Optimised Inductively Coupled Reader Antennas for Smart HF RFID Systems

Soheyl Soodmand

Submitted for the Degree of Doctor of Philosophy from the University of Surrey

Centre for Communication Systems Research
Faculty of Engineering and Physical Sciences
University of Surrey
Guildford, Surrey GU2 7XH, UK

August 2014

@Soheyl Soodmand 2014
Summary

The Internet of things (IoT) refers to uniquely identifiable objects and their virtual representations in an Internet-like structure to be managed and inventoried by computers. Radio-frequency identification (RFID) - a prerequisite for the IoT - is an automatic way for data transaction in object identification and is used to improve automation, inventory control and checkout operations. An RFID system consists of a reader device and one or several tags. Smart reader systems are building blocks for cutting edge applications of RFID and as a subdivision of these systems, RFID smart shelf solutions are started to be implemented for large-scale item-level management where characteristics of reader antennas are critical issue.

This work focuses on designing optimised reader antennas for high frequency (HF) RFID smart shelf systems which operate based on inductive coupling between the tag and the reader antennas and have good performance in crowded environments. Firstly, an approach is presented to increase band-width of HF RFID reader antennas to improve the reception of sub-carrier frequencies. A fabricated enhanced band-width antenna at 13.56 MHz is evaluated for its capability in being used for smart shelf applications. The obtained band-width supports sub-carrier frequencies for all the HF RFID standards to be detected easier and thus leads to increased identification range. It is shown the HF RFID technology is capable of identifying the distance of tagged books based on the received magnetic field intensity.

Secondly, multi turn small self resonant coil (MT SSRC) antennas are introduced and analysed as a new model of inductively coupled reader antennas. Based on the analysis, two turn planar SSRC (TTP SSRC) antennas having similar dimension with the current HF RFID reader antennas are investigated. Fabricated TTP SSRC antenna operating at 13.56 MHz is resulted to optimised Q factor and more uniform near field pattern in comparison with the similar antennas. Also, a number of TTP SSRC antennas operating at a distinct frequency, 13.56MHz, are fabricated on different substrates and it is shown the desired Q factor and antenna dimension can be obtained based on the dielectric characteristics of the substrate.

Key words: RFID Antennas, Near-field Antennas, Internet of Things, Inductive Coupling

Email: S.soodmand@surrey.ac.uk
WWW: http://surrey.ac.uk/ccsr/people/phd_students/soheyl_soodmand/
Acknowledgements

I would like to express my special appreciation and regards to my supervisor Dr. Tim Brown. I would like to thank him for constant encouraging my research and also for his priceless advice and guidance during the entire period of my PhD research program and also preparation of this thesis. I would also like to thank my co-supervisor, Dr. Payam Barnaghi and my previous co-supervisor, Dr. Alex Gluhak for their brilliant comments and suggestions. I can not forget the great help and moral support of Professor Rahim Tafazolli for his inspiring guidance, support and encouragement during my investigation, so many Thanks.

I would also like to thank all of my friends who supported me to strive towards my goal. A special thanks to my family. Words cannot express how grateful I am to my sisters Saharnaz and Solmaz for all of the supports they have given me throughout my life. In the end, I would like express appreciation to my beloved mother Nayyereh and father Mohammad who were always my support in the moments when there was no one to answer my queries. Your prayer for me was what sustained me thus far.

This thesis is dedicated to Mohammad, Nayyeh, Saharnaz, Solmaz and also Tanin, my lovely cute niece who starts school next year. Thanks God.

Soheyl Soodmand
August 2014
Contents

1 Introduction, Motivation and Objectives of Research ................................................................. 1

1.1 On the Relation of Internet of the Things to Smart RFID Systems ..................................... 1

1.2 RFID ........................................................................................................................................ 2

1.2.1 RFID Systems Based on Frequency, Range and Coupling ............................................. 5

1.2.1.1 Inductive Coupled or Near Field RFID Systems ....................................................... 5

1.2.1.2 Long-Range or Far Field RFID Systems .................................................................... 6

1.3 Motivation for Research and Objectives .................................................................................. 7

1.4 Original Contributions to Knowledge ....................................................................................... 7

1.5 List of Publications .................................................................................................................. 8

2 Physical Principles of HF RFID (Inductive Coupling) Systems - Background Theory ................................. 9

2.1 Magnetic Field Strength $H$ .................................................................................................... 9

2.1.1 Path of Field Strength $H(x)$ in Conductor Loops ......................................................... 10

2.1.2 Optimal Antenna Diameter ............................................................................................. 13

2.2 Magnetic Flux $\Phi$ and Magnetic Flux Density $B$ ..................................................................... 13

2.3 Inductance $L$ ................................................................................................................................ 14

2.3.1 Inductance of a Conductor Loop ....................................................................................... 14

2.4 Mutual Inductance $M$ ............................................................................................................. 15

2.5 Coupling Coefficient $k_C$ ...................................................................................................... 16

2.6 Faraday’s Law .......................................................................................................................... 18

2.7 Resonance and Q Factor ......................................................................................................... 20

2.8 Practical Operation of the Transponder .................................................................................. 24

2.8.1 Power Supply to the Transponder ..................................................................................... 24

2.8.2 Voltage Regulation .......................................................................................................... 24

2.9 Interrogation Field Strength $H_{min}$ ..................................................................................... 26

2.9.1 Energy Range of Transponder Systems ......................................................................... 28

2.9.2 Interrogation Zone of Readers ......................................................................................... 30
2.10 Equivalent Circuit Diagram of Reader .......................................................... 32
2.11 Supply Reader Antennas via Coaxial Cable .................................................. 34
2.11.1 The Influence of the Q Factor ...................................................................... 36
2.12 A Comparison - Inductively Coupled RFID Antennas and Long Range RFID Antennas ............................................................................................................. 36
2.12.1 Transition from Near-Field to Far-Field in Conductor Loops ..................... 36
2.13 Placing Antennas In Metal Environments ..................................................... 38
2.13.1 Waveguide Materials ................................................................................... 40
2.14 Summary .......................................................................................................... 40

3 HF RFID Reader Antennas for Smart Shelf Applications - Prior Art 41
3.1 Summary of Challenges on HF RFID Reader Antennas ................................. 42
3.2 Loop Antennas and Metal-Backed Loop Antennas with HF RFID Smart Shelf Application .................................................................................................................. 43
3.2.1 Impedance Matching, Field Response and Resonant Frequency ................. 44
3.2.2 Magnetic-Field Intensity, Field Distribution and Detection Range ............. 45
3.2.3 Application to HF-RFID Smart-Shelf System ............................................. 45
3.3 Summary .......................................................................................................... 50

4 Increased Band-width HF RFID Reader Antennas and its Evaluation For Smart Shelf Applications - Novel Work Undertaken 51
4.1 Approach to Increase Band-width of HF RFID Antennas .............................. 52
4.2 Prototype HF RFID Antenna Operating at 13.56 MHz for Smart Shelf .......... 54
4.3 Simulation and Measurement Results ............................................................... 55
4.4 Summary .......................................................................................................... 58

5 Multi Turn Small Self Resonant Coil (MT SSRC) at HF Band - Novel Work Undertaken 59
5.1 From the Old Model to a New Model of HF RFID Antennas ......................... 60
5.2 Mutli-Turn Small Coils for HF RFID Application ......................................... 61
5.2.1 Magnetic Field ............................................................................................... 62
5.2.2 Q Factor ........................................................................................................ 62
5.2.3 Self Resonance Frequency (SRF) ................................................................. 63
5.3 Multi-Turn Small Self Resonant Coil (MT SSRC) at HF Band ...................... 64
5.3.1 MT SSRC with Non-Insulated Wires .......................................................... 64
5.3.2 MT SSRC with Insulated Wires ................................................................. 66
5.4 Summary .......................................................................................................... 66
6 Two Turn Planar SSRC (TTP SSRC) Antennas for HF RFID Applications - Novel Work Undertaken

6.1 Two Turn Planar SSRC (TTP SSRC) Antennas at HF Band ................................................. 69
6.2 TTP SSRC Antenna Operating at 13.56 MHz .................................................................. 69
6.3 Adjusting Q Factor and Dimension of TTP SSRC Antenna by Dielectric Characteristics of the Substrate .................................................................................. 74
6.4 Summary .......................................................................................................................... 75

7 Conclusions and Future Works ......................................................................................... 77

References ............................................................................................................................ 79
List of Constants

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Constant</th>
<th>Unit and Value *</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\varepsilon_0$</td>
<td>Electric Field Constant</td>
<td>$8.85 \times 10^{-12}$ As/Vm</td>
</tr>
<tr>
<td>$\mu_0$</td>
<td>Magnetic Field Constant</td>
<td>$1.257 \times 10^{-6}$ Vs/Am</td>
</tr>
<tr>
<td>$c$</td>
<td>Speed of Light</td>
<td>$3 \times 10^8$ m/s</td>
</tr>
</tbody>
</table>

* A = Ampere, F = Farad, H = Henry, Hz = Hertz, m = metre, rad = radian, s = second, V = Volt, W = Watts, Ω = Ohm
## List of Variables

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Variable</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\varepsilon$</td>
<td>Permittivity</td>
<td>$\text{A s/Vm}$</td>
</tr>
<tr>
<td>$\varepsilon_r$</td>
<td>Relative Permittivity</td>
<td>-</td>
</tr>
<tr>
<td>$\mu$</td>
<td>Permeability</td>
<td>$\text{V s/Am}$</td>
</tr>
<tr>
<td>$\mu_r$</td>
<td>Relative Permeability</td>
<td>-</td>
</tr>
<tr>
<td>$\Phi$</td>
<td>Magnetic Flux</td>
<td>$\text{V s}$</td>
</tr>
<tr>
<td>$\psi$</td>
<td>Total Magnetic Flux</td>
<td>$\text{V s}$</td>
</tr>
<tr>
<td>$\omega=2\pi f$</td>
<td>Angular Frequency</td>
<td>rad/s</td>
</tr>
<tr>
<td>$\beta=2\pi/\lambda$</td>
<td>Phase Constant</td>
<td>rad/m</td>
</tr>
<tr>
<td>$\lambda$</td>
<td>Wavelength</td>
<td>m</td>
</tr>
<tr>
<td>$\tan \delta$</td>
<td>Magnetic Loss Tangent</td>
<td>-</td>
</tr>
<tr>
<td>$A$</td>
<td>Area</td>
<td>$\text{m}^2$</td>
</tr>
<tr>
<td>$B$</td>
<td>Magnetic Flux Density</td>
<td>$\text{V s/m}^2$</td>
</tr>
<tr>
<td>$C$</td>
<td>Capacitance</td>
<td>$\text{F}$</td>
</tr>
<tr>
<td>$D$</td>
<td>Distance Between Centers of Tag and Reader Antennas</td>
<td>m</td>
</tr>
<tr>
<td>$d$</td>
<td>Distance Between Centres of Two Adjacent Wires of Coil</td>
<td>m</td>
</tr>
<tr>
<td>$E$</td>
<td>Electric Field Strength</td>
<td>$\text{V/m}$</td>
</tr>
<tr>
<td>$f$</td>
<td>Frequency</td>
<td>Hz</td>
</tr>
<tr>
<td>$H$</td>
<td>Magnetic Field Strength</td>
<td>$\text{A/m}$</td>
</tr>
<tr>
<td>$h$</td>
<td>Dielectric Thickness</td>
<td>m</td>
</tr>
<tr>
<td>$I$</td>
<td>Electric Current</td>
<td>A</td>
</tr>
</tbody>
</table>
\[ k = \frac{\text{Length of Coil}}{\lambda} \]

- Electrical Length Factor

- Coupling Coefficient

\[ L \]

- Inductance

\[ M \]

- Mutual Inductance

\[ N \]

- Number of Coil Windings

\[ R \]

- Radius of Coil/Loop

\[ R_{\text{Something}} \]

- Electrical Resistance

\[ S \]

- Power Density

\[ S = j \omega \]

- Complex Variable in Laplace Domain @ Section 4.1

\[ s \]

- Dielectrics Thickness

\[ t \]

- Insulation Thickness

\[ U \]

- Electric Voltage

\[ Z \]

- Impedance
<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Full Form</th>
</tr>
</thead>
<tbody>
<tr>
<td>ETSI</td>
<td>European Telecommunications Standards Institute</td>
</tr>
<tr>
<td>HF</td>
<td>High Frequency</td>
</tr>
<tr>
<td>IEC</td>
<td>International Electrotechnical Commission</td>
</tr>
<tr>
<td>IEEE</td>
<td>Institute of Electrical and Electronics Engineers</td>
</tr>
<tr>
<td>IoT</td>
<td>Internet of Things</td>
</tr>
<tr>
<td>IPv6</td>
<td>Internet protocol version 6</td>
</tr>
<tr>
<td>ISO</td>
<td>International Organization for Standardization</td>
</tr>
<tr>
<td>ISM</td>
<td>Industrial, Scientific and Medical</td>
</tr>
<tr>
<td>ITU</td>
<td>International Telecommunication Union</td>
</tr>
<tr>
<td>LF</td>
<td>Low Frequency</td>
</tr>
<tr>
<td>MMID</td>
<td>Millimetre Wave Identification</td>
</tr>
<tr>
<td>MTP SSRC</td>
<td>Multi Turn Planar Small Self Resonant Coil</td>
</tr>
<tr>
<td>MT SSRC</td>
<td>Multi Turn Small Self Resonant Coil</td>
</tr>
<tr>
<td>NFC</td>
<td>Near Field Communication</td>
</tr>
<tr>
<td>PCB</td>
<td>Printed Circuit Board</td>
</tr>
<tr>
<td>PEEC</td>
<td>Partial Element Equivalent Circuit</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
<tr>
<td>RFID</td>
<td>Radio Frequency Identification</td>
</tr>
<tr>
<td>SMA</td>
<td>Sub Miniature version A connector</td>
</tr>
<tr>
<td>SRF</td>
<td>Self Resonance Frequency</td>
</tr>
<tr>
<td>SSRC</td>
<td>Small Self Resonant Coil</td>
</tr>
<tr>
<td>Acronym</td>
<td>Description</td>
</tr>
<tr>
<td>---------</td>
<td>------------------------------</td>
</tr>
<tr>
<td>TTP SSRC</td>
<td>Two Turn Planar Small Self Resonant Coil</td>
</tr>
<tr>
<td>UHF</td>
<td>Ultra high Frequency</td>
</tr>
<tr>
<td>USID</td>
<td>Ultrasound Identification</td>
</tr>
<tr>
<td>UWB</td>
<td>Ultra Wideband</td>
</tr>
</tbody>
</table>
Chapter 1

1 Introduction, Motivation and Objectives of Research

1.1 On the Relation of Internet of Things to Smart RFID Systems

One major step in the Internet development is to progressively evolve from a network of interconnected computers to a network of interconnected objects, from books to cars, from electrical appliances to food, and thus create an ‘Internet of things’ (IoT) [1]. The term Internet of Things was first used by Kevin Ashton in 1999. IoT refers to uniquely identifiable objects (things) connected through a wireless network and their virtual representations in an Internet-like structure [2]. If all objects of daily life were equipped with radio tags, they can be identified and inventoried by computers [1, 3, 4]. Examples of IoT application rely on several building blocks such as radio-frequency identification (RFID), near field communication (NFC), 2D bar codes, wireless sensor/actuators, Internet protocol version 6 (Ipv6), ultrawide-band or 3/4G, which are all expected to play an important role in future deployments [1]. Most of today’s foreseeable IoT devices are expected to be in the ‘radio frequency’ group (i.e. >100 kHz) and operate with very low power. Table 1.1 presents the mapping between the frequency spectrum and the existing standards and protocols that are used for implementing wireless identifiable devices for IoT applications [5].

RFID technology - a prerequisite for the IoT - is an automatic way for data transaction in object identification without any human intervention or error and has been extensively used to improve automation, inventory control, and checkout operations in stores, factories, and warehouses. An RFID system consists of a reader device and one or several transponders/tags.
The RFID tags have more advantages in comparison with the bar code: RFID tags can be detected when they are soiled or blocked by obstacles and also carry more information than a typical bar code. When considering RFID systems to be used for IoT application, each physical object is accompanied by a rich, globally accessible virtual object that contains both current and historical information on that object’s physical properties, origin, ownership, and sensory context (for example, the temperature at which a milk carton is being stored). When ubiquitous and available in real time, this information can dramatically streamline how we manufacture, distribute, manage, and recycle our goods. This “real-life” context can unlock the door to various business, environmental, personal, and social contexts hitherto inaccessible to Internet applications [6, 7]. Companies would not run out of stock or waste products, as involved parties would know which products are required and consumed [7]. Misplaced and stolen items would be easily tracked and located, as would the people who use them.

Some works have been carried out towards introducing a model for the Internet of things as a system of RFID smart objects - that is, is autonomous physical/digital objects augmented with sensing, processing, and network capabilities. Smart objects carry chunks of application logic that let them make sense of their local situation and interact with human users. They sense, log, and interpret what’s occurring within themselves and the world, act on their own, intercommunicate with each other, and exchange information with people. [8]. As a subdivision of smart RFID reader systems, RFID smart shelf solutions has been started to be implemented by leading retailers for their large-scale item-level management [9]. A summary of challenges on smart shelf systems are mentioned in Chapter 3 of this thesis and some solutions in order to overcome the problems or improve the system efficiency is presented in Chapters 4-6.

1.2 RFID

A wireless system using radio frequency identification (RFID) consists of a reader/interrogator device and one or several transponders/tags. RFID is a technology that communicates through radio waves to transfer data between the reader and the tag attached to an object for the purpose of identification and tracking (Figure 1.1). The tag contains electronically stored information which can be read from up to several metres away, which always function as sleeping markers regardless of the type of RFID system or application. The reader initialises the communication by sending a signal, which is replied to in different ways by the tags. Unlike a bar-code the tag does not need to be within line of sight of the reader and may be embedded in the tracked object.

The reader normally contains an radio frequency (RF) transceiver, a control unit and a coupling element to the tag. Also, many readers are fitted with an additional interface like RS 232 or RS 485 to enable them to forward the data received to another system like PC or robot control system. The transponders/tags, which represent the actual data-carrying device of an RFID
system, normally consist of a coupling element and an electronic microchip, (Figure1.2). Simple tags like the ones used in some anti-theft systems consist of a diode-connected antenna, which reflects harmonics of the transmitted reader signal frequency. In these systems the reader transmits continuously and receives harmonics at the same time. When it detects a harmonic of the signal it sets off the alarm. Other, still very simple tags receive the reader signal and then reply with a data signal containing its identification number or other data stored in the tag. The tags mentioned above are called read tags since they contain information that can be read only, regardless if the information is a block of data, an identification number or simply a reflected signal telling the reader that a tag is within reading range. More advanced tags can also be written to by the reader. These tags are referred to as read/write tags. Examples of simple read/write tags are the ones used in the anti-theft system at libraries which can be activated/deactivated when the book has been registered by the librarian for lending. Some read/write tags that need to process large amounts of data contain a microprocessor. A disadvantage is that such a tag is quite energy consuming.

Table 1.1 Summary of standards predicted for IoT applications.

<table>
<thead>
<tr>
<th>Range</th>
<th>Frequency Range</th>
<th>Wavelength</th>
<th>Frequency</th>
<th>Standard</th>
</tr>
</thead>
<tbody>
<tr>
<td>LF</td>
<td>Low Frequency</td>
<td>30kHz to 300kHz</td>
<td>1km to 1km</td>
<td>39-50kHz, 125/134kHz/156/45kHz, USID ISO/IEC 88000-2, IEEE P9002.1, RuBec ETSI EN 300 330</td>
</tr>
<tr>
<td>MF</td>
<td>Medium Frequency</td>
<td>300kHz to 3MHz</td>
<td>1km to 100m</td>
<td>ISO/IEC 60880-3, ISO/IEC 15993, ISO/IEC 4443, ISO/IEC 18092/NFC ISO/IEC 10536 EPCglobal EPC HF G2 ETSI EN 300 330</td>
</tr>
<tr>
<td>HF</td>
<td>High Frequency</td>
<td>3MHz to 30MHz</td>
<td>100m to 10m</td>
<td>27MHz ISO/IEC 88000-7, ISO/IEC 88000-6 Types A, B, C, D EPCglobal EPC UHF C1G2 IEEE 802.11 ISO/IEC 88000-4 IEEE 802.15 WPAN IEEE 802.15 WPAN Low Rate IEEE 802.15 RFID ETSI EN 300 220 ETSI EN 300 440 ETSI EN 302 208</td>
</tr>
<tr>
<td>VHF</td>
<td>Very High Frequency</td>
<td>30MHz to 300MHz</td>
<td>1m to 1m</td>
<td>125MHz ISO/IEC 88000-7, ISO/IEC 88000-6 Types A, B, C, D EPCglobal EPC UHF C1G2 IEEE 802.11 ISO/IEC 88000-4 IEEE 802.15 WPAN IEEE 802.15 WPAN Low Rate IEEE 802.15 RFID ETSI EN 300 220 ETSI EN 300 440 ETSI EN 302 208</td>
</tr>
<tr>
<td>UHF</td>
<td>Ultra High Frequency</td>
<td>300MHz to 3GHz</td>
<td>1m to 10cm</td>
<td>433MHz ISO/IEC 88000-7, ISO/IEC 88000-6 Types A, B, C, D EPCglobal EPC UHF C1G2 IEEE 802.11 ISO/IEC 88000-4 IEEE 802.15 WPAN IEEE 802.15 WPAN Low Rate IEEE 802.15 RFID ETSI EN 300 220 ETSI EN 300 440 ETSI EN 302 208</td>
</tr>
<tr>
<td>SHF</td>
<td>Super High Frequency</td>
<td>3GHz to 30GHz</td>
<td>10cm to 1cm</td>
<td>3-10.6GHz ISO/IEC 88000-7, ISO/IEC 88000-6 Types A, B, C, D EPCglobal EPC UHF C1G2 IEEE 802.11 ISO/IEC 88000-4 IEEE 802.15 WPAN IEEE 802.15 WPAN Low Rate IEEE 802.15 RFID ETSI EN 300 220 ETSI EN 300 440 ETSI EN 302 208</td>
</tr>
<tr>
<td>EHF</td>
<td>Extremely High Frequency</td>
<td>30GHz to 300GHz</td>
<td>1mm to 1mm</td>
<td>24-125GHz MMID ETSI EN 300 440</td>
</tr>
</tbody>
</table>
Near field RFID technology uses induction. When a current flows through a coil, a magnetic field is generated around it. If another conductor or even better, another coil is placed within this magnetic field a current is induced in it. The reader antenna works as a coil providing a magnetic field, which induces a current in the antenna coil in the tag. This is where RFID differs from classic radio transceivers.

An important distinction criterion of different RFID systems is how the energy supply of the tag works. Most RFID tags are passive since they have no power supply/battery of their own and are powered by the radio waves used to read them. They use the induced current from the magnetic or electromagnetic field generated by the reader to process the information and send a reply. In order to transmit data from the tag to the reader, the field of the reader can be modulated or the tag can intermediately store, for a short time, energy from the field of the reader. This means the energy emitted by the reader is used for data transmission both from the reader to the tag and back to the reader. If the tag is located outside the reader’s range, the tag has no power supply at all and, therefore, will not be able to send signals. Other RFID tags use a local power source. They are active tags that have their own energy supply, e.g. in form of a battery or a solar cell.

RFID tags are used in many industries. An RFID attached to an automobile during production can be used to track its progress through the assembly line. Pharmaceuticals can be tracked through warehouses. Livestock and pets may have tags injected, allowing positive identification of the animal. RFID identity cards can give employees access to locked areas of a building, and RF tags mounted in automobiles can be used to bill motorists for access to toll roads or parking. Since RFID tags can be attached to clothing, possessions, or even implanted within people, the possibility of reading personally-linked information without consent has raised privacy concerns [10].

![Diagram of RFID system components](image)

**Figure 1.1** The reader and transponder are the main components of the RFID system.
1.2.1 RFID systems based on Frequency, Range and Coupling

The most important differentiation criteria for RFID systems are the operating frequency of the reader, the physical coupling method and the range of the system. RFID systems operate at widely differing frequencies, ranging from 135 kHz long-wave to 5.8 GHz in the microwave range. Electric, magnetic and electromagnetic fields are used for the physical coupling. Finally, the achievable range of the system varies from a few millimetres to above 15 m. The different distances the reader and the tags can communicate on are divided into two areas and are given below [10]. RFID systems with a very small range - typically in the region of up to 1 cm - which are known as close coupling systems [10] are not mentioned here.

1.2.1.1 Inductive Coupled or Near Field RFID Systems

Systems with write and read ranges of up to 1m are known by the collective term of Inductive (remote) coupling systems. Almost all remote coupled systems are based upon an inductive (magnetic) coupling between reader and tag. These systems are therefore also known as inductive radio systems. In addition there are also a few systems with capacitive (electric) coupling. At least 90% of all RFID systems currently sold are inductively coupled systems. For this reason there is now an enormous number of such systems on the market. There is also a series of standards that specify the technical parameters of tag and reader for various standard applications e.g animal identification, industrial automation and etc. These also include proximity coupling (ISO 14443, contact-less smart cards) and vicinity coupling systems (ISO 15693, smart label and contact-less smart cards). The following preferences at Table 1.2 exist for the various frequency ranges of inductively coupled systems.

Figure 1.2 Basic layout of the tag i.e. the RFID data-carrying device. Left, inductively coupled tag with antenna coil; right, microwave tag with dipole antenna.
Table 1.2 Preferences exist for the various frequency ranges of inductively coupled systems.

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Application</th>
<th>Specification</th>
</tr>
</thead>
</table>
| <135 kHz  | Large ranges & low-cost transponders | • High level of power available to the transponder.  
• The transponder (tag) has a low power consumption due to its lower clock frequency.  
• Miniaturised tag formats are possible (animal ID) due to the use of ferrite coils in the tag.  
• Low absorption rate or high penetration depth in non-metallic materials and water (the high penetration depth is exploited in animal identification by the use of the bolus, a transponder placed in rumen). |
| 6.78 MHz  | Low-cost and medium-speed transponders | • Worldwide ISM frequency according to ITU frequency plan; however, this is not used in some countries (i.e. licence may not be used worldwide).  
• Available power is a little greater than that for 13.56 MHz.  
• Only half the clock frequency of that for 13.56 MHz. |
| 13.56 MHz | High-speed/high-end and medium-speed/low-end applications | • Available worldwide as an ISM frequency.  
• Fast data transmission (typically between 106 kbit/s and 848 kbit/s).  
• High clock frequency, so cryptological functions or a microprocessor can be realised.  
• Parallel capacitors for transponder coil (resonance matching) can be realised on-chip. |
| 27.125 MHz | Only for special applications (e.g. Eurobalise) | • Not a worldwide ISM frequency.  
• Large band-width, thus fast data transmission (typically 424 kbit/s).  
• High clock frequency, thus cryptological functions or a microprocessor can be realised.  
• Parallel capacitors for transponder coil (resonance matching) can be realised on-chip.  
• Available power somewhat lower than for 13.56 MHz. |

1.2.1.2 Long-Range or Far Field RFID Systems

RFID systems with ranges significantly above 1 m are known as long-range systems. All long range systems operate using electromagnetic waves in the UHF and microwave range. The vast majority of such systems are also known as backscatter systems due to their physical operating principle. In addition, there are also long-range systems using surface acoustic wave tags in the microwave range. These systems operate at the UHF frequencies of 868 MHz (Europe) and 915 MHz (USA) and at the microwave frequencies of 2.5 GHz and 5.8 GHz. Typical ranges of 3 m can now be achieved using passive (battery-free) backscatter tags, while ranges of 15 m and
above can even be achieved using active (battery-supported) backscatter tags. The battery of an active tag, however, never provides the power for data transmission between tag and reader, but serves exclusively to supply the microchip and for the retention of stored data. The power of the electromagnetic field received from the reader is the only power used for the data transmission between tag and reader.

In order to avoid reference to a possibly erroneous range value, only the terms inductively or capacitively coupled system and microwave system or backscatter system are used for classification. RFID systems that use frequencies between approximately 100 kHz and 30 MHz operate using inductive coupling. By contrast, microwave systems at 2.45 GHz and 5.8 GHz frequencies are coupled using electromagnetic waves. Microwave systems have a significantly higher range than inductive systems, typically 2–15 m.

1.3 Motivation for Research and Objectives

The concept of the RFID as a prerequisite for the IoT applications was briefly described in Section 1.1 and also the RFID smart shelf was mentioned as one of the major subdivisions of smart RFID reader systems. In the past two years, renewed interest and rapid progress has been made in the application of high frequency (HF) RFID or inductive methods for the IoT applications [11]. Most the previous works address the RFID applications on the performance of ultra high frequency (UHF) band and yet little have been reported that are relevant to HF RFID applications [10]. Inductive coupled or HF RFID technology, which has operation range of around metre, has good performance in a crowded environment when compared with UHF RFID because the magnetic field is not affected by most of the surrounding dielectric materials $\mu_r \approx 1$[12]. As the characteristics of the reader antenna are critical issue for the successful implementation of a HF RFID smart shelf system, this work is focused on designing optimised reader antennas for such systems.

1.4 Original Contributions to Knowledge

Contribution 1: In the first objective of this research, a theoretical approach is presented to increase the band-width of HF RFID reader antennas to improve the reception of the sub-carrier RFID response. A HF RFID antenna with six times more band-width than similar antennas having the same dimension and the same operating frequency centered at 13.56 MHz is fabricated and designed and also is evaluated by measurement for its capability in being used for smart shelf applications. It is shown that the increased band-width supports sub-carrier frequencies for the all the existing HF RFID standards (ISO/IEC 15693 and ISO/IEC14443) to be detected more easily and thus leads to an increased range of identification. Also, it is shown that in the presence of books as a sample material, the HF RFID technology is capable of identifying the distance of a tag antenna (Position of book) based on the received magnetic field intensity ($H$-Field).
Contribution 2: In the second objective, multi turn small self resonant coil (MT SSRC) antennas are introduced and mathematically analysed as a new model of inductively coupled (HF RFID) reader antennas. Based on the analysis, compact two turn planar SSRC (TTP SSRC) antennas having similar dimensions with the current HF RFID reader antennas are investigated. As a sample, a TTP SSRC antenna operating at 13.56 MHz is fabricated and optimised in $Q$ factor and near field pattern uniformity in comparison with similar HF RFID antennas. In front of a uniform near field pattern the tag could be placed anywhere at the surface of the object without any difference in the detected $H$-field. Also, in front of a uniform near field pattern, narrower aligned objects can be identified as there would be virtual narrow levels of uniform $H$-field in front of the reader antenna. Furthermore, a number of TTP SSRC antennas operating at a distinct frequency, 13.56 MHz, are fabricated on different substrates and it is shown that the desired $Q$ factor and the antenna dimension (Smart shelf dimension) can be obtained based on the dielectric characteristics of the antenna substrate.

1.5 List of Publications


Chapter 2

2 Physical Principles of HF RFID (Inductive Coupling) Systems - Background Theory

In the literature, physical principles of RFID systems are investigated in two areas as discussed in the previous chapter. Firstly, RFID systems that use frequencies between approximately 100 kHz and 30MHz operate using inductive coupling. Therefore, under of the procedures of power and data transfer requires a thorough grounding in the physical principles of magnetic phenomena. Secondly, electromagnetic fields - radio waves in the classic sense - are used in RFID systems that operate at above 30 MHz. The propagation of waves in the far field and the principles of radar technology should be investigated to aid under of these systems.

Because this work is focused on HF RFID antennas, only physical principles of inductive coupling RFID systems are investigated in this chapter. At first physical characteristics of inductive coupling are introduced then basic laws as well as practical operation of tag and reader are investigated here. Electromagnetic fields of RFID systems operating above 30 MHz (Far field RFID) and also capacitive data transmission in close-coupling systems are investigated in detail in [10], which are outside the scope of this work. However, in order to reach good under of inductively coupled antennas a brief comparison between antennas used in inductively coupled RFID systems and antennas used in long range RFID systems (Far Field) is presented in this chapter.

2.1 Magnetic Field Strength $H$

Every flow of current in wires is associated with a magnetic field (Figure 2.1). The magnitude of the magnetic field is described by the magnetic field strength $H$ regardless of the material
properties of the space. The contour integral of magnetic field strength along a closed curve is equal to the sum of the current strengths of the currents within it [10]:

\[ \sum I = \oint \vec{H} \cdot d\vec{s} \]  

(2.1)

According to the (2.1) formula, in a straight conductor the field strength \( H \) along a circular flux line at a distance \( r \) is constant as [10]:

\[ H = \frac{I}{2\pi r} \]  

(2.2)

### 2.1.1 Path of Field Strength \( H(x) \) in Conductor Loops

So-called short cylindrical coils or conductor loops are used as magnetic antennas to generate the magnetic alternating field in the write/read devices of inductively coupled RFID systems. If the measuring point is moved away from the centre of the coil along the coil axis (\( x \) axis), then the strength of the field \( H \) will decrease as the distance \( x \) is increased. A more in-depth investigation shows that the field strength in relation to the radius (or area) of the coil remains constant up to a certain distance and then falls rapidly (Figure 2.3). In free space, the decay of field strength is approximately 60 dB per decade in the near-field of the coil, and flattens out to 20 dB per decade in the far-field of the electromagnetic wave that is generated (a more precise explanation of these effects can be found in [10]).

The following equation can be used to calculate the path of field strength along the \( x \) axis of a round coil (conductor loop) similar to those employed in the transmitter antennas of inductively coupled RFID systems [13]:

\[ H = \frac{I \cdot N \cdot R^2}{2\sqrt{(R^2 + x^2)^3}} \]  

(2.3)

where \( N \) is the number of windings, \( R \) is the circle radius and \( x \) is the distance from the centre of the coil in the \( x \) direction. The following boundary condition applies to this equation: \( d \ll R \) and \( x < \lambda/2\pi \) (the transition into the electromagnetic far field begins at a distance \( >2\pi \) and not investigated here). At the centre of the antenna, the formula can be simplified to [14]:

\[ H = \frac{I \cdot N}{2R} \]  

(2.4)

The field strength path of a rectangular conductor loop, which is often a more potential antenna solution for smart shelf, with edge length \( a \times b \) at a distance of \( x \) is presented in the following equation. This format is often used as a transmitter antenna [10]:

\[ H = \frac{N \cdot I \cdot ab}{4\pi \sqrt{\left(\frac{a}{2}\right)^2 + \left(\frac{b}{2}\right)^2 + x^2}} \cdot \left( \frac{1}{\left(\frac{a}{2}\right)^2 + x^2} + \frac{1}{\left(\frac{b}{2}\right)^2 + x^2} \right) \]  

(2.5)
Figure 2.3 shows the calculated field strength path $H(x)$ for three different antennas at a distance 1mm–10m. The number of windings and the antenna current are constant in each case; the antennas differ only in edge length $a=b$. The calculation is based upon the following values: $H_1: a=b=110$ cm, $H_2: a=b=15$ cm, $H_3: a=b=2$ cm. The calculation results confirm that the increase in field strength flattens out at short distances ($x < a=b$) from the antenna coil. Interestingly, the smallest of the three antennas exhibits a significantly higher field strength at the centre of the antenna (distance = 0), but at greater distances ($x > a=b$) the largest of the three antennas generates a significantly higher field strength. Similar results are obtained for short circular coils having similar dimensions with the rectangular ones [10]. It is vital that this effect is taken into account in the design of antennas for inductively coupled RFID systems.

![Figure 2.1](image1.png)  
**Figure 2.1** Lines of magnetic flux around a current-carrying conductor and a current-carrying cylindrical coil.

![Figure 2.2](image2.png)  
**Figure 2.2** The path of the lines of magnetic flux around a short cylindrical coil, or conductor loop, similar to those employed in the transmitter antennas of inductively coupled RFID systems.
Figure 2.3 Path of magnetic field strength $H$ in the near field of short rectangular coils, or conductor coils, as the distance in the $x$ direction is increased [10].

Figure 2.4 Field strength $H$ of a transmission antenna given a constant distance $x$ and variable radius $R$, where $I = 1$ A and $N = 1$. 
2.1.2 Optimal Antenna Diameter

If the radius \( R \) of the transmitter antenna is varied under the simplifying assumption of constant coil current \( I \) in the transmitter antenna, then field strength \( H \) is found to be at its highest at a certain ratio of distance \( x \) to antenna radius \( R \). This means that for every read range of an RFID system there is an optimal antenna radius \( R \). This is quickly illustrated by a glance at Figure 2.3: if the selected antenna radius is too great, the field strength is too low, even at a distance \( x = 0 \) from the transmission antenna. If, on the other hand, the selected antenna radius is too small, then we find ourselves within the range in which the field strength falls in proportion to \( x^3 \).

Figure 2.4 shows the graph of field strength \( H \) as the coil radius \( R \) is varied. The optimal coil radius for different read ranges is always the maximum point of the graph \( H(R) \). To find the mathematical relationship between the maximum field strength \( H \) and the coil radius \( R \) we must first find the inflection point of the function \( H(R) \), given by Equation 2.3 [15]. To do this we find the first derivative \( H'(R) \) by differentiating \( H(R) \) with respect to \( R \):

\[
H'(R) = \frac{d}{dR} H(R) = \frac{2 \cdot I \cdot N \cdot R}{(R^2 + x^2)^3} - \frac{3 \cdot I \cdot N \cdot R^3}{(R^2 + x^2)^3} \quad (2.6)
\]

The maximum value of \( H(R) \) is found from the zero points of \( dH/dR \):

\[
R_1 = x \cdot \sqrt{2}; \quad R_2 = -x \cdot \sqrt{2} \quad (2.7)
\]

The optimal radius of a transmission antenna is thus twice the maximum desired read range. The second zero point is negative merely because the magnetic field \( H \) of a conductor loop propagates in both directions of the \( x \) axis, Figure 2.2. An accurate assessment of a system’s maximum read range requires knowledge of the interrogation field strength \( H_{\text{min}} \) of the transponder in question (Section 2.9).

2.2 Magnetic Flux \( \Phi \) and Magnetic Flux Density \( B \)

The magnetic field of a (cylindrical) coil will exert a force on a magnetic needle. If a soft iron core is inserted into a (cylindrical) coil - all other things remaining equal - then the force acting on the magnetic needle will increase. The product \( I \times N \) (Section 2.1) remains constant and therefore so does field strength. However, the flux density - the total number of flux lines - which is related to the force generated [13], has increased.

The total number of lines of magnetic flux that pass through the inside of a cylindrical coil, for example, are denoted by magnetic flux \( \Phi \). As shown in Figure 2.5, Magnetic flux density \( B \) is a further variable related to area \( A \) (this variable is often referred to as ‘magnetic inductance \( B' \) in the literature) [16]. Magnetic flux is expressed as:

\[
\Phi = B \cdot A \quad (2.8)
\]

13
The material relationship between flux density $B$ and field strength $H$ (Figure 2.5) is expressed by the material equation [16]:

$$B = \mu_0 \mu_r H = \mu H$$  \hspace{1cm} (2.9)

$\mu_0$ is the magnetic field constant ($\mu_0 = 4\pi \times 10^{-6} \text{Vs/A m}$) and describes the permeability (magnetic conductivity) of a vacuum. The variable $\mu_r$ is called relative permeability and indicates how much greater than or less than $\mu_0$ the permeability of a material is.

### 2.3 Inductance $L$

A magnetic field, and thus a magnetic flux $\Phi$, will be generated around a conductor of any shape. This will be particularly intense if the conductor is in the form of a loop (coil). Normally, there is not one conduction loop, but $N$ loops of the same area $A$, through which the same current $I$ flows. Each of the conduction loops contributes the same proportion $\Phi$ to the total flux $\psi$ [13]:

$$\Psi = \sum N \Phi_N = N \cdot \Phi = N \cdot \mu \cdot H \cdot A$$  \hspace{1cm} (2.10)

The ratio of the interlinked flux $\psi$ that arises in an area enclosed by current $I$, to the current in the conductor that encloses it (conductor loop) is denoted by inductance $L$ (Figure 2.6) [13]:

$$L = \frac{\Psi}{I} = \frac{N \cdot \Phi}{I} = \frac{N \cdot \mu \cdot H \cdot A}{I}$$  \hspace{1cm} (2.11)

Inductance is one of the characteristic variables of conductor loops (coils). The inductance of a conductor loop (coil) depends totally upon the material properties (permeability) of the space that the flux flows through and the geometry of the layout.

#### 2.3.1 Inductance of a Conductor Loop

If we assume that the diameter $d$ of the wire used is very small compared with the diameter $D$ of the conductor coil ($d/D < 0.0001$) a very simple approximation can be used [10]:

$$L = N^2 \mu_0 R \cdot \ln \left( \frac{2R}{d} \right)$$  \hspace{1cm} (2.12)

where $R$ is the radius of the conductor loop and $d$ is the diameter of the wire used.

![Figure 2.5](image) Relationship between magnetic flux $\Phi$ and flux density $B$. 

14
2.4 Mutual Inductance $M$

If a second conductor loop 2 (area $A_2$) is located in the vicinity of conductor loop 1 (area $A_1$), through which a current is flowing, then this will be subject to a proportion of the total magnetic flux flowing through $A_1$. The two circuits are connected together by this partial flux or coupling flux (Figure 2.7). The magnitude of the coupling flux $\Psi_{21}$ depends upon the geometric dimensions of both conductor loops, the position of the conductor loops in relation to one another, and the magnetic properties of the medium (e.g. permeability) in the layout.

Similarly to the definition of the (self) inductance $L$ of a conductor loop, the mutual inductance $M_{21}$ of conductor loop 2 in relation to conductor loop 1 is defined as the ratio of the partial flux $\Psi_{21}$ enclosed by conductor loop 2, to the current $I_1$ in conductor loop 1 [13]:

$$M_{21} = \frac{\Psi_{21}(I_1)}{I_1} = \oint_{A_2} \frac{B_2(I_1)}{I_1} \cdot dA_2$$

Similarly, there is also a mutual inductance $M_{12}$. Here, current $I_2$ flows through the conductor loop 2, thereby determining the coupling flux $\Psi_{12}$ in loop 1. The following relationship applies:

$$M = M_{12} = M_{21}$$

Mutual inductance describes the coupling of two circuits via the medium of a magnetic field. Mutual inductance is always present between two electrical circuits. Its dimension and unit are the same as for inductance.

The coupling of two electrical circuits via the magnetic field is the physical principle upon which inductively coupled RFID systems are based. Figure 2.8 shows a calculation of the mutual inductance between a transponder antenna and three different reader antennas, which differ only in diameter. The calculation is based upon the following values: $M_1$: $R = 55$ cm, $M_2$: $R = 7.5$ cm, $M_3$: $R = 1$ cm, transponder: $R = 3.5$ cm, $N = 1$ for all reader antennas.
The graph of mutual inductance shows a strong similarity to the graph of magnetic field strength $H$ along the $x$ axis. Assuming a homogeneous magnetic field, the mutual inductance $M_{12}$ between two coils can be calculated using Equation (2.13). It is found to be [13]:

$$M_{12} = \frac{B_2(I_1) \cdot N_2 \cdot A_2}{I_1} = \frac{\mu_0 \cdot H(I_1) \cdot N_2 \cdot A_2}{I_1} \quad (2.15)$$

Due to the relationship $M = M_{12} = M_{21}$ the mutual inductance can be calculated as follows for the case $A_2 \geq A_1$: (If $A_2 \leq A_1$ then at the denominator of the equation 2.16, $R_2$ is replaced with $R_1$) [13]:

$$M_{21} = \frac{\mu_0 \cdot N_1 \cdot R_2^2 \cdot N_2 \cdot R_1^2 \cdot \pi}{2 \sqrt{(R_2^2 + x^2)^3}} \quad (2.16)$$

### 2.5 Coupling Coefficient $k_C$

Mutual inductance is a quantitative description of the flux coupling of two conductor loops. The coupling coefficient $k_C$ is introduced so that we can make a qualitative prediction about the coupling of the conductor loops independent of their geometric dimensions. The following applies [16]:

$$k_C = \frac{M}{\sqrt{L_1 \cdot L_2}} \quad (2.17)$$

The coupling coefficient always varies between the two extreme cases $0 \leq k_C \leq 1$.

- $k_C = 0$: Full decoupling due to great distance or magnetic shielding.
- $k_C = 1$: Total coupling. Both coils are subject to the same magnetic flux $\Phi$. The transformer is a technical application of total coupling, whereby two or more coils are wound onto a highly permeable iron core.

An analytic calculation is only possible for very simple antenna configurations. For two parallel conductor loops centred on a single $x$ axis the coupling coefficient according to [17] can be approximated from the following equation. However, this only applies if the radii of the conductor loops fulfil the condition $r_{\text{Transp}} \leq r_{\text{Reader}}$. The distance between the conductor loops on the $x$ axis is denoted by $x$.

$$k_C(x) \approx \frac{r_{\text{Transp}}^2 \cdot r_{\text{Reader}}^2}{\sqrt{r_{\text{Transp}}^2 \cdot r_{\text{Reader}}^2 \cdot \left(\sqrt{x^2 + r_{\text{Reader}}^2}\right)^3}} \quad (2.18)$$

Due to the fixed link between the coupling coefficient and mutual inductance $M$ and because of the relationship $M = M_{12} = M_{21}$, the formula is also applicable to transmitter antennas that are smaller than the transponder antenna. Where $r_{\text{Transp}} \geq r_{\text{Reader}}$, we write [10]:

$$k_C(x) \approx \frac{r_{\text{Transp}}^2 \cdot r_{\text{Reader}}^2}{\sqrt{r_{\text{Transp}}^2 \cdot r_{\text{Reader}}^2 \cdot \left(\sqrt{x^2 + r_{\text{Transp}}^2}\right)^3}} \quad (2.19)$$
The coupling coefficient \( k_c(x) = 1 \ (= 100\%) \) is achieved where the distance between the conductor loops is zero \((x = 0)\) and the antenna radii are identical \( (r_{\text{transp}} = r_{\text{reader}}) \), because in this case the conductor loops are in the same place and are exposed to exactly the same magnetic flux \( \psi \). In practice, however, inductively coupled transponder systems operate with coupling coefficients that may be as low as 0.01 \(<1\%)\) (Figure 2.9).

**Figure 2.7** The definition of mutual inductance \( M_{12} \) by the coupling of two coils via a partial magnetic flow.

**Figure 2.8** Graph of mutual inductance between reader and transponder antenna as the distance in the \( x \) direction increases.
2.6 Faraday’s Law

Any change to the magnetic flux $\Phi$ generates an electric field strength $E_i$. This characteristic of the magnetic field is described by Faraday’s law. The effect of the electric field generated in this manner depends upon the material properties of the surrounding area. Figure 2.10 shows some of the possible effects [13]:

- Vacuum: in this case, the field strength $E$ gives rise to an electric rotational field. Periodic changes in magnetic flux (high-frequency current in an antenna coil) generate an electromagnetic field that propagates itself into the distance.

- Open conductor loop: an open-circuit voltage builds up across the ends of an almost closed conductor loop, which is normally called induced voltage. This voltage corresponds with the line (path) integral of the field strength $E$ that is generated along the path of the conductor loop in space.

- Metal surface: an electric field strength $E$ is also induced in the metal surface. This causes free charge carriers to flow in the direction of the electric field strength. Currents flowing in circles are created, so-called eddy currents. This works against the exciting magnetic flux (Lenz’s law), which may significantly damp the magnetic flux in the vicinity of metal surfaces. This effect is undesirable in inductively coupled RFID systems (installation of a transponder or reader antenna on a metal surface) and must therefore be prevented by suitable countermeasures (For more information see literatures about ferrite shielding).
In its general form Faraday’s law is written as follows [13]:

\[ u_i = \oint E_i \cdot ds = -\frac{d\Psi(t)}{dt} \quad (2.20) \]

For a conductor loop configuration with \( N \) windings, we can say \( u = N \cdot d\psi/dt \) [13]. To improve under of inductively coupled RFID systems we will now consider the effect of inductance on magnetically coupled conduction loops. A time-variant current \( i_1(t) \) in conduction loop \( L_1 \) generates a time-variant magnetic flux \( d\Psi(i_1)/dt \). In accordance with the inductance law, a voltage is induced in the conductor loops \( L_1 \) and \( L_2 \) through which some degree of magnetic flux is flowing. We can differentiate between two cases:

- Self-inductance: the flux change generated by the current change \( di_n/dt \) induces a voltage \( u_n \) in the same conductor circuit.
- Mutual inductance: the flux change generated by the current change \( di_n/dt \) induces a voltage in the adjacent conductor circuit \( L_m \).
Figure 2.11 shows the equivalent circuit diagram for coupled conductor loops. In an inductively coupled RFID system $L_1$ would be the transmitter antenna of the reader. $L_2$ represents the antenna of the transponder, where $R_2$ is the coil resistance of the transponder antenna. The current consumption of the data memory is symbolised by the load resistor $R_L$. A time varying flux in the conductor loop $L_1$ induces voltage $u_2$ in the conductor loop $L_2$ due to mutual inductance $M$. From the circuit, $u_2$ is obtained as [10]:

$$u_2 = \frac{j \omega M \cdot i_1}{1 + \frac{j \omega L_2 + R_2}{R_L}}$$

$R_L \rightarrow \infty : u_2 = j \omega M \cdot i_1$

$R_L \rightarrow 0 : u_2 \rightarrow 0$

(2.21)

### 2.7 Resonance and Q Factor

The voltage $u_2$ induced in the transponder coil is used to provide the power supply to the data memory (microchip) of a passive transponder (see Section 2.8.1). In order to significantly improve the efficiency of the equivalent circuit illustrated in Figure 2.11, an additional capacitor $C_2$ is connected in parallel with the transponder coil $L_2$ to form a parallel resonant circuit with a resonant frequency that corresponds with the operating frequency of the RFID system in question*. The resonant frequency of the parallel resonant circuit can be calculated using the Thomson equation [9]:

$$f = \frac{1}{2 \pi \sqrt{L_2 \cdot C_2}}$$

(2.22)

In practice, $C_2$ is made up of a parallel capacitor $C'_2$ and a parasitic capacitance $C_p$ from the real circuit. $C_2 = (C'_2 + C_p)$. The required capacitance for the parallel capacitor $C'_2$ is found using the Thomson equation, taking into account the parasitic capacitance $C_p$:

$$C'_2 = \frac{1}{(2 \pi f)^2 L_2} - C_p$$

(2.23)

Figure 2.12 shows the equivalent circuit diagram of a real transponder. $R_2$ is the natural resistance of the transponder coil $L_2$ and the current consumption of the data carrier (chip) is represented by the load resistor $R_L$.

If a voltage $u_{Q2}=u_i = \omega . k_c \cdot \sqrt{L_1 \cdot L_2} \cdot i_1$ is induced in the coil $L_2$, the following voltage $u_2$ can be measured at the data carrier load resistor $R_l$ in the equivalent circuit shown in Figure 2.12 [10]:

$$u_2 = \frac{j \omega \cdot k_c \cdot \sqrt{L_1 \cdot L_2} \cdot i_1}{1 + (j \omega L_2 + R_2) \cdot \left( \frac{1}{R_L} + j \omega C_2 \right)}$$

(2.24)

*However, in 13.56 MHz systems with anti-collision procedures, the resonant frequency selected for the transponder is often 1–5 MHz higher to minimise the effect of the interaction between transponders on overall performance.
or in the non-complex form [18]:

\[
    u_2 = \frac{\omega \cdot k_c \cdot \sqrt{L_1 L_2} \cdot i_1}{\sqrt{\left(\frac{\omega L_2}{R_L} + \omega R_2 C_2\right)^2 + \left(1 - \omega^2 L_2 C_2 + \frac{R_2}{R_L}\right)^2}}
\]  

(2.25)

where \( C_2 = (C'_2 + C_p) \) and \( k_c \) is coupling coefficient. Figure 2.13 shows the simulated graph of \( u_2 \) with and without resonance over a large frequency range for a possible transponder system. The current \( i_1 \) in the transmitter antenna (and thus also \( \Phi(i_1) \)), inductance \( L_2 \), mutual inductance \( M \), \( R_2 \) and \( R_L \) are held constant over the entire frequency range.

Figure 2.12  Equivalent circuit diagram for magnetically coupled conductor loops. Transponder coil \( L_2 \) and parallel capacitor \( C_2 \) form a parallel resonant circuit to improve the efficiency of voltage transfer. The transponder’s data carrier is represented by the grey box [10].

Figure 2.13  Plot of voltage at a transponder coil in the frequency range 1–100 MHz, given a constant magnetic field strength \( H \) or constant current \( i_1 \). A transponder coil with a parallel capacitor shows a clear voltage step-up when excited at its resonant frequency (\( f_{\text{res}} = 13.56 \) MHz).
It is seen that the graph of voltage $u_2$ for the circuit with the coil alone (Figure 2.11) is almost identical to that of the parallel resonant circuit (Figure 2.12) at frequencies well below the resonant frequencies of both circuits, but that when the resonant frequency is reached, voltage $u_2$ increases by more than a power of ten in the parallel resonant circuit compared with the voltage $u_2$ for the coil alone. Above the resonant frequency, however, voltage $u_2$ falls rapidly in the parallel resonant circuit, even falling below the value for the coil alone. For transponders in the frequency range below 135 kHz, the transponder coil $L_2$ is generally connected in parallel with a chip capacitor ($C_2' = 20 \sim 220$ pF) to achieve the desired resonant frequency. At the higher frequencies of 13.56 and 27.125 MHz, the required capacitance $C_2$ is usually so low that it is provided by the input capacitance of the data carrier together with the parasitic capacitance of the transponder coil.

Let us now investigate the influence of the circuit elements $R_2$, $R_L$ and $L_2$ on voltage $u_2$. To gain a better understanding of the interactions between the individual parameters we will now introduce the $Q$ factor (Section 2.11.1). The $Q$ factor is a measure of the voltage and current step-up in the resonant circuit at its resonant frequency ($Q = f_r / \Delta f$ where $f_r$ is the resonant frequency and $\Delta f$ is the band-width). Its reciprocal $1/Q$ denotes the expressively named circuit damping $d$. In Figure 2.12 that $\omega$ is the angular frequency ($\omega = 2\pi f$) of the transponder resonant circuit [18]:

$$Q = \frac{1}{R_2 \cdot \sqrt{\frac{C_2}{L_2} + \frac{1}{R_L} \cdot \sqrt{\frac{L_2}{C_2}}} = \frac{1}{\frac{R_2}{\omega L_2} + \frac{\omega L_2}{R_L}}$$  \hspace{1cm} (2.26)$$

Equation (2.26) shows that when $R_2 \to \infty$ and $R_L \to 0$, the $Q$ factor tends towards zero. Also, when the transponder coil has a very low coil resistance $R_2 \to 0$ and $R_L \to \infty$ (corresponding with a high load resistor e.g. very low transponder chip power consumption), very high $Q$ factors can be achieved. Now, The voltage $u_2$ is proportional to the quality of the resonant circuit, which means that the dependency of voltage $u_2$ upon $R_2$ and $R_L$ is clearly defined. Voltage $u_2$ thus tends towards zero where $R_2 \to \infty$ and $R_L \to 0$. At a very low transponder coil resistance $R_2 \to 0$ and a high value load resistor $R_L \gg 0$, on the other hand, a very high voltage $u_2$ can be achieved (compare Equation 2.25). It is interesting to note the path taken by the graph of voltage $u_2$ when the inductance of the transponder coil $L_2$ is changed, thus maintaining the resonance condition (i.e. $C_2 = 1/\omega^2 L_2$ for all values of $L_2$). It is seen that for certain values of $L_2$, voltage $u_2$ reaches a clear peak (Figure 2.14).
If we now consider the graph of the $Q$ factor as a function of $L_2$ (Figure 2.15), then we observe a maximum at the same value of transponder inductance $L_2$. The maximum voltage $u_2 = f(L_2)$ is therefore derived from the maximum $Q$ factor, $Q = f(L_2)$, at this point. This indicates that for every pair of parameters ($R_2$, $R_L$), there is an inductance value $L_2$ at which the $Q$ factor, and thus also the supply voltage $u_2$ to the data carrier, is at a maximum. This should always be taken into consideration when designing a transponder, because this effect can be exploited to optimise the energy range of an inductively coupled RFID system. However, we must also bear in mind that the influence of component tolerances in the system also reaches a maximum in the $Q_{\text{max}}$ range. This is particularly important in systems designed for mass production. Such systems should be designed so that reliable operation is still possible in the range $Q \ll Q_{\text{max}}$ at the maximum distance between transponder and reader. $R_L$ should be set at the same value as the input resistance of the data carrier after setting the ‘power on’ reset, i.e. before the activation of the voltage regulator, as is the case for the maximum energy range of the system [10].

![Image](Figure 2.14) Plot of $u_2$ for different values of transponder (tag) inductance $L_2$. The resonant frequency of the tag is equal to transmission frequency of reader for all values of $L_2$ ($i_i = 0.5 \text{ A}, f = 13.56 \text{ MHz}, R_2 = 1\Omega$).

![Image](Figure 2.15) Graph of the $Q$ factor as a function of transponder inductance $L_3$, where the resonant frequency of the transponder is constant ($f = 13.56 \text{ MHz}, R_3 = 1\Omega$).
2.8 Practical Operation of the Transponder

2.8.1 Power Supply to the Transponder

Transponders are classified as active or passive depending upon the type of power supply they use. Active transponders incorporate their own battery to provide the power supply to the data carrier. In these transponders, the voltage $u_2$ is generally only required to generate a ‘wake up’ signal. As soon as the voltage $u_2$ exceeds a certain limit this signal is activated and puts the data carrier into operating mode. The transponder returns to the power saving ‘sleep’ or ‘standby mode’ after the completion of a transaction with the reader, or when the voltage $u_2$ falls below a minimum value.

In passive transponders the data carrier has to obtain its power supply from the voltage $u_2$. To achieve this, the voltage $u_2$ is converted into direct current using a low loss bridge rectifier and then smoothed. A simple basic circuit for this application is introduced in [10].

2.8.2 Voltage Regulation

The induced voltage $u_2$ in the transponder coil very rapidly reaches high values due to resonance step-up in the resonant circuit. Considering the example in Figure 2.13, if we increase the coupling coefficient $k_C$, possibly by reducing the gap between reader and transponder, or the value of the load resistor $R_L$, then voltage $u_2$ will reach a level much greater than 100 V. However, the operation of a data carrier requires a constant operating voltage of 3-5V (after rectification).

In order to regulate voltage $u_2$ independently of the coupling coefficient $k_C$ or other parameters, and to hold it constant in practice, a voltage-dependent shunt resistor $R_S$ is connected in parallel with the load resistor $R_L$. The equivalent circuit diagram for this is shown in Figure 2.16.

As induced voltage $u_{Q2}=u_i$ increases, the value of the shunt resistor $R_S$ falls, thus reducing the quality of the transponder resonant circuit to such a degree that the voltage $u_2$ remains constant.

To calculate the value of the shunt resistor for different variables, we refer back to Equation (2.25) and introduce the parallel connection of $R_L$ and $R_S$ in place of the constant load resistor $R_L$. The equation can now be solved with respect to $R_S$. The variable voltage $u_2$ is replaced by the constant voltage $u_{\text{transp}}$ - the desired input voltage of the data carrier - giving the following equation for $R_S$ [10]:

$$R_S = \left| \frac{1}{\left( j\omega \cdot k_C \cdot \sqrt{L_1} \cdot L_2 \cdot i_1 \right)} - 1 \right| + \frac{1}{R_L} \left( j\omega C_2 - \frac{1}{R_L} \right) |u_2 - u_{\text{transp}}| > u_{\text{Transp}}$$  (2.27)
Figure 2.16 Operating principle for voltage regulation in the transponder using a shunt regulator.

Figure 2.17 Example of the path of voltage $u_2$ with and without shunt regulation in the transponder, where the coupling coefficient $k_C$ is varied by altering the distance between transponder and reader antenna. (The calculation is based upon the following parameters: $i_1 = 0.5$ A, $L_1 = 1$ $\mu$H, $L_2 = 3.5$ $\mu$H, $R_L = 2$ k$\Omega$, $C_2=1/\omega^2L_2$).

Figure 2.17 shows the graph of voltage $u_2$ when such an ‘ideal’ shunt regulator is used. Voltage $u_2$ initially increases in proportion with the coupling coefficient $k_C$. When $u_2$ reaches its desired value, the value of the shunt resistor begins to fall in inverse proportion to $k$, thus maintaining an almost constant value for voltage $u_2$. Figure 2.18 shows the variable value of the shunt resistor $R_S$ as a function of the coupling coefficient. In this example the value range for the shunt resistor covers several powers of ten. This can only be achieved using a semiconductor circuit, therefore so-called shunt or parallel regulators are used in inductively coupled transponders. These terms describe an electronic regulator circuit, the internal resistance of which falls disproportionately sharply when a threshold voltage is exceeded. A simple shunt regulator based upon a Zener diode [19] is shown in Figure 2.19.
The value of the shunt resistor $R_s$ must be adjustable over a wide range to keep voltage $u_2$ constant regardless of the coupling coefficient $k_C$ (parameters as Figure 2.17).

![Figure 2.19 Example circuit for a simple shunt regulator.](image)

### 2.9 Interrogation Field Strength $H_{\text{min}}$

We can now use the results obtained in Section 2.7 to calculate the interrogation field strength of a transponder, which will depend on field, antenna design as well as distance away from it. This is the minimum field strength $H_{\text{min}}$ (at a maximum distance $x$ between transponder and reader) at which the supply voltage $u_2$ is just high enough for the operation of the data carrier. However, $u_2$ is not the internal operating voltage of the data carrier (3 or 5 V) here; it is the RF input voltage at the terminal of the transponder coil $L_2$ on the data carrier, i.e. prior to rectification.

The voltage regulator (shunt regulator) should not yet be active at this supply voltage. $R_L$ corresponds with the input resistance of the data carrier after the ‘power on reset’, $C_2$ is made up of the input capacitance $C_p$ of the data carrier (chip) and the parasitic capacitance of the transponder layout $C_2'=C_2=(C_2'+C_p)$.

$$u_1 = \mu_0 \cdot A \cdot N \cdot \omega \cdot H_{\text{eff}} \quad (2.28)$$
where $H_{\text{eff}}$ is the effective field strength of a sinusoidal magnetic field, $\omega$ is the angular frequency of the magnetic field, $N$ is the number of windings of the transponder coil $L_2$, and $A$ is the cross-sectional area of the transponder coil.

We now replace $u_{q_2}=u_j=\omega M \cdot i_1$ from Equation (2.24) with Equation (2.28) and thus obtain the following equation for the circuit in Figure 2.12:

$$u_2 = \frac{j \omega \cdot \mu_0 \cdot H_{\text{eff}} \cdot A \cdot N}{j \omega \left( \frac{L_2}{R_L} + R_2 C_2 \right) + \left( 1 - \omega^2 L_2 C_2 + \frac{R_2}{R_L} \right)} \quad (2.29)$$

We now solve this equation for $H_{\text{eff}}$ and obtain the value of the complex form. This yields the following relationship for the interrogation field $H_{\text{min}}$ in the general case [10]:

$$H_{\text{min}} = \frac{u_2 \cdot \sqrt{\left( \frac{\omega L_2}{R_L} + \omega R_2 C_2 \right)^2 + \left( 1 - \omega^2 L_2 C_2 + \frac{R_2}{R_L} \right)^2}}{\omega \cdot \mu_0 \cdot A \cdot N} \quad (2.30)$$

A more detailed analysis of Equation (2.30) shows that the interrogation field strength is dependent upon the frequency $\omega = 2\pi f$ in addition to the antenna area $A$, the number of windings $N$ (of the transponder coil), the minimum voltage $u_2$ and the input resistance $R_2$. This is not surprising, because we have determined a resonance step-up of $u_2$ at the resonant frequency of the transponder resonant circuit. Therefore, when the transmission frequency of the reader corresponds to the resonant frequency of the transponder, the interrogation field strength $H_{\text{min}}$ is at its minimum value.

To optimise the interrogation sensitivity of an inductively coupled RFID system, the resonant frequency of the transponder should be matched precisely to the transmission frequency of the reader. Unfortunately, this is not always possible in practice. First, tolerances occur during the manufacture of a transponder, which lead to a deviation in the transponder resonant frequency. Second, there are also technical reasons for setting the resonant frequency of the transponder a few percentage points higher than the transmission frequency of the reader (for example in systems using anti-collision procedures to keep the interaction of nearby transponders low).

Some semiconductor manufacturers incorporate additional smoothing capacitors into the transponder data chip to smooth out frequency deviations in the transponder caused by manufacturing tolerances. During manufacture the transponder is adjusted to the desired frequency by switching individual smoothing capacitors on and off [20].
In Equation (2.30) the resonant frequency of the transponder is expressed as the product \( L_2 C_2 \). This is not recognisable at first glance. In order to make a direct prediction regarding the frequency dependency of interrogation sensitivity, we rearrange Equation (2.22) to obtain:

\[
L_2 C_2 = \frac{1}{(2\pi f_0)^2} = \frac{1}{\omega_0^2}
\]  
(2.31)

By substituting this expression into the right-hand term under the root of Equation (2.30) we obtain a function in which the dependence of the interrogation field strength \( H_{\text{min}} \) on the relationship between the transmission frequency of the reader \( \omega \) and the resonant frequency of the transponder \( \omega_0 \) is clearly expressed. This is based upon the assumption that the change in the resonant frequency of the transponder is caused by a change in the capacitance of \( C_2 \) (e.g. due to temperature dependence or manufacturing tolerances of this capacitance), whereas the inductance \( L_2 \) of the coil remains constant. To express this, the capacitor \( C_2 \) in the left-hand term under the root of Equation (2.30) is replaced by \( C_2 = (\omega_0^2 L_2)^{-1} \) [10]:

\[
H_{\text{min}} = \frac{u_2 \cdot \omega^2 (\frac{L_2}{R_L} + \frac{R_2}{\omega_0^2 L_2})^2 + \left( \frac{\omega_0^2 - \omega^2}{\omega_0^2} + \frac{R_2}{R_L} \right)^2}{\omega \cdot \mu_0 \cdot A \cdot N}
\]  
(2.32)

Therefore a deviation of the transponder resonant frequency from the transmission frequency of the reader will lead to a higher transponder interrogation field strength and thus to a lower read range. An example about the transponder resonant frequency detuning is shown in Figure 2.20.

2.9.1 Energy Range of Transponder Systems

If the interrogation field strength of a transponder is known, then we can also assess the energy range associated with a certain reader. The energy range of a transponder is the distance from the reader antenna at which there is just enough energy to operate the transponder (defined by \( u_{2\text{ min}} \) and \( R_L \)). However, the question of whether the energy range obtained corresponds to the maximum functional range of the system also depends upon whether the data transmitted from the transponder can be detected by the reader at the distance in question.

Given a known antenna current \( I \), radius \( R \), and number of windings of the transmitter antenna \( N_1 \), the path of the field strength in the \( x \) direction can be calculated using Equation (2.3) (see Section 2.1.1). If we solve the equation with respect to \( x \) we obtain the following relationship between the energy range and interrogation field \( H_{\text{min}} \) of a transponder for a given reader [16]:

\[
x = \sqrt[3]{\frac{3}{2} \left( \frac{I \cdot N_1 \cdot R^2}{2 \cdot H_{\text{min}}} \right)^2} - R^2
\]  
(2.33)

**If the antenna current of the transmitter antenna is not known it can be calculated from the measured field strength \( H(x) \) at a distance \( x \), where the antenna radius \( R \) and the number of windings \( N_1 \) are known (see Section 2.1.1).**
Figure 2.20  Interrogation sensitivity of a contactless smart card where the transponder resonant frequency is detuned in the range 10–20 MHz ($N = 4, A = 0.05 \times 0.08 \text{ m}^2, \ u_2 = 5 \text{ V}, L_2 = 3.5 \mu \text{H}, R_2 = 5 \Omega, R_L = 1.5 \text{k}\Omega$). If the transponder resonant frequency deviates from the transmission frequency (13.56 MHz) of the reader an increasingly high field strength is required to address the transponder. In practical operation this results in a reduction of the read range.

As an example (see Figure 2.21), let us now consider the energy range of a transponder as a function of the power consumption of the data carrier ($R_L = u_2/i_2$). The reader in this example generates a field strength of $0.115 \text{ A/m}$ at a distance of $80 \text{ cm}$ from the transmitter antenna (radius $R$ of transmitter antenna: $40 \text{ cm}$). This is a typical value for RFID systems in accordance with ISO 15693.

As the current consumption of the transponder (lower $R_L$) increases, the interrogation sensitivity of the transponder also increases and the energy range falls. The maximum energy range of the transponder is determined by the distance between transponder and reader antenna at which the minimum power supply $u_{2\text{min}}$ required for the operation of the data carrier exists, even with an unloaded transponder resonant circuit (i.e. $i_2 \to 0, R_L \to \infty$). Where distance $x = 0$ the maximum current $i_2$ represents a limit, above which the supply voltage for the data carrier falls below $u_{2\text{min}}$, which means that reliable operation of the data carrier can no longer be guaranteed in this operating state.
The energy range of a transponder also depends upon the power consumption of the data carrier ($R_L$). The transmitter antenna of the simulated system generates a field strength of 0.115 A/m at a distance of 80 cm, a value typical for RFID systems in accordance with ISO 15693 (transmitter: $I = 1\text{A}$, $N_1 = 1$, $R = 0.4\text{m}$. Transponder: $A = 0.048 \times 0.076 \text{m}^2$ (smart card), $N = 4$, $L_2 = 3.6\mu\text{H}$, $u_{z\min} = 5\text{V}/3\text{V}$).

### 2.9.2 Interrogation Zone of Readers

In the calculations above the implicit assumption was made of a homogeneous magnetic field $H$ parallel to the coil axis $x$. A glance at Figure 2.22 shows that this only applies for an arrangement of reader coil and transponder coil with a common central axis $x$. If the transponder is tilted away from this central axis or displaced in the direction of the $y$ or $z$ axis this condition is no longer fulfilled.

If a coil is magnetised by a magnetic field $H$, which is tilted by the angle $\vartheta$ in relation to the central axis of the coil, then, in very general terms, the following applies [10]:

$$u_{0\vartheta} = u_0 \cdot \cos(\vartheta)$$

(2.34)

Where $u_0$ is the voltage that is induced when the coil is perpendicular to the magnetic field. At an angle $\vartheta = 90^0$, in which case the field lines run in the plane of the coil radius $R$, no voltage is induced in the coil.

Figure 2.21 The energy range of a transponder also depends upon the power consumption of the data carrier ($R_L$). The transmitter antenna of the simulated system generates a field strength of 0.115 A/m at a distance of 80 cm, a value typical for RFID systems in accordance with ISO 15693 (transmitter: $I = 1\text{A}$, $N_1 = 1$, $R = 0.4\text{m}$. Transponder: $A = 0.048 \times 0.076 \text{m}^2$ (smart card), $N = 4$, $L_2 = 3.6\mu\text{H}$, $u_{z\min} = 5\text{V}/3\text{V}$).
As a result of the bending of the magnetic field lines in the entire area around the reader coil, here too there are different angles $\theta$ of the magnetic field $H$ in relation to the transponder coil. This leads to a characteristic interrogation zone (Figure 2.23, grey area) around the reader antenna. Areas with an angle $\theta = 0^\circ$ in relation to the transponder antenna – for example along the coil axis $x$, but also to the side of the antenna windings (returning field lines) – give rise to an optimal read range. Areas in which the magnetic field lines run parallel to the plane of the transponder coil radius $R$ – for example, exactly above and below the coil windings – exhibit a significantly reduced read range. If the transponder itself is tilted through $90^\circ$ a completely different picture of the interrogation zone emerges (Figure 2.23, dotted line). Field lines that run parallel to the $R$-plane of the reader coil now penetrate the transponder coil at an angle $\theta = 0^\circ$ and thus lead to an optimal range in this area.

![Figure 2.22](image)

**Figure 2.22** Cross-section through reader and transponder antennas. The transponder antenna is tilted at an angle $\theta$ in relation to the reader antenna.

![Figure 2.23](image)

**Figure 2.23** Interrogation zone of a reader at different alignments of the transponder coil.
2.10 Equivalent Circuit Diagram of Reader

Figure 2.24 shows the equivalent circuit diagram for a reader (the practical realisation of this circuit configuration can be found in Section 2.11). The conductor loop necessary to generate the magnetic alternating field is represented by the coil $L_1$. The series resistor $R_1$ corresponds to the ohmic losses of the wire resistance in the conductor loop $L_1$. In order to obtain maximum current in the conductor coil $L_1$ at the reader operating frequency $f_{TX}$, a series resonant circuit with the resonant frequency $f_{RES} = f_{TX}$ is created by the serial connection of the capacitor $C_1$. The resonant frequency of the series resonant circuit is calculated using the Thomson equation (2.22). The operating state of the reader can be described by [9]:

$\frac{1}{2\pi \sqrt{L_1 \cdot C_1}}$ \quad (2.35)

Because of the series configuration, the total impedance $Z_1$ of the series resonant circuit is the sum of individual impedances, i.e.:

$Z_1 = R_1 + j\omega L_1 + \frac{1}{j\omega C_1}$ \quad (2.36)

At the resonant frequency $f_{RES}$, however, the impedances of $L_1$ and $C_1$ cancel each other out. In this case the total impedance $Z_1$ is determined by $R_1$ only and thus reaches a minimum [10].

$\frac{1}{j\omega C_1} = 0 \quad (\omega = 2\pi f_{RES}) \Rightarrow Z_1(f_{RES}) = R_1$ \quad (2.37)

The antenna current $i_1$ reaches a maximum at the resonant frequency and is calculated (based upon the assumption of an ideal voltage source where $R_i = 0$) from the source voltage $u_0$ of the transmitter high level stage, and the ohmic coil resistance $R_1$ [10]:

$u_1(f_{RES}) = \frac{u_0}{Z_1(f_{RES})} = \frac{u_0}{R_1}$ \quad (2.38)

$u_1$ and $u_{C_1}$ cancel each other out at the resonant frequency because $i_1$ is the same (Figure 2.24) i.e. $u_1$ and $u_{C_1}$ are in antiphase. The individual values for $u_1$ and $u_{C_1}$ may be very high. $u_0$ is a low source voltage, usually just a few volts, but figures of a few hundred volts can be reached at $L_1$ and $C_1$. Designs for conductor loop antennas for high currents must therefore incorporate sufficient voltage resistance in the components, in particular the capacitors, because otherwise these would easily be destroyed by arcing. An example of voltage step-up at resonance is shown in Figure 2.25. Despite the voltage may reach very high levels, it is completely safe to touch the voltage-carrying components of the reader antenna. The series resonant circuit is rapidly detuned by the additional capacitance of the hand, thus reducing the resonance step-up of voltage [10].
Figure 2.24 Equivalent circuit diagram of a reader with antenna $L_1$. The transmitter output branch of the reader generates the RF voltage $u_0$. The receiver of the reader is directly connected to the antenna coil $L_1$.

Figure 2.25 Voltage step-up at the coil and capacitor in a series resonant circuit in the frequency range 10–17 MHz ($f_{\text{res}} = 13.56$ MHz, $u_0 = 10$ V, $R_1 = 2.5\Omega$, $L_1 = 2$ $\mu$H, $C_1 = 68.8$ pF). The voltage at the conductor coil and series capacitor reaches a maximum of above 700 V at the resonant frequency. Because the resonant frequency of the reader antenna of an inductively coupled system always corresponds to the transmission frequency of the reader, components should be sufficiently voltage resistant.
2.11 Supply Reader Antennas via Coaxial Cable

Reader antennas in inductively coupled RFID systems generate magnetic flux, which is used for the power supply of the transponder and for sending messages between the reader and the transponder. This gives rise to three fundamental design requirements for a reader antenna:

- maximum current $i_1$ in the antenna coil, for maximum magnetic flux;
- power matching so that the maximum available energy can be used for the generation of the magnetic flux;
- sufficient band-width for the undistorted transmission of a carrier signal modulated with data.

Connecting the antenna coil using a long, unshielded two-core wire in the RF range would therefore lead to undesired effects, such as power reflections, impedance transformation and parasitic power emissions, due to the oscillations of an RF voltage. Because these effects are difficult to control when they are not exploited intentionally, shielded cable - so-called coaxial cable - is normally used in radio technology. Sockets, plugs and coaxial cable are uniformly designed for a cable impedance of $50 \, \Omega$ and, being a mass produced product, are correspondingly cheap. RFID systems generally use $50 \, \Omega$ components.

The block diagram of an inductively coupled RFID reader system using $50 \, \Omega$ technology in Figure 2.26 shows the most important RF components. The antenna coil $L_1$ represents an impedance $Z_L$ in the operating frequency range of the RFID system. To achieve power matching with the $50 \, \Omega$ system, this impedance must be transformed to $50 \, \Omega$ (matched) by a passive matching circuit. Power transmission from the reader output module to the matching circuit is achieved (almost) without losses or undesired radiation by means of a coaxial cable.

A suitable matching circuit can be realised using just a few components. The circuit illustrated in Figure 2.27, which can be constructed using just two capacitors, is very simple to design [21]. This circuit is used in practice in various 13.56MHz RFID systems. The impedance of a real antenna coil is generated by the serial connection of the coil inductance $L_S$ with the ohmic wire resistance $RL_s$ of the wire. The serial connection from $XL_s$ and $RL_s$ can also be represented in the impedance level.

Figure 2.28 shows a reader with an integral antenna for a 13.56MHz system. Coaxial cable has not been used here, because a very short supply line can be realised by a suitable layout (strip-line). The matching circuit is clearly visible on the inside of the antenna coil (SMD component) and capacitors with flat characteristic impedance at HF band are preferred to be used at the matching circuit in order to overcome parasitic effects.
Figure 2.26 Connection of an antenna coil using 50 Ω technology.

Figure 2.27 Simple matching circuit for an antenna coil.

Figure 2.28 Reader with integral antenna and matching circuit (MIFARE R-reader, reproduced by permission of Philips Electronics N.V.).
2.11.1 The Influence of the \(Q\) Factor

A reader antenna for an inductively coupled RFID system is characterised by its resonant frequency and by its \(Q\) factor. A high \(Q\) factor causes high current in the antenna coil and thus improves the power transmission to the transponder. In contrast, the transmission band-width of the antenna is inversely proportional to the \(Q\) factor. A low band-width, caused by an excessively high \(Q\) factor, can therefore significantly reduce the modulation side-band received from the transponder, which is undesirable when a strong detection of sub-carrier frequencies is preferred. However, based on the specifications of the HF RFID system, it should be an acceptable trade-off between the level of the transferred power and the data band-width. The \(Q\) factor of an inductive reader antenna can be calculated from the ratio of the inductive coil resistance to the ohmic loss resistance and/or series resistance of the coil [10,13]:

\[
Q = \frac{2\pi \cdot f_0 \cdot L_{\text{coil}}}{R_{\text{total}}}
\]  

(2.39)

The band-width of the antenna can be simply calculated from the \(Q\) factor [15]:

\[
BW = \frac{f_0}{Q}
\]  

(2.40)

For many systems, the optimal \(Q\) factor is 10–30. However, it is impossible to generalise here because the \(Q\) factor depends upon the required band-width and thus upon the modulation procedure used (e.g. coding, modulation, sub-carrier frequency).

2.12 A Comparison - Inductively Coupled RFID Antennas and Long Range RFID Antennas

When designing antennas for RFID systems several conditions need to be met. Antennas used for close coupled and inductively coupled systems are designed using completely different criteria than antennas used in long range systems. Different sorts of RFID systems are briefly investigated in the term of frequency and range in sections 1.1.1 and 1.1.2. Field regions of an antenna are shown in Figure 2.29 [22].

The antennas used in close coupled and inductively coupled systems are not really antennas in the classic radio meaning. These antennas are operating in the near field. The electric component (E-field) in the electromagnetic field is not used for communication in these systems. Instead, the magnetic component (\(B\) - field) is used through modulation of the load. The antennas in this type of communication are actually coils. A magnetic field is generated by the reader, inducing a current in the transponder antenna coil. As mentioned before in the previous sections of this chapter, the current induced in the transponder needs to be strong enough to support the transponder circuit with power. The important parameters to consider when designing the coils for
this type of system are maximum reading range and the minimum amount of power needed in the transponder for it to be operable.

The antennas used in long range RFID systems are operating in the far field and are therefore designed in a more classic antenna matter than the ones used for close and inductively coupled systems. The RFID long range transponder antenna is used for:

- Receiving the signal from the reader.
- Absorbing enough power to supply the transponder circuit with power.
- Transmitting signals back to the reader.

Apart from this, the circuits used for long range systems are often used in systems keeping track of goods. To keep costs at a low level the circuits should be small, cheap and being operable in sometimes shaded environments, e.g., warehouses. Which type of antenna that is used in general for RFID applications is impossible to say. In some services using RFID, a reader is placed at a fixed position and detects transponders passing by. One example of such a system is the ones used at toll roads to register payment for vehicles passing by. In this situation the transponders in the cars will always approach the reader from the same direction. These types of systems use a directional antenna to avoid waste of energy. Very commonly used antennas are the loop antenna and the dipole antenna. The antennas are often in form of micro-strip or patch antennas. The patch antennas can be constructed either as loop antennas, dipole antennas or folded dipole antennas [22]. Another useful antenna in readers for some services is the Yagi-Uda antenna [10].

![Figure 2.29](image) Field regions of an antenna.
2.12.1 Transition from Near-Field to Far-Field in Conductor Loops

As the magnetic field generated by a conductor loop propagates, an electric field increasingly also develops by induction (Section 2.1.1 and Figure 2.10). When going from a near field region to a far field region, as is depicted in Figure 2.29, the field, which was originally magnetic, is thus continuously transformed into an electromagnetic field. Moreover, at a distance of $\lambda/2\pi$ the electromagnetic field begins to propagate from the antenna into the space in the form of an electromagnetic wave. The area from the antenna to the point where the electromagnetic field forms is called the near-field of the antenna. The area after the point at which the electromagnetic wave has fully formed and propagated from the antenna is called the far-field.

A propagated electromagnetic wave can no longer retro-act upon the antenna that generated it by inductive or capacitive coupling. For inductively coupled RFID systems this means that once the far-field has begun, an inductive coupling is no longer possible. The beginning of the far-field (the radius $= \lambda/2\pi$ can be used as a rule of thumb) around the antenna thus represents an insurmountable range limit for inductively coupled systems [10].

2.13 Placing Antennas in Metal Environments

Metal surfaces, as the ones used in bookshelf systems, introduces difficulties for antennas in systems using radio communication in the far field and also for near field antennas in inductively coupled systems. This is a big issue since RFID antennas often need to be placed on metal, a HF RFID bookshelf as an example. The most simple resolution is to allow some spacing between the antenna and the metal surface. For 13.56 MHz, 2-3 cm of air spacing between antenna and metal is sufficient to assure practically no negative effects from the surrounding metal.

Several phenomena occur when an antenna coil is placed close to metal. The metal decreases the inductance of the coil causing the $Q$ factor to drop and self-resonance frequency to change. As an example, a Phillips Mifare 1k card changed from having $Q = 22$ and $f_{res} = 18.9$ MHz with only air surrounding to having $Q = 13$ and $f_{res} = 28.1$ MHz when placed upon a metal surface and measured with a network analyser. The other major effect, having the worst impact on the communication in metal environment is that the magnetic field induces eddy currents in the metal. The eddy currents create a counteracting magnetic field according to Lenz’s law, see Figure 2.30. This creates a minimum close to the metal surface and prevents communication.

The effect of eddy currents is commonly illustrated in basic physics or electromagnetic field theory courses by letting a magnet fall through both a metal tube and a plastic tube. When the magnet falls through the plastic tube, it is only affected by gravity. When it falls through the metal tube, the counteracting field created by eddy currents cause the fall time of an equal distance to be several times longer than in the case with no metal surroundings. This implies that eddy currents cause a significant difference and has to be included in design calculations [23].
Eddy currents create a counteracting \( B \)-field.

Illustration of the \( B \)-field in non-metal environment.

Illustration of the \( B \) field when a ferrite wave-guide is placed between the target antenna and the metal.
2.13.1 Wave-guide Materials

To avoid that the magnetic field induces eddy currents which create a counteracting field and prevent communication when metal is present, a highly permeable material with high resistivity can be used to guide the magnetic field away from the metal. Soft ferrite materials have good characteristics for this purpose. Since these materials already have been very useful within other radio areas than RFID (e.g., to reduce SAR values in cell phones), several well suiting products are available on the market. These products usually consist of resin layer mixed with powdered ferrite. This solution makes the material soft and form-able instead of hard and fragile. The magnetic field is propagated in the material and the high resistivity prohibits the formation of eddy currents. Therefore, no counteracting field is produced and the communication is not hindered. The magnetic field when no metal is present is shown in Figure 2.31. A piece of ferrite material is simply placed between the antenna and the metal to guide the B field past the metal without inducing any eddy currents as illustrated in Figure 2.32.

2.14 Summary

In order to design inductively coupled HF RFID antennas, under of the power and data transfer mechanisms are required thorough studying the physical principles of magnetic phenomena. In this chapter, physical characteristics of inductive coupling such as magnetic field strength $H$, magnetic flux $\Phi$, magnetic flux density $B$, inductance $L$, mutual inductance $M$, coupling coefficient $k_c$ and Faraday’s law were introduced, and then practical operation of tag and reader were investigated. A brief comparison between antennas used in inductively coupled RFID systems and antennas used in long range RFID systems (Far Field) was also presented in this chapter. All the reviewed concepts prepare a good background theory to design HF RFID antennas. For the scope of this thesis, the application of HF RFID to smart applications in Internet of things will be studied further whereby background theory presented in this chapter will be applied to new antenna solutions presented in subsequent chapters.
Chapter 3

3  HF RFID Reader Antennas for Smart Shelf Applications - Prior Art

The concept of RFID as a prerequisite for the IoT applications was briefly described in Section 1.1 and also RFID smart shelf was mentioned as one of the major subdivisions of smart RFID reader systems. In the recent years, renewed interest and rapid progress has been made in the application of high frequency (HF) RFID or inductive methods for the IoT applications [11]. Also more recently, the RFID smart-shelf system has received much attention, because of increasing demands for large-scale item-level management of such items as grocery products in the retail supply chain, large volumes of books in libraries, bottles in pharmaceutical industry, and important documentation in offices [24-27]. In the retail industry, keeping adequate stock on the shelves and ensuring product placement within the storage container or on the shelf at the appropriate time are critical. The RFID smart-shelf system can help here. It can also go a step further, by automatically sending stock requests to the back room, warehouse, or supplier. Suppliers and retailers can also collect data and keep track of the time a product stays at a specific location in the retail supply chain. Even home users will find such a system useful, as it helps to identify items already at home to prevent purchase duplication and avoid waste.

All the previous works address the RFID application in terms of the performance of ultra high frequency (UHF) band and yet little have been reported that are relevant to HF RFID applications [10]. Inductive coupled or HF RFID technology, which has operation range of around one metre, has good performance in a crowded environment when compared with UHF RFID because the magnetic field is not affected by most of the surrounding dielectric materials \( \mu_r \approx 1 \) [12]. A summary of previews works by others on HF RFID reader antennas is presented here, and characteristics of a practical HF RFID reader antenna for smart shelf applications as an example is shown. Technical points with more details will be explained in next chapters parallel with the new achievements of this research.
3.1 Summary of Challenges on HF RFID Reader Antennas

The antenna is one of the key factors in RFID systems [28-32]. The detection range and accuracy of an RFID system are directly dependent on the performance of the reader and tag antennas. In addition, optimised antenna design benefits the RFID system with longer range, better accuracy, and lower antenna fabrication cost, as well as simpler system configuration and implementation.

Loop antennas have been widely used in near-field RFID applications at low frequencies (LF: 30 to 400 kHz) or high frequencies (HF: 3 to 30 MHz), where the power transfer and communication between the RFID reader and tags is achieved by inductive coupling [10, 33-35]. Of all antennas that can be used to excite a magnetic field, a circular or rectangular loop is clearly the best choice in case the current on the entire antenna is in phase. As the distance from the point where the tag is located to all current carrying parts of the antenna is equal in this case, the contributions of all parts of the antenna arrive in phase at the tag, resulting in constructive interference. For a larger loop, where the current over the loop can not be supposed constant, it is less obvious [36]. The characteristics of a perfectly conducting loop have been studied extensively in free space [37-42], and above a conducting or dielectric interface [43-50]. These works have addressed the radiation characteristics of loop antennas, such as the gain, directivity, and radiation patterns for antenna dimensions of the order of $\beta b \approx 1$ (where $\beta = 2\pi/\lambda$, with $\lambda$ being the wavelength at the corresponding operating frequency, and $b$ being the radius of a circular loop antenna or the side of a square or rectangular loop antenna). These results are not applicable to the loop antennas used in LF and HF RFID applications, where the loop antenna is electrically tiny ($\beta b \approx 0.01$ or less) and the radiation characteristics of the antenna are not of interest. Instead, the loop antenna is usually characterised by parameters such as the resonant frequency, magnetic-field intensity, and quality factor ($Q$ factor).

A HF RFID reader antenna has to provide the tag with a field that is sufficient to power up the hardware in the tag. Hence, the antenna will be designed in such a way that the magnetic field at a certain read out distance $r_d$ is large enough. The read out distance can not be infinite, because inductive coupling from the tag to the reader implies that the tag must be in the near field of the reader. Indeed, if the tag receives a travelling wave instead of a quasi-static field, the modifications to the field due to the tag will never travel back to the reader. This means that the tag must be located in the reactive near field [51] around the antenna, in this case $r_d \ll \lambda/2\pi \approx 4$. Furthermore, if the total wire length of the loop becomes a considerable part of the wavelength, the loop can not be considered as a lumped element. Length increments will cause several resonances at lower frequencies and decrease the total field. In such case the loop radius should be decreased and the current increased even more.
In [52] equivalent circuit models of inductive coupled RFID antennas are extracted by means of the partial element equivalent circuit (PEEC) method where the antenna impedance is analysed separately regarding frequency dependent behaviour including skin and proximity effects as well as parasitic capacitances. The design of a HF RFID reader antenna for usage at 13.56 MHz is described in [36], which presents the theory with emphasis on the effect of the read out distance on the antenna design. To compute the power delivered to the RFID tag as a function of the mutual coupling between the loop antennas of the tag and reader, an analytical method is introduced in [12]. The effect of the variable mutual coupling on the power delivery is discussed and an adaptive impedance matching circuit implemented on the reader antenna is proposed as a solution to mitigate the detrimental effect of the mutual load on the power transfer efficiency. Ink-jet printed HF RFID antennas using commercially available silver nano-particle ink are presented in [53], and quality factors of 5.3 and 9.4 are obtained for 13.56 MHz.

Energy transmission of inductive coupled systems is investigated, enabling an optimised system design in [54] and a transmission of 80 mW over a distance of up to 7.6 cm is obtained by 275 mW reader output power. The design study of a HF RFID system in [54] suggests that the $Q$ of a reader coil antenna should be as small as possible for some standards when reliable data transfer is aimed. This observation is followed in [36, 55], where the a single turn coil reader antenna is proposed, which in fact had the minimum achievable $Q$ suitable for reliable data transmissions. To enhance the $H$-field of a single turn reader antenna with low $Q$, an approach is applied to exploit the internal area of the antenna by adding several inward turns in [56], but the approach for unconstrained $Q$, [56], will be seen as an impractical solution and is at the expense of a 25% reduction in the $Q$ factor (increases the antenna loss and lowers its efficiency, [57]).

According to the above summary, an optimised design with proper $H$-field and $Q$ performance is necessary to investigate.

3.2 Loop Antennas and Metal-Backed Loop Antennas with HF RFID Smart Shelf Application

As a good example of applicability of a loop antenna to the HF RFID smart shelf in particular, Figures 3.1 and 3.2 show the configuration of the loop antenna [9, 55]. The antenna is considered as a basis for the novel work undertaken in Chapter 4 of this thesis. As the antenna is used to identify book position here, the dimensions of the antenna is considered a bit bigger than the dimensions of a normal book while to identify other objects, different dimensions for the loop-antenna/smart-shelf should be considered. However, the loop antenna with dimensions of 250 mm $\times$ 250 mm, was designed and fabricated on a low-cost printed circuit board (PCB) with dimensions of 280 mm $\times$ 280 mm and a thickness of 0.5 mm. The width of the strips was 10 mm. The metal-backed loop antenna was configured by adding a metal plate (280 mm $\times$ 280 mm) at one side of
the loop antenna. The metal plate was positioned parallel to the loop antenna with a separation $S=10$ mm. The supporting material between the antenna and the plate was Styrofoam with a dielectric constant of $\sim 1.07$. The output stubs of the antenna were connected to the inner/outer conductors of a sub miniature version A (SMA) connector, respectively. An impedance matching circuit [33, 58] with capacitors was designed to match the antenna to a 50 Ω feed line, and to tune the operating frequency of the antenna to 13.56 MHz.

**Figure 3.1** The configuration of the metal-backed loop antenna. Reprinted from Characteristics of a Metal-Backed Loop Antenna and its Application to a High-Frequency RFID Smart Shelf, by X. Qing and Z. N. Chen, 2008, Antennas and Propagation Magazine, IEEE, Volume 51, page 26-38. Reprinted with permission.

### 3.2.1 Impedance Matching, Field Response and Resonant Frequency

In order to match the antenna to 50Ω input both $PI$ or $T$ matching circuits can be chosen [33]. Both $PI$ and $T$ matching networks, have an extra degree of freedom in comparison with $L$ matching networks that allows one to control the bandwidth of the match. However, $PI$ networks can be converted to $T$ networks and in vice versa using standard impedance transformations [33, 70]. Also, as $PI$ network is more convenient to connect to a balance structure, it is chosen here. When choosing lumped elements to be used at the matching circuit, since the tolerance of the inductance value becomes the variation in the matching, an inductor with less tolerance may be required to satisfy the necessity. The actual inductance value of the inductor is not a continued
value like 1, 1.1 or 1.2 nH while there are more degrees of freedom to choose proper value of capacitors and also choosing variable capacitors. However, the loop antenna was matched to 50 Ω at 13.56 MHz by using the PI circuit (Figure 3.1) with the following capacitors: $C_{S1} = 47$ pF, $C_{S2} = 82$ pF, $C_{P1} = 390$ pF, $C_{P2} = 7$ to 50 pF, and $C_{P3} = 120$ pF. The variable capacitor in the circuit was used to tune the frequency to the desired 13.56 MHz, with a return loss, $|S_{11}|$, of -15 dB which yields to a small bandwidth. A measured relative magnetic-field intensity of -34.8 dB was observed when the probe was placed at the centre of the loop antenna with a distance of 50 mm. The separation of the metal plate had a significant effect on the resonant frequency and the impedance matching. Furthermore, the metal plate on the shelf placed close to the loop antenna led to a large resonant frequency shift, which was caused by the occurrence of the eddy current in the metal plate.

### 3.2.2 Magnetic-Field Intensity, Field Distribution and Detection Range

Figure 3.3 exhibits the simulated magnetic-field distribution of the metal-backed loop antennas at corresponding resonant frequencies. The loop was located in the $x$-$y$ plane at $z = 0$ mm, and the size of the metal plate was 280 mm × 280 mm. The metal plate was placed at $z = 15$, -10, -15, -25, and -35 mm. As it is seen, the closer the metal plate was placed, the weaker was the field intensity.

Figure 3.4 shows the variation of the measured magnetic-field intensity of the matched antennas as a function of distance at 13.56 MHz. As expected, the loop antenna produced an identical magnetic-field intensity variation at both sides, with the strongest magnetic-field intensity near $d = 0$ mm. Also, the metal-backed loop antennas showed the asymmetrical characteristics of the magnetic field intensity, which was much weaker at the left side (negative $z$ axis, with metal plate) than that at the right side (positive $z$ axis, without metal plate). The peak magnetic field occurred at a few centimetres away from the antenna, and weakened towards the metal plate.

When the separation was larger ($s = 25$ mm), the tag can be detected at 40 cm away at the side without the metal plate, while the distance was reduced to 30 cm on the opposite side. As the metal plate moved closer toward the loop antenna, the detection range showed a decreasing trend on both sides. The tag can only be detected with a detection range of 25 cm at the side without the metal plate of the metal-backed loop antenna when $s$ was 5 mm.

### 3.2.3 Application to HF-RFID Smart-Shelf System

The above antenna was developed in the Institute for Infocomm Research, Singapore, to monitor and track the location of the books on a tier base in real time in a library. Figure 3.5 illustrates a typical antenna configuration in an HF-RFID smart-shelf system. Two loop antennas are
vertically positioned in each tier to detect the tagged items on the tier. Positioning the loop antennas vertically on the racks is the most direct method for book detection, as the tags affixed inside the books are oriented in such a manner that this allows the most of the magnetic flux to pass through the antennas. System implementation and cost are important considerations as well. From a detection-accuracy point of view, the coupling zones of the antennas should be restricted to their own tiers: any extended coupling to other tiers will provide wrong information, and degrade the detection accuracy. The conventional antenna configuration, as shown in Figure 3.5, suffers from severe interference between the antennas in the adjacent tiers. This greatly degrades the detection performance due to wrongly reading books in other tiers. It is a big challenge to control the coupling zone of the loop antenna, because the antennas are required to offer enough field intensity for enough coverage in a specific tier, while the dimensions of the antennas are constrained to fit the shelves.

In addition, using antennas that are of different dimensions in the shelves is not effective from a system point of view, because it makes the system more complex and costly. One way to constrain the interference between the antennas is to install shielding plates between the tiers. However, this solution is not preferable in practical applications, since it greatly increases the cost of system implementation. Furthermore, modifying existing shelves is not allowed in most practical applications. The metal-backed loop antenna offers great promise and flexibility for solving the problems mentioned above. The asymmetry of the coupling zone of the antenna provides extra flexibility for the system design and implementation.

![Image](image.png)

**Figure 3.2** The prototype metal-backed loop antenna. Reprinted from Characteristics of a Metal-Backed Loop Antenna and its Application to a High-Frequency RFID Smart Shelf, by X. Qing and Z. N. Chen, 2008, Antennas and Propagation Magazine, IEEE, Volume 51, page 26-38. Reprinted with permission.
Figure 3.3 The simulated magnetic-field distribution of the metal-backed antenna with different separations of the metal plate. The loop was located in the x-y plane at z = 0 mm, and the field distribution was calculated in the x-y plane at z = 50 mm. The size of the metal plate was 280 mm × 280 mm: (a) s =5mm, (b) s =10 mm, (c) s =15 mm, (d) s=25 mm, (e) s=35mm, (f) loop antenna. Reprinted from Characteristics of a Metal-Backed Loop Antenna and its Application to a High-Frequency RFID Smart Shelf, by X. Qing and Z. N. Chen, 2008, Antennas and Propagation Magazine, IEEE, Volume 51, page 26-38. Reprinted with permission.
Figure 3.6 demonstrates an antenna configuration in an HF RFID smart-shelf system wherein the loop antennas and the metal backed loop antennas are alternately positioned in adjacent layers to constrain the interference between the antennas. The antennas are positioned in as follows:

- Antennas on adjacent tiers in adjacent shelves (such as Tier 2 in Shelf 1, Shelf 2, and Shelf 3): The left and right tiers in Shelf 1 and Shelf 3 can be covered by using two loop antennas, respectively. The middle tier in Shelf 2 is designed to have one loop antenna positioned in the centre, to provide the major coverage of the tier, and two fully metal-backed loop antennas at the ends of the tier, to respectively offer unidirectional detection. The magnetic-field intensity of the metal backed loop antenna is required to be strong enough on one side to provide adequate detection capability for coverage of their own tier, whereas the magnetic-field intensity is also required to be weak enough on the other side to avoid detection of the books in other adjacent tiers.

- Antennas on adjacent tiers in the same shelf (such as Tier 1, Tier 2, and Tier 3 in Shelf 1): The antennas are mounted with specific distance offsets on the upper and lower tiers.

- Antennas on adjacent tiers in opposite shelves (such as Tier 2 in Shelf 1/Shelf 4): The antennas are located with adequate distance offsets.

An HF-RFID smart-shelf system using such an antenna configuration has been successfully implemented in the Singapore National Library in 2008, as shown in Figure 3.7, for real-time monitoring and tracking of the location of the books on a tier base. The detection accuracy of the system was up to 95%.
Figure 3.5  The antenna configuration of the RFID smart-shelf system using the loop antenna. The interference between the antennas degraded the system-detection accuracy.

Figure 3.6  The antenna configuration of the RFID smart-shelf system using the loop antennas and metal-backed loop antennas to constrain interference.
3.3 Summary

A summary of previous challenges on HF RFID reader antennas investigated by others was presented in this chapter. Also, a practical HF RFID reader antenna for smart shelf applications as an example was discussed as a base for next chapter and also to provide a characterized understanding of the issue.
Chapter 4

4 Increased Band-width HF RFID Reader Antennas and its Evaluation For Smart Shelf Applications - Novel Work Undertaken

Firstly, a new theoretical approach is presented in this chapter as a novel work to increase the band-width of RFID reader antennas operating at the HF band to improve the reception of the backscattered RFID response. It is shown here that, a wider band-width will support sub-carrier frequencies for the all the existing HF RFID standards (ISO/IEC 15693 and ISO/IEC14443) to be detected more easily and thus leads to increase the range of identification. A prototype antenna is designed and evaluated by measurement for its capability in being used for smart shelf applications. A wider receiving range is obtained for the ISO/IEC standards about 6 times more than previous antennas. This causes better receiving of the sub-carrier frequency with a minimum optimisation of 12% and maximum optimisation of 30% for the ISO/IEC standards in comparison with the previous antennas. Also, it is shown that in the presence of books, the HF RFID technology is capable of identifying the distance of a tag antenna or the position of book based on the received $H$-field, which can be used in a smart shelf management system.
4.1 Approach to Increase the Band-width of HF RFID Antennas

A procedure similar to the design of bandpass filters is presented to increase the bandwidth of the antennas operating around the S band (2 - 4 GHz) [59], for example 0.5λ or 0.25λ resonators are used in the form of open stubs at the input of micro-strip antennas to increase their band-width [59-62] similar to the sample shown in Figure 4.1. The mechanism of increasing bandwidth is shown in Figure 4.2. The stub resonator adds a new resonance to the antenna resonance and this two resonances together form an increased bandwidth. The characteristic impedance of an open ideal 0.25λ stub, $Z_{0.25\lambda}$, at the frequency of $f$ based on transmission line theory is simulated by ADS software and is shown in Figure 4.3.

![Figure 4.1](image1.png) Left: Layout of the antenna using resonator stubs; Right: Layout of reference antenna [59].

![Figure 4.2](image2.png) Mechanism of band-width increment

![Figure 4.3](image3.png) Characteristic impedance of an open ideal 0.25λ stub at frequency of f.
As the 0.5λ or 0.25λ resonator stubs have large length values at the HF band, it is nearly impossible to use them in the input of the practical HF RFID antennas in order to increase their band-width. For example for a HF RFID antenna operating at 13.56 MHz, the wavelength (λ) is about 22.5 m while the dimension of conventional HF RFID antenna is about 25×25 cm² [9, 55]. Also, using long stubs causes unwanted radiations. An equivalent circuit of the open 0.25λ resonator stub is already presented as a series LC circuit [63] in order to implement RF filters, as shown in circuit A in Figure 4.4.

In our work, circuit A is considered to be added to the input of HF RFID antenna, after some modifications as circuit B (Figure 4.4), in order to achieve a band-width increment. Circuit B is obtained by adding a capacitor, $C_p$, parallel with the inductor $L$ at the circuit A. The input impedance of circuit B is calculated as:

$$Z_B = \frac{LS^2(C_s + C_p) + 1}{(LC_pS^2 + 1)C_sS}$$  \hspace{1cm} (4.1)$$

where $S = j\omega = j2\pi f$. It is obvious that for $C_p << C_s$, $Z_B \approx Z_A$ ($Z_A$ is the input impedance of circuit A). However, the resonant frequencies of circuit A and circuit B are $f_A = 1/2\pi \sqrt{LC_s}$ and $f_B = 1/2\pi \sqrt{L(C_s + C_p)}$, respectively. Considering $C_p = 0.2C_s$, $f_B = 0.92f_A$ is obtained. This means that by using a small $C_p$ capacitor, in par- comparison with the $C_s$, the resonance frequency of $f_B$ is around $f_A$. Circuit B with a small $C_p$ is used because it is less sensitive to impedance variations due to cable movement than the circuit A. For example, when moving the cable connected to an 13.56 MHz HF RFID antenna attached with the circuit A instead of the circuit B in Figure 4.5 there is a maximum 25% frequency shift due to the cable movement in the enhanced bandwidth, while with the circuit B with a small $C_p$ there isn’t any frequency shift. The antenna in Figure 4.5 will be investigated with more detail in Section 4.2. According to the above discussion, by choosing in-range values for $C_p$, $C_s$ and $L$ in circuit B added to the input of HF RFID antenna, considering $f_B = 1/2\pi \sqrt{L(C_s + C_p)}$ as the operating frequency of the HF RFID antenna and $C_p<0.2C_s$ condition, an increased band-width is expected to be obtained with the minimum frequency shift due to the cable movement.

![Figure 4.4](image.png)

**Figure 4.4** (A) Equivalent series LC circuit of an open 0.25λ stub used for RF filter design [63] (B) modified LC circuit of circuit A.
4.2 Prototype HF RFID Antenna Operating at 13.56 MHz for Smart Shelf

A HF RFID loop antenna operating at 13.56 MHz is already characterised and investigated in terms of impedance matching, resonant frequency, magnetic-field intensity/field distribution, quality factor, and detection range in [9, 55]. Here, a square loop antenna having the same dimensions as the antenna introduced in [9, 55] (side = 250 mm) is considered in order to increase its band-width considering the conditions presented in section 4.1. Values of the lumped elements of circuit B in Figure 4.4 are calculated and are added to a HF input impedance matching circuit as shown in Figure 4.5. $C_p = 27 \, \text{pF}$, $C_s = 133 \, (100 || 33) \, \text{pF}$ and $L = 1 \, \mu\text{H}$ are chosen as proper values to satisfy the conditions in Section 4.1 where $f_{HF} = f_B = 13.56 \, \text{MHz}$ is considered. An impedance matching circuit [33, 58] with capacitors was designed to match the antenna to a 50 $\Omega$ feed line, and to tune the central frequency of the antenna to 13.56 MHz with $C_{P1} = 470 \, \text{pF}$, $C_{P2} = 165 \, (150 || 15) \, \text{pF}$, $C_{S1} = 33 \, \text{pF}$ and $C_{S2} = 27 \, \text{pF}$ as shown in Figure 4.5. The antenna is fabricated and mounted on a supporting flexible substrate with a dielectric constant of 1.07 and 0.1 mm thickness. The equivalent circuit of the whole antenna structure containing the loop, matching circuit and circuit B is shown in Figure 4.6. The loop is modelled as an inductor $L_A$ having a very low resistance $R_A \approx 0$ [10].

**Figure 4.5** Configuration of the enhanced band-width HF RFID loop antenna.
4.3 Simulation and Measurement Results

The return loss measurements are carried out by connecting the antenna to a calibrated Network Analyser. The simulated and measured return loss of the antenna in Figure 4.5, is presented in Figure 4.7. The small difference between measurement and simulation results is due to less than 20% difference between the $Q$ factor of the simulated and real lumped elements. However, it is seen that the obtained $S_{11} < -10$ dB band-width is about 800 kHz which is about 6 times more than the antenna presented in [9, 55] which has less than 100 kHz band-width. The return loss of the enhanced band-width HF RFID antenna presented here for the different 13.56 MHz HF RFID standards are compared in Table 4.1 with the antenna of [9,55], where 13.56 MHz+$f_s$ kHz is the band-width of the standard and $f_s$ is the sub-carrier frequency. For both ISO/IEC 15693 standards with $f_s =$212 and 424 kHz while the ISO/IEC14443 standard with $f_s = 848$ kHz [10], a better reception of the sub-carrier frequency with a minimum increase optimisation of 12% and maximum increase optimisation of 30% is obtained for the standards in comparison with the HF RFID antenna in [9, 55]. Figure 4.8 shows the simulated magnetic-field distribution of the enhanced band-width loop antenna at three frequencies around the start, middle and end of the band-width (13.2, 13.5 and 13.8 MHz) using a 2 dBm power source. The simulated loop is located in the $x$-$y$ plane at $z = 0$ mm and the magnetic-field distribution in the $x$-$y$ plane at $z = 50$ mm is shown. $S_{12}$ parameter parameter is considered as channel coefficient here where the reader antenna and the tag antenna are connected to the two ports of the calibrated Network Analyser. Figure 4.9 shows the simulated and the measured channel coefficients over the whole band-width of the enhanced band-width antenna at 20 cm and 30 cm in front of the antenna at free space and also with books as a sample material in front of the antenna using a 27 dBm power source. Figure 4.10 shows the measured $H$-field of the loop antenna, at the centre and corner of the antenna, versus distance with and without books at 13.56 MHz using a 27 dBm power source.
is fixed on the surface of a stable large cardboard box with negligible dielectric effects, which is considered as the x-y plane. Three 13.56 MHz probe tags are used to do measurements as shown in Figure 4.11 (Two rectangular and one circular). The probes are connected separately to a spectrum analyser and are moved along the z axis at the corner and the centre of the antenna to carry out measurements. Also, the measurements are carried out again with books as a sample material aligned in front of the antenna. Results presented in Figure 4.10 show that the designed antenna does have a characteristic in how the magnetic $H$-field decays over distance within 80 cm and that the effect of the books maintains such a characteristic wherever the tag is positioned on the book. The $H$-field is dependent on $1/r^3$ at the identification distance as expected for the near field of the antenna. As for ISO/IEC 14443 and ISO/IEC 15693 standards the minimum $H$-field is 1.5 A/m and 0.115 A/m, respectively, the measured $H$-field is within an appropriate range. However, based on the Biot-Savart law, the $H$-field is expected to be increased with a larger input power. The measurement setup is shown in Figure 4.12.

![Figure 4.7 The measured and simulated return loss of the enhanced band-width loop antenna vs [9,55].](image)

**Table 4.1** Maximum return loss (Figure 4.6) of the enhanced band-width antenna for 13.56 MHz HF RFID standards

<table>
<thead>
<tr>
<th>Standard</th>
<th>$f_0$ (kHz)</th>
<th>Minimum return loss at $13.56 \pm f_0$ MHz band-width</th>
<th>Optimization percentage</th>
</tr>
</thead>
<tbody>
<tr>
<td>ISO/IEC 15693</td>
<td>212</td>
<td>0.5 dB [9,55] 15 dB Enhanced BW antenna</td>
<td>15/0.5= 30.00%</td>
</tr>
<tr>
<td></td>
<td>424</td>
<td>0.5 dB [9,55] 9.5 dB Enhanced BW antenna</td>
<td>20.00%</td>
</tr>
<tr>
<td>ISO/IEC 14443</td>
<td>848</td>
<td>0.5 dB [9,55] 6 dB Enhanced BW antenna</td>
<td>12.00%</td>
</tr>
</tbody>
</table>

56
Figure 4.8 The simulated magnetic-field distribution of the enhanced band-width loop antenna at 13.2, 13.5 and 13.8 MHz. The loop was located in the x-y plane at \( z = 0 \) mm, and the field distribution was calculated in the x-y plane at \( z = 50 \) mm.

Figure 4.9 The measured and simulated channel coefficient distributions of the loop antenna.

Figure 4.10 The measured \( H \)-field of the loop antenna versus distance with and without books at 13.56 MHz using a 27 dBm power source.
Figure 4.11 13.56 MHz tags which are used for the measurements.

Figure 4.12 The measurement setup of the enhanced band-width antenna with and without books.

4.4 Summary

A method to increase the band-width of HF RFID reader antennas was presented in this chapter and a prototype was subsequently produced suitable for smart shelf applications with increased reading range. The results presented in this chapter show that the smart shelf antenna designed with increased band-width does accommodate the ISO/IEC standards for NFC/RFID at the HF band does have a defined characteristic in how the magnetic $H$-field decays over distance within 1m. The effect of the books maintains such a characteristic wherever the tag is positioned on the book. However, the magnitude of the $H$-field at a fixed distance from the antenna is not uniform when a tag is placed in a different position in the book. For example a different $H$-field will be received when the tag is placed in the centre of the book compared to the top or bottom of the book. Further work is required to ensure there is uniformity at a fixed distance for smart shelf capabilities, which will be addressed in the next chapters.
Chapter 5

5  Multi Turn Small Self Resonant Coil (MT SSRC) at HF Band - Novel Work Undertaken

A new model of inductively coupled HF RFID reader antennas is presented in this chapter based on the idea of transmitting at the self resonance frequency (SRF) of a small multi turn HF coil as a HF RFID reader antenna. Multi-turn small coils for HF RFID application are reviewed briefly in section 5.2. The introduced new model of multi turn small self resonant coil (MT SSRC) antenna operating at its SRF is analysed mathematically at the HF band in terms of SRF, number of turns, dimensions and dielectric characteristics of the insulation if present in Section 5.3. The analysis shows that the adjacent turns would be physically in contact in the structure of a non-insulated MT SSRC, which causes an electrical short and also leads to a very sensitive structure to small mechanical stresses. To get a practical structure, the insulated MT SSRC should be chosen to be investigated as a HF RFID antenna. This chapter is a backbone to the next chapter, Chapter 6, where two turn planar SSRC (TTP SSRC) antennas having better characteristics for HF RFID applications in comparison with the similar antennas in the literature, are introduced.
5.1 From the Old Model to a New Model of HF RFID Antennas

In previous inductively coupled/HF RFID systems the reader coil/loop antenna has been modelled as an inductor, \( L_1 \), [9, 10, 33, 36, 52, 54, 55] where the series resistor, \( R_1 \), corresponds to the ohmic losses of the wire resistance of the coil which are insignificant for electrically small coils. In this classic model as shown in Figure 5.1(a), by parallel connection of a lumped capacitor \( C_1 \) to the inductor \( L_1 \), a resonant circuit is created and the resonance frequency/operating frequency of the reader circuit is calculated using the Thomson equation (2.22) as:

\[
 f_{TX} = \frac{1}{2\pi \sqrt{L_1 C_1}} \tag{5.1}
\]

and an impedance matching circuit at the input of the reader antenna is designed to match it to a 50 \( \Omega \) feed source [10, 33, 54, 55]. Also, the previous attempts [33, 36, 54, 64] to design multi turn HF RFID coils with more than one turn to be used in the model of Figure 5.1(a) were carried out considering the \( f_{TX} < \text{self resonance frequency (SRF)} \) condition, where SRF is due to the parallel resonance of the inductance of the loop and the parasitic capacitance which is due to the electric field between the turns of wire which are at slightly different potentials. The reason for this condition is discussed at Section 5.2 where it is shown that a single loop HF RFID antenna is the best choice to be used with the old model of Figure 5.1(a).

A new model of inductively coupled HF RFID reader antennas is presented here based on the idea of transmitting at the SRF of a small multi turn HF coil, \( f_{TX} = \text{SRF} \), as shown in Figure 5.1(b). In this new model, the inherent capacitance between turns of the coil instead of the external lumped capacitor of the old model of Figure 5.1(a), is used to make resonance with the inductance of the coil i.e SRF. The new multi turn small self resonant coil (MT SSRC) antenna operating at its SRF is analysed mathematically in Section 5.3.

![Image](https://example.com/image.png)

**Figure 5.1** Circuit diagram of the reader antenna of the inductive coupled system as an inductor coil (a) operating below its self resonance frequency (b) operating at its self resonance frequency.
5.2 Mutli-Turn Small Coils for HF RFID Application

For a reader antenna, the circular shape coil/loop shown in Figure 5.2, is the best choice to excite a magnetic alternating field in an inductively coupled system because the contributions of the current carrying parts of the circular loop arrive in phase at the tag location, resulting in constructive interference. $R$, $N$ and $D$ represent the radius of the reader loop antenna, number of coil turns and the distance between tag-reader coils, respectively. From the size point of view, the coils/loops used in the inductively coupled RFID systems are non-standard antennas in which their total wire length is small compared to the wavelength of the frequency of operation, $f_{TX}$, (Total wire length $< 0.1 \lambda$). This is because if the total wire length becomes a considerable part of the wavelength, the coil can't be considered as a lumped element; and length increment will cause resonances at lower frequencies and the total field will be decreased [10- 36]. Based on previous literature, [10, 36] and [64-67], the effects of increasing $N$ on magnetic field intensity, $Q$ factor and self resonance frequency (SRF) of HF multi turn small coils are reviewed briefly here and it will be shown that the single loop (one turn coil) is the best choice to be used in model of Figure 5.1(a).
5.2.1 Magnetic Field

Based on the Bio-Savart law of magneto-statics the current can be assumed to be constant over the electrically small coils shown in Figure 5.2 so the amplitude of the magnetic field (H-field) in the direction perpendicular to the tag, at the distance $D$ is found as [68]:

$$
|\vec{H}_z| = \frac{NIR^2}{2\sqrt{(R^2 + D^2)^3}}
$$

(5.2)

It has been shown that for the loop with radius $R$ the magnetic field of $|\vec{H}_z|$ at a distance $D \leq R/\sqrt{2}$ from the centre is still around its maximal/optimal value [10, 69]. However, $|\vec{H}_z|$ is plotted as a function of distance in the $D$ direction from 1mm to 10 m in Figure 5.3. The current ($I=1$A) is constant in each case; the coils differ in radius $R$ and number of turns $N$. Based on (5.2) and Figure 5.3, in the case the coil is fed with a current source, it is tempting to think that a high $N$ will result in a high magnetic field at the tag location when $R$ is constant, but it is shown in [36] that more turns are only advantageous for smaller $D$ ($D < 1$m). The fact is that by increasing $N$, the total wire length of the loop which results cannot be regarded as small compared to the wavelength and for bigger $D$ the contributions of all parts of the loop at the tag location will not be in phase and therefore partial cancellation will occur over the loop. Also, when the coil is fed with a voltage source, calculations in [36] for constant $R$ results in maximum value of $|\vec{H}_z|$ when $N = 1$. Increasing $N$ will cause increased ohmic resistance as well, so the current in the loop wire would be even more limited than would be expected due to the skin effect. However, the $H$-field can be boosted by increasing the voltage or current of the feed source.

5.2.2 Q Factor

The performance of an HF RFID antenna is related to its quality factor ($Q$ factor). In general, the higher the $Q$ factor, the higher the power output for a particular sized antenna but too high a $Q$ factor may change the bandpass characteristics of the reader and the amplified oscillating can create problems in the protocol bit timing. For these reasons the $Q$ factor of the HF RFID antenna is normally decreased to a value around $15 \sim 30$ by adding a resistor, $R_Q$, parallel with the inductor as shown in Figure 5.1(a), depending upon the required band-width and thus upon the modulation procedure used [10, 33, 54]. The calculations of $R_Q$ are presented in [33] and an optimised system design based on numerical methods is presented in [54] for inductively coupled RFID systems with high power demand. The main problem for a HF RFID coil with more than one turn is that a high $Q$ factor will result [10, 33, 36, 54]. It is shown in [36] that the $Q$ factor of a two turn coil reader antenna is high in comparison with single turn coil as even at very small distances no communication can take place without using a resistor $R_Q$. 
5.2.3 Self Resonance Frequency (SRF)

When the perimeter of a conducting loop is about half a wavelength, the loop will resonate. If the loop has multiple turns, resonance occurs at a lower frequency due to the parallel resonance of the inductance of the loop and the parasitic capacitance (due to the electric field between the turns of wire which are at slightly different potentials). When designing a coil with \( N > 1 \) turns, to use in the model of Figure 5.1(a), the SRF of the coil should be checked to ensure it is higher than the reader operating frequency, i.e. \( f_{TX} < \text{SRF} \), because above its SRF the coil starts to behave as a capacitor and the Thomson equation of (5.1) can't to be used. SRF of HF coils are reviewed here and the formulas are used in the sections to follow. The capacitance between two turns of the coil at HF band is calculated as [65]:

\[
C_{tt} = \frac{2\pi^2 R\epsilon_0}{\ln\left(\frac{d}{2r} + \sqrt{\left(\frac{d}{2r}\right)^2 - 1}\right)} \quad (5.3)
\]

where \( R, r, d \) and \( \epsilon_0 \) are loop radius, wire radius, distance between the centres of the two wires and permittivity of free space, respectively. If the wire has an insulation with a dielectric constant of \( \epsilon_r \neq 1 \) the formula for coil at HF band is calculated as [66]:

\[
C_{tt} = \frac{2\pi^2 R\epsilon_0\epsilon_r}{\ln\left(1 + \frac{t}{r}\right)} \quad (5.4)
\]

where the insulation of adjacent turns is supposed to touch; without any air gap between them. The insulation thickness around the wire is represented by \( t \). Substituting \( \epsilon_r = 1 \) into (5.4) does not result in (5.3) as different approximations are made in [66] that led to (5.4).

The equivalent circuit of an inductor with multiple turns is shown in Figure 5.4. The equivalent capacitance is found as the series circuit of all turn-to-turn capacitances. As the capacitances between non-adjacent turns have a slight effect on the SRF, they are negligible and are not shown. However, for a uniformly wound HF coil, \( C_1 = C_2 = \ldots = C_N = C_n \) and \( L_1 = L_2 = \ldots = L_N = L_{1,N=1} = L_{1,N} / N \). \( L_{1,N} \) and \( C_n \) indicate the inductance (in the presence of the other turns) and capacitance between two turns, respectively. \( L_{1,N=1} \) is the inductance of a single turn in the absence of all other turns and is calculated as [67]:

\[
L_{1,N=1} = \mu_0 R\ln\left(\frac{8R}{r}\right) - 2 \quad (5.5)
\]

The SRF is calculated as [36]:

\[
\text{SRF} = \frac{1}{2\pi\sqrt{L_{1,N}C_n}} = \frac{1}{2\pi\sqrt{N L_{1,N=1}C_n}} \quad (5.6)
\]
5.3 Multi-Turn Small Self Resonant Coil (MT SSRC) at HF Band

Based on the discussion in Section 5.2 it is seen that the one turn coil ($N = 1$) is the best choice to use in the model of Figure 5.1(a), while the previous attempts, [10, 33, 36, 54, 64], to design multi turn HF RFID coils using this model considered $f_{TX} < $ SRF and used the Thomson equation to define operation frequency. Here, a new model as shown in Figure 5.1(b) is presented on the idea of transmitting at the SRF of the small HF coil ($f_{TX} = $ SRF) with more than one turn. Therefore the inherent capacitance between the turns of the multi turn coil is used instead of a lumped capacitor of model of Figure 5.1(a) to make self-resonance with the inductance of the coil. The coil is still considered to be small so the effect of increasing $N$ on magnetic field intensity is the same as in Section 5.2.1. Here, multi-turn small self-resonant coil (MT SSRC) antennas operating at their SRF are introduced and analysed mathematically at the HF band considering insulated and non-insulated wires in their structure in terms of SRF, number of turns, dimension and dielectric constant (where insulated wire is used). Also it is shown mathematically that because of the possibility of an electrical short circuit, a non-insulated structure can't be used and an insulated structure is required in order to get a practical structure for the MT SSRC antenna.

5.3.1 MT SSRC with Non-Insulated Wires

By replacing (5.3) and (5.5) at (5.6), knowing the fact that for transmission of the magnetic field at SRF, $\lambda_{SRF} = C/\text{SRF} = 1/\text{SRF} \cdot \sqrt{\varepsilon_0 \mu_0}$ where $c$ is speed of light in free space, the following equation is obtained for the wavelength of the SRF of the small HF coil having $N$ turns:

$$\lambda_{SRF} = \pi^2 R \sqrt{8 N \frac{\ln(\frac{8 R}{r}) - 2}{\ln\left(\frac{d}{2r} + \sqrt{\left(\frac{d}{2r}\right)^2 - 1}\right)}}$$ (5.7)
The total wire length of a compressed coil is well approximated as \(2\pi NR\). In order to get the following equations in terms of the electrical length of the coil, \(k = 2\pi NR / \lambda_{\text{SRF}}\) and it is obvious that for \(k < 0.1\) the antenna coil is small. However, replacing \(R\) with \(k\lambda_{\text{SRF}} / 2\pi N\) in (5.7) gives:

\[
\lambda_{\text{SRF}} = \frac{\pi r N}{4k} \exp \left(2 \left( \frac{d}{2r} + \sqrt{\left( \frac{d}{2r} \right)^2 - 1} \right) \right) \frac{N}{\pi r^2}
\]

(5.8)

Based on (5.8), \(\lambda_{\text{SRF}}\) of MT SSRC versus \(d/2r\) is plotted for coils with \(N = 2, 3, 16\) and \(r = 0.1, 1\) mm. The plots show \(k = 0.02\) and \(0.06\) in Figure 5.5 (a) and (b) respectively. It is seen that for \(k = 0.02\) and \(0.06\), which both result in small dimensions for the MT SSRC, \(d/2r\) is obtained around 1 with a maximum of 1.05 at the HF band. Therefore \(d = 2r\) would be a proper approximation for the dimension of the MT SSRC to have its SRF at the HF band considering the above specifications. The \(d = 2r\) condition forces the adjacent turns to be physically touching, resulting in an electrical short. Also, the maximum value of \(d = 1.05 (2r)\) as the distance between the turns results in a very sensitive structure susceptible to very small mechanical stresses.

![Figure 5.5](image)

**Figure 5.5** \(\lambda_{\text{SRF}}\) vs \(d/2r\) for non-insulated MT SSRC (a) \(k=0.02\), (b) \(k=0.06\).
5.3.2 MT SSRC with Insulated Wires

If the coil is made of insulated wire with a relative permittivity of $\varepsilon_r$ and a thickness of $t$ for the insulation, then by replacing equations (5.4) and (5.5) at (5.6) the $\lambda_{SRF}$ is obtained as:

$$\lambda_{SRF} = \pi^2 R \left( \frac{8 N \varepsilon_r}{\ln(\frac{8 R}{r}) - 2} \right) \ln\left(\frac{8 R}{r} - 2\ln\left(1 + \frac{t}{r}\right)\right)$$

(5.9)

As mentioned in Section 5.2.3, equation (5.9) is valid when the insulation of the two adjacent turns are in contact without any air gap between them. As in Section 5.3.1, $R$ is replaced in (5.9) with $k \lambda_{SRF} / 2 \pi N$ so $\lambda_{SRF}$ is obtained as:

$$\lambda_{SRF} = \frac{\pi r N \exp(2)}{4k} \left(1 + \frac{t}{r}\right)^{\frac{N}{2\pi r k^2}}$$

(5.10)

Based on (5.10), $\lambda_{SRF}$ of insulated MT SSRC versus insulation thickness of $t$ is plotted in logarithmic scale when $N$ is equal to 2 and 16; $r$ is fixed at 0.1 and 1 mm and $\varepsilon_r$ has values of 2 and 5 where $k$ is consequently equal to 0.02 and 0.06 at the HF band as shown in Figure 5.6. In this case, there is no possibility of an electrical short circuit due to the insulated turns. For example for the insulated MT SSRC having specifications of $N = 2$, $k = 0.06$ and $\varepsilon_r = 2$ when $r$ is 0.1 mm or 1 mm, the insulation thickness should be around $t = 0.07$ mm or 0.4 mm, respectively, to achieve an SRF at 13.56 MHz ($\lambda_{SRF} = 22 \text{m}$).

5.4 Summary

The previous attempts to design multi turn coils using the classic model of HF RFID reader antennas considered $f_{TX} < SRF$ and it was shown that a single loop HF RFID antenna is the best choice to be used in the classic model. A new model of inductively coupled HF RFID reader antennas was presented in this chapter based on the idea of transmitting at the SRF of a small multi turn coil as a HF RFID reader antenna, $f_{TX} = SRF$. The multi-turn small self-resonant coils (MT SSRCs) operating at their SRF were analysed mathematically at the HF band for both insulated and non-insulated structures. The analysis showed that non-insulated MT SSRC structure causes electrical shorts and also results in a very sensitive structures even to very small mechanical stresses. Therefore to get a practical structure, insulated MT SSRC structure should be chosen for HF RFID antennas. This chapter is a backbone to the next chapter where planar two turn SSRC (PTT SSRC) antennas having more uniform patterns, an optimised/controllable $Q$ factor, controllable dimensions and a stronger $H$-field in comparison with the previous model, are introduced.
Figure 5.6 $\lambda_{SRF}$ vs $t$ for insulated MT SSRC (a) $k=0.02$, (b) $k=0.06$. 
Chapter 6

6 Two Turn Planar SSRC (TTP SSRC) Antennas for HF RFID Applications - Novel Work Undertaken

To avoid an electrical short and also to obtain a similar structure with the current HF RFID reader antennas by using the analysis done in Chapter 5 as an approximated procedure, a compact two turn planar version of the insulated MT SSRC, i.e. a two turn planar SSRC (TTP SSRC), is proposed in this chapter. The dependency of the SRF on the antenna dimension will be observed. A TTP SSRC with \( f_{TX} = SRF \) has an optimised \( Q \) factor nearer to the specifications of the HF RFID standards (ISO /IEC 15693 and ISO/IEC 14443). It also has a near field (\( H \)-field) pattern with more uniformity in comparison with the single loop. It is shown that two adjacent tags affects the measured magnetic field by about 1 dB when placed 1cm distance form each other. At greater separations no effect on the measured magnetic field is observed. Finally, it is shown that the \( Q \) factor and dimension of a TTP SSRC antenna operating at 13.56 MHz for HF RFID can be controlled and adjusted based on the characteristics of the dielectric substrate. A number of TTP SSRC antennas are fabricated to verify this. The measured and simulated results of the TTP SSRC antenna show good agreement with the mathematical analysis from Chapter 5 and confirm that the TTP SSRC antenna prototypes have beneficial application to HF RFID smart shelf applications.
6.1 Two Turn Planar SSRC (TTP SSRC) Antennas at HF Band

As seen in Section 5.3.1, using non-insulated wires in the structure of a MT SSRC causes electrical short and a sensitive mechanical structure. To avoid this and also to have a similar structure with the current HF RFID reader antennas, a compact planar version of insulated MT SSRC antennas having two turns is proposed here based on the analysis of Section 5.3.2 as an approximated design procedure. Based on the results shown in Figure 5.6(b), an insulated MT SSRC antenna with $k = 0.06$, $N = 2$ and dielectric constant of $\varepsilon_r = 5$ for the insulation is considered as an example to be tried in a planar structure. For these specifications from Fig.6(b) it is seen that when $r = 0.1 \sim 1$ mm, the insulation thickness of $t$ should be at $0.2 \sim 2$ mm range to get the SRF at the HF band (3MHz ~ 30MHz) for the MT SSRC antenna with $N=2$. An FR4 substrate with a dielectric constant of $\varepsilon_r = 4.4$ and thickness of $h = 2t = 0.8$ mm, which are similar characteristics to those mentioned above for the insulation, is selected for the two turn planar SSRC (TTP SSRC) with $k = 0.06$. Because adjacent turns without any air gap are considered for the insulated MT SSRC, as mentioned in Section 5.3.2, the dielectric thickness of the substrate is approximated with the twice the insulation thickness i.e $h = 2t$. The TTP SSRC antenna structure is mounted as two circular copper strip loops having the same radius of $R$ and the same strip width of 10mm on the top and bottom of the FR4 substrate, as shown in Figure 6.1. In order to obtain a semi coil structure, the two loops are connected together by via hole, $a$, in the substrate while via $b$ connects the bottom loop to the antenna feed in the top layer. Figure 6.2 shows the input impedance of TTP SSRC antennas, which is simulated by HFSS, with radius $R$ values of 40, 50, 100 and 200 mm as a function of SRF. The corresponding $k$ values are provided, as calculated from the formula of Section 5.3.2. As seen in Figure 6.2 the real impedance peak value at the all of SRFs is about 6 k$\Omega$. Simulations not shown here depict that for $w = 5$ mm and 15 mm there is about 15% shift in the SRFs with the real impedance peak values of 8 k$\Omega$ and 5.5 k$\Omega$, respectively.

6.2 TTP SSRC Antenna Operating at 13.56 MHz

Following the above section, a TTP SSRC antenna having $k = 0.06$ was fabricated on FR4 substrate with $\varepsilon_r = 4.4$ and $h = 0.8$ mm, Figure 6.3(b), to operate at 13.56 MHz. As the real part of the impedance peaks at the SRF at around 7 k$\Omega$, this is too far from 50 $\Omega$, to match the antenna with a $PI$ or $T$ matching network as a very high $Q$ factor would result [70]. In order to solve this problem the operating frequency of 13.56 MHz is considered at the beginning of the real impedance curve when SRF = 14.5 MHz ($R = 90.2$mm) as shown in Figure 6.4. The impedance of the TTP SSRC antenna at 13.56 MHz is around 50+615j $\Omega$ when $SRF = 14.5$ MHz that the inductive part of 615j $\Omega$ is eliminated by just adding two series capacitors of 39 pF to match the antenna to 50 $\Omega$ line, as shown in Figure 6.3(b).
Also, based on the old model shown in Figure 5.1 (a), a planar single loop HF RFID antenna having the same radius of \( R = 90.2 \) mm with the TTP SSRC having \( k = 0.06 \) dimension is designed and fabricated to operate at 13.56 MHz. Following the Thomson formula by adding a 390 pF capacitor to the input of the single loop a resonance LC circuit is created to operate at 13.56 MHz and also a 50 \( \Omega \) matching circuit is designed as described in [9, 55] and illustrated in Figure 6.3(a). From the measurement and simulation results of return loss shown in Figure 6.5, the \( Q \) factor of both antennas are calculated using the \( f_{\text{centre}}/(f_{\text{high-3dB}} - f_{\text{low-3dB}}) \) expression and are obtained as 8 and 25 for TTP SSRC antenna and single loop antenna, respectively. It is seen that the \( Q \) factor of the TTP SSRC antenna following the new model of Figure 5.1(b), \( f_{RX} = SRF \), is nearer to the specifications of the standard 13.56 MHz systems at 15 ~ 30 range ([10, 33, 54], ISO 14443, ISO 15693).

![Figure 6.1](image1.png)  
**Figure 6.1** Geometry of the TTP SSRC antenna ; (a) Top and bottom layers, (b) Side view.

![Figure 6.2](image2.png)  
**Figure 6.2** Impedance characteristics of the TTP SSRC antenna for different values of \( R \).
Figure 6.3 Geometries of the (a) Planar single loop antenna (b) TTP SSRC antenna (both are designed to operate at 13.56 MHz and are matched to 50 Ω input).

* All of the capacitors are in the top layer

Figure 6.4 Impedance characteristics of the TTP SSRC antenna for $R = 90.2$ mm, Figure 6.3(b).
The relative magnetic field intensity for the single loop and the TTP SSRC antenna is measured at different fixed distances away from the antenna while the probe is adjusted to three different positions A, B and C, in the transverse direction. The fixed distances range up to 1 metre as shown in Figure 6.6. From Figure 6.6 (b) it is seen that the measured $H$-field is more uniform and has less changes in front of the TTP SSRC antenna in the three points than the single loop planar antenna in Figure 6.6 (a). It is seen than that the maximum difference of the measured $H$-field in the three points is changed from 8 dB in the old antenna to about 1 dB in the new antenna. Therefore tagged objects can be identified even with changing the position of the tag at the surface of the object and also the distance from the loop can be predicted with more accuracy in front of the TTP SSRC antenna. This is crucial for the smart shelf application such that it can determine the position of the item along the shelf wherever the tag is placed within the item, for example a book. This will be crucial in a reliable system ascertaining the order in which a set of books on a shelf are stacked. Also, it is seen that the measured relative magnetic field intensity of the TTP SSRC antenna is about 5 dB stronger than the single loop antenna in front of the three points at the constant distance levels that follows the discussion of Section 5.2.1.

As mentioned before in Section 2.4, if a conductor loop is located in the vicinity of the other one, through which a current is flowing, there would be a mutual inductance between the loops. The effect of the mutual inductance on the induced voltage is analysed in Section 2.6. Here, the effect of mutual inductance is measured in terms of the changes on the measured magnetic field intensity. In order to measure relative magnetic intensity when adjacent tags are placed in front of the TTP SSRC reader antenna, a set up as shown in Figure 6.7 is considered. Tag 1 is connected to
a spectrum analyser and placed at distance $M$ in front of the TTP SSRC antenna while tag 2 is placed at distance $P$ from tag 1, towards the reader antenna. In four individual measurements, tag 1 is fixed at $M = 10, 15, 20, 30$ cm and the position of tag 2 is changed in the $0 \sim 10$ cm range. The measurement results obtained by the spectrum analyser are shown in Figure 6.7 with about 1 dB difference for points A and C. It is seen that the adjacent tag 2 affects the measured magnetic field by $<1$ dB when is placed less than 1.5 cm distance from the tag 1 and it has no effect on the measured magnetic field far away from this distance. This could be used to identify many tagged objects in front of the reader antenna where the measurements are affected less than 1 dB due the mutual inductance between the tags together and also the reader antenna.

![Figure 6.6](image)

**Figure 6.6** The measured relative magnetic field intensity for the single loop antenna, (a), and the TTP SSRC antenna, (b), at different distances in front of three points of A, B and C, up to 1 metre distance.
6.3 Adjusting Q Factor and Dimension of TTP SSRC Antenna by Dielectric Characteristics of the Substrate

It is shown in [10] that the Q factor of the single loop is dependent on the loop radius $R$. In this section dependency of the Q factor of the TTP SSRC antenna to the dielectric characteristics of the substrate is investigated and it is shown that the Q factor of the TTP SSRC antenna can be controlled based on these characteristics. Seven TTP SSRC antennas all operating at 13.56 MHz are designed and fabricated on substrates having different dielectric characteristics. Figure 6.8 shows one of the measurement setups and the measured results are summarised in Table 6.1. It is seen that the dominant parameter to control the Q factor is loss tangent or dissipation factor of the dielectric which decreases the Q factor as it increases. This observation agrees with the effect for the lumped capacitors [71]. Also, the thickness of the dielectric affects the Q factor, for example the Q factor of the antennas fabricated on RT / Duroid 5880 and RT / Duroid 5870 have about the same loss tangent but different thickness. This could be explained as the converse relevancy between the volume between two conductor with different potentials and Q factor [65,66]. The dielectric constant also affects the TTP SSRC antenna dimension (dielectric length) and as is seen in Table 6.1 higher dielectric constant results in a smaller dimension or $R$. It is seen that loss tangent, dielectric thickness and dielectric constant all can be considered to control the Q factor and dimension of the TTP SSRC antenna operating in a defined frequency, 13.56 MHz.
Table 6.1 Measured $Q$ factor and dimensions of seven TTP SSRC antennas, all having the same operating frequency of 13.56 MHz, fabricated on substrates with different loss tangent, thickness and dielectric constant parameters.

<table>
<thead>
<tr>
<th>loss tangent (tanδ)</th>
<th>$h$ (mm)</th>
<th>$Q$</th>
<th>-10 dB BW (kHz)</th>
<th>$\varepsilon_r$</th>
<th>$R$ (mm)</th>
<th>Substrate</th>
</tr>
</thead>
<tbody>
<tr>
<td>$4 \times 10^{-4}$</td>
<td>0.25</td>
<td>670</td>
<td>6.5</td>
<td>2.2</td>
<td>82</td>
<td>RT / Duroid 5880</td>
</tr>
<tr>
<td>$5 \times 10^{-4}$</td>
<td>1.6</td>
<td>275</td>
<td>17</td>
<td>2.3</td>
<td>149</td>
<td>RT / Duroid 5870</td>
</tr>
<tr>
<td>$1.4 \times 10^{-3}$</td>
<td>0.5</td>
<td>265</td>
<td>20</td>
<td>3</td>
<td>93.3</td>
<td>Arlon /AD 300A</td>
</tr>
<tr>
<td>$2 \times 10^{-3}$</td>
<td>0.64</td>
<td>225</td>
<td>24</td>
<td>10.2</td>
<td>64</td>
<td>RT / Duroid 6010</td>
</tr>
<tr>
<td>$1.75 \times 10^{-2}$</td>
<td>0.8</td>
<td>25</td>
<td>120</td>
<td>4.55</td>
<td>92</td>
<td>FR4</td>
</tr>
<tr>
<td>$1.75 \times 10^{-2}$</td>
<td>1.6</td>
<td>20</td>
<td>130</td>
<td>4.55</td>
<td>112</td>
<td>FR4</td>
</tr>
<tr>
<td>$4.5 \times 10^{-2}$</td>
<td>1.6</td>
<td>16</td>
<td>260</td>
<td>4.2</td>
<td>110</td>
<td>Econoboard' CEM/1</td>
</tr>
</tbody>
</table>

Figure 6.8 Set up to measure the characteristic of the TTP SSRC antenna.

6.4 Summary

By using the analysis of multi-turn SSRC antennas in Chapter 5 as an approximated procedure in this chapter, a two turn planar SSRC (TTP SSRC) antenna was proposed in this chapter and the dependency of the SRF to the antenna dimension was observed. A TTP SSRC with SRF=13.56 MHz following the new model and a single loop antenna following the classic model having same dimension and operation frequency were fabricated and tested. It was shown that the new TTP SSRC antenna has an optimised $Q$ factor nearer to the current specifications in the HF RFID standards (ISO/IEC 15693 and ISO/IEC 14443) and also a near field ($H$-field) pattern with more uniformity was obtained for the TTP SSRC antenna in comparison with the single loop following the classic model. Also, it was shown that the mutual inductance between two adjacent tags affected the measured $H$-field less than 1 dB. Finally, it was shown that the $Q$ factor and dimension of a TTP SSRC antenna operating at 13.56 MHz could be controlled and adjusted based on the characteristics of the dielectric substrate where a number of TTP SSRC antennas
were fabricated. All the measured and simulated results of the TTP SSRC antenna showed good agreement with the mathematical analysis from Chapter 5. This confirmed that the TTP SSRC antenna prototypes have beneficial application to HF RFID smart shelf applications.
Chapter 7

7 Conclusions and Future Works

Recently, renewed interest and rapid progress has been made in the application of inductive coupling HF RFID methods to IoT applications. HF RFID smart shelf solutions as one of the major subdivisions of smart reader systems, which are building blocks for cutting edge applications of RFID, are beginning to be implemented for IoT purposes where the characteristics of the reader antennas are a critical issue.

In the first objective of this research, a theoretical approach was presented to increase the band-width of the HF RFID reader antennas. An antenna having about 6 times more band-width than similar antennas at 13.56 MHz was designed and fabricated using the approach to improve the reception of the sub-carrier RFID response. The increased band-width supports sub-carrier frequencies for the all the existing HF RFID standards (ISO/IEC 15693 and ISO/IEC14443) to be detected easier than the similar antennas and thus leads to increased range of identification. Also, it was shown that in the presence of book(s) as a sample material, the HF RFID technology is capable of identifying the distance of a tagged book based on the received magnetic field intensity ($H$-Field) for its capability in being used for smart shelf applications.

In the second objective, multi-turn small self-resonant coil (MT SSRC) antennas were introduced and mathematically analysed as a new model of HF RFID reader antennas. Based on the analysis, compact two turn planar SSRC (TTP SSRC) antennas having similar dimension with the current HF RFID reader antennas were investigated. As a sample, a TTP SSRC antenna operating at 13.56 MHz was fabricated and resulted in optimised $Q$ factor and more uniform near field pattern in comparison with the similar HF RFID antennas. Also, a number of prototype TTP SSRC antennas operating at 13.56MHz were fabricated on different substrates and it was shown that the desired $Q$ factor and the antenna dimension (Smart shelf dimension) can be obtained based on the dielectric characteristics of the antenna substrate.
This project considered low cost improved inductively coupled reader antennas to be used in practical smart HF RFID systems, and good results were obtained i.e. more band-width, controllable dimension of antenna, controllable/optimized $Q$ factor, more uniform pattern and also stronger $H$-field. More band-width will cause to better range for receiving backscatter signals from the tag, thereby increasing detection range in the smart shelf system. By controlling the antenna dimension, according to the dimension of the material to be detected, dimension of the smart shelf can be chosen. Also, as an HF RFID system has different identification ranges in the 13.56 MHz ISO/IEC standards, fixing the $Q$ factor in a desired value relevant to the chosen standard is important. Having a more uniform magnetic field pattern provides more degrees of freedom to change the position of the tag on the object's surface to be identified without error in a smart shelf system. Also narrower objects can be aligned in front of the antenna to be detected easily.

Some of the topics as a proposed future work are listed below:

**Future work 1:** The approach to increase band-width in the first objective can be tested for the other types of antennas operating at high frequency. For example, a wider band for the HF inductive coupled antennas used in biomedical systems, can support more data coming from the heart battery system.

**Future work 2:** Research on other types of inductively coupled antennas like the antennas used in wireless power transfer (WPT) systems and inductive charging. The procedures presented in this thesis can be tried in order to increase the efficiency of the wireless WPT systems in terms of transferring high amount of power.

**Future work 3:** Researches on other prototype implementations on other materials and also other measurements necessary with files not books. The measurement approach used in our research can be tried to assess the ability of HF RFID systems to identify objects consisting of different materials like foodstuffs at retail stores.

**Future work 4:** Using a sensitive detection circuit at the reader part in order to detect the position of the nearest and the furthest tag on a smart shelf would be a useful exploitation of this research as an HF RFID radio development. A step attenuator would be an appropriate low cost choice of implementation at the front end circuit of the reader antenna in such a detection system. Therefore a higher attenuation would detect the nearest tag and as it steps down progressively to lower attenuations, it would be possible that tags at further distances away would be gradually detected.

**Future work 5:** As the results of the TTP SSRC antenna, presented in the second objective, satisfy the analysis of MT SSRC antennas, it opens a research gate to multi-turn planar SSRC (MTP SSRC) antennas. As multi turn coil has more $Q$ factor in comparison with coils having fewer turns, research on MTP SSRC antennas to transfer high power near field would be useful.
References

[1] Commission of the European communities, “Communication from the commission to the European parliament, the council, the European economic and social committee and the committee of the regions - Internet of Things - An action plan for Europe” Brussels, 2009.


