropic point source and is given by:

$$Gain = 4\pi r^2 \left(\frac{S}{P_{in}} \right) \tag{10}$$

The far field power density S (equation 11) can be calculated using the equation 10:

$$S = \frac{Pin \cdot Gain}{4\pi d^2} \quad \left[\frac{W}{m^2}\right] \tag{11}$$

Pin ……Average Power at Antenna [W];

Gain ….Numerical Gain;

d……….Distance from Antenna [m].

More generally the gain is related to the directivity of the antenna by the efficiency of the antenna e

$$G(\theta,\phi) = eD(\theta,\phi) \tag{12}$$

where *D* is given by

$$D(\theta,\phi) = \frac{\text{power radiated in a fixed direction}(\theta,\phi)}{\text{total power radiated by the antenna}} \tag{13}$$

In national UHF frequency regulations, radiated power is defined with different terms in the different countries. The two most common terms are EIRP and ERP.

EIRP (Equivalent Isotropic Radiated Power) – mainly used in the US. Fig 9 shows how a directive antenna is related to an isotropic antenna.

ERP (Effective Radiated Power) – mainly used in Europe. The EIRP is related to the ERP by the following relation:

EIRP = ERP * 1.64

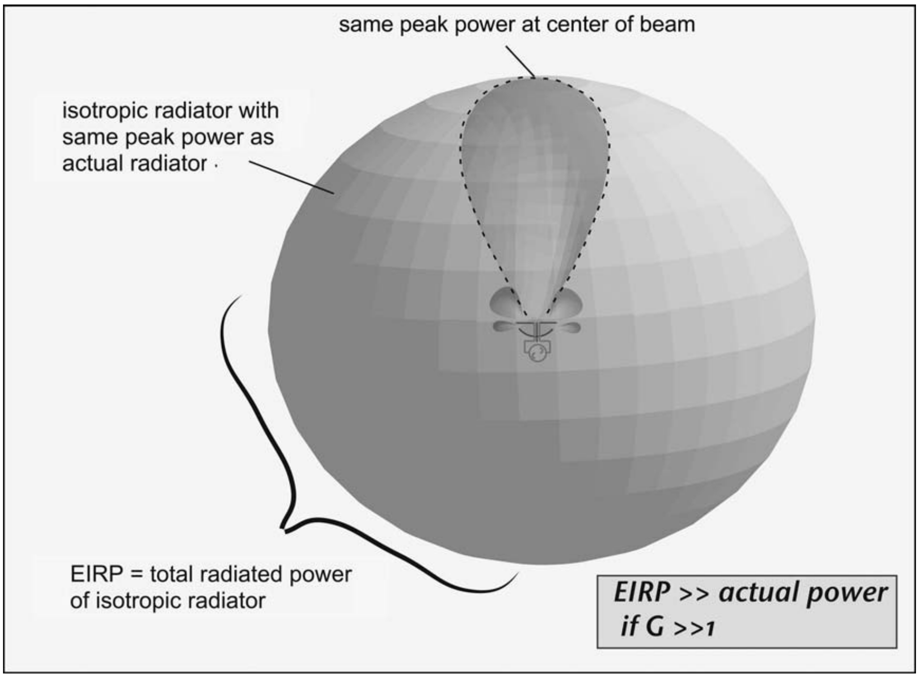

Fig 9. Definition of effective isotropic radiated power

3.6 Efficiency

The efficiency of an antenna takes into account the influence of additional losses, which may occur at the input terminals or within the structure of the antenna.

It is the percentage of the power delivered to the antenna that actually gets radiated as opposed to being absorbed or reflected.

For an ideal, lossless antenna, the gain equals the directivity (equation 12).

Practically, these can be losses due to reflection (mismatching between the IC and the antenna), or conduction and/or dielectric losses.

3.7 Bandwidth and Q

The bandwidth can be considered to be the range of frequencies, on either side of a center frequency (for example the resonance frequency of a dipole), where the antenna characteristics (such as input impedance, pattern, beam width, polarization, side lobe level, gain, beam direction, radiation efficiency) are within an acceptable value of those at the center frequency.

An antenna simultaneously stores charge (capacitance), opposes changes in current (inductance), and radiates power into the wide world (resistance). From the electrical point of view, an antenna looks like an R-L-C circuit. The configuration of the circuit depends on the antenna type. A dipole that is short compared to the wavelength looks like a series combination of an inductor and capacitor with some resistance (see Fig

10). The inductance and capacitance are both roughly proportional to the length of the dipole;

In general, the bandwidth is inversely proportional to the Q factor, which is the ratio of the total reactance to the resistance. For a simple series resonant circuit, Q is about twice the voltage amplification factor .Thus, voltage amplification must be traded against bandwidth. Antennas with large reactance (that is, large values of inductance and small values of capacitance) and small values of resistance may be adjusted to be matched to the tag at one frequency, with good power transfer and voltage multiplication, but performance will degrade at other frequencies. Antennas with small reactance will provide better performance over frequency.

$$BW \sim \frac{1}{Q}$$

3.8 Electrical equivalent circuits

For UHF antenna design, it is necessary to investigate the impedances of the system components, meaning, IC, assembly, and tag antenna. The following section describes the impedances from the IC to the tag antenna.

3.8.1 Equivalent circuit of the IC

Electrically, a passive UHF RFID IC can be depicted as a complex, capacitive impedance. Fig 10 shows the equivalent circuit of the input impedance. The impedance value of a UCODE IC is given in the datasheet and also corresponds to the shown serial equivalent circuit.

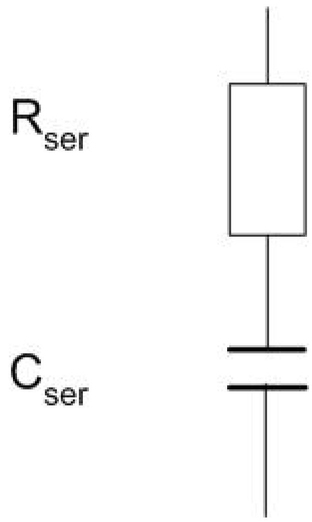

Fig 10. Equivalent circuit of IC input impedance

Note: For the parallel circuit, the values of the real- and imaginary part would change.

3.8.2 Equivalent circuit of the assembled IC

At each assembly process, parasitic capacitances and resistors are inherited, which result in an additional parallel impedance. The value depends on assembly process and parameters.

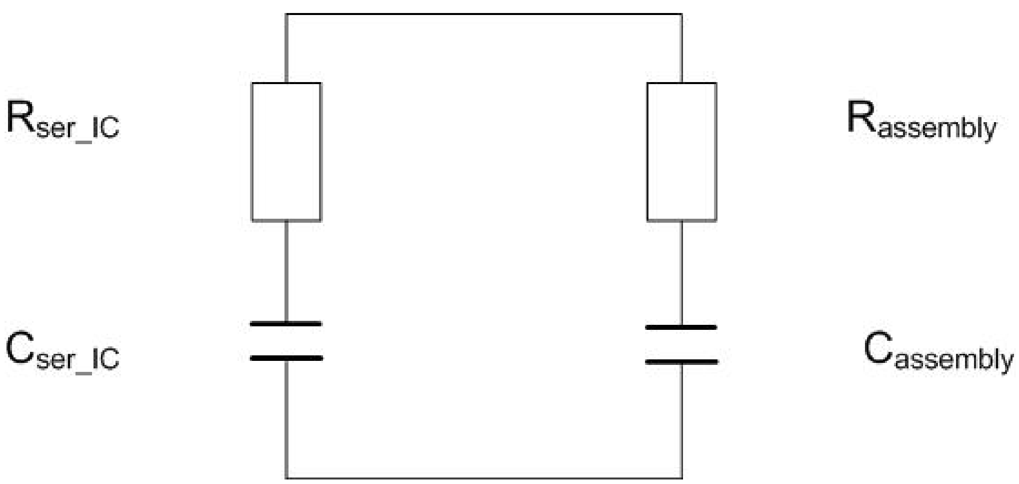

Fig 11. Serial equivalent circuit of assembled IC

3.8.3 Equivalent circuit of a loop antenna

Since the magnetic loop is quite small compared to the wavelength, its near-field components are mainly magnetically. Due to that fact, its reactance is quite insensitive to materials with different electrical permittivity in its close proximity.

The reactance of the magnetic loop can be simply modeled by an inductor. Together with the input capacitance of the IC, an LC-resonating circuit is formed as shown in Fig 12.

The real part of the impedance of the magnetic loop consists of its radiation resistance and an additional resistance representing the losses of the loop antenna. The radiation of a loop-antenna depends on its area relative to the wavelength at the operating frequency. Assuming an loop area of about 2cm² and the wavelength at 915MHz of about 33cm, the radiation resistance can be calculated by following formula (equation 14) to be:

$$R_{loop,radiation} = 31171 \cdot \frac{A^2}{\lambda^4} = 0.1 Ohm \quad (14)$$

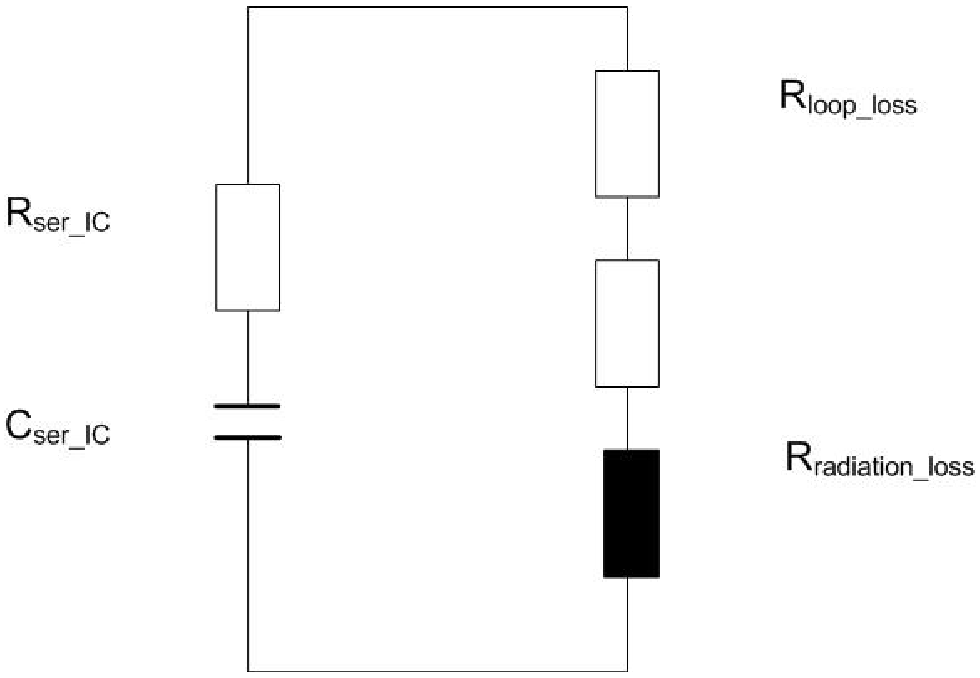

Fig 12. Equivalent circuit magnetic loop

3.8.4 Equivalent circuit of a broadband dipole

A conductor always has a capacitance and an inductance. Hence a dipole in a homogenous electromagnetic field can be represented by a series-L-C resonant circuit.

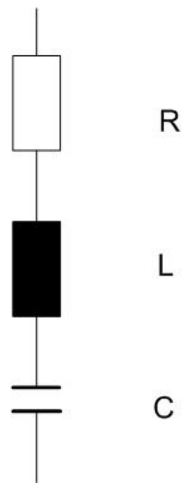

Fig 13. Equivalent circuit of a dipole

The resonance frequency of such a circuit is determined by:

$$\omega_{res} = \frac{1}{\sqrt{LC}} \quad (15)$$

At a given resonance frequency, the product of L and C will result in a certain fixed value. Although the product is fixed the relation of L/C can be chosen arbitrarily. As an example, doubling C and halving L will result in the same resonance frequency. This gives some freedom in the design of the dipole, which means that it is possible to design various kinds of dipoles with different shapes, which all have the same resonance frequency.

The difference between 2 dipoles with the same resonance frequency but a different L/C-ratio is the quality and hence the bandwidth, since

$$Q \sim \sqrt{\frac{L}{C}} \quad (16)$$

And

$$BW \sim \frac{1}{Q} \quad (17)$$

Therefore the L/C-ratio has to be minimized in order to obtain a high bandwidth of the dipole. As an example increasing the thickness of a dipole can achieve this. The increased thickness reduces the self-inductance and the increased surface of the dipole results in an increased capacitance of the dipole. Chapter 4.3 gives further insight on this matter.

3.9 Matching

The input impedance is the parameter, describing the antenna input behavior as a circuit element. As usual in electronic circuit design it is important to match this input impedance ($\underline{Z}_{antenna}$) to a given source impedance, which is in case of RFID applications the chip impedance of the transponder IC ($\underline{Z}_{chip}$). The maximum power delivered from the source to the antenna is given if the antenna input impedance is complex conjugate to the chip impedance:

$$\underline{Z}_{antenna} = \underline{Z}^{*}_{chip} \tag{18}$$

Separated into real- and imaginary parts we get the following conditions:

$$R_{antenna} = R_{chip} \tag{19}$$

$$X_{antenna} = -X_{chip} \tag{20}$$

For designing an efficient antenna it is necessary to match the real part and the conjugate imaginary part of the source impedance (practically this takes into account the IC impedance AND the parasitic assembly impedance). Assuming complex conjugate impedance matching between antenna and chip (assembled) and further assuming the case of a receiving antenna the maximum power delivered from the antenna ($P_{antenna,\, max}$) is:

$$P_{antenna,\, \max} = \frac{\left|\underline{V}_{antenna}\right|^{2}}{4R_{antenna}} \tag{21}$$

Where $V_{antenna}$ is the voltage generated by the label antenna due to receiving an electro-magnetic wave and $R_{antenna}$ is the resistance of the label antenna.

In order to estimate the variation of the read range of a given transponder if the loss impedance is being modified (caused by modifications of the flip chip mounting parameters) the following approach can be followed:

The received (tag) power at an arbitrary point is:

$$P_{Tag} = A_{Tag} \cdot S \tag{22}$$

with: A_{Tag} = Effective Area of the (tag) antenna (derived from radar technique) and S=power density.

Further it is:

$$A_{Tag} = \frac{\lambda_0^2}{4\pi} \cdot G_{Tag} \tag{23}$$

with: λ_0 = wave length ($\lambda_0 = c/f_0$) and

G_{Tag} = gain of tag antenna

And

$$S = \frac{P_{Reader} \cdot G_{Reader}}{4\pi \cdot D^2} \tag{24}$$

with: P_{reader} = reader/scanner power,

G_{reader} = gain of reader antenna and

D = maximum distance between tag and reader.

Using Equation 22 and Equation 23 in Equation 24 and solving for the distance D yields:

$$D = \frac{\lambda_0}{4\pi} \cdot \sqrt{\frac{P_{Reader}}{P_{Tag}} \cdot G_{Reader} G_{Tag}} \tag{25}$$

Assuming the following data applicable for UHF labels:

λ_0 = 32.8 cm (f_0= 915 MHz),
P_{Reader} = 30 dBm (= 1 W),
G_{Reader} = 2.15 dBi (= 1,642, 100% efficient dipole)
P_{Tag} = -15.23 dBm (= 30 µW, minimum operating power of tag),
G_{Tag} = 2.15 dBi (= 1,642, 100% efficient dipole)

The maximum theoretical distance between transponder and reader can be calculated to be:

D_0 = 7.811 m.

No we can calculate the new distance (which is reduced since energy is short circuited by the loss impedance parallel to the chip) in relation to the maximum theoretical distance. Defining the relative change of the distance yields:

$$\Delta D = \frac{D - D_0}{D_0} \tag{26}$$

Using Equation 25 with P_{Tag} = P0 and P_{Tag}= P respectively yields:

$$\Delta D = \sqrt{\frac{P_0}{P}} - 1 \tag{27}$$

Further the total power received by the tag consists of three terms:

$$P_{Tot} = P_{Chip} + P_{loss} + P_{antenna} \tag{28}$$

The related efficiency for the IC will result in:

$$\eta_{Chip} = \frac{P_{Chip}}{P_{tot}} = \frac{P_{Chip}}{P_{Chip} + P_{loss} + P_{antenna}} \tag{29}$$

Regarding the total received tag power (P_{tot}) we have to differ two cases:

1.) Ideal case

Theoretical maximum distance with no losses and optimum impedance matching between antenna and chip, which yields:

$$P_0 = P_{tot} = P_{Chip} + P_{loss} + P_{antenna} \tag{30}$$

With P_{loss} = 0 (no losses) and P_{chip} = $P_{antenna}$ (power matching) we get:

$$P_0 = 2 \cdot P_{Chip} \tag{31}$$

2.) Real case

Taking into consideration the parallel loss impedance, which yields:

$$P = P_{tot} = \frac{P_{Chip}}{\eta_{Chip}} \tag{32}$$

Using equation 29 and equation 30 in equation 26 yields the change in the distance between label and reader in relation to the theoretical maximum range:

$$\Delta D = \sqrt{2 \cdot \eta_{Chip}} - 1 \tag{33}$$

The matching system is shown by a simplified equivalent network consisting of the chip, the loss impedance, and the antenna (including an ideal voltage source). Fig 14 shows the related network. The elements of the transponder are substituted by series circuits of resistors and capacitors (chip and loss impedance) and a resistor and an inductance (antenna). All elements are parallel connected. The voltage source V_Antenna represents the voltage of the antenna generated by the received electromagnetic wave.

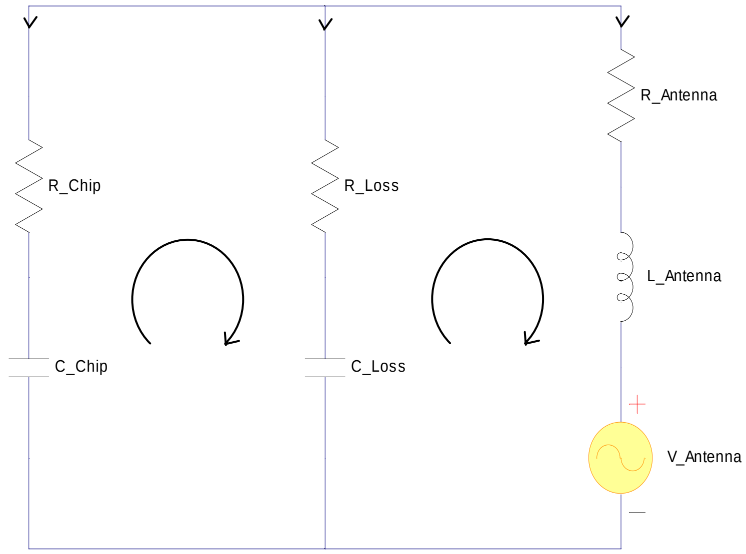

Fig 14. Equivalent electrical circuit of a transponder

3.10 Reflection coefficient

Especially at high frequencies such as e.g. UHF the mismatch between a source and a load (or in general between to elements) is represented by the reflection coefficient $\underline{\Gamma}$. Based on transmission line theory this reflection coefficient is defined by the ratio of the reflected wave to an incident wave. Hence the reflection coefficient is simply a measure of the quality of the match between the source- and the load impedance. Therefore the dimensionless complex reflection coefficient is defined by:

$$\underline{\Gamma} = \frac{\underline{Z} - \underline{Z}_0^*}{\underline{Z} + \underline{Z}_0} \qquad (34)$$

Where $\underline{Z}$ is the measured impedance and
$\underline{Z}_0$ is the normalizing impedance

Very often the imaginary part of $\underline{Z}_0$ is zero (or can be neglected) and the real part is fixed

$$\underline{\Gamma} = \frac{\underline{Z} - R_0}{\underline{Z} + R_0} \qquad (35)$$

to R_0 = 50 Ω, which results in a reflection coefficient of:

Further more the reflection coefficient is usually given as a logarithmic data (in dB):

$$\Gamma_{dB} = 20 \cdot \mathbf{lg}\left(\left|\underline{\Gamma}\right|\right) \qquad (36)$$

The above equation has to be used for amplitudes like e.g. voltages. Whereas the factor 20 has to be replaced by the factor 10 if powers i.e. square of amplitudes are considered.

Another derivation of the reflection at a port of a device is based on the network analysis theory of a two-port device (or network) and it's related scattering parameters. Two-port s-parameters are defined by considering a set of voltage waves. Those voltage waves are being divided by the square root of reference impedance. If further the square of the magnitude of these new variables is being considered one gets traveling power waves:

$|\underline{a}_1|^2$ = incident power wave at the network input

$|\underline{b}_1|^2$ = reflected power wave at the network input

$|\underline{a}_2|^2$ = incident power wave at the network output

$|\underline{b}_2|^2$ = reflected power wave at the network output

These variables are related in the following equation system:

$$\begin{aligned} \underline{\boldsymbol{b}}_1 &= \underline{\boldsymbol{a}}_1 \cdot \underline{\boldsymbol{s}}_{11} + \underline{\boldsymbol{a}}_2 \cdot \underline{\boldsymbol{s}}_{12} \\ \underline{\boldsymbol{b}}_2 &= \underline{\boldsymbol{a}}_1 \cdot \underline{\boldsymbol{s}}_{21} + \underline{\boldsymbol{a}}_2 \cdot \underline{\boldsymbol{s}}_{22} \end{aligned} \qquad (37)$$

A two-port network with its related power waves is presented in Fig 15.

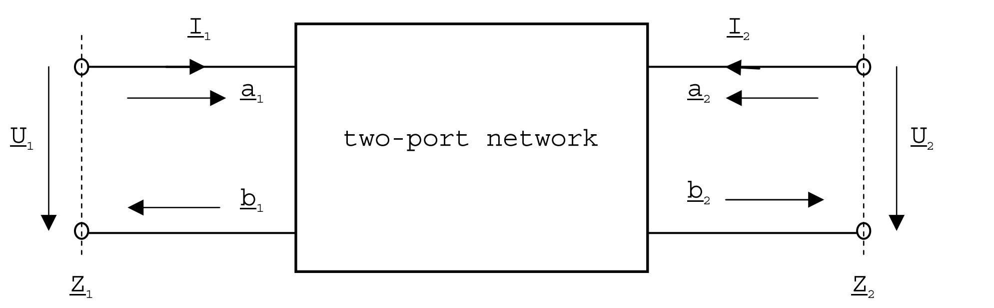

Fig 15. Two-port network with wave-parameter

Terminating the output of a two-port device with the reference impedance causes no reflection at the output and hence $\underline{a}_2$ is zero, then we get:

$$\underline{s}_{11} = \left.\frac{\underline{b}_1}{\underline{a}_1}\right|_{a_2=0} \equiv \text{input reflection coefficient}$$
$$\underline{s}_{21} = \left.\frac{\underline{b}_2}{\underline{a}_1}\right|_{a_2=0} \equiv \text{forward transmission coefficient} \quad (38)$$

Driving the two-port device (or network) from the output terminal and terminating the input port with the reference impedance causes no reflection at the input and hence $\underline{a}_1$ is zero, then we get:

$$\underline{s}_{22} = \left.\frac{\underline{b}_2}{\underline{a}_2}\right|_{a_1=0} \equiv \text{output reflection coefficient}$$
$$\underline{s}_{12} = \left.\frac{\underline{b}_1}{\underline{a}_2}\right|_{a_1=0} \equiv \text{reverse transmission coefficienta} \quad (39)$$

Per definition the scattering parameter $\underline{s}_{11}$ is equal to the reflection coefficient $\underline{\Gamma}$ as described above. Usually the scattering parameters are also given in decibel (dB).

The traditional way to determine the reflection coefficient ρ is to measure the standing wave caused by the superposition of the incident wave and the reflected wave. The ratio of the maximum divided by the minimum is the Voltage Standing Wave Ratio (VSWR). The VSWR is infinite for total reflections because the minimum voltage is zero. If no reflection occurs the VSWR is 1.0. VSWR and reflection coefficient are related as follows:

$$VSWR = \frac{(1+|\rho|)}{(1-|\rho|)} \quad (40)$$

Most present instrumentation measures the reflection coefficient and calculates the VSWR.

4. Label Antenna Design

The antenna design is one of the most critical point in passive UHF RFID systems. As described in the previous sections the antenna transfers the radiated power from the reader antenna to the IC. In order to achieve the maximum power transfer, the antenna impedance (Z_{Ant}) must be the complex conjugate of the IC impedance (Z_{IC}):

$$Z_{Ant} = Z^{*}_{IC} \tag{41}$$

The antenna is directly connected to the chip, which typically presents a high-capacitive input impedance. The mechanical process of direct attaching the IC to the antenna introduces further parasitic capacitance between the IC and the antenna. The parasitic capacitance due to the assembling process are ideally in parallel to the IC, therefore they are added to the IC capacitance (in case of equivalent parallel model of the IC).

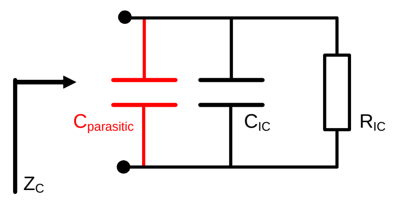

Fig 16. Equivalent circuit of the IC including parasitic.

By taking into account the parasitic effects of the mechanical assembly process, the new impedance of the assembled IC (calculated and or measured) is denoted as Z_C .

To achieve the best matching between the antenna and the IC the following equation must be satisfied.

$$Z_{Ant} = Z^{*}_{C} \tag{42}$$

Of course due to physical restriction the exact matching most of time in real life application can not be achieved.

A way to obtain the best matching between the antenna and the IC is to minimize the power reflection coefficient $|s|^2$ in the following equation:

$$|s|^2 = \left| \frac{Z_{Ant} - Z_C^*}{Z_{Ant} + Z_C} \right|^2 \tag{43}$$

By knowing the tag antenna parameter, it is possible to calculate the RFID tag performance by following:

$$Range = \frac{\lambda}{4\pi} \sqrt{\frac{P_{eirp} G_r}{P_{th}} p\left(1 - |s|^2\right)} \tag{44}$$

Where λ is the wavelength, P_{eirp} is the equivalent isotropic radiated power transmitted by the reader, G^r is the tag antenna gain, P_{th} is the minimum power required to activate the chip (as provided in the datasheet), and p is the polarization loss factor.

4.1 Fixed parameters

For every label antenna we have to consider fixed parameters that can’t be changed and variable parameters which can be influenced and tuned to meet the requirements of the design.

The design starts by specifying the fixed parameters such as Q value, Impedance, and power consumption which are considered as given and constant.

In order to outline the variable parameters the NXP Antenna Questionnaire described in the next section can be used.

4.2 Variable parameters (Antenna questionnaire)

Before starting to design an antenna, certain boundary conditions need to be fixed.

Details like assembly process, antenna material, but also required performance of the antenna design are vital inputs for designing an antenna.

Following questionnaire shows an example of which parameters should be defined upfront.

Size		
Label Size:	4*0,5	
Antenna Size:	95*10	
Additional Information:		
Antenna Substrate		
Material:	PET	
Thickness:	50 µm	
Additional Information:		
Lamination glue (if applicable)		
Material:	ACP	
Thickness:	3,72*E07 S/m	
Additional Information:	20 µm	
Antenna Conductor		
Material:		
Estimated conductivity (special materials):		
Thickness:		
Additional Information:		
Minimum feature size:		
Line width:		
Gap width:		
Nonstandard fiducial marks required	☐	
Filename for fiducial marks:		
Performance		
Required Frequency Bands		
US:	☑	
EU:	☑	
Japan:	☑	
Others:	☐	
Specification:		
Application and required read range for all bands		Min. Read Range
Free Air	☐	5 m
Cardboard/Corrugated	☐	5 m
Plastic	☐	5 m
High Loss material	☐	3 m
Others	☐	
Specification		
Material or		
Epsilon r	8	
Loss factor		
Stacked tags	☐	
Max. no. of tags	5	
Min. distance of tags	2 cm	
Required Read Range for stacked tags		> 3m
Additional Information:		
Chip Attachment		
Direct Attach:	☑	
Assembly Machine	Mühlbauer	
Expected Assembly Capacitance		
Low Force Thermodes	☐	
FCP 2	☐	
Attachment process for FCP2:		
Crimping	☐	
Others	☐	
Specification:		

Fig 17. Antenna questionnaire

4.3 Design Flow examples

The design described in the following was approached according the steps discussed earlier in section 4.1 and 4.2.

According section 4.1 the following design constants were fixed:

Quality factor Q = 9,

Impedance Z = 22 – j195 Ω

Minimum operating power supply Pmin = -15 dBm

According the antenna questionnaire in section 4.2 the following requirements were defined: With respect to performance, the design is broadband and works in all RFID frequency regulations. The special design target was a solid performance on different carrier materials. It is defined for applications on carrier materials with a dielectric constant of 8. The dielectric constant is a scalar value As it is a far field design, an optimized performance in stacked mode is not required. Direct chip attach was used as assembly process.

4.3.1 Far Field Antenna

This section will describe how to design a Far Field antenna and the basic rules to tune the antenna to the correct frequency band for optimization.

Due to the fact that the antenna shall perform optimal in far field applications (distance between reader and tag antenna > 1m), the antenna design shall be able to receive energy from the radiated electric field. The electric component is the dominant part of the radiated electromagnetic wave in the far field.

In the far field condition as described in the previous section the ratio between the E and H is equal to wave impedance Z_0 .

$$Z_0 = \frac{E}{H} \tag{45}$$

$$Z_0 = \sqrt{\frac{\mu_0}{\varepsilon_0}} = 377\Omega \tag{46}$$

Derived from this equation it is obvious that the E field is the dominant part of the radiated wave in far field condition.

The propagation path loss and antenna gain play a crucial role regarding the system performance. Taking into account these environmental influences, he antenna design shall be able to power up the IC at distances higher than 1 meter from the reader antenna.

4.3.2 How to proceed

To fulfill the requirements of a far field design (strong E field behavior) we will focus on a dipole type based antenna. A dipole is able to get maximum coupling with the radiated E-Field and hence harvest most energy from the field. Using a loop as a matching network the dipole is then matched to the IC

The use of a matching network is necessary due to the fact that a dipole based antenna is mainly a capacitive resonant circuit and the IC with is a capacitive load need to be compensated to match the dipole antenna and satisfy the equation:

$$Z_{Ant} = Z_{c}^{*} \tag{47}$$

In our case a capacitive IC impedance (with a negative imaginary part) should be compensated. This refers to the model depicted in Fig 16.

The two parallel capacitances and the parallel resistance that an equivalent impedance in the following form:

$$Z_{c} = X - jY \tag{48}$$

With both X and Y > 0

It is obvious, that a positive imaginary part will be required, in order to get a matched system. The conjugate complex of a capacitive impedance is an inductive impedance (positive imaginary part)

Thus, our antenna will be based on a loop matching network coupled with a dipole.

The loop dimension is influencing the absolute value of the imaginary part of the impedance. The impedance value of the loop shall match as exact as possible the impedance of the IC, regarding the imaginary part as well as the real part

The positive imaginary part of the loop impedance increases the inductance of the loop (size). The resistance is increased by reducing the width of the trace but this will also influence the bandwidth of the antenna.

The resonance frequency of the dipole is defined by setting the length of the dipole

Following section shows an equivalent model of the dipole

4.3.2.1 Loop

The loop based part of the antenna can be schematized as described in Fig 18. The real part of the impedance of the loop consist of a part given by the radiation resistance (Radiation) and a part given by the losses given by the loop conductor (R loss). In general for a small circular loop the radiation resistance is given by the following equation (49)

$$R_{loop,radiation} = 20\pi^{2}\frac{C^{4}}{\lambda^{4}} \tag{49}$$

The equation shows clearly that the impedance (real part) is strictly related to the area of the loop.

The first limit to the design (based on a matching loop) is given by the chip impedance. In our case it will be matched to the impedance of the UCODE G2X. It requires a loop having a big area to compensate the real part and to generate the necessary inductance to compensate the imaginary part of the chip impedance.

The inductance of a small loop is given (approximately) by the equation (50). Where a denotes the loop radius and b the wire radius (with the assumption that the loop is a wire)

Main purpose of equation 50 is to show the dependence of the inductance from the loop dimension.

$$L_A = \mu_0 a \left[\ln\left(\frac{8a}{b}\right) - 2 \right]$$
$$L_i = \frac{l}{\omega P}\sqrt{\frac{\omega\mu_0}{2\sigma}} \qquad (50)$$

The second limit is the bandwidth. In this case the chip is an electrical system with a low Q (Q<7). This will lead a broad band loop.

Note: Narrowband loops can reduce the performance of the system by limiting the whole bandwidth.

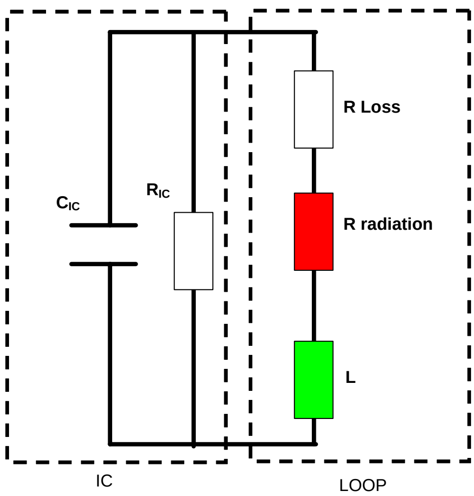

Fig 18. Electrical scheme of a loop antenna.

4.3.2.2 Dipole

The typical representation of a dipole is given by a series-L-C resonant circuit showed in Fig 19.

The resonance frequency of the dipole is function of the inductance and capacitance. For a series L-C resonant structure the resonance frequency is given by the equation (51).

$$\omega_{res} = \frac{1}{\sqrt{LC}} \tag{51}$$

The quality factor and the bandwidth of the dipole as described in Fig 19 are functions of $\sqrt{L/C}$ and of $1/Q$ respectively.

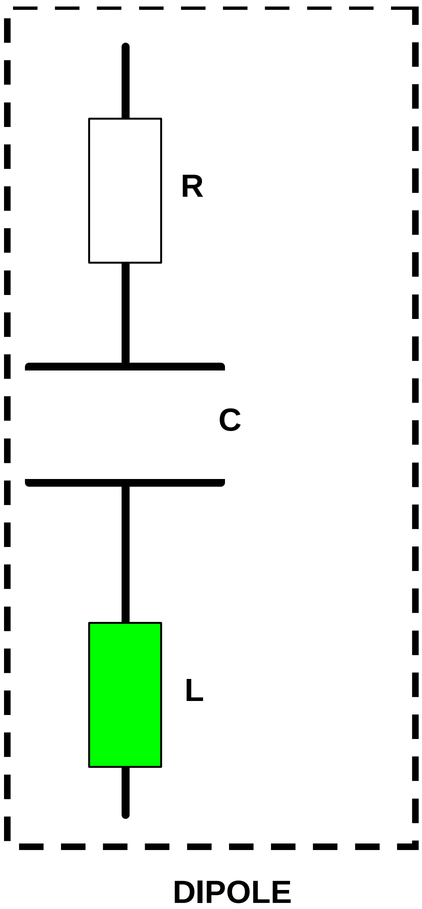

Fig 19. Electrical scheme of a dipole antenna.

4.3.2.3 Dipole – Loop Coupling

The main aspect of the system based on a loop and a dipole antenna is how to couple two radiating structures. To this purposes there are two main approaches:

- Inductive and/or Capacitive coupling
- Conducted coupling

The two methods can be described schematically in Fig 20 (magnetic coupling with ratio 1: N function of the distance between loop and antenna) and Fig 16 (conducted coupling with the ration 1:1).

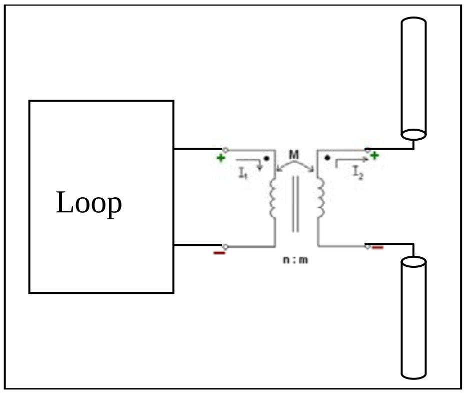

Fig 20. Magnetic coupling.

In the investigated design the loop and the dipole is connected using conducted coupling.

4.4 Example Far Field Reference Antenna Design

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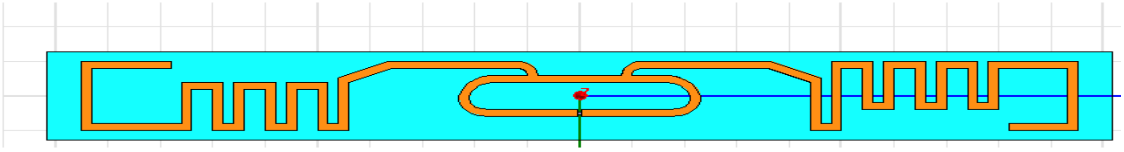


Fig 21. FF95-8 G2X Aluminum Reference Design

4.4.1 FF95-8 CAD Drawing

In the following section shows the 2D CAD design of the label with quotes. The dipole part of the antenna is depicted in Fig 22, the loop area of the antenna in Fig 23.

Main purposes of the CAD drawing is to verify the integrity of the eventual dxf files containing the antenna drawing that may be send together with the this application notes.

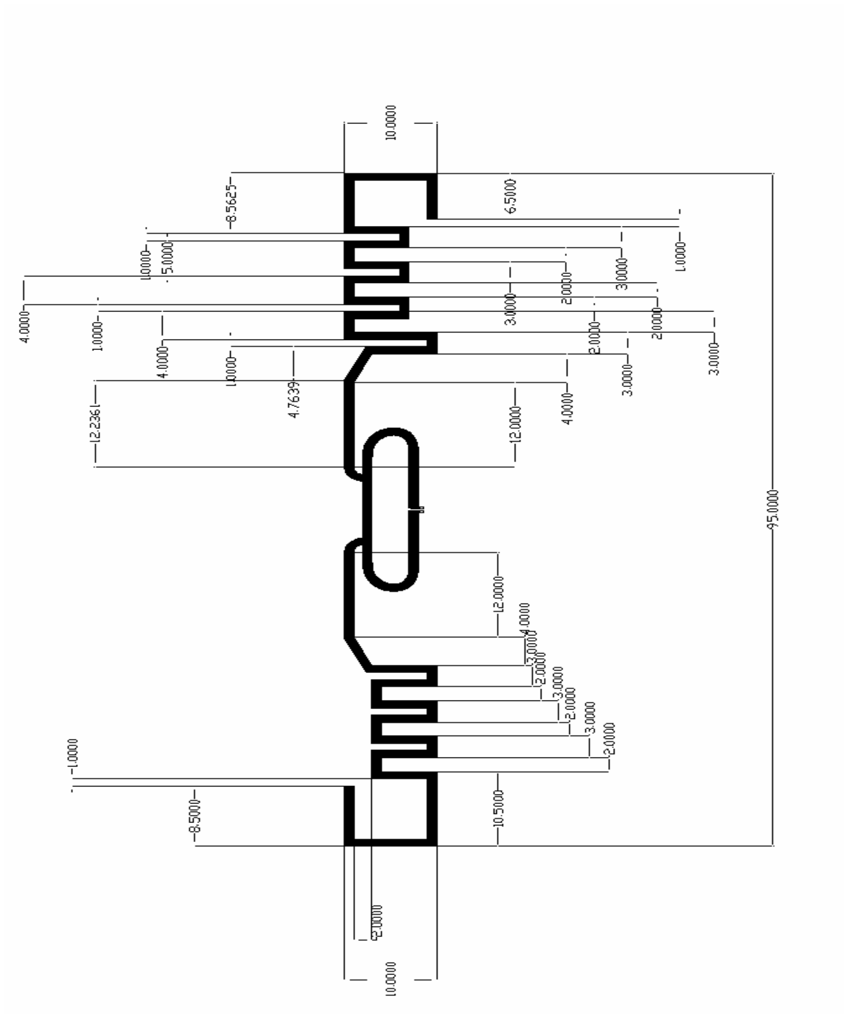


Fig 22. FF95-8 Dipole CAD DIPOLE Area [mm]

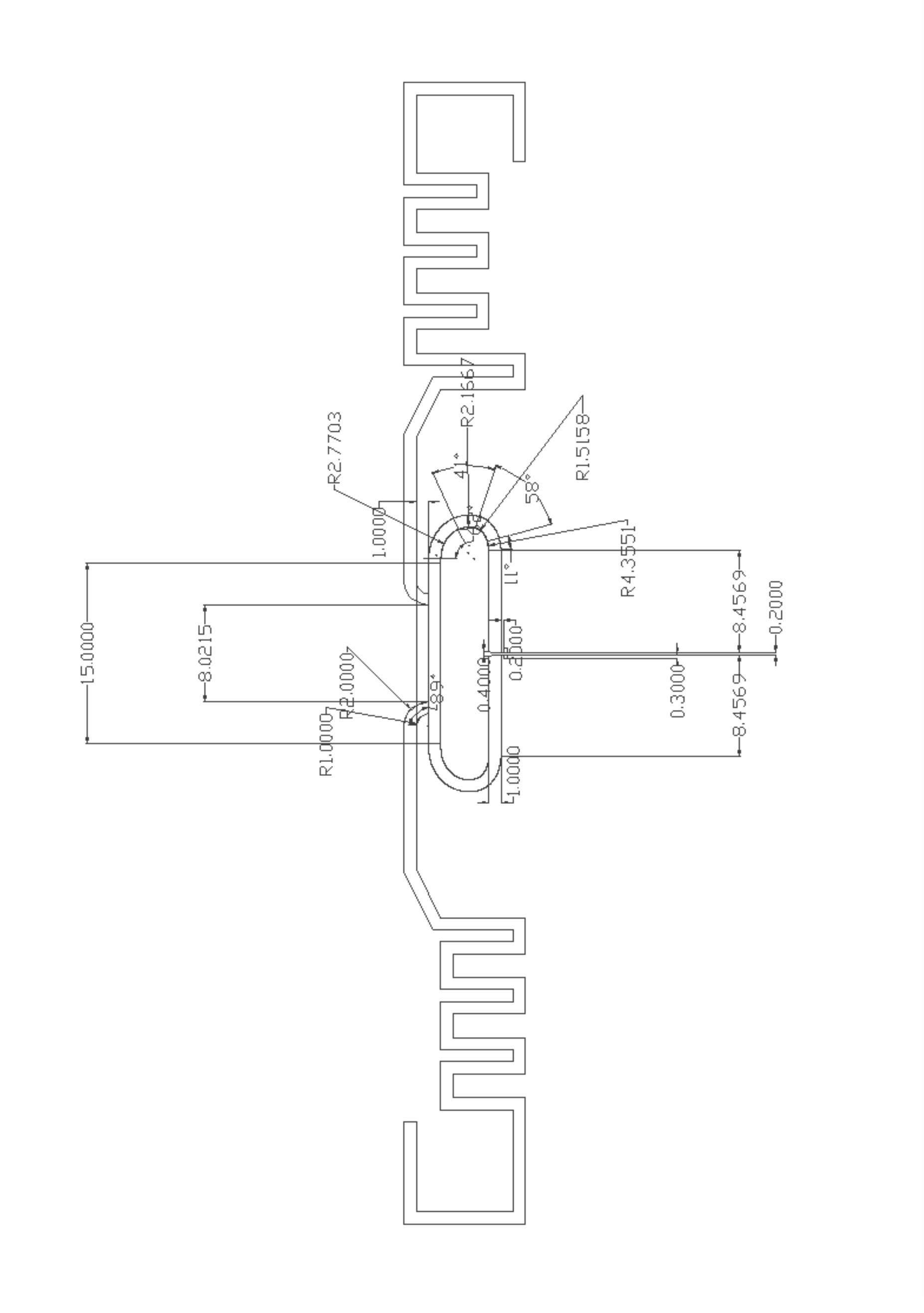
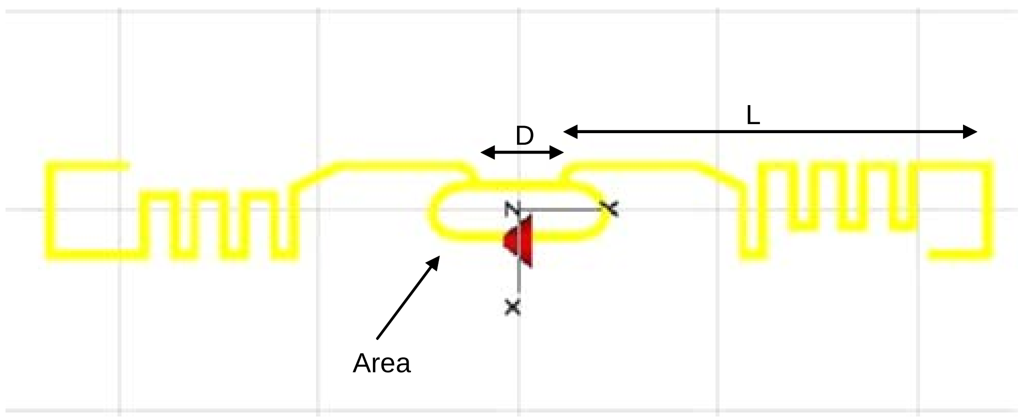


Fig 23. FF95-8 Dipole CAD Loop Area [mm]

4.4.2 Tuning of an antenna

Taking into account the above described matching technique and the impact of each part on the matching between antenna and IC, following parameters can be used to tune the antenna:

- Loop size
- Dipole length
- Bandwidth



1 Loop size: Area

2 Dipole length: L (menader line structure)

3 Dipole Distance: D

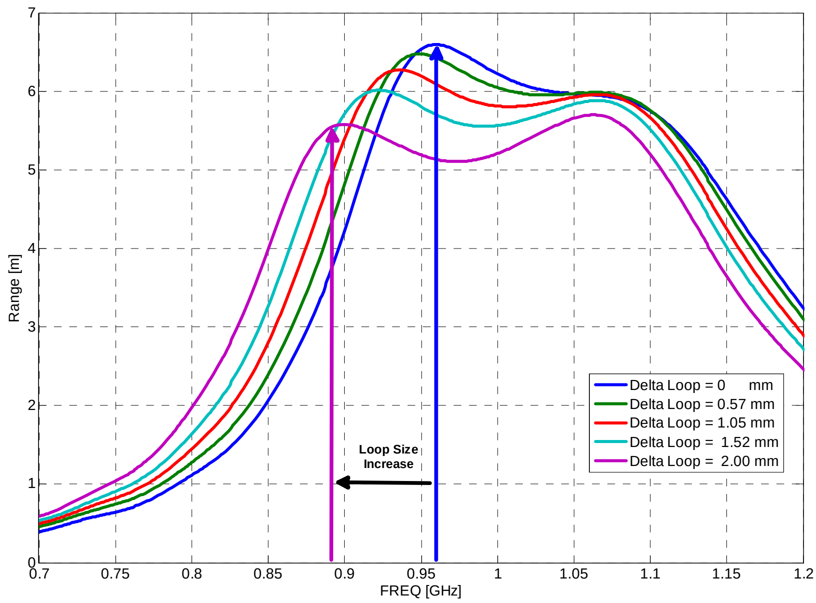
Fig 24. FF95-8 Dipole CAD [mm]

Impact of Loop Size on the matching and performances

The following picture shows the parametric read range of the tag in function of the delta dipole dimension.

It can be observed how the tuning of the antenna will change by increasing the loop dimension.

Obviously an increasing of the loop dimension will decrease the lower resonance frequency (loop “resonance”) of the tag antenna.



1 Loop size: Area

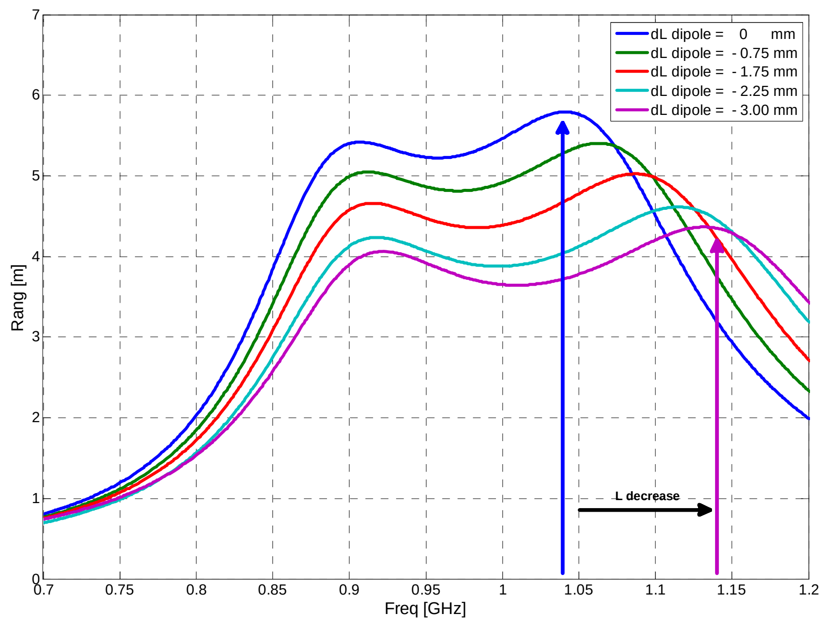
Fig 25. Read range vs. change of the loop area

Impact of dipole length L on the matching and performances

Following picture shows a parametric read range of the tag as a function of the delta length of the dipole.

It can be observed how the tuning of the antenna will change by reducing the total length of the dipole (obtained by modifying the meander line).

From the picture can be easily understood that a reduction of the total length L of the dipole will increase the upper resonance frequency (dipole resonance) of the tag antenna.



2 Dipole length: L (meander line structure)

Fig 26. Read range in meters vs. dipole length variation

Impact of dipole distance D on the antenna matching and performance

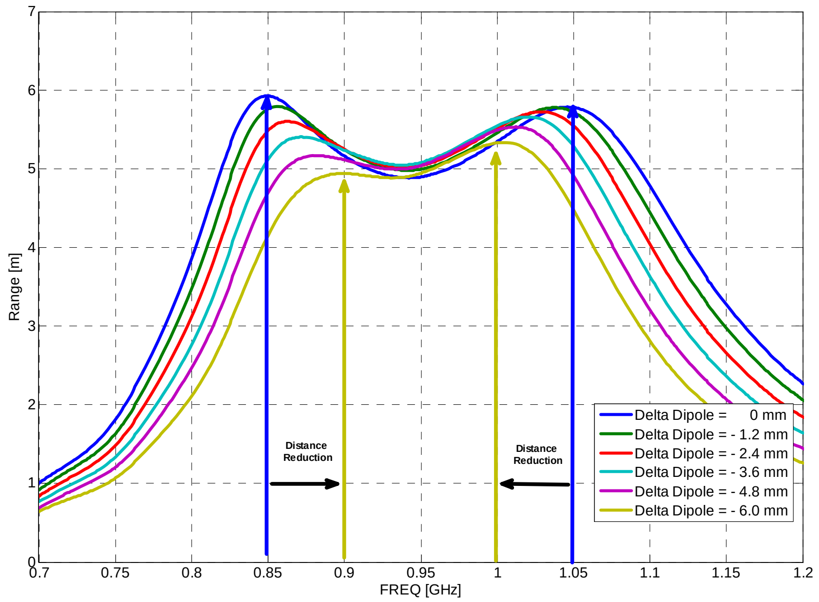
The picture in the following shows a parametric read range of the tag in function of the distance D (see Fig 27).

Variation on the distance D between the connection points of the two half dipoles to the loop will allow to vary the in frequency the two main resonances of the tag antenna (loop and dipole, by increasing or reducing the antenna bandwidth).

A small distance between the dipole arms implies close resonance frequencies and small bandwidth.

High separation distances imply higher separation in frequency between the two resonances and higher bandwidths.

Note that a high distance can create a low performing frequency region between the two resonance frequencies.

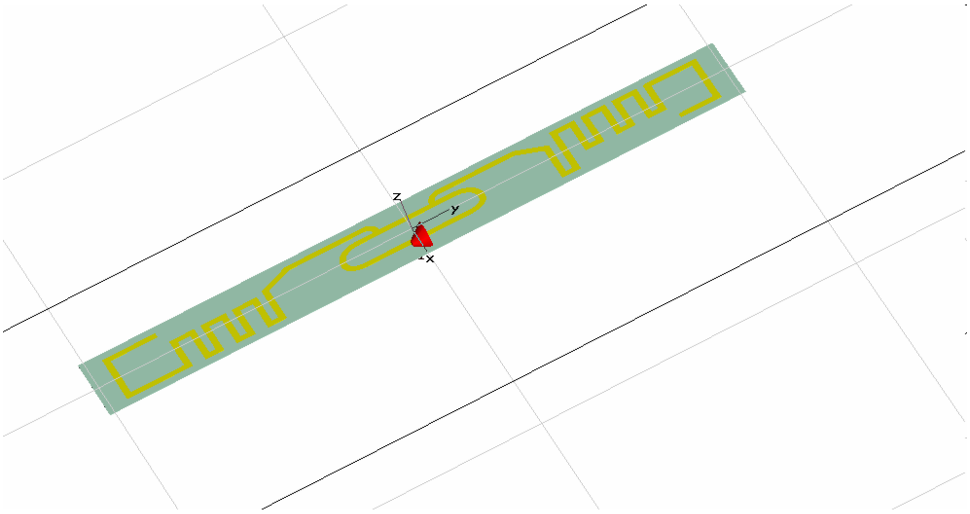


3 Bandwidth: D

Fig 27. Read range in meters vs. dipole distance (D) variation

4.4.3 Far Field Simulation Results at Fixed Geometrical Configuration

Electromagnetic Simulation



ANTENNA LAYOUT MODEL

Fig 28. Antenna layout model – EM simulation

The following simulations are solved using CST MWS 2008, a commercial Full 3-D FDTD solver for electromagnetic field.

Simulated Reflection Coefficient

The reflection coefficient describes either the amplitude or the intensity of a reflected wave relative to an incident wave and hence describes the ratio of the reflected wave to the amplitude of the incident wave. A low reflection coefficient is an indication of a good matching between chip impedance and antenna impedance.

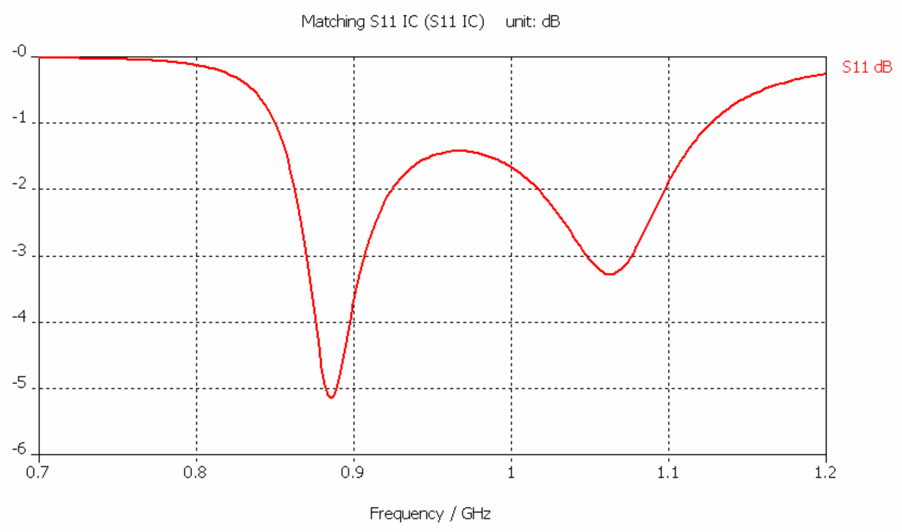


Fig 29. Reflection Coefficient S11

Simulated Antenna Gain Pattern

In the following sections the simulated 3-D antenna radiation pattern at 868 MHz and 915 MHz are shown. The dipole structure of the label antenna is positioned in the x-y-plane with arms pointing in y-direction.

Total Gain @ 868 MHz

The simulated 3-D characteristics of the label antenna show a typical dipole behavior and a gain of maximal ~3 dB in x and in z direction.

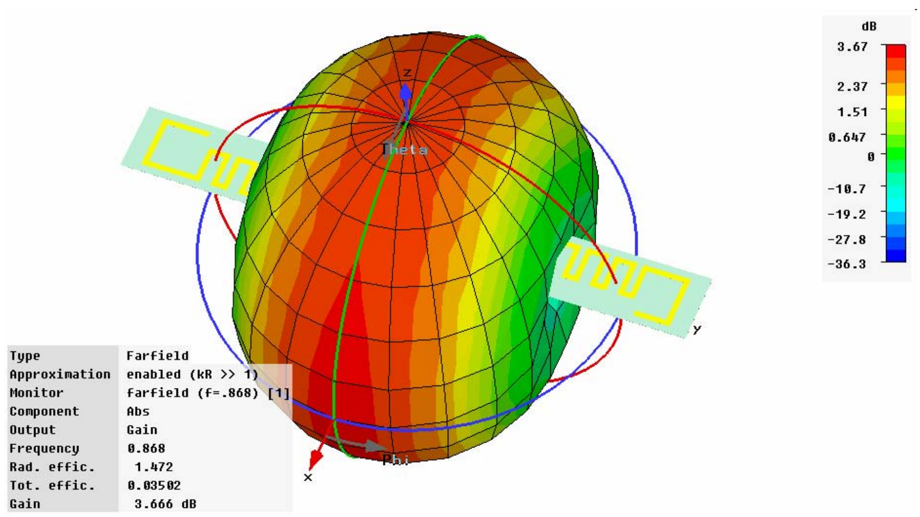


Fig 30. Total Gain @ 868 MHz

Φ Gain @ 868 MHz

3-D radiation pattern of the antenna, which would be seen by an ideal linear polarized reader antenna, moving around the y-axis rotated with the angel Φ. Φ denotes the rotation angle around the z-axis.

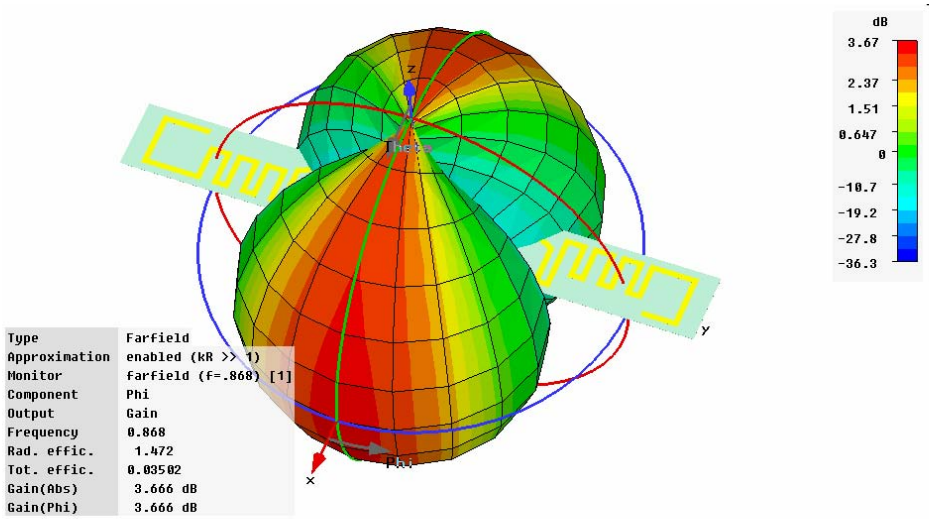


Fig 31. Φ Gain @ 870 MHz

θ Gain @ 868 MHz

3-D radiation pattern of the antenna, which would be seen by an ideal linear polarized reader antenna, moving around the x-axis rotated with the angel θ. θ denotes the angle in the x-z-plane.

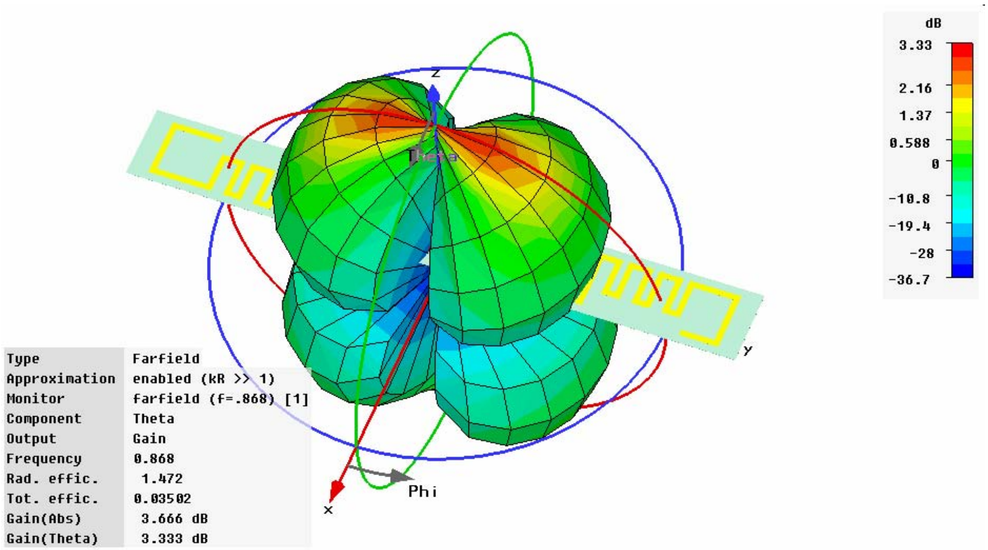


Fig 32. Θ Gain @ 868 MHz

Total Gain @ 915 MHz

The following figures show the radiation patterns of the antenna under US frequency regulations at 915MHz.

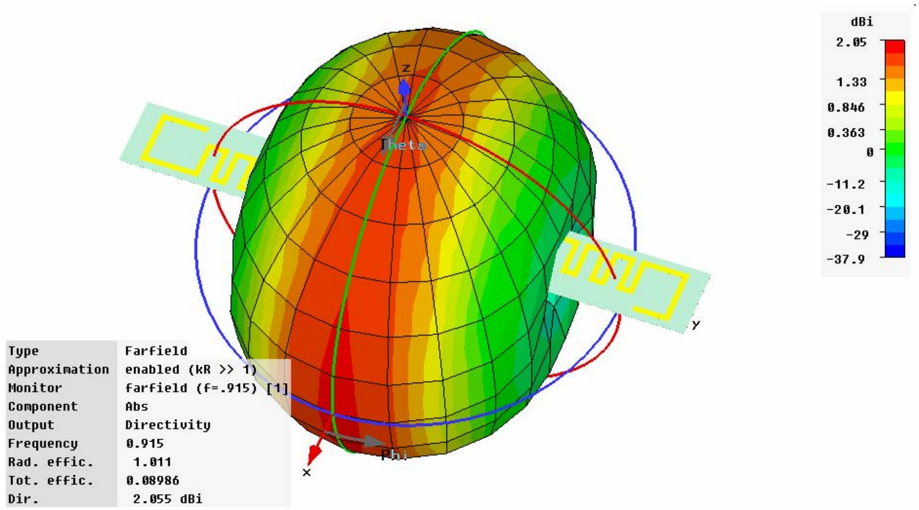


Fig 33. Total Gain @ 915 MHz

Φ Gain @ 915 MHz

3-D radiation pattern of the reference antenna, which would be seen by an ideal linear polarized reader antenna, moving around the y-axis rotated with the angel Φ. Φ denotes the rotation angle around the z-axis.

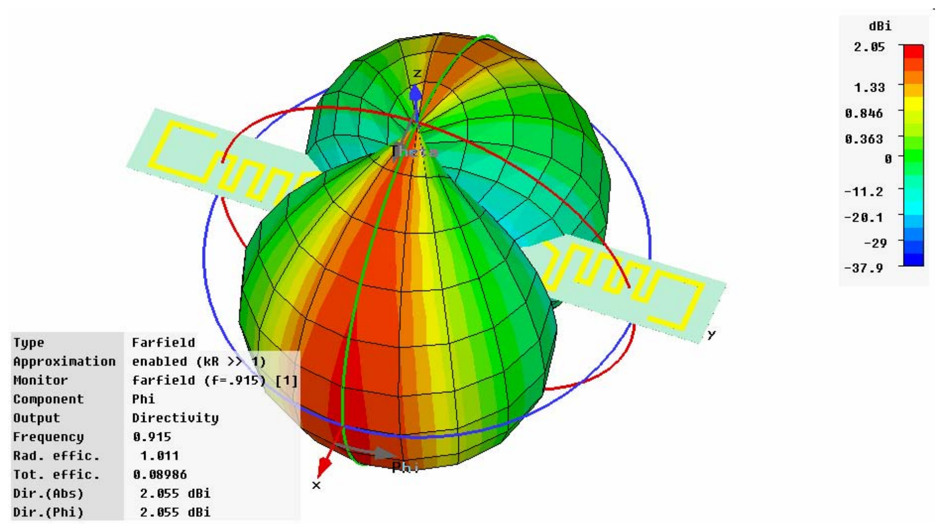


Fig 34. Φ Gain @ 915 MHz

Θ Gain @ 915 MHz

3-D radiation pattern of the reference antenna, which would be seen by an ideal linear polarized reader antenna, moving around the x-axis rotated with the angel θ. θ denotes the angle in the x-z-plane.

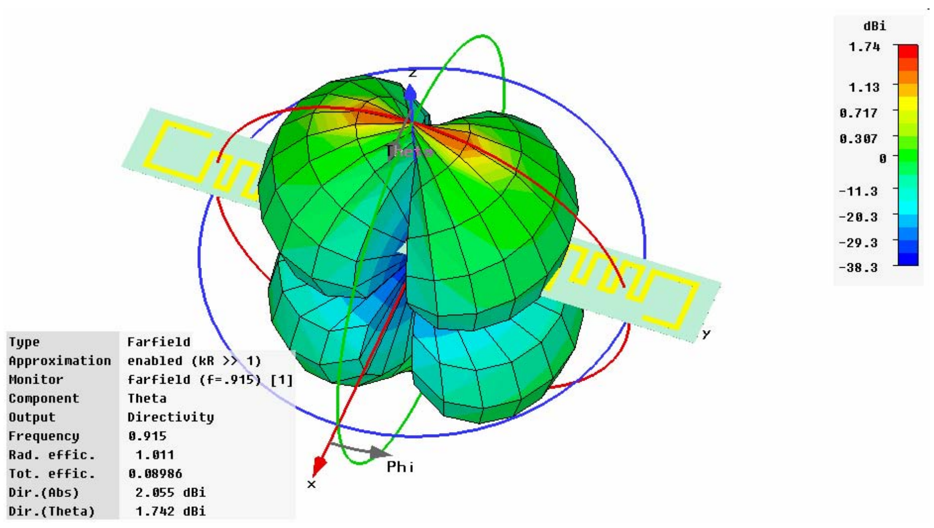


Fig 35. Θ Gain @ 915 MHz

Simulated antenna impedance

Fig 36 shows the simulation of the real part of the antenna impedance. The peak of the curve corresponds to the resonance frequency of the pure antenna. When the antenna is attached to the Chip (Assembly), the resonance peak will shift down due to the impedance chances (IC and parasitics). The overall resonance frequency will be then located in the RFID frequency band around 900 MHz.

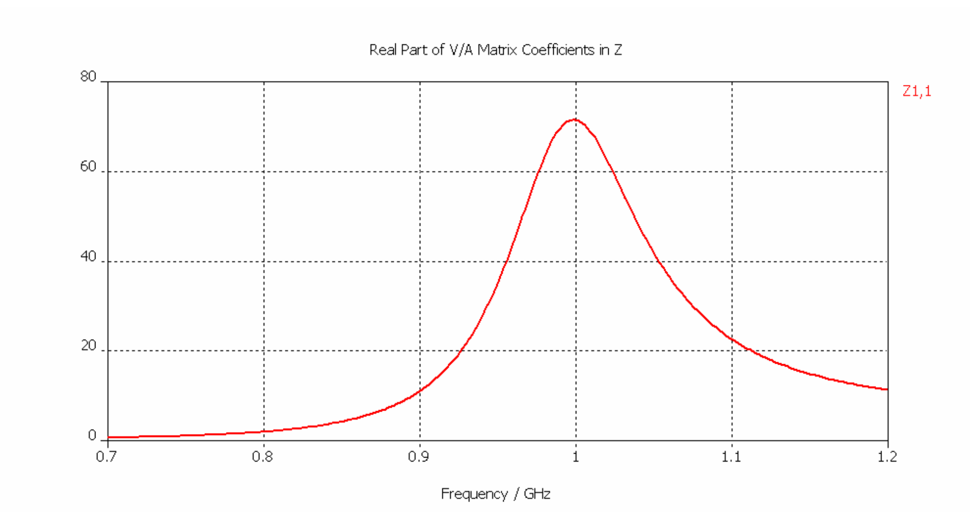


Fig 36. Real part of the Input impedance

Fig 37 shows the simulation of the imaginary part of the antenna. The value matches the IC impedance (conjugate complex).

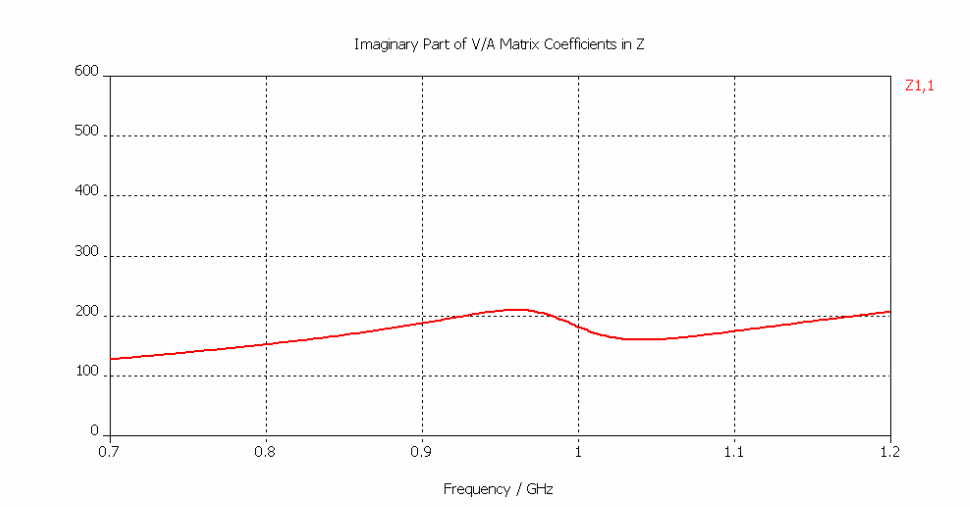


Fig 37. Imaginary part of the Input impedance

Fig 38 shows the Smith chart of the IC impedance. In order to match the impedance well, the antenna should have a complex conjugate impedance value which is located within the red circle. The circle is created around the chip impedance value normalized to 50 Ohms.

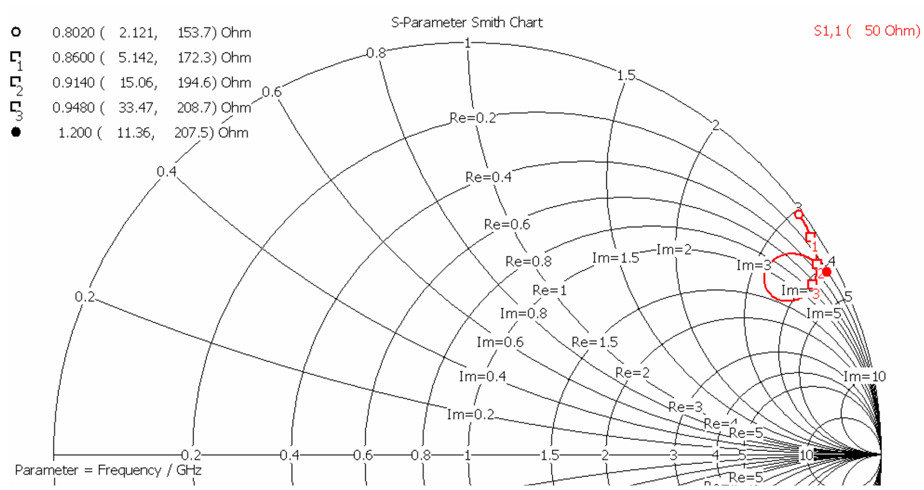


Fig 38. Smith Chart

Estimated Matching and Read Range.

Fig 39 shows the simulation of the reflection coefficient S11. The lower the value, the less power is reflected and the better is the matching. The best matching point is located around 890 MHz

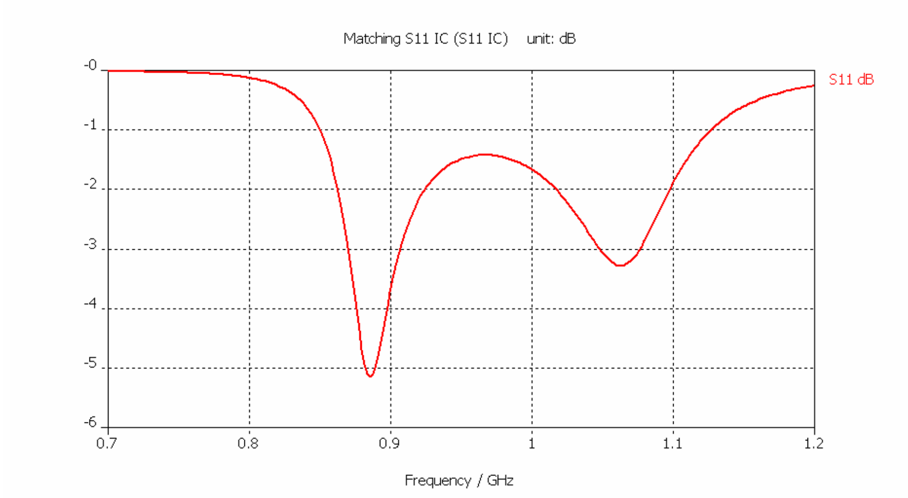


Fig 39. Reflection coefficient - Matching S11

Fig 40 directly results from Fig 39, it is the estimated read range. The best read range performance can be obtained at 890 MHz

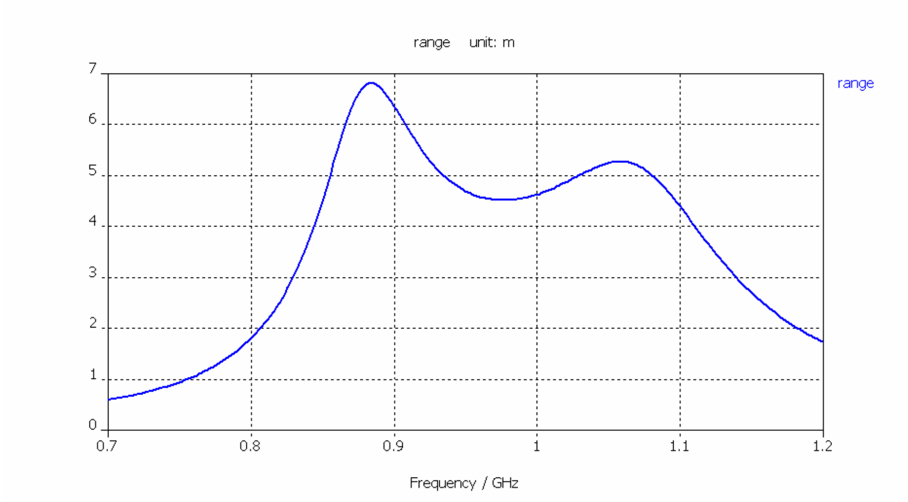


Fig 40. Estimated Read Range

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Notice: All referenced brands, product names, service names and trademarks are property of their respective owners.

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6. References

[1] Antenna Theory, Constantine A Balanis

[2] The RF in RFID, Passive UHF RFID in Practice, Daniel M. Dobkin

7. Contents

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