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Coupling Relations in RFID Systems

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ABSTRACT

A proper understanding of electromagnetic coupling between interrogators and labels is important for successful design and optimisation of antennas in RFID systems.

As a foundation for this understanding we provide an overview of the fundamental laws of electrodynamics, and the principal theorems deriving therefrom upon which antenna theory is based. The retarded potential solutions of Maxwells equations are recited and the resulting expressions for near and far fields for infinitesimal electric and magnetic dipoles are stated. The influence of electromagnetic compatibility constraints on the choice of operating frequency and antenna style is explored.

A number of field creation structures for near and far fields are illustrated, and label antenna forms and parameters for use in various object geometries are derived. Circuit models for antennas are provided, along with a set of experimental results and empirical equations for performance parameters of useful antennas. Future work will be concerned with experimental evaluation of further model parameters for proposed field creation structures and labels.

Coupling theory for electric and magnetic field sensitive antennas is developed at length and the relation to far field antenna theory is established. Some conclusions about antenna optimisation are drawn.

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Coupling Relations in RFID Systems

Biographies

Peter Cole
Research Director & Professor

Dr. Cole, Professor of Radio Frequency Identification (RFID) systems in the Department of Electrical and Electronic Engineering at the University of Adelaide, has been selected to head a new RFID study in Australia. Dr. Cole's current research covers the industrial applications of electromagnetic identification and tracking systems, the design of multi-function microcircuits, the design of signaling methodologies for simultaneous high-speed reading of multiple electronic labels, and the development of international standards for RFID systems. Dr. Cole will be working closely with both the MIT and Cambridge Labs. He will be focusing his research and expertise on the EPC™ concept within the silicon chips currently being prototyped by Center sponsors.

Behnam Jamali
Doctoral Candidate & Research Assistant

Graduated with a degree in Electronics Engineering at The University of Adelaide, Australia. Currently, Behnam is a PhD candidate carrying out research at The University of Adelaide's Auto-ID Centre. Research interests include low power analogue CMOS circuits, as well as design of efficient rectifiers.

Damith Ranasinghe
Doctoral Candidate & Research Assistant

Graduated with a degree in Information Technology and Telecommunications Engineering at The University of Adelaide. Currently, Damith is a PhD candidate working at The Auto-ID Centre research group at The University of Adelaide. Current research interest include electromagnetic coupling, encryption, and authentication in RFID.

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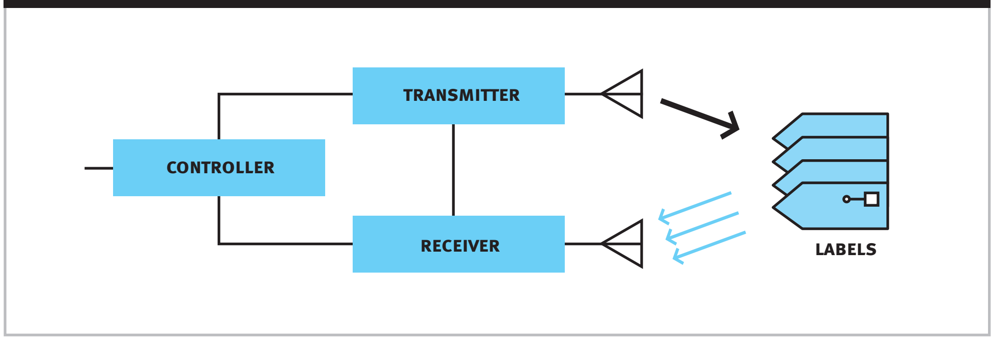
1. INTRODUCTION

Radiofrequency identification systems, illustrated briefly in Figure 1, are evolving rapidly as a result of:

a) Increased awareness of the technology;
b) development of improved techniques for multiple tag reading;
c) realisation in the business community of the benefits of widespread adoption in the supply chain;
d) adoption by designers of sensible concepts in the arrangement of data between labels and databases;
e) development of efficient data-handling methodologies in the relevant supporting communication networks;
f) appreciation of the need for cost reduction, and
g) development of new manufacturing techniques that will achieve manufacture of billions of labels at acceptable costs.

Figure 1

AN RFID SYSTEM



These developments should be seen as integrated and mutually supporting. However, most RFID systems that exploit these developments are desired by the users to be **passive**. i.e. to exploit microcircuits which can both power themselves and provide power for a reply signal from the small amount of energy that they can extract from an interrogation signal.

In this effort they are subject to what must be seen as laws of physics that are not amenable to manipulation, but are amenable to exploitation when correctly understood. The importance of correct understanding is underscored by the fact that passive RFID labelling systems, in contrast to electromagnetic broadcasting or communication systems that are powered at both ends of the communication link, have no margin for inefficiency: if the electromagnetic propagation aspects are not understood or not optimised, the systems will just not work to acceptable performance levels.

This paper is designed to provide support to RFID system designers in what is probably the least understood aspect of system design, that of the electromagnetic propagation to and from the label.

2. OBJECTIVES

We pursue in this paper the following objectives.

1. To provide an analysis of the electromagnetic coupling between interrogators and labels in radio frequency identification systems that will be useful across frequency ranges from the LF to UHF, and cover both large and small antennas operating in the near field and the far field, and exploiting coupling to electric fields, magnetic fields, or electromagnetic fields.

2. To encourage ways of thinking that the authors believe are fruitful in understanding and designing electromagnetic coupling systems for RFID labels.

3. To assemble in one place the principal equations on which sound design may be based.

4. To report briefly on experiments being conducted on the practical realisation of the ideas in this paper.

3. OUTLINE

We will begin with a formal statement of the laws of electrodynamics, and use them to derive boundary conditions to be noted where fields interact with materials. Such relations are amplified by introduction of useful concepts of demagnetising and depolarising factors. The retarded potential solutions of Maxwell's equations and their utility in developing near and far radiated fields are developed, especially for infinitesimal electric and magnetic dipoles. The influence of electromagnetic compatibility constraints on the choice of operating frequency and antenna style is explored. A number of field creation structures for near and far fields are described, and label antenna forms and parameters for use in various object geometries are illustrated. Coupling volume theory for electric field and magnetic field sensitive antennas is developed at length and the relation to far field antenna theory is established. Some conclusions about antenna optimisation are drawn.

4. NOTATION

The notation and nomenclature used in this paper for physical quantities will be as defined in ISO 1000, [1]. Sinusoidally varying quantities will be represented by peak (not r.m.s.) value phasors. Lower case variables will be used for instantaneous values of scalars, and bold calligraphic characters for instantaneous values of field vectors. Upper case and upright variables will be used for phasors representing sinusoidally varying quantities, and bold upright Roman characters will be used for phasors representing sinusoidally varying field vectors. There are some traditional exceptions to these rules where Greek and Roman upper case symbols do not differ sufficiently.

We will reduce the results of our analysis to dimensionless ratios where possible, and try to employ concepts that can be seen as having physical meaning.

5. ELECTROMAGNETIC THEORY

5.1. The Complete Laws

The complete laws of electrodynamics were first assembled correctly by Maxwell in the form enunciated in words below.

Faraday's law
The circulation of the electric field vector $\mathcal{E}$ around a closed contour is equal to minus the time rate of change of magnetic flux through a surface bounded by that contour, the positive direction of the surface being related to the positive direction of the contour by the right hand rule.

Ampere's law as modified by Maxwell
The circulation of the magnetic field vector $\mathcal{H}$ around a closed contour is equal to the sum of the conduction current and the displacement current passing through a surface bounded by that contour, with again the right hand rule relating the senses of the contour and the surface.

Gauss' law for the electric flux
The total electric flux (defined in terms of the $\mathcal{D}$ vector) emerging from a closed surface is equal to the total conduction charge contained within the volume bounded by that surface.

Gauss' Law for the magnetic flux
The total magnetic flux (defined in terms of the $\mathcal{B}$ vector) emerging from any closed surface is zero.

With the aid of Gauss' and Stokes' laws of mathematics and the definitions

$$\mathcal{D} = \varepsilon_0\, \mathcal{E} + \mathcal{P} \quad \text{and} \quad \mathcal{B} = \mu_0 (\mathcal{H} + \mathcal{M})$$

these laws may be expressed, when the fields are spatially continuous, in the differential form

$$\nabla \times \mathcal{E} = -\frac{\partial \mathcal{B}}{\partial t}$$
$$\nabla \times \mathcal{H} = \mathcal{J} + \frac{\partial \mathcal{D}}{\partial t}$$
$$\nabla \cdot \mathcal{D} = \rho$$
$$\nabla \cdot \mathcal{B} = 0$$

5.2. Source and Vortex Interpretation

We remain firmly committed to the source and vortex interpretation of those equations, [2]. In that interpretation, the above equations state that the electric field vector $\mathcal{E}$ can have vortices caused by changing magnetic flux, the magnetic field $\mathcal{H}$ can have vortices caused by conduction or displacement currents; the electric flux density $\mathcal{D}$ can have sources caused by conduction charge density; and the magnetic flux density vector $\mathcal{B}$ can have no sources.

In linear media, some of the statements about $\mathcal{D}$ and $\mathcal{B}$ can be extended to $\mathcal{E}$ and $\mathcal{H}$ as well, but when non-uniform fields and boundaries are considered, it can be shown that $\mathcal{E}$, $\mathcal{D}$, and $\mathcal{H}$ can have both sources and vortices, but $\mathcal{B}$ is alone in that it can have no sources.

Figures 2 and 3 below provide archetypical illustrations of the source nature of the electrostatic field and the vortex nature of a magnetodynamic field, as well as illustrations of two of the most important boundary conditions that apply when any electric field $\mathcal{E}$ or a magnetodynamic field $\mathcal{H}$ approaches a conducting surface.

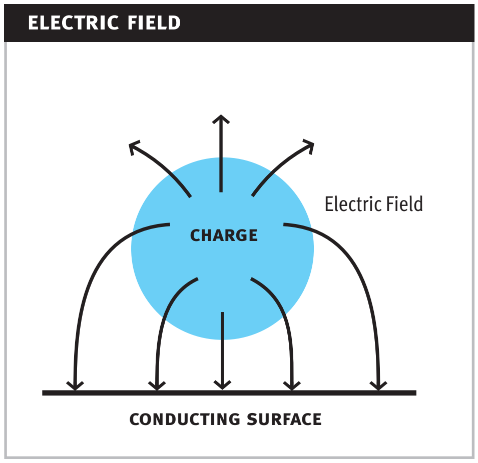


Figure 2: Electric field near a conducting surface

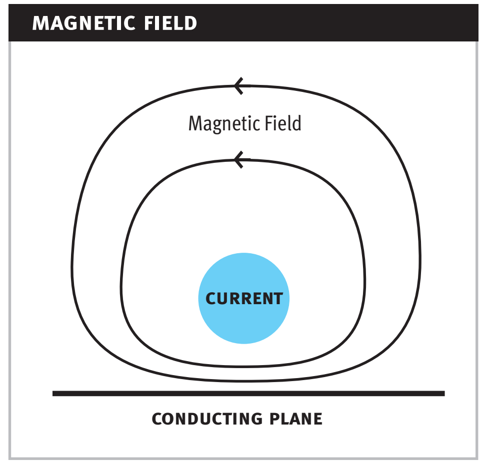


Figure 3: Oscillating magnetic field near a conducting surface

5.3. Boundary Conditions

We have already given simplified illustrations for the most important cases of sinusoidal electric and magnetic fields occurring adjacent to conductors.

A full statement of the electromagnetic boundary conditions is provided in the two paragraphs below. All of those results may be derived from the statement in words of the basic laws given earlier.

What we can deduce from those laws, without taking to account the properties on any materials involved, is that the tangential component of $\mathcal{E}$ is continuous across any boundary; the normal component of $\mathcal{B}$ is continuous across such a boundary; the normal component of $\mathcal{D}$ may be discontinuous across a boundary, with the discontinuity being equal to any conduction charge density ρ_v^c per unit area on the surface; and the tangential component of $\mathcal{H}$ may be discontinuous across a boundary, with the discontinuity being equal to in magnitude and at right angles in direction to a surface current density flowing on the surface.

When we take into account the restrictions imposed by the properties of the materials which may exist on one or other side of the boundary, we may further conclude that: the electric field is continuous across the boundary for all materials and time variations; that there are no electric fields or fluxes, or time-varying magnetic field or flux densities inside a good conductor; that a surface current density can exist only on the surface of a perfect conductor; and that time-varying charge density cannot exist on the surface of a perfect insulator, although in a feline world, a static surface charge density can.

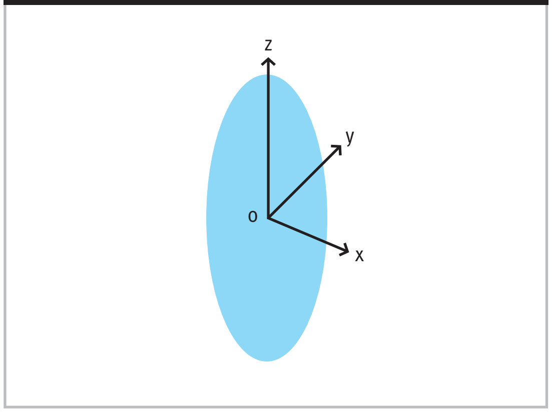
5.4. Demagnetising and Depolarising Factors

The theory of demagnetising and depolarising factors, [3] is useful in gaining an appreciation of how the interior fields of an object differ from the fields exterior to that object, and for calculating useful properties of magnetic cores.

When an ellipsoidal shape as illustrated in Figure 4 below of dielectric or soft magnetic material is introduced into a region in which there was previously a spatially uniform electric field $\mathcal{E}$ or magnetic field $\mathcal{H}$ caused by some distribution of sources (charges or currents), and these sources do not vary as a result, the material becomes polarized with a polarisation vector $\mathcal{P}$ or magnetised with a magnetisation vector $\mathcal{M}$. That polarisation or magnetisation causes on the surface of the shape, induced surface charge densities or magnetic pole densities which make an additional contribution $\mathcal{E}^d$ or $\mathcal{H}^d$ to the fields interior to the shape. These fields are in a direction opposite to the original applied fields $\mathcal{E}^a$ or $\mathcal{H}^a$ and tend to depolarise or demagnetise the material, and hence reduce (but not reverse in direction) the polarisation or magnetisation in the interior of the shape. The internal fields $\mathcal{E}$ and $\mathcal{H}$ are also reduced.

Figure 4: An ellipsoid of dielectric or magnetic material

ELLIPSOID OF DIELECTRIC OR MAGNETIC MATERIAL



The depolarising or demagnetising fields are given by:

$$\begin{bmatrix} \mathcal{E}_x^d \\ \mathcal{E}_y^d \\ \mathcal{E}_z^d \end{bmatrix} = -\frac{1}{\varepsilon_0}\begin{bmatrix} N_x & 0 & 0 \\ 0 & N_y & 0 \\ 0 & 0 & N_z \end{bmatrix}\begin{bmatrix} \mathcal{P}_x \\ \mathcal{P}_y \\ \mathcal{P}_x \end{bmatrix}$$

$$\begin{bmatrix} \mathcal{H}_x^d \\ \mathcal{H}_y^d \\ \mathcal{H}_z^d \end{bmatrix} = -\begin{bmatrix} N_x & 0 & 0 \\ 0 & N_y & 0 \\ 0 & 0 & N_z \end{bmatrix}\begin{bmatrix} \mathcal{M}_x \\ \mathcal{M}_y \\ \mathcal{M}_x \end{bmatrix}$$

where N_x, N_y or N_z are known as depolarising or demagnetising factors. These dimensionless constants depend upon the shape of the ellipsoid, and are subject to the condition

$$N_x + N_y + N_z = 1.$$

This constraint, and relations between N_x, N_y and N_z for certain symmetrical shapes, allow conclusions to be drawn for the cases of the sphere, a long thin rod, or a flat thin desk.

5.5. Retarded Potentials

The retarded potentials listed below may be regarded as integral solutions for Maxwell's equations which are available when charge and current distribution are known.

For the calculation of electric and magnetic fields at a point r_2 caused by a distribution of charge and current at points r_1 over a volume v we may make use of the retarded potentials

$$\Phi(r_2) = \frac{1}{4\pi\varepsilon_0}\int_v \frac{\rho(r_1)e^{-j\beta r_{12}}}{r_{12}}dv$$

and

$$\mathbf{A}(r_2) = \frac{\mu_0}{4\pi}\int_v \frac{\mathbf{J}(r_1)e^{-j\beta r_{12}}}{r_{12}}dv$$

The fields in the sinusoidal steady-state can be derived from these potentials by the equations

$$\mathbf{E} = -\mathbf{grad}\Phi - j\omega\mathbf{A}$$

$$\mathbf{B} = \mathbf{curl}\ \mathbf{A}$$

These formulae may be used in the calculation of the electromagnetic fields generated by oscillating infinitesimal electric or magnetic dipoles as reported below.

5.6. Reciprocity

The integral form of the Lorenz reciprocity relation for two solutions $\mathbf{E_1}, \mathbf{H_1}$ and $\mathbf{E_2}, \mathbf{H_2}$ of Maxwell's equations in the same region and at the same angular frequency ω is under appropriate conditions

$$\int_S (\mathbf{E}_1 \times \mathbf{H}_2 - \mathbf{E}_2 \times \mathbf{H}_1) \cdot \mathbf{ds} = \int_v (\mathbf{J}_1 \cdot \mathbf{E}_2 - \mathbf{J}_2 \cdot \mathbf{E}_1)\, dv$$

where the integral is over a closed surface S bounding a volume v. In the derivation, which proceeds from Maxwell's equations, it is allowed that the material parameters be characterized by complex (to allow for losses) and possibly tensor (to allow for anisotropy) dielectric permittivities and magnetic permeabilities, but gyromagnetic behaviour of magnetic materials is not permitted.

In certain cases the right-hand side becomes zero. These cases include those where the surface is a conducting surface, where the surface encloses all sources, where the surface encloses no currents, and where currents flow only by the mechanism of drift, (but not by diffusion, as they can in a semiconductor).

The theorem has in electromagnetic theory, and on the simplification of electromagnetic theory known as lumped circuit theory, some profound consequences. These include: the symmetry of impedance, admittance and suitably defined scattering matrices; the equivalence of transmitting and receiving antennas; the gain of a lossless transmitting antenna being related to the effective area of the same antenna used as a receiver; and the propagation loss from an interrogator to a label being equal to propagation loss from the label to the receiver when antennas have single ports. Use of these results will be made in data sections dealing with coupling between interrogators and labels.

6. INFINITESIMAL DIPOLE FIELDS

6.1. Electric Dipole

In spherical polar coordinates at a point $P(r,\theta,\phi)$ the non-zero field components of an oscillating small electric dipole of length L carrying a current I and of moment P directed along the z axis where $j\omega\mathrm{P} = \mathrm{I}L$

$$\mathrm{E}_r = \frac{\beta^2 j\omega\mathrm{P}\eta}{4\pi}\left(\frac{2}{(\beta r)^2} - \frac{2j}{(\beta r)^3}\right)e^{-j\beta r}\cos\theta$$

$$\mathrm{E}_\theta = \frac{\beta^2 j\omega\mathrm{P}\eta}{4\pi}\left(\frac{j}{(\beta r)} + \frac{1}{(\beta r)^2} - \frac{j}{(\beta r)^3}\right)e^{-j\beta r}\sin\theta$$

$$\mathrm{H}_\phi = \frac{\beta^2 j\omega\mathrm{P}}{4\pi}\left(\frac{j}{(\beta r)} + \frac{1}{(\beta r)^2}\right)e^{-j\beta r}\sin\theta$$

6.2. Magnetic Dipole

In spherical polar coordinates at a point $P(r,\theta,\phi)$ the non-zero field components of an oscillating small magnetic dipole of moment $\mathrm{M} = \mathrm{I}A$ are

$$\mathrm{H}_r = \frac{\beta^2 j\omega\mu_0 \mathrm{M}}{4\pi\eta}\left(\frac{2}{(\beta r)^2} - \frac{2j}{(\beta r)^3}\right)e^{-j\beta r}\cos\theta$$

$$\mathrm{H}_\theta = \frac{\beta^2 j\omega\mu_0 \mathrm{M}}{4\pi\eta}\left(\frac{j}{(\beta r)} + \frac{1}{(\beta r)^2} - \frac{j}{(\beta r)^3}\right)e^{-j\beta r}\sin\theta$$

$$\mathrm{E}_\phi = \frac{\beta^2 j\omega\mu_0 \mathrm{M}}{4\pi}\left(\frac{j}{(\beta r)} + \frac{1}{(\beta r)^2}\right)e^{-j\beta r}\sin\theta$$

6.3. Characteristics of Near and Far Fields

The above expressions show that the distance $r = 1/\beta = \lambda/(2\pi)$ is of significance in determining the nature of the fields surrounding the dipoles. Within this distance the dominant fields may be recognised as being the same as the energy storage fields with which we are familiar for electrostatic or magnetostatic dipoles. This region is known as the near-field region, and the dominant fields therein are called the near fields, and simply store energy that periodically emerges from, and later disappears back into, the dipole. Outside of this distance, the dominant fields are those associated with energy propagation by electromagnetic waves away from the sources. This region is known as the far-field region, and the dominant fields therein are called the far fields, and transport energy continuously away from the dipoles.

7. ELECTROMAGNETIC COMPATIBILITY CONSTRAINTS

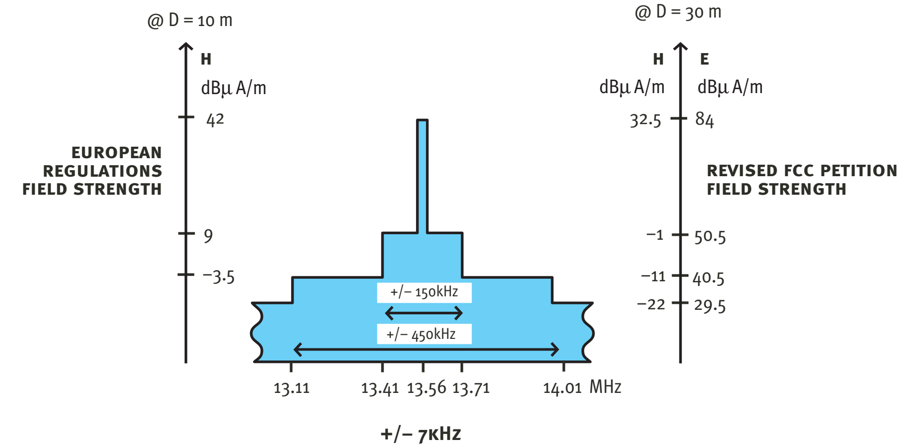
In understanding the effect of electromagnetic compatibility constraints that are normally applied to a RFID systems, we first note that all of the regulations, whether at UHF or at HF, are enforced in the far field. However, as shown in equations of 6.1, near and far fields scale differently with distance, and in particular the near field energy density per unit volume decreases as the inverse sixth power of distance from the antenna. The result is that close to the antenna, substantial energy densities may be obtained, but these diminish very quickly as distance increases.

Thus although the HF electromagnetic compatibility regulations, shown in Figure 5 below, allow only minimal (5 mW) radiation, well inside the radian sphere distance of $r = 2\pi/\lambda$ it is practicable at HF to obtain a sufficient energy storage field for the operation of a RFID label. However, as efforts to increase reading distance are made, the previously mentioned inverse sixth power of the reactive power density is generally sufficient to reduce the label energising signal to a level below that for practical operation before the boundary of the far field, where its less severe inverse square dependence of energy density, is reached. Thus under current regulations operation of HF systems is almost entirely confined to the near field and short distances.

In an endeavour to improve the range of HF RFID systems, efforts have been made, through the design of clever quadrupole antenna systems, to minimise far field radiation while enhancing close-in near fields. While these efforts do achieve some improvement, maintaining the necessary balance between antenna elements which are intended to produce far field cancellation is difficult to achieve in practice.

Figure 5

HF ELECTROMAGNETIC COMPATIBILITY REGULATIONS



This focus on minimizing far-field radiation does tend to divert attention from an alternative productive approach to gaining long interrogation range. This approach is that of using greater label antenna sizes. To the extent that this approach is successful, ranges will increase to approach the mid field distance. Once this happens, it is reasonable to assume that many antenna fields will, in the ratio of mid field to far field amplitude, replicate the value that can be estimated from the dipolar field expressions given in Section 6.1. Once this is done, we conclude that we cannot really have a strong mid field without a related strong value for the far field, i.e. without some radiation, and this conclusion does not significantly change with interrogator antenna size. In consequence, we can see that the rational approach to achieving long interrogation range includes the step of arguing for a greater allowed radiation from HF interrogator antennas, at least in respect of their carrier level, if not in the signalling sideband level.

It may be noted that the central portion of the spectrum shown in Figure 5, which regulates the operation of RFID equipment in the HF region, is an industrial scientific and medical (ISM) band, in which there is no limit to the radiation that is allowed for equipment which receives the classification industrial, scientific or medical. It may be argued that a substantial increase in HF radiation over the narrow central portion of the band shown in Figure 5, could be allowed with substantial technological benefits.

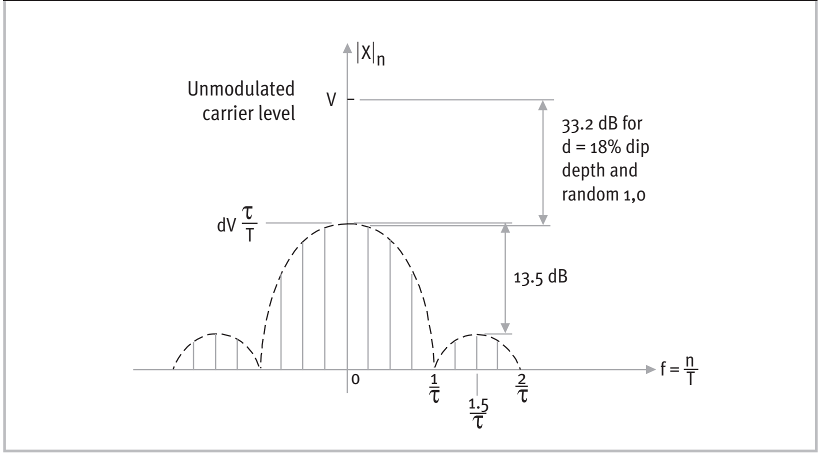
In Figure 5, the sidebands come from interrogator signalling in “reader talks first” systems. Reasonably frequent interrogator signalling is required in systems that manage large label populations. The result can be significant interrogator signalling sideband levels in such systems.

Interrogator signalling is generally by means of dips in interrogator signal amplitude. Dips of 18%, with a 1 in 4 duty cycle, and a symbol set which requires one dip for a binary one and no dip for a binary zero have recently been adopted by the Auto-ID Centre.

Calculation of the sideband signalling spectrum resulting from these parameters is easily performed from the properties of Fourier transforms of single pulses and Fourier series for periodic functions. The result is illustrated in Figure 6 below for the above parameters and for random 1,0 signalling. Conformance to the regulations of Figure 5 is evident. However, if regulations are revised to permit an increased un-modulated carrier level but provide for no increase in sideband signalling levels, techniques to reduce the relative proportion of sideband signalling levels will have to be found. Fortunately, such techniques are available, and have been built into Auto-ID specifications for HF labels as an option.

Figure 6

ILLUSTRATION OF ELECTROMAGNETIC COMPATIBILITY CALCULATIONS AT HF



8. FIELD CREATION STRUCTURES

8.1. Near Field Creation

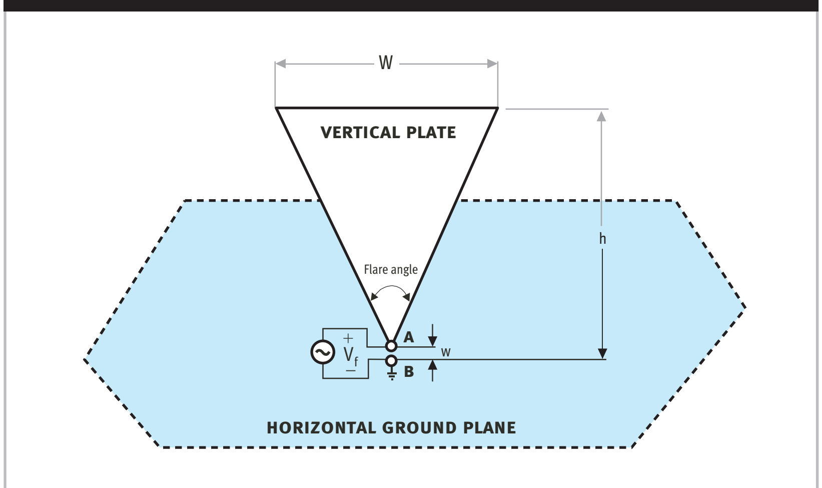
We first illustrate in Figures 7–10 and Figure 13 below structures which may be useful in the HF region. Such structures can generate either electric or magnetic fields, either in the near-field, or the mid-field, that being the field at the boundary between the near field and the far field.

As a consequence of the value (22 m) of the electromagnetic wavelength at 13.56 MHz, the structures are always in practice electrically small. The first three can be considered as creating in the near-field mainly electric field, but as a consequence of the expressions given earlier for dipole fields, which apply to some extent to this situation, the structures will also create some lesser value of near magnetic field.

In the far-field these structures will create electric and magnetic fields in equal proportion, in the sense that for radiated fields in the far-field $|\mathbf{E}| = \eta\,|\mathbf{H}|$.

Figure 7: A monopole plate antenna above a ground plane

A MONOPOLE WEDGE ABOVE GROUND ANTENNA

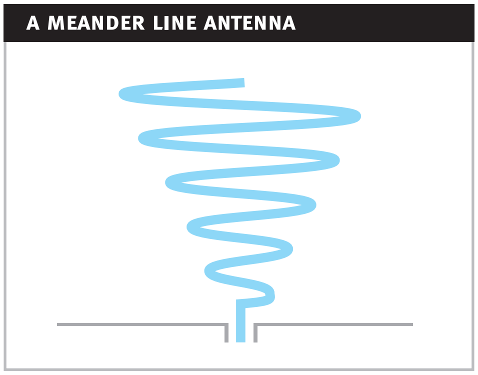


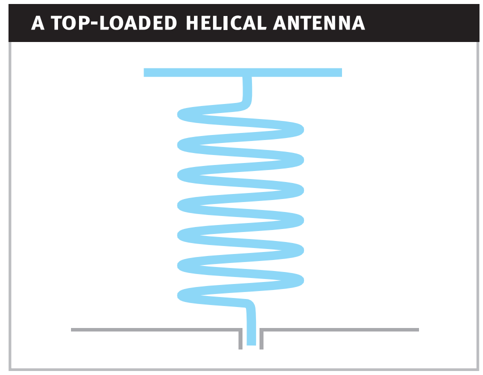
The driving point impedance properties of the structure are given in Section 10.7.

An alternative is the meander line structure illustrated below. This and the next structure are designed to have a similar field creation properties to the former, with small variations in the impedance properties to provide simpler driving requirements.

Below we illustrate a top-loaded helical structure which may be useful for creating large volume near-fields.

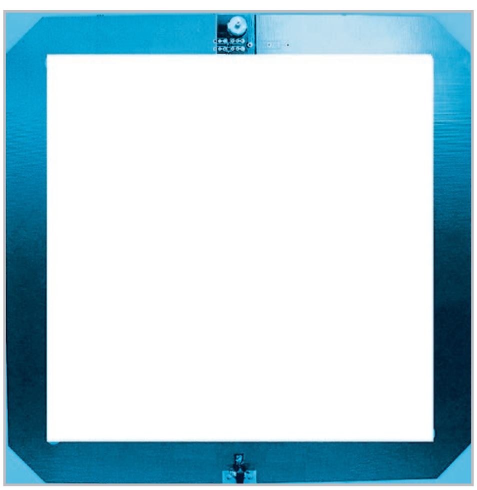
Figure 8, Figure 9





As HF tags often couple to magnetic fields because they are less easily stopped by conducting materials than are electric fields, there is in fact a more common interest in the creation of strong near-field magnetic fields. The usual structure by means of which this is achieved is a small current carrying loop such as is illustrated in Figure 10. This loop has tuning and matching elements at the top end. A strip line transmission line (not visible in the picture) on the underside of the right hand half conveys the driving signals from the connecting point at the centre of the bottom to the driving point terminals at the centre of the top.

Figure 10: A small loop antenna

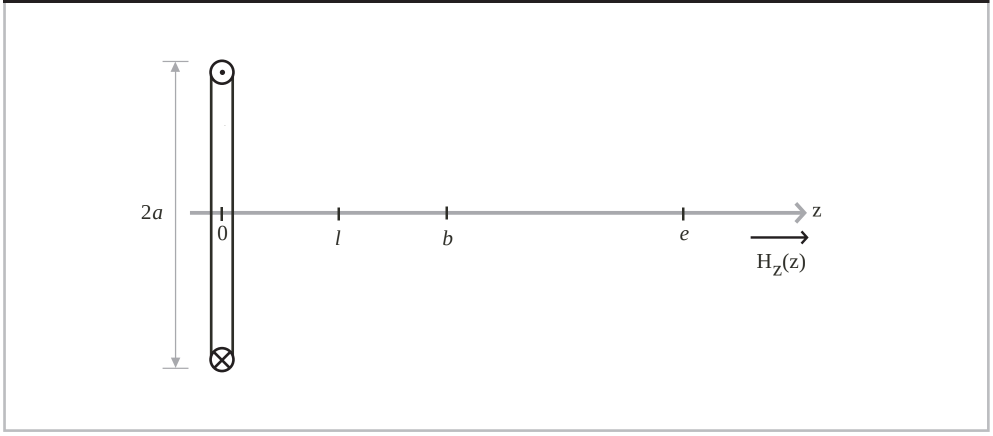


As the algebra of Sections 6.1 and 6.2 shows, magnetic dipoles are superior to electric dipoles in the creation of magnetic fields in the near-field.

As is clear from the diagram, the antenna is made from metallic strip. In the mathematical analysis of such a structure, formulae for coils made from round wires are often used, with the diameter of the wire being set to half the strip with.

Figure 11

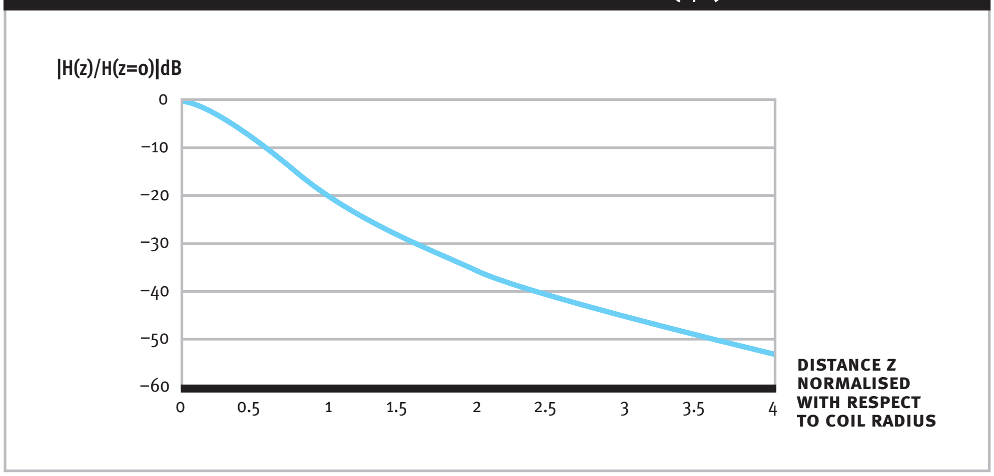
GEOMETRY FOR CALCULATING LOOP ANTENNA NEAR FIELDS



The task of exciting a label in the near-field is substantially one of creation of a stored energy density per unit volume in the volume served by the antenna. As shown in Figure 12, below, the smaller of the loop, the more the field is reasonably confined to the volume near the antenna, and the less the range of the antenna, but less total stored energy that needs to be provided.

Figure 12: Variation of loop field with distance along axis

RELATIVE MAGNETIC FIELD STRENGTH AT A NORMALISED DISTANCE (Z/A) FROM A CIRCULAR COIL



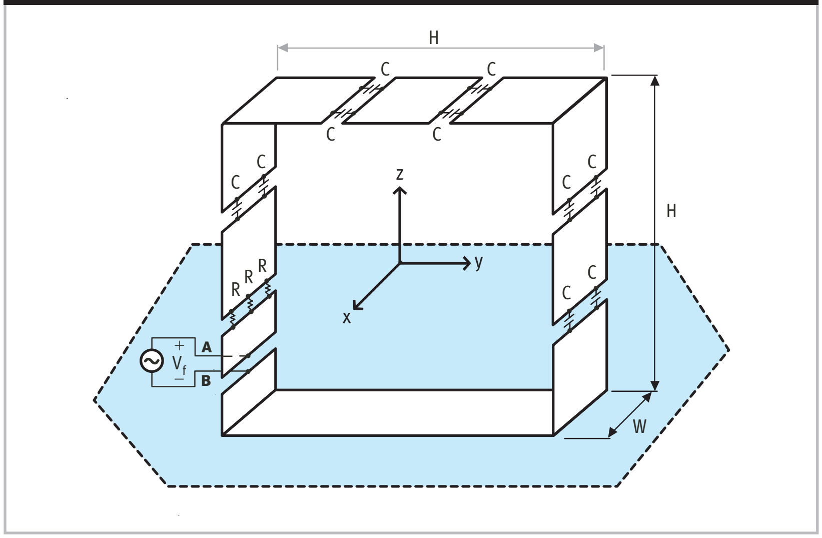
But there is another feature of small loops that is of interest. If we take our task as that of maximizing the magnetic field at the label position l in the near field for a given electromagnetic compatibility enforcement distance e in the far field it can be shown that the optimum loop size is one where the radius a tends to zero.

Of course this is not a practical solution, as it has neglected important practical consequences and practical constraints. One of them is that as the antenna size is made small, an increasing concentration of very high and unused (because the label is further away) energy density per unit volume is created closer to the loop centre than is the label, with the results that the total stored energy becomes very large, the quality factor of the antenna becomes unreasonably high, and also the real power density required to drive it becomes too large.

For this reason, attention passes to the construction of larger loops such as shown in the Figure 13. In such structures, and effort has been made to counteract the tendency of the current distribution on long wire antennas to become sinusoidal by breaking the loop resonating capacitance into several series elements that are distributed around the loop. Such structures have been used by ourselves to allow substantially increased range from high frequency labels.

Figure 13

A LARGE LOOP ANTENNA



8.2. Far Field Creation

The almost universal choice for the creation of far-fields, particularly in the UHF region, is the patch antenna illustrated below.

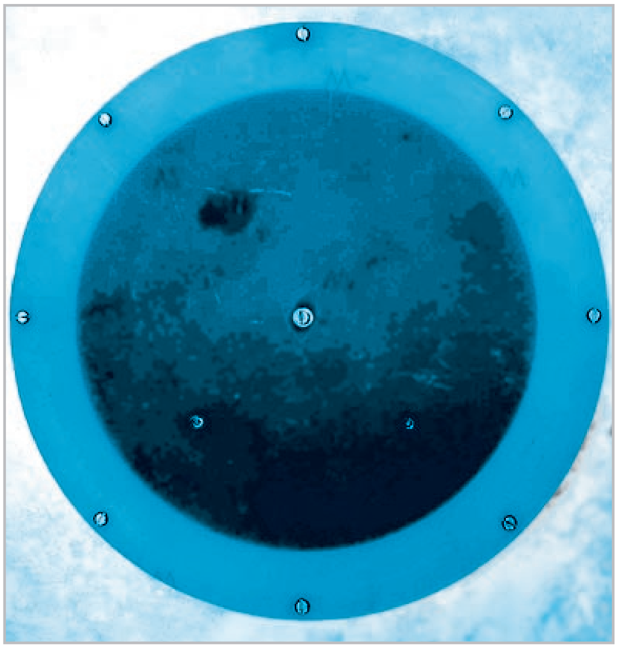


Figure 14: A patch antenna

9. ANTENNA SIZE EFFECTS

RFID label antennas may be electrically large or small depending upon frequency band. Interrogator antennas may also be electrically large or small depending upon frequency band. We will consider in this section some of the issues that arise.

9.1. Measures of Label Excitation

Appropriate measures of label excitation in the near field are energy storage or oscillating reactive power per unit volume, and in the far field energy propagating per unit time per unit area. At HF, we will model tag antennas using **coupling volume concepts** to be defined Section 10. In that modelling we can recognise both electric field sensitive and magnetic field sensitive designs.

9.2. Magnetic Field Sensitive Antennas

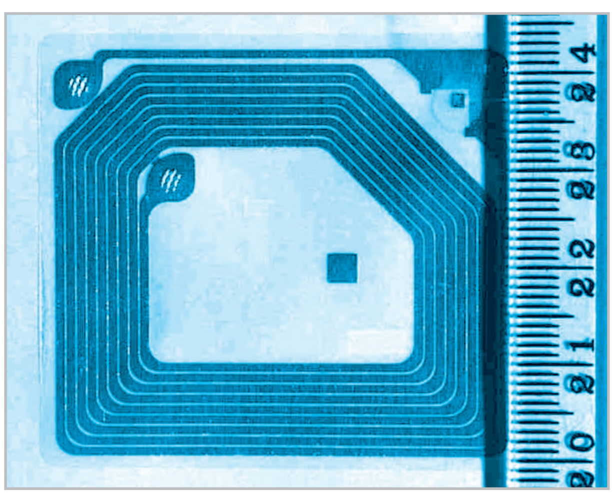
A common example of a magnetic field sensitive HF label is shown in Figure 15 below. The label is 42 mm wide by 47mm high.

The label is designed to have sufficiently many turns to provide the resonating inductance for the microcircuit input capacitance, as well as a flux collecting area in the interior which is as large as practicable consistent with the size requirement for the label.

Advantages of working in the near field at HF rather than at LF is that the number of turns required to resonate the microcircuit capacitance is small enough for low resolution lithography to be used in antenna construction, and no additional external resonating capacitance is required.

When a larger space is available for the label, a larger coil area should be used. As shown in Figure 16, fewer turns are then needed to obtain the required tuning inductance. It will be shown later that the figure of merit (the coupling volume) for a planar coil operating in the near field varies as the third power of size. Thus the antenna of Figure 16 is about 18 times more sensitive than that of Figure 15. Unfortunately, this increased sensitivity does not translate to a corresponding increase in range, as small coil interrogator antennas have an inverse sixth power decrease in energy density per unit volume as distance from the interrogator increases.

Figure 15: A magnetic field sensitive HF label antenna



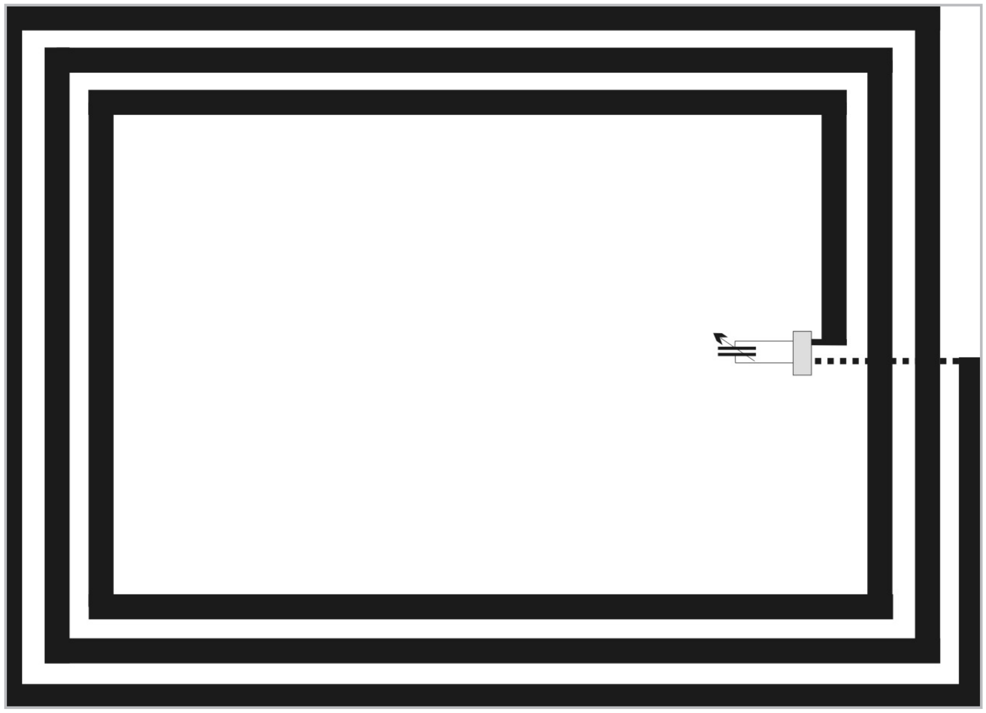
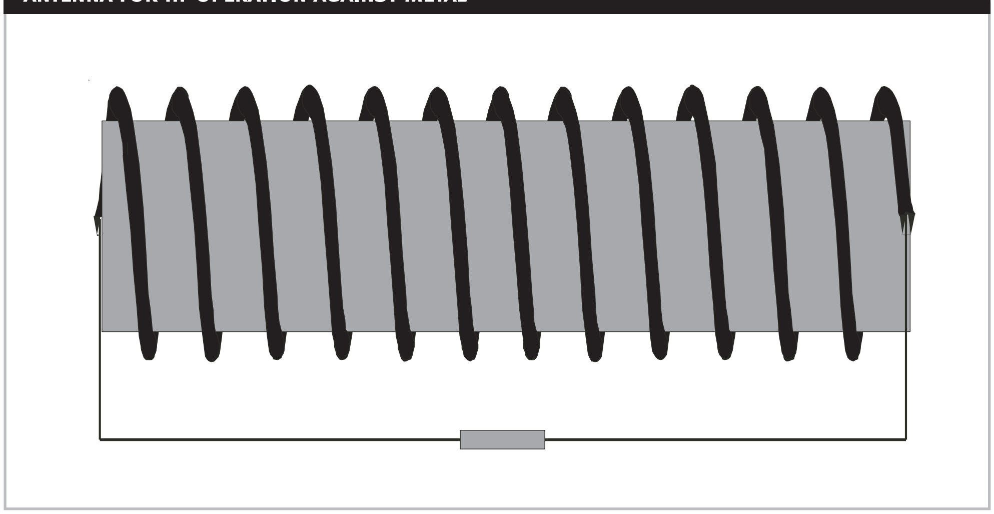


Figure 16: A larger loop antenna for an HF label

Clearly, both of the designs illustrated above are unsuitable for being placed flat against metal, as the boundary conditions shown in Figure 3 will not allow a normal component of magnetic flux density at the metal surface. For this situation, the label antenna employing a solenoid with a magnetic core design shown in Figure 17, is employed.

Figure 17

ANTENNA FOR HF OPERATION AGAINST METAL



It will be shown in Section 10.3 that without the magnetic core the coupling volume of a long solenoid is just the physical volume, but when a magnetic core is inserted, the coupling volume increases by factor equal to the effective permeability defined in that later section.

This behaviour may be contrasted with that of electric field labels, in which in Section 10.6 we will find that the inclusion of dielectric material into the interior of the label is not helpful.

9.3. Electric Field Sensitive Antennas

Two varieties of electric field sensitive antennas are shown in Figures 18 and 19 below. Figure 18 shows a small bow tie antenna that is intended to be sensitive to electric fields in the horizontal direction.

Figure 18

AN ELECTRIC FIELD SENSITIVE LABEL

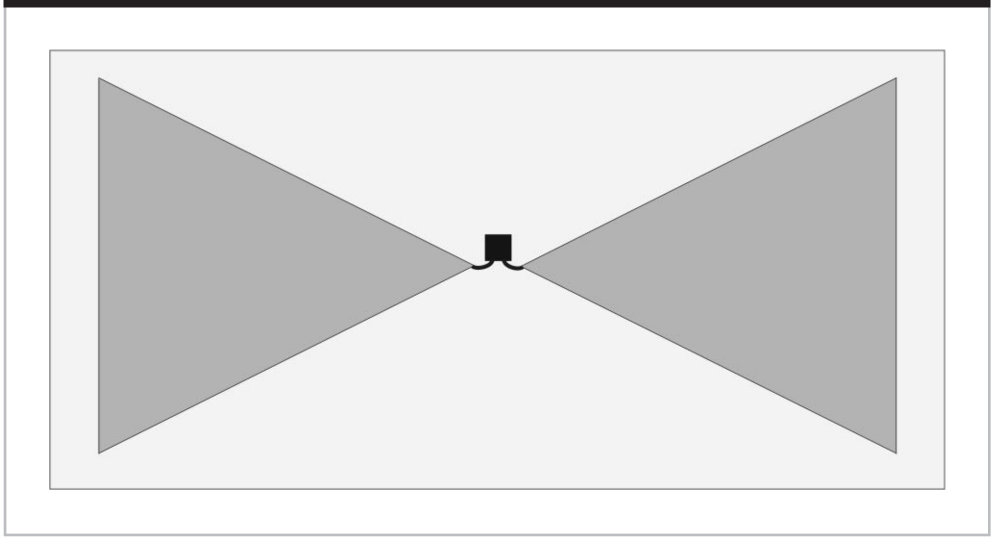
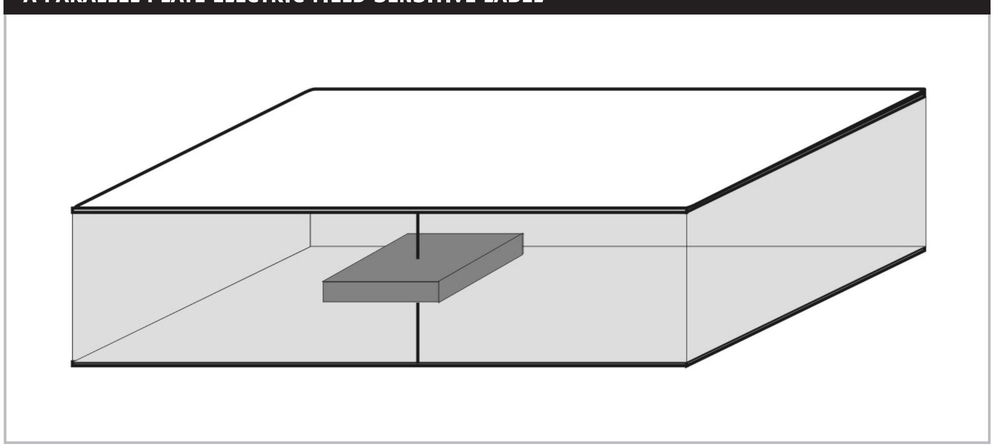


Figure 19 shows an electric field sensitive antenna that is suitable for placement against a horizontal metal plate.

Figure 19

A PARALLEL PLATE ELECTRIC FIELD SENSITIVE LABEL



Analysis of both of these structures is provided in a later section. The analysis has a common feature that the figure of merit for these antennas, when placed in the energy storage electric field is a coupling volume, in the second case equal to the physical volume of the structure, and in the first case a volume derived more from the label dimensions, even though the antenna itself has no physical volume.

Both of the antennas will also have an effective area, but this area should not be confused with the effective area concept of a radiating antenna or a far field antenna. The effective area for a near field electric field sensitive antenna describes the extent to which the antenna can extract current from the displacement current density of the driving electric field.

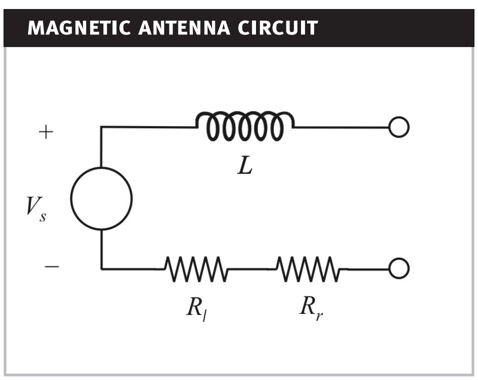
9.4. Label Antenna Equivalent Circuits

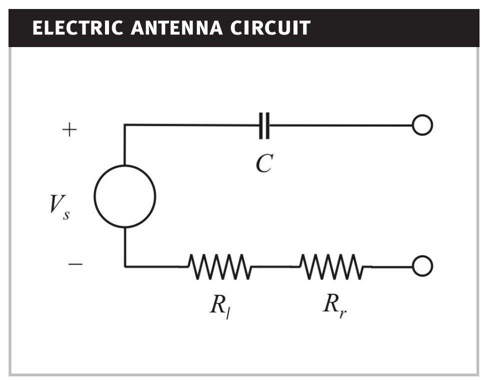
Equivalent circuits for small magnetic field sensitive antennas and electric field sensitive antennas are shown in Figures 20 and 21 below. The range of validity of these equivalent circuits is where the reactance properties of the antenna may be described by a single parameter, L or C. When the antenna is a larger, as will be reported in a later section, reactance properties might well be described by an appropriate mixture of L and C.

In Figure 20, the source voltage is the voltage induced in the flux collecting area of the coil by magnetic fields other than those fields which resulted from current flowing within the coil itself. Those induced voltages are represented by the voltage drop in the inductor L. The parameters R_l and R_r are loss and radiation resistances respectively.

Figure 20: Equivalent circuit for a small magnetic field sensitive antenna

Figure 21: Equivalent circuit for a small electric field sensitive antenna





In Figure 21, the source voltage is the voltage developed across the self capacitance of the antenna as a result of the current injected into it, when it is short-circuited, by the displacement current density of the electric field in which the antenna is in immersed. The parameters R_l and R_r are loss and radiation resistances respectively.

Calculation of the parameters of the magnetic field sensitive antenna is straightforward, the relevant formulae being obvious or contained in the Appendix. For the electric field antennas, determination of the relevant parameters is sometimes not quite so simple, as electrostatic field solutions for the relevant geometries are not readily available. Therefore empirical results or numerical modelling are more commonly employed for useful shapes. Results will be reported in a later section for some significant structures.

9.5. Electromagnetic Field Antennas

We consider an antenna as an electromagnetic field antenna on a couple of different bases. Firstly, if the antenna is capable of responding to both electric and magnetic fields we would consider it to be an electromagnetic field antenna. It is almost invariably true that unless the antenna is very small, it does have this property. Proper analysis requires that it be analysed using the full set of Maxwell's equations, rather than the subset or simplified versions that pertain to electrostatic or magnetostatic problems. A good example of this phenomenon is provided by the electromagnetic field sensitive antenna shown in Figure 22, in which there is no obvious effort to couple to either electric or magnetic field alone.

Figure 22: An electromagnetic field sensitive antenna

Such electromagnetic antennas are generally useful for operation in the far field, because far field interrogation systems have shorter wavelengths, and antennas of acceptable size can no longer be considered to be electrically very small, but are merely small.

Despite this distinction, there are electromagnetic label reading environments in the UHF region in which, through reflections, either the electric or magnetic field is emphasised at the expense of the other. For such situations it is normally useful to take into account the nature of the driving fields in antenna design, and to shape the design so that it is recognisably attuned to one or other of those fields.

10. COUPLING RELATIONS

10.1. Near and Far fields

In the analysis of the performance of RFID systems it is important to consider whether the labels are placed in the far (propagating) or near (energy storage) fields of the interrogator antenna. When that antenna is of small gain, the distance which divides the near and far fields is the size or the radian sphere of radius $r = \lambda\backslash 2\pi$, where λ is the free space electromagnetic wavelength at the operating frequency.

10.2. Field Measures

For a linearly polarised magnetic field described by a peak value phasor $\mathbf{H}$ we may develop the two measures of the exciting field

$$\text{Radial component of Poynting vector } S_r = \frac{\eta\,|\mathrm{H}|^2}{2} \text{ in Wm}^{-2}$$

$$\text{Volume density of reactive power } W_V = \frac{\omega\mu_0|\mathrm{H}|^2}{2} \text{ in VAm}^{-3}$$

For a linearly polarised magnetic field described by a peak value phasor $\mathbf{E}$ we may develop the two measures of the exciting field

$$\text{Radial component of Poynting vector } S_r = \frac{|\mathrm{E}|^2}{2\eta} \text{ in Wm}^{-2}$$

$$\text{Volume density of reactive power } W_V = \frac{\omega\varepsilon_0|\mathrm{E}|^2}{2} \text{ in VAm}^{-3}$$

In both cases the latter expression is ω times the peak value of stored magnetic or electric energy per unit volume. We can easily show that if β is the propagation constant at the frequency used

$$W_v = \beta S_r .$$

The last expression assumes that we are in the far field, i.e. there are no near-field reactive energy storage fields to augment W_v without contributing to S_r.

10.3. Near Field Operation – Magnetic Field

For near field operation, the fields which excite the label are basically energy storage fields for which consideration of power flow is inappropriate, and for which pre-Maxwell versions of electrodynamics can give correct calculations. Both the interrogation field creation means and the label field detection means can be considered as weakly coupled inductors of self-inductances L_1 and L_2 and mutual inductance M. When both those coils are tuned to resonance with respective quality factors Q_1 and Q_2, it can be shown that the power P_2 dissipated in the losses of the label coil is related to the power P_1 dissipated in the losses of the "transmitter'" coil by

$$\frac{P_2}{P_1} = k^2 Q_1 Q_2 \text{ where } k = \frac{M}{\sqrt{\mathrm{L}_1 L_2}}$$

This equation is useful to show the role of the quality factor, Q, of the resonances in both the label and interrogator coils in promoting power transfer, but it is not useful in separately optimising the properties of those two widely dissimilar elements.

For that purpose we can focus first on the **energy storage** measure of exciting field, which is the reactive power per unit volume in the field created by the interrogator at the label position. In terms of that measure, we can define, [4] a figure of merit of a label antenna as the ratio

$$V_c = \frac{[\text{Reactive power flowing in the untuned label coil when it is short circuited}]}{[\text{Volume density of reactive power created by the interrogator at the label position}]}$$

which clearly has the dimensions of volume, and is for this reason called the **coupling volume** of the label antenna. For the performance of the interrogator antenna, we can define the companion concept of **dispersal volume** V_d given by

$$V_d = \frac{[\text{Reactive power flowing in the inductor of the interrogator field creation coil}]}{[\text{Volume density of reactive power created by the interrogator at the label position}]}$$

When both antennas are tuned it is possible to show

$$\frac{P_2}{P_1} = \frac{V_c}{V_d} Q_1 Q_2$$

The benefit of this formulation is that the coupling volume is a property of the label parameters alone, and the dispersal volume is a property of the interrogator antenna parameters alone (and of the label position), and separate optimisation becomes possible, whereas k^2 is a complex function of the entire system geometry.

Coupling volumes for various label antenna are readily determined. For an air cored solenoidal antenna it is approximately the volume of that antenna, and when a magnetic core is in place that volume becomes multiplied by the relative effective permeability

$$\mu_{er} = \frac{\mu_{ir}}{1 + N(\mu_{ir} - 1)}$$

where μ_{ir} is the relative intrinsic permeability of the core material and N is the demagnetising factor in the direction of the interrogator field.

For a planar coil, which in its idealised state has no physical volume, the coupling volume is given by

$$V_c = \frac{\mu_0 A^2}{L}$$

where A is the flux-collecting area (incorporating by summation an area for each turn) of the coil, and L is the self inductance.

10.4. Near Field Operation – Electric Field

In RFID systems, the coupling can be via the magnetic field or the electric field. However it is almost always by way of the magnetic field. Nevertheless it is possible to couple to the electric field whether in the far field or the near field.

The energy transfer is provided by the electric flux terminating on the antenna surface and inducing a charge on the antenna. The induced charge will oscillate as the field around it oscillates. The induced charges will thus produce a current.

The issues of understanding and optimising coupling to the electric field are important. The coupling volume theory developed here for the electric field case assists in that effort and allows comparisons between different antenna structures for their effectiveness and efficiency in terms of their actual physical volume and the coupling volume. It is also of interest in making comparisons with the magnetic field case.

When we can focus on the **energy storage** measure of exciting field, which is the reactive power per unit volume in the field created by the interrogator at the label position, we can define a figure of merit of a label antenna as the ratio

$$V_c = \frac{[\text{Reactive power flowing in the untuned label capacitor when it is open circuited}]}{[\text{Volume density of reactive power created by the interrogator at the label position}]}$$

which clearly has the dimensions of volume, and is for this reason also called the **coupling volume** of the label antenna.

For the performance of the interrogator antenna, we can define the companion concept of **dispersal volume** V_d given by

$$V_d = \frac{[\text{Reactive power flowing in the capacitance of the interrogator field creation electrodes}]}{[\text{Volume density of reactive power created by the interrogator at the label position}]}$$

When both antennas are tuned it is possible to show

$$\frac{P_2}{P_1} = \frac{V_c}{V_d} Q_1 Q_2$$

The benefit of this formulation is that the coupling volume is a property of the label parameters alone, and the dispersal volume is a property of the interrogator antenna parameters alone (and of the label position), and separate optimisation becomes possible.

Coupling volumes for various label antennas are readily determined, as will be done in the next section.

10.5. Coupling Volume of a General Shape

In order to derive the coupling volume of antenna structures it is important to define a number of concepts. For a given antenna we can define an **electric flux collecting area** as the area in space required by the antenna to generate from the displacement current of the exciting field an oscillating current I when the antenna is placed in an oscillating electric field of flux density D. This area may or may not be equivalent to the physical area. This electric flux collecting area will be denoted by the symbol A_f will be referred to as effective area.

The formula for the phasor representing the current I from an antenna placed in an electric field with a flux density represented by the phasor D and oscillating angular frequency w, using the effective area A_f of the antenna structure is

$$\mathrm{I} = j\omega \mathrm{D} A_f$$

The reactive power flowing in the antenna with a self-capacitance C when the antenna is open circuit is given by

$$\text{Reactive power} = \frac{|\mathrm{I}|^2}{2\omega C}$$

The reactive power W_v per unit volume in the field is obtained by

$$W_V = \frac{\varepsilon_0 |\mathrm{E}|^2 \omega}{2}$$

Thus the coupling volume of an antenna structure with a self-capacitance of C can be obtained as

$$V_c = \frac{\varepsilon_0 A_f^2}{C}$$

This formula provides the coupling volume of any antenna with a self capacitance of C and an effective are of A_f. It should be pointed out that A_f can easily be determined experimentally by measuring the current flowing from an antenna placed in an oscillating electric field of known strength.

$$A_f = \frac{|\mathrm{I}|}{\varepsilon_0 |\mathrm{E}| \omega}$$

10.6. Coupling Volume of a Rectangular Capacitor

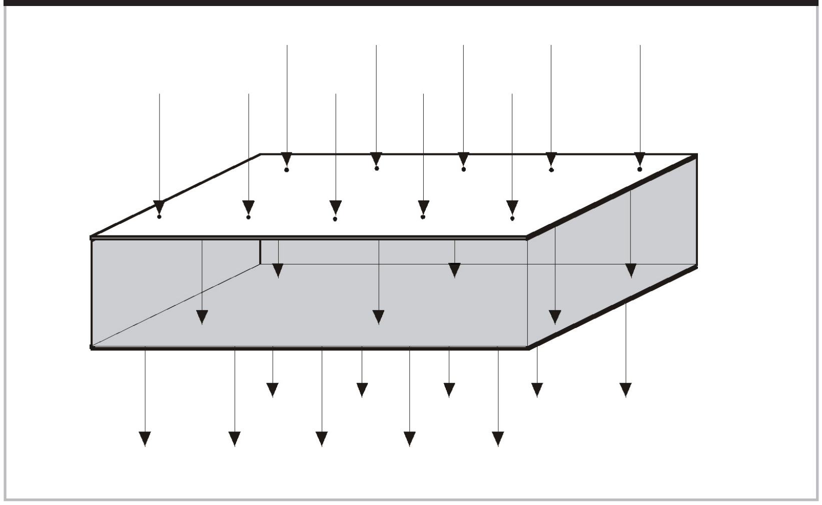
The derivation discussed below will prove that the coupling volume of an air dielectric parallel plate capacitor is equal to the physical volume of the capacitor.

Consider a parallel plate capacitor antenna, as shown in Figure 23, consisting of two plates, each having an area A and immersed in an external electric field E. Then the charge Q distributed on the plate is given by

$$Q = \varepsilon_0 A \,|\,\mathrm{E}\,|$$

Figure 23

FIELD CONFIGURATION FOR A PARALLEL PLATE ELECTRIC FIELD ANTENNA



The capacitance C of a parallel plate capacitor is given by

$$C = \frac{\varepsilon_0 \varepsilon_r A}{d}\,.$$

Now the flux density between the plates is the same as the flux density outside, but the field strength between the plates is $1/\varepsilon_r$ times the field outside. So the voltage V between the plates is

$$V = \frac{|\,\mathrm{E}\,|\,d}{\varepsilon_r}$$

The energy stored in the capacitor is obtained as

$$\frac{1}{2}\frac{\varepsilon_0 A \,|\,\mathrm{E}\,|^2\, d}{\varepsilon_r}$$

As the energy stored per unit volume in an electric field is $\frac{1}{2}\varepsilon_0\,|\,\mathrm{E}\,|^2$, we have $V_c = \frac{Ad}{\varepsilon_r}$.

Examination of the above formula reveals a number of results. Coupling volume has a dependence on the value of the relative permittivity ε_r, and a dielectric if present will reduce the coupling volume. This last result stands in contrast to the case for a rectangular magnetic coil, where effective relative permeability μ_{er} multiplies the coupling volume, not divides it. However, for an air dielectric capacitor, the coupling volume is equal to the physical volume.

It can be observed that the coupling volume of a general structure reduces to the coupling volume of a parallel plate capacitor when

$A_f = A$ (area of a plate) and $C = \frac{\varepsilon_0 \varepsilon_r A}{d}$.

10.7. Properties of the Bow Tie Antenna

Calculating the self-capacitance as depicted by the field lines of Figure 24 of the bow tie antenna by seeking analytical solutions to Laplace's equation presents a difficult problem. Nevertheless a numerical solution is both tractable and much simpler under the present circumstances and has been performed. The method of moments provides a suitable numerical approximation to the self capacitance of a bow tie antenna.

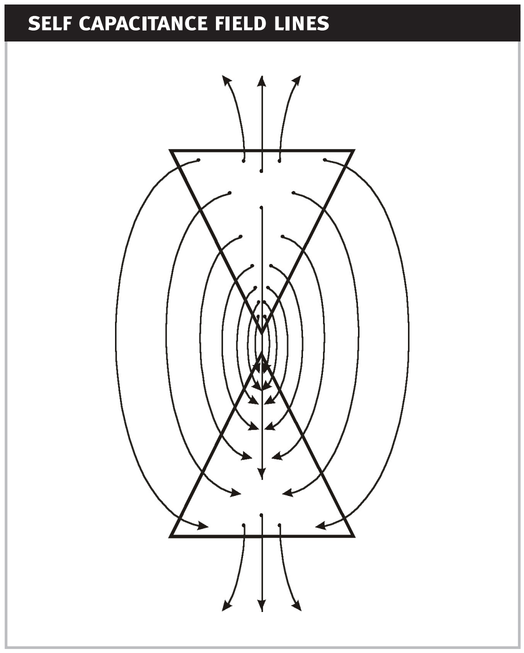


Figure 24: Field configuration for self capacitance calculation of bow tie antenna

In addition, a detailed study undertaken of results first published by Woodward, [5] provided the following empirical model for an equivalent circuit shown in Figure 25 of a monopole wedge above ground plane. The model derived from Woodward's results for the reactance $X(\omega)$ of a bow tie antenna provides the values of a capacitor, whose value is that of the self-capacitance of the bow tie, and an inductor placed in the series circuit shown.

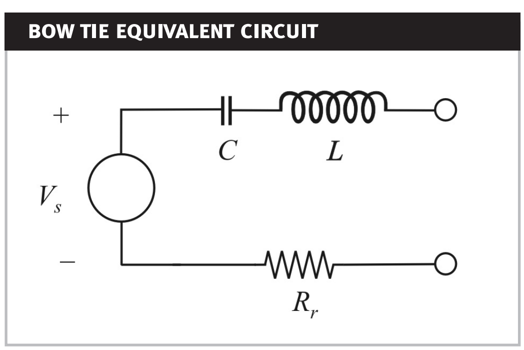


Figure 25: Three parameter equivalent circuit for a bow tie antenna

$$jX = \frac{1}{j\omega C} + j\omega L .$$

This model is only suitable for electrically small antennas obeying the strict limit of

$$h << \frac{\lambda}{6}$$

where h is the height of the antenna and λ is the wavelength, measured in meters.

The model parameters obtained will vary for different flare angles of the bow tie antenna. However the capacitance can be expected to scale up with increasing height for any constant flare angle.

The following results give the capacitor and inductor model values for a bow tie antenna with a flare angle of 90°.

$$C = K_C \varepsilon_0 h \quad \text{with} \quad K_c = 3.8$$

$$L = K_L \mu_0 h \quad \text{with} \quad K_L = 0.4270$$

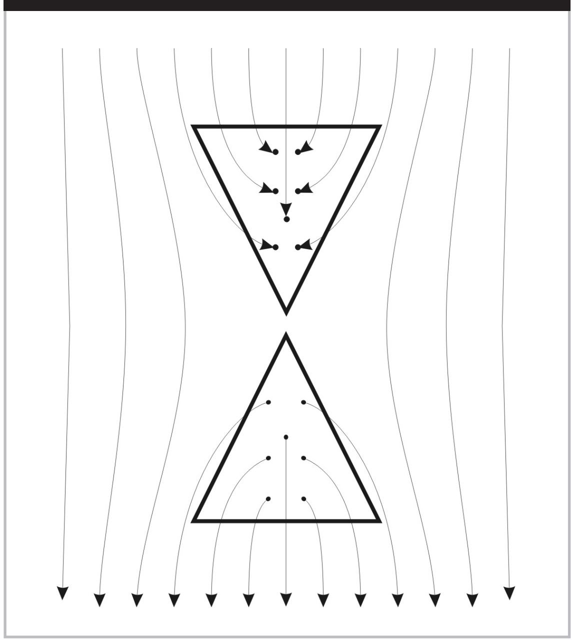
where h is the height of the bow tie antenna.

The significant finding is that, as expected, the low frequency impedance of a bow tie antenna is mainly capacitive and this value can thus be obtained by calculating the self-capacitance of the bow tie antenna.

Figure 26: Field configuration for calculation of effective area of bow tie antenna

Table 1: Table of flare angles and radiation resistance constants for a bow tie antenna of height h

INDUCED CHARGE FIELD LINES



FLARE ANGLE	K_R IN Ω
5	30
10	35.4
30	45.2
40	50
50	52.8
90	60

Woodward's results and experimental results of radiation resistance of monopole wedge above ground antennas have been used to produce a value for the radiation resistance R_r identified in Figure 25 for different antenna heights and flare angles. The following generic formula can be applied to evaluating the radiation resistance R_r of a bow tie antenna antenna.

$$R_r = K_R (\beta h)^2 \quad \Omega$$

where the constant K_r, measured in Ω, is a specific value that depends on the flare angle of the bow tie antenna. Table 1 provides a set of values for K_r derived from Woodward's results [5].

The radiation resistance obtained has significance in two ways. It allows the amount of radiated power to be calculated for a transmitting antenna, and also provides for a label antenna a means of calculating, using the reciprocity theorem, the effective electric flux collecting area, as depicted in Figure 26, of the antenna.

All of these results agree with the results of our own direct measurements and our own numerical analysis performed using the method of moments.

10.8. Far Field Relations

For calculation of the power coupled in the far field to a label with a lossless receiving antenna the usual approach is to derive the available source power from the label antenna from the formula

$$P_r = S_r A_e = \frac{g_r \lambda^2}{4\pi} S_r$$

where A_e and g_r are the effective area and gain of the label antenna, related as shown above by reciprocity considerations. If the label has losses, or is not conjugately matched, the power delivered to the external load is derived from simple modifications of the above formula. It should be noted that the effective area for the far field is a concept unrelated to either the magnetic flux collection area of a loop or the electric flux collection area of a plate.

The power flow per unit area is given by

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

wherein g_t is the gain of the transmitter antenna and P_t the power which it transmits, and r is the distance from the transmitter antenna to the label position. In using this formula, we are implicitly assuming that the label has been placed in the direction of strongest radiation from the interrogator antenna.

The power P_r which may be extracted under optimum conditions of tuning and matching by a lossless label antenna placed at the above position is given by

$$P_r = A_{er} \times \text{Power flow per unit area}$$

wherein A_{er} is a property of the label known as its **effective area**. It is unrelated to the physical area of the antenna, (which if it is just a piece of thin wire, does not have a physical area), but has the desirable property that we may imagine the label antenna collects all of the radiated power that flows through that effective area that may be thought of as surrounding the label antenna.

The **Lorenz reciprocity theorem** of electrodynamics may be used to show that the effective area of a receiving antenna is related to the gain g_r it would have in a transmitting role by the equation

$$A_{er} = \frac{g_r \lambda^2}{4\pi} .$$

$$\frac{P_r}{P_t} = g_r g_t \left(\frac{\lambda}{4\pi r} \right)^2$$

and

$$\frac{P_r}{P_t} = \frac{A_{et} A_{er}}{\lambda^2 r^2} .$$

11. RELATIONS BETWEEN FORMULATIONS

11.1. General Remarks

As the Poynting vector-effective area formulation and the coupling volume-dispersal volume formulation are so apparently dissimilar, it is useful to show that they are equivalent, but useful in different contexts where different approximations may be made. It will be shown that the basic difference between the formulations is whether they emphasise the **radiation resistance** or the **internal losses** of the label antenna.

11.2. Power Matching Considerations

However we regard our label, we expect to have, in its series equivalent circuit, as well as a reactance, a radiation resistance R_r, a loss resistance R_c, and a load resistance R_L, and possibly a matching reactance. Considerations of maximum power transfer to the load will always require that at optimum we set

$$R_L = R_r + R_c$$

11.3. Choice of Formulations

Because labels are mostly small in relation to a wave length, we will frequently (but not always) have $R_r << R_c$. When the label becomes particularly small, we will focus attention on R_c and neglect R_r entirely. Also because labels are mostly small we will have, in their tuning and matching, quality factors which are large.

Our analysis will be performed for the magnetic field sensitive antenna consisting of a small loop. We will leave it to the reader to produce a similar analysis for the electric field sensing antenna.

11.4. Analysis of a Small Loop

In the case of a small loop of area A and self inductance L, with loss and radiation resistances as described above, excited by a magnetic field, the radiation resistance is

$$R_r = \left(\frac{\eta}{6\pi}\right)(\beta)^4 A^2$$

Assuming the loop has the gain g_d of a small dipole it has an effective area A_e of

$$A_e = g_d \frac{\lambda^2}{4\pi}$$

and neglecting losses the available source power P_a when the loop is in a field of Poynting vector S_r is

$$P_a = g_d \frac{\lambda^2}{4\pi} S_r$$

This is the power which a lossless antenna would deliver to a load $R_L = R_r$. For comparison with coupling volume theory, which we apply when R_r is negligible with respect to the losses R_c, it is appropriate, in view of the definition in Section 10.3, of coupling volume, to calculate the power delivered to the losses R_c without any external load yet having been added. This power P_c is given by

$$P_c = \left(\frac{4R_r}{R_c}\right) P_a$$

If we now substitute for R_r and P_a from equations earlier in this section and S_r from Section 10.2 and use the value 1.5 for g_d, we obtain

$$P_c = \frac{\eta^2 \beta^{\ 2} |\mathrm{H}|^2 \ A^2}{2R_c}$$

To manipulate this into a more familiar form, we replace R_c by $\omega L/Q$ and $\eta\beta$ by $\omega\mu_0$ and obtain

$$P_c = \frac{(\omega\mu_0 \ |\mathrm{H}|^2)}{2}\left(\frac{\mu_0 A^2}{L}\right)Q$$

In this relation we find the familiar factors W_v in the first bracket and the coupling volume V_c in the second bracket. Thus we can rewrite the above equation as

$$P_c = QW_v V_c$$

which is the standard form of the result from coupling volume theory for coils coupling to the magnetic field.

What we have done is to show that the effective area and coupling volume formulations of loop antenna behaviour are entirely equivalent.

11.5. Available Power From a Small Loop

When an optimum load resistor $R_L = R_c$ is added to a small loop, the power which can be delivered to that load is one quarter of that given by the equation immediately above, if we continue to interpret Q as defined by the loop losses and its inductance. If however we redefine Q to be the new and lowered Q, determined by the sum of the loop losses and the damping of the external load, then the power we can deliver to a matched load is now half that given by that equation.

11.6. Creation of an Exciting Field

We next consider issues concerned with the generation of label exciting fields. We can distinguish in the provision of exciting field a number of different situations such as:

a) a limited and enclosed volume;
b) a limited but not enclosed volume in which it is practicable to construct an antenna with dimensions of the order of the scanned space, and (c) interrogation over a distance large in relation to the dimensions of any practicable antenna.

We will consider the last of these situations below.

11.7. Interrogation at a Large Distance

By choice of frequency we may be able to place the large distance in the near field or in the far field. In the near field we are not concerned with radiation, and normally try to minimise it. We also minimise the losses and obtain a total reactive power U_t for a power P_t delivered to a field creation structure of quality factor Q_t given by

$$U_t = Q_t P_t$$

The energy density per unit volume at the label position is obtained by introducing the dispersal volume V_d, so

$$W_v = \frac{Q_t P_t}{V_d}$$

A good example to take is that of a circular coil of diameter D from which we wish to create a label exciting field at a distance R. Since we are considering here the large distance case, analysis of the field properties of such a loop leads, to a good approximation when $R >> D$, to

$$V_d = F\left(\frac{4R^2}{D}\right)^3$$

where R is the distance from the interrogator to the label, and F is a factor of the order of 2. Substituting this value into the second equation of Section 11.7 gives

$$W_v = \frac{Q_t P_t D^3}{F(2R)^6}$$

which is a result we will compare with another about to be derived. Now let us consider a different frequency for which we have the label in the far field of the transmitter antenna, which in this case will be an efficient radiator. If we have a transmitter gain g_t, the power density per unit area in the Poynting vector is

$$S_r = \frac{g_t P_t}{4\pi R^2}$$

and in view of the last equation of Section 10.2 we have

$$W_v = \frac{\beta g_t P_t}{4\pi R^2}$$

Replacement of β by $2\pi/\lambda$ gives

$$W_v = \frac{g_t P_t}{2\lambda R^2}$$

Comparing this result with that obtained earlier in this section, we see that we have lost the advantage of the high Q_t but gained the advantage that λ in the denominator of the equation just obtained is significantly smaller than the corresponding R in the seventh equation of this section. This remark applies when $R = D$; at greater values of R the situation gets rapidly worse. Furthermore, the factor 2^6 in the denominator of the seventh equation of this section tends to eliminate the benefit of any practical value of Q_t, and we also have in the equation immediately above the advantage (probably modest) of the transmitter antenna gain g_t.

Clearly the case for propagating communication becomes overwhelming when $R > D$. However, when we consider fields at the centre of a portal, it may be shown that $V_d = F D^3$ and our factor of 2^6 has disappeared. This is just showing us how relatively strong is the field at the centre of a portal in relation to the field a distance D away.

11.8. Operation under EMC Constraints

The preceding analysis has explored the efficiency of power transfer between a transmitter and a label and is effectively drawn from the point of view of minimising transmitter power. A more real constraint is provided by electromagnetic compatibility (EMC) licensing considerations that are directed not to transmitter power but equivalent isotropic radiated power at stated distances. What this results in, is a direct constraint on S_r at the label site, whether or not the interrogation distance is the same as the EMC constraint distance. Our attention then changes to the problem of receiving power from a given area power density, with a label of size limited by the application.

As long as the antenna matching is efficient we find the power transfer is enhanced by lowering the frequency so the effective area of the label, considered as a receiving antenna, is increased. Eventually, however, as the frequency is lowered too far, the radiation resistance drops below the loss resistance of the antenna, and does so sufficiently rapidly for the increase in effective area of the label antenna to be more than offset by the mismatch efficiency. This conclusion is sustained for both electric dipole and magnetic loop antennas, although the details of the loss mechanisms vary between the two cases.

As a result of this analysis we are led to conclude that: the optimum frequency for operation of an RFID system in the far field is the lowest frequency for which a reasonable match to the radiation resistance of the label antenna can be achieved, at the allowed size of label, without the label or matching element losses intruding.

12. CONCLUSIONS

The fundamental principles governing electromagnetic coupling between an interrogator and its labels in an RFID system have been outlined, and a set of concepts suitable for describing coupling in the near and far fields, using electric field, magnetic field or electromagnetic field sensitive antennas, have been defined.

Some of the properties of electric field and magnetic field antennas have been derived.

Some important theorems about optimising antenna sizes and operating frequencies subject to electromagnetic compatibility regulations have been derived. Optimisation policies which can be pursued in the near field, the mid field or the far field have been identified.

A proposal for relaxation of electromagnetic compatibility regulations in a narrow band in the HF region is investigated, and some of the interesting consequences for HF operation at substantially improved distances have been identified. These results, while attractive, are unlikely, without substantial increase in label antenna sizes, to challenge the supremacy of far field systems for long-range RFID system operation.

13. REFERENCES

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APPENDIX – USEFUL FORMULAE

In addition to the formulae quoted in the body of the paper, the following relations are often useful.

A.1. Inductance Calculations

A.1.1. Planar Circular Coil

The self inductance of a single-turn circular coil of diameter D made from wire of diameter d is, when the currents flow on the surface, given by

$$L = \frac{\mu_o D}{2}\left[\log_e\left(\frac{8D}{d}\right) - 2\right]$$

An empirical but useful formula for the self inductance of a thin wire solenoidal coil of N turns wound over a length of l on a former of diameter of $2r$ is

$$L = \frac{\mu_0 \pi r^2 N^2}{l + 0.9r}$$

A.1.2. Twin Wire Line

The self inductance L of a twin-wire line in which the conductors have diameter d and separation s is given by

$$L = \frac{\mu_o}{\pi} \operatorname{arc\,cosh}\left(\frac{s}{d}\right)$$

$$\approx \frac{\mu_o}{\pi} \log\left(\frac{2s}{d}\right) \qquad \textit{when } s \gg d$$

A.2. Axial Field of a Circular Coil

In the magnetostatic approximation, the field at a point at a distance z along the axis of a single turn circular coil of radius a carrying a current I is given by

$$H_z(0,0,z) = \frac{Ia^2}{2\left(a^2 + z^2\right)^{\frac{3}{2}}}$$

A.3. Skin Effect

Skin depth δ in a metal at an angular frequency ω is given by

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

The surface resistivity R_S per square due to skin effect is

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

and the wave impedance η at the surface is

$$\eta = (1 + j)R_s$$

A4. Radiation Resistances

A.4.1. Electric Dipole

The radiation resistance of a short electric dipole of length L, operating at a frequency for which the free space propagation constant has magnitude β, is given in Ohms by

$$R_r = 20(\beta L)^2$$

A.4.2. Magnetic Dipole

The radiation resistance of a small current loop of radius a, operating at a frequency for which the free space propagation constant has magnitude β, is given in Ohms by

$$R_r = 20\pi^2 (\beta a)^4$$

Small loops of other shapes but the same area have the same radiation resistance.