Design and Optimization of Passive UHF RFID Tag Antenna for Mounting on or inside Material Layers

Dissertation

Presented in Partial Fulfillment of the Requirements for the Degree Doctor of Philosophy in the Graduate School of The Ohio State University

By

Shuai Shao, B.S., M.S.

Graduate Program in Electrical and Computer Engineering

The Ohio State University

2015

Dissertation Committee:
Professor Robert J. Burkholder, Advisor
Professor John L. Volakis, Advisor
Professor Fernando L. Teixeira
Abstract

There is great desire to employ passive UHF RFID tags for inventory tracking and sensing in a diversity of applications and environments. Owing to its battery-free operation, non-line-of-sight detection, low cost, long read range and small form factor, each year billions of RFID tags are being deployed in retail, logistics, manufacturing, biomedical inventories, among many other applications. However, the performance of these RFID systems has not met expectations. This is because a tag’s performance deteriorates significantly when mounted on or inside arbitrary materials. The tag antenna is optimized only for a given type of material at a certain location of placement, and detuning takes place when attached to or embedded in materials with dielectric properties outside the design range. Thereby, different customized tags may be needed for identifying objects even within the same class of products. This increases the overall cost of the system. Furthermore, conventional copper foil-based RFID tag antennas are prone to metal fatigue and wear, and cannot survive hostile environments where antennas could be deformed by external forces and failures occur. Therefore, it is essential to understand the interaction between the antenna and the material in the vicinity of the tag, and design general purpose RFID tag antennas possessing excellent electrical performance as well as robust mechanical structure.

A particularly challenging application addressed here is designing passive RFID tag antennas for automotive tires. Tires are composed of multiple layers of rubber
with different dielectric properties and thicknesses. Furthermore, metallic plies are embedded in the sidewalls and steel belts lie beneath the tread to enforce mechanical integrity. To complicate matters even more, a typical tire experiences a 10% stretching during the construction process. This dissertation focuses on intuitively understanding the interaction between the antenna and the material in the proximity and designing broad band and mechanically robust RFID tag antennas for elastic materials.

As a first step, the effects of dielectric materials on an antenna’s impedance match and radiation pattern are investigated. The detuning effect is quantified based on the theoretical frequency scaling and effective permittivity of a dielectric material of finite thickness. Using simple formulas, the operational range of a tag can be predicted without intensive full-wave simulations of different materials. Next, a spectral domain Green’s function is applied to compute the antenna pattern when the tag is mounted on or inside a layered medium. The optimal placement of the tag is found based on the focusing effect that the material has on the gain pattern of the antenna. For tires, the steel ply in the sidewall of a tire looks like a periodic wire grating. The performance of an antenna placed close to a wire grating is predicted using Floquet theory. The results indicate that steel plies embedded in the tire can be utilized as a reflector to further focus the gain pattern and increase the read range of a tag.

Using these design tools and theoretical analysis, several broadband RFID tag antennas are designed for multi-layered materials. A novel stretchable conductive textile (E-fiber) based tag antenna is also developed for placement in elastic materials. Prototype antennas are fabricated and embedded in a tire during the tire manufacturing
process. Experimental results indicate that tags with the new antennas achieve significant improvement compared with commercially available tags.
I am greatly indebted to my advisor Prof. Burkholder for many years of guidance, support and encouragement. I am motivated by his dedication to research and admire his intelligence and integrity. As an educator he does not only spend time exploring students’ potentials, encouraging them to be independent thinkers but also create the best environment for students to do research and finish their education successfully. Through past several years he has spent countless hours teaching and guiding me to transform from a student to a self-motivated researcher. I would also like to sincerely thank my co-advisor Prof. Volakis for his support in both my personal and academic life. At ESL he is distinguished by his endless energy and the pursuit of excellence. I learned a lot from him. Not just how to write good papers and make better presentations but also how to be a person, how to survive in the real world and how to achieve goals in my life and career.

I would also like to thank my committee member Prof. Teixeira for his review and suggestions on this work. I would like to thank my sponsor Bridgestone tires. This dissertation would not be possible without their support. I would like to thank my colleague Dr. Asimina Kiourti for textile-based antenna fabrications and her support in my research, papers and this dissertation. It has been a wonderful and awarding experience. I would also like to thank all my colleagues and friends at ESL, OSU for all the interesting discussions and companionship.
I reserve the last paragraph to thank my parents. Nothing would be possible without them.
Vita

September 8, 1987 . . . . . . . . . . . . . . . . . . . . . . . . . . Born - China

2010 . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . B.S. Electronic and Communication Engineering, City University of Hong Kong

2014 . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . M.S. Electrical and Computer Engineering, The Ohio State University

2010-present . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . Graduate Research Associate, The Ohio State University.

Publications

Research Publications


vii


**Fields of Study**

Major Field: Electrical and Computer Engineering
# Table of Contents

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>Abstract</td>
<td>ii</td>
</tr>
<tr>
<td>Acknowledgments</td>
<td>v</td>
</tr>
<tr>
<td>Vita</td>
<td>vii</td>
</tr>
<tr>
<td>List of Tables</td>
<td>xii</td>
</tr>
<tr>
<td>List of Figures</td>
<td>xiii</td>
</tr>
<tr>
<td>1. Introduction</td>
<td>1</td>
</tr>
<tr>
<td>1.1 What Is RFID?</td>
<td>1</td>
</tr>
<tr>
<td>1.2 RFID State of the Art</td>
<td>5</td>
</tr>
<tr>
<td>1.3 New RFID Tag Antenna Designed for Elastic Materials</td>
<td>6</td>
</tr>
<tr>
<td>1.4 Organization of This Thesis</td>
<td>8</td>
</tr>
<tr>
<td>2. RFID Tag Antenna Design Approach-Impedance Matching</td>
<td>12</td>
</tr>
<tr>
<td>2.1 Introduction</td>
<td>12</td>
</tr>
<tr>
<td>2.2 RFID Tag Antenna Formulas</td>
<td>14</td>
</tr>
<tr>
<td>2.3 Antenna Design</td>
<td>18</td>
</tr>
<tr>
<td>2.4 Experimental Results</td>
<td>22</td>
</tr>
<tr>
<td>2.4.1 Rubber Sample Test</td>
<td>22</td>
</tr>
<tr>
<td>2.4.2 On-Tire Experiment</td>
<td>25</td>
</tr>
<tr>
<td>2.5 Summary of Chapter 2</td>
<td>26</td>
</tr>
<tr>
<td>3. RFID Tag Antenna Design Approach-Antenna Pattern</td>
<td>28</td>
</tr>
<tr>
<td>3.1 Introduction</td>
<td>28</td>
</tr>
<tr>
<td>3.2 Antenna Radiation: Construction of the Equations</td>
<td>30</td>
</tr>
</tbody>
</table>
3.3 Computed Gain and Patterns .................................................. 35
3.4 Experimental Validation ..................................................... 38
  3.4.1 Antenna Design ......................................................... 40
  3.4.2 Measurement on Dielectric Slabs .................................. 41
  3.4.3 Measurement on the Tire ............................................. 42
3.5 Summary of Chapter 3 ....................................................... 43

4. RFID Tag in the Vicinity of a Wire Grating .............................. 44
  4.1 Introduction ................................................................. 44
  4.2 EM Theory of a Periodic Structure .................................... 45
    4.2.1 TM Polarization: dipole parallel to the wire grating ...... 46
    4.2.2 TE Polarization: dipole transverse to the wire grating ... 56
  4.3 Antenna Design ............................................................ 67
    4.3.1 Measurements and Results ....................................... 70
  4.4 Summary of Chapter 4 .................................................... 71

5. UHF RFID Tag Antennas for Automotive Tire Applications ............ 72
  5.1 Flexible and Stretchable UHF RFID Tag Antenna for Automotive
      Tire Sensing ................................................................. 72
    5.1.1 Tag Antenna Design Approach .................................. 75
    5.1.2 Experimental Results .............................................. 79
  5.2 RFID Tag Antenna for Mounting on Metallic Objects ................. 81
    5.2.1 Antenna Design Approach and Proposed Tag Antenna ...... 82
    5.2.2 Experimental Results ............................................. 86
  5.3 Summary of Chapter 5 .................................................... 87

6. Textile Based RFID Tag Antenna .......................................... 88
  6.1 Broadband Textile-Based Passive UHF RFID Tag Antenna for Elas-
      tic Material ............................................................... 88
    6.1.1 Proposed Tag Antenna ............................................. 90
    6.1.2 Textile Based Tag Antenna ...................................... 95
    6.1.3 Experimental Results ............................................ 96
  6.2 Investigation of Broadband Textile Based Antenna Configurations
      for Passive UHF RFID Tags ............................................. 100
    6.2.1 Antenna Design .................................................... 101
    6.2.2 Embroidered Antenna ............................................. 103
  6.3 Summary of Chapter 6 .................................................... 104

x
7. Contributions and Future Work ............................................. 106

7.1 Looking Ahead ............................................................... 107

Bibliography .......................................................................... 112
## List of Tables

<table>
<thead>
<tr>
<th>Table</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1 Dielectric Properties of Rubber Materials Under Test</td>
<td>24</td>
</tr>
<tr>
<td>2.2 Rubber Sample Threshold Power Test Results</td>
<td>24</td>
</tr>
<tr>
<td>2.3 Threshold Power Vs Thickness</td>
<td>24</td>
</tr>
<tr>
<td>2.4 On Tire Threshold Power Measurement Results</td>
<td>26</td>
</tr>
<tr>
<td>3.1 Threshold Power of the T-slot on the Slab</td>
<td>41</td>
</tr>
<tr>
<td>3.2 On-Tire Threshold Power Measurement Results</td>
<td>43</td>
</tr>
<tr>
<td>4.1 Threshold Power Vs Thickness for a T-slot Tag Mounted on the Side-wall of a Tire with and without a Rubber Pad Substrate.</td>
<td>71</td>
</tr>
<tr>
<td>5.1 Read Range of the RFID Tags on Dual-Wheel (ft)</td>
<td>80</td>
</tr>
<tr>
<td>5.2 Threshold Power of the Antenna Embedded in the Fixture</td>
<td>81</td>
</tr>
<tr>
<td>5.3 Threshold Power of the Tags Under Test</td>
<td>87</td>
</tr>
<tr>
<td>6.1 Read Range of the RFID Tags on Dual-Wheel (ft)</td>
<td>98</td>
</tr>
<tr>
<td>6.2 Threshold Power Test of the Designed Tag Fabricated with Different Material (dBm)</td>
<td>98</td>
</tr>
<tr>
<td>6.3 Stretch Test of the Copper Wire Tag</td>
<td>99</td>
</tr>
<tr>
<td>6.4 Threshold Power of the Tags Under Test</td>
<td>103</td>
</tr>
<tr>
<td>6.5 On-Tire Threshold Power Measurement Results</td>
<td>104</td>
</tr>
</tbody>
</table>
## List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1</td>
<td>RFID system is comprised by a computer, a reader, a reader antenna and RFID tags.</td>
<td>3</td>
</tr>
<tr>
<td>1.2</td>
<td>(a) Diagram of a passive RFID tag. (b) Backscattering modulation is achieved by switching the chip’s input impedance.</td>
<td>5</td>
</tr>
<tr>
<td>2.1</td>
<td>Equivalent circuit model of a passive UHF RFID tag in receiving mode.</td>
<td>15</td>
</tr>
<tr>
<td>2.2</td>
<td>A center-fed dipole is placed on a dielectric slab with relative permittivity $\epsilon_r$ and thickness h. (a) Electric field distribution. (b) Equivalent capacitance.</td>
<td>16</td>
</tr>
<tr>
<td>2.3</td>
<td>Effective relative permittivity of an infinite homogeneous dielectric medium vs. relative permittivity of a dielectric slab of varying thickness.</td>
<td>18</td>
</tr>
<tr>
<td>2.4</td>
<td>Structure of the modified end-loaded meander-line (ELML) dipole antenna for a passive UHF RFID tag.</td>
<td>19</td>
</tr>
<tr>
<td>2.5</td>
<td>Power reflection coefficient of the ELML in free space and on the slab with $\epsilon_r = 10.63$ and $h = 2$ mm.</td>
<td>19</td>
</tr>
<tr>
<td>2.6</td>
<td>Reflection coefficient of the ELML on a 2 mm dielectric slab with $3 &lt; \epsilon_r &lt; 13$.</td>
<td>20</td>
</tr>
<tr>
<td>2.7</td>
<td>Reflection coefficient of the ELML on the surface of a material with $\epsilon_r = 6$ and varying thickness.</td>
<td>21</td>
</tr>
<tr>
<td>2.8</td>
<td>Passive UHF RFID tags used in the experiments. (a) Fabricated ELML tag. (b) Squiggle tag from Alien Technology.</td>
<td>22</td>
</tr>
</tbody>
</table>
2.9 Calculated operational ranges of the MLA-type RFID tags. (a) Fabricated ELML. (b) Alien Squiggle®. ................................. 23

2.10 Cross-section of a truck tire. ..................................................... 25

3.1 Full wave simulation of the T-slot type RFID tag antenna gain, (a) in free space, (b) mounted on a finite dielectric slab residing in the xy plane. ......................................................... 29

3.2 A current source embedded in a dielectric slab with finite thickness. . 32

3.3 (a) Illustration of a short dipole placed on a dielectric slab of finite thickness. (b) Possible mode inside the slab that traps power. ....... 35

3.4 Peak gain of a short dipole on both half space vs the thickness of the dielectric slab ($\epsilon_r = 10.63$ and $\tan\delta = 0$), (a) $0 < \text{thickness} < 5\lambda_g$ (b) $0 < \text{thickness} < 0.5\lambda_g$. .................................................. 36

3.5 Peak gain of a short dipole on both half space vs the thickness of the dielectric slab ($\epsilon_r = 10.63$ and $\tan\delta = 0.163$), (a) $0 < \text{thickness} < 5\lambda_g$ (b) $0 < \text{thickness} < 0.5\lambda_g$. .................................................. 37

3.6 Peak gain in the half space above the slab vs the depth of placement. 38

3.7 Layout of the T-slot antenna. .................................................... 39

3.8 Impedance of the T-slot antenna mounted on a dielectric slab with $\epsilon_r = 10$. ................................................................. 39

3.9 Power reflection coefficient of the T-Slot antenna in free space and placed on the slab having $\epsilon_r = 6$ and 10. ......................... 39

3.10 Operational range of the tag (shaded region), (a) designed T-slot, (b) Alien Squiggle. ......................................................... 40

3.11 T-slot is placed on a dielectric slab. ........................................ 41

3.12 (a) Experiment Setup six positions are labeled on the sidewall of the tire. (b) The T-slot antenna is fabricated and connected to a RFID chip. 42

4.1 Radial metallic plies are embedded in the sidewall of a truck tire. . 45
4.2 Top view of the wire grating: the steel plies can be modeled as an infinite array of circular cylinders. .................................................. 46

4.3 TMz plane wave incident at angle $\theta$ on the wire grating, E field is parallel to the cylinders. .................................................. 47

4.4 TMz bistatic scattering vs frequency computed using the eigenfunction (101 modes) and MoM methods. The angle of the incident wave $\theta = 0$ degrees. The angle of the observation point is $\phi = 90$ degrees. ................. 50

4.5 Real and imaginary values of the integral $\frac{2}{k_y} \int_0^\pi \cos[ak_xn \cos \phi']e^{-jk_yn_x a \sin \phi'}d\phi'$ vs mode number n. $a = 0.75$ mm, $d = 2.5$ mm, $\theta = 45$ degrees. ......................... 53

4.6 Reflection and transmission coefficients of the $0^{th}$ order Floquet mode, (a) real part, (b) imaginary part. $a = 0.75$ mm, $d = 2.5$ mm, and $\theta = 45$ degrees. ......................... 54

4.7 Real and imaginary values of the coefficients vs the angle of the incident wave. $f = 915$ MHz, $a = 0.75$ mm, and $d = 2.5$ mm. (a) Reflection coefficient, (b) transmission coefficient. ......................... 55

4.8 TEz plane wave incident at angle $\theta$ on the wire grating, H field parallel to the cylinders. .................................................. 56

4.9 Geometry of the cylinder and self-term contour $\Delta C$ for two-dimensional MFIE. .................................................. 59

4.10 Scattering width $\sigma_{2-D}/\lambda$ computed using the eigenfunction and MoM methods. $a = 0.75$ mm. The angle of the incident wave is $\theta = 0$ degrees. The angle of the observation position is $\phi = 180$ degrees . . . 61

4.11 Real and imaginary value of the integrals vs mode number a) $\frac{2a}{\lambda}e^{j\psi} \int_0^{2\pi} [\hat{R}_{in} \cdot \hat{n}_n'] H_1^{(2)}(kR_{in})d\phi_n'$ in $Z_{ij}$, b) $\frac{2a}{\lambda}e^{j\psi} \int_0^{2\pi} [\hat{R}_{in} \cdot \hat{n}_n'] e^{-j\phi'} H_1^{(2)}(kR_{in})d\phi_n'$ in $Z_{ia}$, c) $\frac{2a}{\lambda}e^{j\psi} \int_0^{2\pi} [\hat{R}_{in} \cdot \hat{n}_n'] e^{j\phi} H_1^{(2)}(kR_{in})d\phi_n'$ in $Z_{ir}$. Here $a = 0.75$ mm, $d = 2.5$ mm, $f = 915$ MHz, and $\theta = 45^\circ$. ................. 64

4.12 Reflection and transmission coefficient of the H field scattered from the wire grating, (a) real part value, (b) imaginary part value. $a = 0.75$ mm, $d = 2.5$ mm, and the incidence angle $\theta = 45^\circ$. ................. 66
4.13 (a) Structure of the tire in the simulation. (b) Radiation pattern of a
dipole source in free space, maximum gain is 1.76 dB. .......................... 68

4.14 Radiation pattern of the dipole source placed transverse to the steel
plies (a) 3D radiation pattern (b) 2D radiation pattern $\phi = 90^\circ$ (Gain
$= 1.27$ dB $\theta = 90^\circ$). ....................................................... 69

4.15 Radiation pattern of the dipole source placed transverse to the steel
plies, (a) 3D radiation pattern, (b) 2D radiation pattern $\phi = 90^\circ$ (Gain
$= 9.1$ dB $\theta = 90^\circ$). ....................................................... 70

5.1 Tires on dual-wheel situation, existing tags cannot be detected when
mounted on sidewall S3 and S4 of the inner tire. ................................. 73

5.2 Cross section of a tire: a tire is composed of multiple layers of rubber,
whose thickness changes along the sidewall. ........................................ 74

5.3 Cu lines printed directly on the pre-stretched and relaxed rubber samples. 76

5.4 Simple meander line antenna fabricated with planar copper foil and
copper wire. .................................................................................. 76

5.5 (a) T-match structure: a center-fed smaller dipole connected to a larger
dipole. (b) Equivalent circuit schematic of the T-match connection. .. 77

5.6 Proposed flexible broadband RFID tag antenna with diamond shaped
loops. .......................................................................................... 78

5.7 Power reflection coefficient of the diamond shaped tag antenna in free
space and mounted on a rubber sample. ............................................. 79

5.8 Fixture used to simulate the environment of an embedded tag. ....... 80

5.9 Cross section of a truck tire. The crown region consists of two parts
1) tread region 2) carcass. ............................................................ 82

5.10 (a) Power reflection coefficient of the T-slot. (b) Radiation pattern of
the T-slot. The length of the antenna is 60 mm, the length of the slot
is 24 mm. .................................................................................... 84
5.11 (a) Power reflection coefficient of the T-slot. (b) Radiation pattern of the T-slot. The length of the antenna is 100 mm, the length of the slot is 80 mm.

5.12 Current distribution of the T-slot antenna (a) small T-slot (b) large T-slot.

5.13 The tags are tested on a fixture which consists of a 2 mm thick dielectric pad and a large metal plate.

6.1 Design of the proposed RFID tag antenna.

6.2 Power reflection coefficient between the RFID chip and antenna. The red curve is in free space and the blue curve is when the tag is mounted on a 4 mm thick material with $\varepsilon_r = 10, 13, 16$.

6.3 Equivalent circuit model for (a) the chip impedance, and for (b) the dipole antenna loaded with a short stub.

6.4 Power reflection coefficient of the designed antenna: (a) with vs. without end-loading, (b) non-stretched vs. stretched by 10%, (c) with original vs. elongated end-loading loops, and (d) with end-loading loops vs. patches.

6.5 Fabrication procedure and resulting flexible single E-fiber constructed RFID tag antenna.

6.6 Copper wire version prototype embedded in polymer.

6.7 (a) The surfaces of the dual-wheel are labeled as S1 to S4. (b) Six positions are labeled as P1 to P6 on the exterior surface.

6.8 The fabricated tag with textile antenna is embedded in the sidewall of a tire.

6.9 Structure of the sparse T-slot antenna.

6.10 Solid patch vs Sparse. (a) Power reflection coefficient. (b) Antenna pattern.

6.11 The fabricated RFID tag with sparse T-slot antenna.
6.12 Fabricated textile based T-slot antennas, (a) solid patch T-slot fabricated with double layer E-fiber, (b) sparse T-slot fabricated with triple thread E-fiber.

7.1 Proposed system architecture of RFID-Based sensor integration.

7.2 How tire manufacturers benefit from using RFID technologies.
Chapter 1: Introduction

RFID (Radio Frequency Identification) technology is rapidly becoming an indispensable solution for automatic item identification, assets tracking and passive sensing. Owing to its battery-free operation, non-line-of-sight detection, low cost, and small form factor, each year billions of RFID tags are being deployed in retail, logistics, manufacturing and biomedical inventories among others. However, key technical challenges need to be addressed before RFIDs can reach their full potential. In this work we study the tag antenna’s performance in the vicinity of dielectric materials and other challenging environments and design/optimize tag antennas based on our analysis.

This chapter introduces the technological aspects of RFID system, including hardware, standards and regulations. The advantages and challenges of implementing RFIDs will be discussed. The state of the art in RFID related research will be presented.

1.1 What Is RFID?

Radio Frequency IDentification (RFID) is a wireless communication technology which includes a set of protocols, frequency allocations, and hardware requirements. The advantage of RFIDs over standard UPC bar codes is that line-of-sight is not
required to read the tag information, and multiple RFID tags can be accessed simultaneously with a single reader. Further, besides product number, most commercial RFID tag chips are able to store information such as manufacturing data, and end to end history. These information could be accessed via the Internet, allowing for real-time item tracking from anywhere in the world which could increase inventory tracking efficiency and reduce labor costs. RFID is therefore a major enabler in the Internet of Things.

As opposed to other popular wireless technologies such as WIFI, Bluetooth, GSM, etc, RFIDs were designed to have low data rate which reduces the chip’s power consumption. Passive RFID tags have no internal power source, they depend on the incoming signal to power the circuitry. The tag antenna is used for both communication and wireless power harvesting. Therefore a good tag antenna design is of paramount importance in RFIDs.

As shown in Fig. 1.1, the RFID system is comprised by a computer, one or more RFID readers, RFID reader antennas and RFID tags. The computer controls each RFID reader which sends out a signal through the reader antenna. An RFID tag modulates the signal with a code unique to the tag number. The modulated reflected signal is picked up and decoded by the reader. The data are then sent back to the host computer. In such a system, there would be one or multiple readers and numerous RFID tags attached to the objects to be identified.

RFID systems use frequencies varying from around 100 KHz to over 5 GHz. Instead of operating across the whole spectrum, RFIDs are mostly activated in some fairly narrow bands. The most commonly encountered frequency bands are 125/134 KHz, which is in the low-frequency band (LF), 13.56 MHz, which is in the high
Figure 1.1: RFID system is comprised by a computer, a reader, a reader antenna and RFID tags.

frequency band (HF), 860-900 MHz and 2.4-2.45 GHz, which are both in the ultra-high-frequency band. Usually the readers and tags operating in the 900 MHz region are referred to as UHF devices, while 2.4 GHz systems are known as microwave devices. The selection of the operating frequency bands is based on the application of the RFID system, and determines not only how much information one can send over a link but also the selection of the devices such as readers and tags. Therefore the selection of the bands is closely related to the performance and cost of the system. UHF RFID systems have a relatively longer read range and a moderate memory at a small form factor. They have been widely used for inventory checking and asset
tracking in warehouses, storerooms, manufacturing plants and sales floor. In this work, we focus on designing UHF RFID tag antennas.

RFID tags can be categorized into three types based on the source of power and the existence of transmitter. A passive tag does not have an internal power source, or a transmitter. It depends on the power transmitted from the reader to operate its circuitry. A semi-passive tag has a battery, but no transmitter. An active tag has both an internal power source and a transmitter such that it can be configured as a conventional bidirectional radio communication device. As passive tags have the simplest structure and lowest cost, they are used widely. In this work, we will also focus on the passive UHF RFID tag antenna design.

Fig. 1.2 (a) shows the diagram of a passive RFID tag. The tag consists of an antenna and a microchip. The tag chip extracts power to operate the circuitry. As a passive tag chip does not have a transmitter, signal modulation is achieved by switching the chip’s input impedance, as shown in Fig. 1.2 (b). Specifically, the chip’s input impedance \( Z_c \) is switched between a value which matches the antenna impedance \( Z_a \) and some other impedance which makes an open circuit presented to the antenna.

The tag antenna’s radar cross section (RCS) can be computed as [1, 2]

\[
\sigma_i^{tag} = \frac{\lambda^2}{4\pi} G_{tag}^2 |1 - \Gamma_i|^2, \quad i = 1, 2
\]  

(1.1)

here, \( G_{tag} \) is the gain of the tag antenna, and \( \Gamma_i \) is the reflection coefficient between the terminals of the antenna and the chip. The power received by the reader antenna is

\[
P_{rd}^r = \frac{P_{rd}' G_{rd}^2 G_{tag}^2 \lambda^4}{(4\pi R)^4} |1 - \Gamma_i|^2
\]  

(1.2)
where $P_{rd}$ is the output power of the reader and $G_{rd}$ is the gain of the reader antenna.

Eq. (1.2) illustrates that a change in the tag antenna’s RCS provides a significant difference in the received signal.

### 1.2 RFID State of the Art

RFID research to date can be summarized into three categories: 1) Antennas and propagation: this includes reader and tag antenna and antenna array design [3, 4], antenna measurement techniques [5, 6], UWB based RFIDs [7] and chipless RFIDs [8, 9]. 2) Communication and protocols: this includes novel modulation schemes [10, 11], security [12, 13], and anti-collision algorithms [14, 15]. 3) Circuits and IC design: this includes RFIC design [16, 17], RF front-end architectures [18, 19] and non-silicon circuit design [20]. RFID based localization [21, 22] and RFID integrated sensors [23, 24] have also been actively researched. They are the driving force of all aforementioned technical developments. This work is focused on enhancing the performance of passive RFID systems by designing novel RFID tag antennas.
1.3 New RFID Tag Antenna Designed for Elastic Materials

Passive UHF RFID tag antennas have been researched extensively for inventory tracking and sensing in a diversity of applications and environments. Current researches on tag antenna designs mostly address the problem of mounting tags on extreme environments, such as on metallic objects and bio-tissues [25, 26]. However, the performance of existing RFID tags has not met expectations. This is because a tag’s performance deteriorates significantly when mounted on or inside arbitrary materials. The tag antenna is optimized only for a given type of material at a certain location of placement. Therefore, detuning takes place when the antenna is attached to or embedded in materials with dielectric properties outside the design range. Thereby, different customized tags may be needed for identifying different objects that might even belong to the same class of products. This increases the overall cost of the system. Furthermore, conventional copper foil-based RFID tag antennas are prone to metal fatigue and wear, and cannot survive hostile environments where antennas could be deformed by external forces and failures occur. Therefore, it is essential to understand the interaction between the antenna and the material in the vicinity of the tag, and design general purpose RFID tag antennas possessing excellent electrical performance as well as robust mechanical structure.

A particularly challenging application addressed here is designing passive RFID tag antennas for automotive tires. Tires are composed of multiple layers of rubber with different dielectric properties and thicknesses. Furthermore, metallic plies are embedded in the sidewalls and steel belts lie beneath the tread to enforce mechanical integrity. To complicate matters even more, a typical tire experiences a 10% stretching during the construction process. This dissertation focuses on intuitively
understanding the interaction between the antenna and the material in the proximity, and designing broadband and mechanically robust RFID tag antennas for elastic materials. The key contributions of this work are:

- Investigated the effect of dielectric materials on an antenna’s impedance match. The detuning effect is quantified based on theoretical frequency scaling and the effective permittivity of a dielectric material of finite thickness. Using simple formulas, the operational range of a tag can be predicted without intensive full-wave simulations of different materials.

- Investigated the effect of dielectric materials on an antenna’s radiation pattern. Applied the spectral domain Green’s function to compute the antenna pattern when the tag is mounted on or inside a layered medium. Found the best placement of the tag that extends the tag’s read range.

- Analyzed the performance of a tag antenna placed close to a wire grating using Floquet theory. Utilized the steel plies embedded in the tire as a reflector to further focus the gain pattern and increase the read range of a tag.

- Designed broadband and mechanically robust copper wire-based tag antenna and metal mountable antenna for automotive tire applications.

- Designed broadband flexible textile-based antennas for elastic materials. Proposed novel textile-based antenna design approach.

It is worth pointing out that even though this work is focused on designing tag antennas for RFID applications, the proposed design approach and analysis could also be applied to antenna design for several other applications.
1.4 Organization of This Thesis

This dissertation is organized as follows.

Chapter 2 investigates the effect of surrounding materials on antenna performance, with a focus on antenna tuning. It is well known that an antenna will be detuned with the presence of dielectric media in close proximity. In this chapter, this detuning effect is quantified based on theoretical frequency scaling and the effective permittivity of a dielectric material. Using the derived formulas, one can compute the operation frequency and bandwidth of an RFID tag antenna given a range of surrounding medium parameters and thicknesses. Thus intensive full-wave simulations are not required. Also, the operational range of a tag can be predicted using the proposed analysis, allowing one to select tags applicable to materials having different electrical and physical properties. An end-load meander-line dipole RFID antenna is designed and tested as an example to demonstrate the proposed design approach. The experimental results indicate that the designed antenna is more robust to different media as compared to commercially available general purpose RFID tag.

Chapter 3 discusses how the antenna pattern is affected by the materials at proximity. It is observed that, even when the antenna is matched, its broadside gain is reduced significantly when the tag is placed on the top surface of a dielectric slab. This is because the presence of dielectric material alters the antenna’s pattern by redistributing the radiated power. In this chapter, the spectral domain Green’s function is used to analyze the antenna’s performance when the tag is mounted in or on layered media. The theoretical result illustrates that the decrease of broadside gain is mainly due to certain modes activated inside the dielectric layer. As a result, a substantial amount of power is confined in the near zone and does not radiate. A
broadband T-slot antenna is designed and tested as an example to demonstrate the proposed analysis. Experimental results validate the theory. The optimal location of placement is found, which further increases the tag’s read range.

Chapter 4 studies the antenna’s behavior when the tag is close to a wire grating. This question is inspired by designing RFID tag antennas for automotive tires. Steel plies are embedded in the truck tires to increase their mechanical integrity. As the plies are in radial direction, they can be modeled as an array of PEC cylinders within a section on the tire sidewall. To study a short dipole radiation with the presence of the wire grating, a theoretical analysis is performed in 2D. The problem is addressed in TM and TE polarization respectively. In each polarization, wave scattering from a thin cylinder is first derived using the Method of Moments (MoM) solution and compared with the eigenfunction solution. Then the plane wave reflection and transmission coefficients are computed for an infinite periodic array of thin cylinders. Free space Green’s function is replaced with its plane wave spectral expansion. The periodicity constraint is imposed to reduce the solution to a sum of Floquet modes. The theoretical results indicate that the wire grating acts as a solid PEC plate and free space when the antenna is placed parallel and transverse to the cylinder array, respectively. A T-slot antenna is optimized to validate the theory. Experimental results indicate that the wire grating can be utilized as a reflector to increase antenna gain. In doing so, the tag’s read range is further increased.

Chapter 5 presents RFID tag antennas designed for automotive tire applications. In order to improve vehicle safety, RFID tags may be mounted on or inside the tires to monitor their condition (e.g., pressure and temperature) and provide history data.
In this chapter, two kinds of tag antennas are designed based on the location of placement. For the tag embedded in the sidewall, the antenna is designed and optimized based on the design approach proposed in Chapter 2 and 3. Special attention is given to making the antenna mechanically flexible and durable. The designed antenna exhibits broad impedance bandwidth, which enables it to maintain tuned when placed on dielectrics of varying permittivity and thickness. The second half of the chapter discusses RFID tags embedded in the crown region. Because a steel belt is embedded in this region, the presence of metallic objects on an antenna’s performance is studied. It is found that with the inherent broadband characteristic, the T-slot antenna functions well even when placed close to a PEC plate. The simulation result illustrates that the T-slot antenna operates in two different modes, and maximum gain is exhibited in the second mode. The optimized T-slot antenna is fabricated and connected to the RFID chip. The experimental result shows that the designed tag achieved exceptional read range when mounted 2 mm above the PEC plate.

Chapter 6 introduces RFID tag antennas designed and fabricated with conductive textile material. As RFID tags or RFID integrated sensors would be mounted in hostile environments, it is essential to design tag antennas that are highly flexible and stretchable. In doing so, the tag would survive in challenging surroundings. Conductive textile (E-fiber) is a perfect candidate for such applications due to its light weight, high strength and flexibility. The first half of this chapter presents a broadband textile-based passive UHF RFID tag antenna for elastic materials. The antenna delivers a bandwidth of 263 MHz (free space) for the purpose of maintaining performance over a wide range of materials. The antenna is fabricated using a single E-fiber thread which makes the tag less prone to fatigue and wear, and favors its
application in elastic materials. The second half of the chapter discusses redesigning conventional solid patch antennas for fabrication using the minimum amount of textile threads. With less conductive material, the cost of the antenna is reduced, and flexibility of the antenna is increased. A T-slot antenna was fabricated with both solid surface and multiple thread E-fiber. Experimental results indicate that the multiple thread antenna achieved equally good performance as its solid surface counterpart.

Chapter 7 concludes the dissertation with a summary of major contributions and a potential research of combining RFID with sensor technologies. Future research regarding RFID technology will also be discussed.
Chapter 2: RFID Tag Antenna Design Approach-Impedance Matching

Perhaps the most important characteristic of a passive RFID tag is the read range – the maximum distance an RFID reader is able to detect the modulated backscattered signal and successfully identify a tag. The read range may be roughly calculated using the Friis transmission formula,

\[ R_{\text{max}} = \frac{\lambda}{4\pi} \sqrt{\frac{P_t G_t G_r (1 - |\Gamma|^2)}{P_{th}}} \]  \hspace{1cm} (2.1)

where \( \lambda \) is the wavelength, \( P_t \) is the transmitted power, and \( G_t \) and \( G_r \) are the gains of the reader antenna (transmitter) and tag antenna (receiver), respectively. \( P_{th} \) is the minimum threshold power which can activate the circuitry. \( \Gamma \) is the voltage reflection coefficient. Here, it is assumed that \( G_r \) is the gain of the tag antenna under perfect matching conditions (e.g., about 1.5 dBi for a short dipole antenna). As would be expected both \( \Gamma \) and \( G_r \) are affected by the presence of dielectric materials around the RFID tag.

2.1 Introduction

Passive UHF RFID tag antennas have been researched extensively for inventory tracking and sensing in a diversity of applications and environments. It is well known
that the presence of unknown materials can detune the antenna, thereby degrading its performance, and reducing read range. Generally speaking an RFID tag antenna is designed for a given type of material to optimize its impedance match, gain and efficiency. Examples include metallic surface mountable tags [27, 28, 29], tags designed for near body applications [25], tags for bottled water [30], paper-based applications [31], and tags for specific products [32]. However, a tag antenna designed for a specific application (mounted on a certain material) may not perform well when placed upon or near other materials, even within the same class of products. This is mainly because antennas have limited bandwidth, as a result detuning takes place when the antennas are attached to or embedded in materials with dielectric properties outside the design range. The thickness of the material also affects the tuning. Hence, it is desirable to design a general purpose tag antenna with sufficient bandwidth to operate on a plurality of materials for ubiquitous applications.

There has been a strong interest to develop broadband and multi-band passive UHF RFID tag antennas for various purposes. Dual-band and multi-band tag antennas have been designed to cover different frequency bands [33, 34] because authorized frequency allocations vary from country to country: 902-928 MHz in North America, 865-868 MHz in Europe, 920-925 MHz in China. However, if these antennas are designed for a certain material, they would still be detuned when the material changes. Cho et al. [35] designed a tag antenna with a bandwidth of 8.5%. However, the antenna was not fabricated and tested to verify its material insensitive characteristics. Xi et al. [36] designed and tested a wide band tag antenna with a bandwidth of 9.3%. However, the dielectric properties and thicknesses of materials under test were not given. Thus, it is hard to tell whether the tag operates well on objects with high
permittivity. Tentzeris [37] studied the permittivity of different kinds of paper and designed tag antennas accordingly. However, based on our research, it is found that there may not be an observable change in the tag’s performance when the thickness of the medium $h << \lambda$, and its permittivity is in the range from 1 to 4.

In this work, we first deduce the frequency band requirement for an RFID tag antenna that can operate upon a given range of material parameters and thicknesses. This is done based on frequency scaling and the effective permittivity of a dielectric slab. Then, a material-insensitive tag antenna is proposed based on existing commercially available tag designs. The modified antenna is fabricated and tested. Experimental results show that the new tag antenna achieved much better performance on dielectric media, with medium to high permittivity and varying thickness, than existing general purpose RFID tags.

2.2 RFID Tag Antenna Formulas

A passive UHF RFID tag is composed of an antenna connected to an IC chip. The equivalent circuit of a tag is shown in Fig. 2.1. The antenna matching is quantified in terms of the voltage reflection coefficient $\Gamma$ between the chip and antenna terminals given by (2.2) [38],

$$\Gamma = \frac{Z_c - Z_a^*}{Z_c + Z_a} \quad (2.2)$$

where $Z_c$ and $Z_a$ are the complex impedances of the RFID chip and antenna, respectively. To achieve maximum read range, the antenna impedance should be conjugate-matched to the chip impedance, $Z_c \approx Z_a^*$. In doing so $\Gamma \approx 0$ over the band of operation, and maximum RF power is transferred to the chip.
Generally speaking, an RFID tag antenna is designed for operation in free space or to be mounted on a specific surface. However, in the real world RFID tags are often attached to or embedded in objects with unknown and widely varying dielectric properties. As pointed out in [39], for a metallic antenna immersed in an infinite dielectric material, the input impedance as a function of frequency scales according to,

\[ Z_a(f; \epsilon_r) = Z_a(f \sqrt{\epsilon_r^2/\epsilon_1}; \epsilon_1) \sqrt{\epsilon_r/\epsilon_1} \]  

(2.3)

where \( \epsilon_1, \epsilon_2 \) are the (real) relative permittivity of the dielectric media 1 and 2 respectively. We see that both the frequency and amplitude are scaled. Eq. (2.3) illustrates that if the impedance curve of an antenna immersed in a dielectric medium with permittivity of \( \epsilon_r^1 \) is known, then the modified impedance curve can be predicted when the antenna is immersed in any other dielectric media with known permittivity \( \epsilon_r^2 \). Using (2.2) and (2.3) together, one may calculate the shift of an antenna’s operational band in frequency when subjected to miscellaneous materials.

Eq. (2.3) is based on the assumption that an antenna is immersed in an infinite homogeneous material, but for most RFID applications tags are attached to the
surface of objects with finite thickness. The problem can be addressed by defining an equivalent effective homogeneous permittivity for a given thickness. In the case where antennas are placed on the surface of a dielectric slab, the average permittivity at the interface can be calculated as:

\[ \epsilon_e = \frac{\epsilon_r + 1}{2}. \]  

(2.4)

However, it has been found that the frequency shift of \( \Gamma \) calculated by (2.4) and used in (2.2)-(2.3) is significantly larger than simulation results indicate. Abbosh [40] derived a more accurate formula for calculating the effective permittivity seen by a centered-fed dipole mounted on a dielectric slab. As shown in Fig. 2.2, a center-fed dipole with length \( l_d \) is placed on a dielectric slab with thickness \( h \) and permittivity \( \epsilon_r \).

![Figure 2.2: A center-fed dipole is placed on a dielectric slab with relative permittivity \( \epsilon_r \) and thickness \( h \). (a) Electric field distribution. (b) Equivalent capacitance.](image)

16
The gap between the two arms of the dipole is $s$. By considering the electric field distribution and associated parallel capacitances, the effective relative permittivity is obtained as,

$$
\epsilon_e = \alpha \left( \frac{\epsilon_r - 1}{2} \right) + 1
$$

(2.5)

where $\alpha$ is a unitless coefficient,

$$
\alpha = \frac{K(k_2)K'(k_1)}{K'(k_2)K(k_1)}.
$$

(2.6)

Here, $K(k_i)$ is the complete elliptical integral of the first kind, and $K'(k_i) = K \left( \sqrt{1 - k_i^2} \right)$. $k_1$ and $k_2$ are given by,

$$
k_1 = \frac{l_d - s}{l_d + s},
\quad
k_2 = \frac{\sinh \left[ \frac{\pi (l_d - s)}{4h} \right]}{\sinh \left[ \frac{\pi (l_d + s)}{4h} \right]}.
$$

(2.7)

Note that as $\alpha \rightarrow 1$, the effective permittivity approaches the result of (2.4) which is what we expect for a tag mounted on a semi-infinite half-space. Fig. 2.3 plots the effective permittivity vs. the permittivity of the dielectric slab for thicknesses up to two times the length of the dipole. Eq. (2.4) is valid if the thickness of the dielectric slab is equal to or larger than this value (red line).

As RFID tags often incorporate dipole-based structures, (2.5) can be applied to tag antenna design. Eq.(2.4) transforms a dielectric slab with finite thickness into an infinite homogeneous medium with an effective permittivity. Thereby using (2.3) and (2.2) we may estimate the bandwidth requirement for a RFID tag antenna applicable to a variety of materials and different thicknesses, without extensive simulations.
2.3 Antenna Design

An end-load meander-line (ELML) dipole antenna is used as an example to demonstrate the design approach. As shown in Fig. 2.4, the ELML antenna incorporates structures of existing classes of antennas to take advantage of their wide bandwidth, small size, low cost, tunability, and mechanical flexibility. The Squiggle® tag manufactured by Alien Technologies [41] is a popular general purpose passive UHF RFID tag of this same basic layout.

The power reflection coefficient of the RFID tag shown in Fig. 2.4 is plotted in Fig. 2.5. The plot shows that ELML has a wide bandwidth of 220 MHz with $|\Gamma|^2$ null
Figure 2.4: Structure of the modified end-loaded meander-line (ELML) dipole antenna for a passive UHF RFID tag.

Figure 2.5: Power reflection coefficient of the ELML in free space and on the slab with $\epsilon_r = 10.63$ and $h = 2$ mm.

at 1.23 GHz in free space. The center frequency of the US RFID band is 915 MHz, so we are relying on the derived formulas above to shift the $|\Gamma|^2$ null when the tag is placed on or in the dielectric. As shown in Fig. 2.5, when the antenna is placed on the surface of a dielectric with $\epsilon_r = 10.63$ the null is shifted down to 0.93 GHz. More importantly, the antenna maintains a wide bandwidth of 205 MHz.

Simulations were conducted using Ansys HFSS to fine-tune and evaluate the performance of the designed antenna mounted on the surface of a 2 mm thick dielectric slab with varying permittivity. Theoretical results are computed using (2.2), (2.3)
and (2.5) based on the impedance curve of the antenna placed on the dielectric slab with $\varepsilon_r = 6$. The impedance of the chip is $16 - j148\Omega$ and it is hereafter assumed to be constant. Fig. 2.6 compares the computed and theoretical reflection coefficient $\Gamma(f)$ for the ELML. The curves are shifted downward in frequency as the permittivity of the dielectric slab increases. The theoretical calculations predict this behavior, although the total shift is slightly small. It is also seen that the minima of the curves increase as the permittivity increases. This is due to the amplitude scaling factor of the antenna impedance in (2.3), which affects how well the antenna stays matched to the chip via (2.2).
Figure 2.7: Reflection coefficient of the ELML on the surface of a material with $\epsilon_r = 6$ and varying thickness.

Fig. 2.7 plots the HFSS simulation and theoretical reflection curves for the ELML on a dielectric slab with $\epsilon_r = 6$ and varying thickness. For the theoretical curves, the effective relative permittivity is calculated from (2.5) and used in (2.2) and (2.3) for the theoretical curves, starting again with the curve for the 2 mm slab. The plot shows that as the thickness of the dielectric slab increases, the curves are shifted downward in frequency. This is consistent with Fig. 2.3 and (2.5) which predict that the effective dielectric increases with thickness.
2.4 Experimental Results

The antenna prototypes were fabricated with copper foil and connected to G2XM RFID chips [42]. All experiments were conducted in a large indoor RFID laboratory. Tag performance was quantified by adjusting the output power of the reader at a fixed distance to determine the minimum threshold power for detecting the tag. An Alien Squiggle® tag was tested under the same conditions to serve as a benchmark [41].

Figure 2.8: Passive UHF RFID tags used in the experiments. (a) Fabricated ELML tag. (b) Squiggle tag from Alien Technology.

Applying (2.2), (2.3) and (2.5), the calculated operational ranges of the fabricated ELML tag (red) and the Alien Squiggle® tag (blue) are plotted in Fig. 2.9. It is observed that the ELML tag is tuned for thicker dielectrics with higher permittivity than the Squiggle®. This is verified in the following experimental results.

2.4.1 Rubber Sample Test

The application of interest here is to mount RFID tags on or inside truck tires. Tires are composed of multiple layers of rubber of varying thickness, permittivity and
loss tangent. The dielectric properties of rubber samples common in truck constructions are measured using a material analyzer and recorded in Table 2.1. Each rubber sample is a 20 by 20 cm square with a thickness of 2 mm.

The tags of Fig. 2.8 were first tested on these rubber samples. The threshold power is recorded for each combination in Table 2.2. A lower threshold power indicates a longer possible read range, hence the better performance of the antenna.

It can be seen that the commercial Alien Squiggle® tag is well tuned for low permittivity applications. However, as the permittivity of the sample increases, the
Table 2.1: Dielectric Properties of Rubber Materials Under Test

<table>
<thead>
<tr>
<th>Material</th>
<th>$\epsilon_r$</th>
<th>Tan$\delta$</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>3.66</td>
<td>0.005</td>
</tr>
<tr>
<td>B</td>
<td>6.66</td>
<td>0.046</td>
</tr>
<tr>
<td>C</td>
<td>10.63</td>
<td>0.162</td>
</tr>
<tr>
<td>D</td>
<td>13.12</td>
<td>0.303</td>
</tr>
</tbody>
</table>

Table 2.2: Rubber Sample Threshold Power Test Results

<table>
<thead>
<tr>
<th>Material (2mm)</th>
<th>free space</th>
<th>A</th>
<th>B</th>
<th>C</th>
<th>D</th>
</tr>
</thead>
<tbody>
<tr>
<td>ELML</td>
<td>23 dBm</td>
<td>23 dBm</td>
<td>21 dBm</td>
<td>22 dBm</td>
<td>23 dBm</td>
</tr>
<tr>
<td>Squiggle</td>
<td>19 dBm</td>
<td>19 dBm</td>
<td>19 dBm</td>
<td>23 dBm</td>
<td>29 dBm</td>
</tr>
</tbody>
</table>

tag’s performance deteriorates. The ELML tag is designed to work on a wide range of materials. The performance is not quite as good as the Squiggle for low permittivity materials, but is fairly consistent over all the materials. It is expected that the ELML antenna could achieve lower threshold power with some additional fine-tuning.

Using the same experiment setup the effect of thickness of the dielectric media is investigated. The threshold power was measured and tabulated in Table 2.3 for the ELML and Squiggle tags mounted on material A with thickness of 2 mm and 8 mm. It can be seen from the table that the Alien tag has better performance on the

<table>
<thead>
<tr>
<th>Thickness</th>
<th>2mm</th>
<th>8mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>ELML</td>
<td>23 dBm</td>
<td>22 dBm</td>
</tr>
<tr>
<td>Squiggle</td>
<td>19 dBm</td>
<td>24 dBm</td>
</tr>
</tbody>
</table>

Table 2.3: Threshold Power Vs Thickness
2 mm sample because the thinner material has a lower effective permittivity. This is demonstrated in Fig. 2.3. The ELML tag has better performance on the 8 mm sample, and also works well on the 2 mm sample.

### 2.4.2 On-Tire Experiment

A typical tire is composed of multiple layers of rubber of varying electrical parameter and thickness. Here we test the effect of placement of an RFID tag on the outer sidewall surface. Fig. 2.10 shows six positions labeled P1-P6 on the exterior surface of the tire. Tags were tested at each position. The threshold power results are recorded in Table 2.4.

![Cross-section of a truck tire](image)

Figure 2.10: Cross-section of a truck tire.
<table>
<thead>
<tr>
<th>Position</th>
<th>P1</th>
<th>P2</th>
<th>P3</th>
<th>P4</th>
<th>P5</th>
<th>P6</th>
</tr>
</thead>
<tbody>
<tr>
<td>ELML</td>
<td>27 dBm</td>
<td>25 dBm</td>
<td>23 dBm</td>
<td>22 dBm</td>
<td>21 dBm</td>
<td>20 dBm</td>
</tr>
<tr>
<td>Squiggle®</td>
<td>29 dBm</td>
<td>28 dBm</td>
<td>27 dBm</td>
<td>25 dBm</td>
<td>21 dBm</td>
<td>21 dBm</td>
</tr>
</tbody>
</table>

Table 2.4: On Tire Threshold Power Measurement Results

As seen from Table 2.4 the placement of tags on the tire surface is critical for achieving the best performance. This is because at different positions the materials used to construct the tire and their thicknesses could be different, as seen in Fig. 2.10. The Alien Squiggle® tag exhibits the strongest dependence on position, and must be placed at P5 or P6 to get the best performance. The fabricated ELML tag has less dependence on position and lower or similar threshold power on all positions as compared to the Alien tag. This position-insensitive feature is desirable for tire applications because tire construction varies significantly for different models and brands, not to mention that the electrical properties of rubber can change with age.

2.5 Summary of Chapter 2

Design formulas have been introduced in this paper for RFID tag antennas mounted on materials with a prescribed range of permittivity and thickness. An end-loaded meander line tag antenna was modified and tested to validate the design approach. The theoretical and experimental results show that the designed tag is more robust to different media compared with a commercially available general-purpose tag. For the automotive tire application, the modified tag was found to be relatively insensitive to the mounting location on the sidewall surface. It is expected that the performance
can be further improved by fine-tuning the handmade antennas to achieve better matching with the chip.
Chapter 3: RFID Tag Antenna Design Approach-Antenna Pattern

According to (2.1) read range of an RFID tag is closely related to the tag antenna’s gain under perfect matching condition. Chapter 2 explained the effect of dielectric layers on the tag antenna’s impedance matching. In this chapter the effect of dielectric materials on an antenna’s gain pattern will be investigated.

3.1 Introduction

The bandwidth and impedance matching of an antenna could be improved via proper dielectric loading [43, 44]. However, the performance of a tag mounted on a dielectric layer will be compromised even when its impedance is matched. As shown in Fig. 3.1, full wave simulations illustrate that the broadside gain of an RFID tag antenna mounted on a dielectric slab is greatly reduced as compared to that in free space.

The performance of an antenna in the presence of dielectrics has been studied. For example the work in [45] compares the antenna’s performance for seven different materials. The work in [46] studies the antenna’s performance when it is placed close to a half space. The work in [47] studies the effect of materials with different
thicknesses and permittivities on the antenna gain pattern. However, most of the previous work merely states observations instead of explaining the reasons.

In this chapter we study the antenna’s performance in the presence of layered dielectrics using Green’s function. The theoretical approach indicates that the reduction of the broadside gain is not only due to material losses, but mainly due to certain modes activated inside the slab. Specifically, a substantial amount of power is confined in the near zone and does not radiate. Based on the analysis, the optimal position of antenna placement is found. As would be expected the read range of a tag is further improved when mounted on the advantageous location.

The developed tool set is then used to design a T-slot antenna. This antenna is used as an example to demonstrate the proposed design approach. This T-slot is connected to a RFID tag chip and tested within the rubber of a tire. Collected measurements validate the theoretical analysis. A best location is also found to further increase the tag’s read range.

Figure 3.1: Full wave simulation of the T-slot type RFID tag antenna gain, (a) in free space, (b) mounted on a finite dielectric slab residing in the xy plane.
3.2 Antenna Radiation: Construction of the Equations

For most RFID applications, the tags are mounted on or inside materials of finite thickness. The analysis of antennas mounted on or inside a dielectric medium is rather complex. In this chapter, we employ spectral domain method to reduce numerical analysis. This method is taken from [48] where it was used to study the dielectric half space and one layer grounded substrate. Here, a generalized spectral domain Green’s function for multilayer dielectric substrates is employed. In [49] the spectral domain Green’s function was derived by introducing spectral domain vector potentials. The spatial domain Green’s function was then computed using asymptotic method. We note that closed-form approximations of the Green’s function in planar stratified media have been given in [50]. More recently, the spectral analysis was used to compute the radiated fields of dipoles in the presence of planar stratified materials [51, 52]. Robust and rapid convergent methods for computing the fields using the spectral Green’s function have also been proposed in [53, 54]. In this work, we derive the spectral Green’s function in layered media using a more intuitive method without vector potentials.

The first step to apply spectral domain analysis is to define the Fourier transform pairs.

\[
\tilde{f}(k_x, k_y) = \mathcal{F}\{f(x, y)\} = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} f(x, y) e^{-j(k_x x + k_y y)} \, dx \, dy \tag{3.1}
\]

\[
f(x, y) = \mathcal{F}^{-1}\left\{\tilde{f}(k_x, k_y)\right\} = \frac{1}{(2\pi)^2} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \tilde{f}(k_x, k_y) e^{j(k_x x + k_y y)} \, dk_x \, dk_y. \tag{3.2}
\]

From (3.2), the field in spatial domain can be interpreted as an infinite summation (integration) of plane waves. Extending (3.1) to the EM field components, the corresponding pairs are
\[ <E_x, \tilde{E}_x>, <E_y, \tilde{E}_y>, <E_z, \tilde{E}_z>, \]
\[ <H_x, \tilde{H}_x>, <H_y, \tilde{H}_y>, <H_z, \tilde{H}_z>. \]

In case of differentiation, we have,

\[ \mathcal{F} \left\{ \frac{\partial}{\partial x} f(x, y) \right\} = jk_x \tilde{f}(k_x, k_y) \] (3.3)

\[ \mathcal{F} \left\{ \frac{\partial}{\partial y} f(x, y) \right\} = jk_y \tilde{f}(k_x, k_y) \] (3.4)

\[ \mathcal{F} \left\{ \frac{\partial}{\partial x \partial y} f(x, y) \right\} = -k_x k_y \tilde{f}(k_x, k_y) \] (3.5)

\[ \mathcal{F} \left\{ \frac{\partial^2}{\partial x^2} f(x, y) \right\} = -k_x^2 \tilde{f}(k_x, k_y). \] (3.6)

As shown in Fig. 3.2, the problem is to determine the radiated field of a current source embedded in a dielectric slab with finite thickness. The field in the source free region is decomposed into \(TE^z\) \((E_z = 0)\) mode and \(TM^z\) \((H_z = 0)\) mode. The total field is, of course, the summation of the modes.

In the source free region, the wave equation is

\[ \nabla^2 E_z + k^2 E_z = \left( \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + \frac{\partial^2}{\partial z^2} \right) E_z + k^2 E_z = 0 \] (3.7)

\[ \nabla^2 H_z + k^2 H_z = \left( \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + \frac{\partial^2}{\partial z^2} \right) H_z + k^2 H_z = 0. \] (3.8)

Applying Fourier transform to these equations, we have

\[ (-k_x^2 - k_y^2 + \frac{\partial^2}{\partial z^2}) \tilde{E}_z + k^2 \tilde{E} = \frac{\partial^2}{\partial z^2} \tilde{E}_z + k^2 \tilde{E}_z = 0 \] (3.9)

\[ (-k_x^2 - k_y^2 + \frac{\partial^2}{\partial z^2}) \tilde{H}_z + k^2 \tilde{H} = \frac{\partial^2}{\partial z^2} \tilde{H}_z + k^2 \tilde{H}_z = 0. \] (3.10)
Fig. 3.2 shows that the computational domain can be divided into four regions. For each region we have the field representatives

1) \( z > t_1 \)

\[
\tilde{E}_z = T_1^{TM} E_2(k_x, k_y)e^{-jk_1z}
\]

\[
\tilde{H}_z = T_1^{TE} H_2(k_x, k_y)e^{-jk_1z}
\]

2) \( 0 < z < t_1 \)

\[
\tilde{E}_z = E_2(k_x, k_y)(e^{-jk_2z} + \Gamma_1^{TM} e^{jk_2z})
\]

\[
\tilde{H}_z = H_2(k_x, k_y)(e^{-jk_2z} + \Gamma_1^{TE} e^{jk_2z})
\]

3) \( -t_2 < z < 0 \)

\[
\tilde{E}_z = E_3(k_x, k_y)(e^{-jk_2z} + \Gamma_2^{TM} e^{jk_2z})
\]

\[
\tilde{H}_z = H_3(k_x, k_y)(e^{-jk_2z} + \Gamma_2^{TE} e^{jk_2z})
\]
4) $z < -t_2$

\[ \tilde{E}_z = T_{2}^{TM} E_3(k_x, k_y)e^{-jk_1z} \]  
\[ \tilde{H}_z = T_{2}^{TE} H_3(k_x, k_y)e^{-jk_1z} \]  

(3.14)

where $k_i^2 = k_0^2\epsilon_i - k_x^2 - k_y^2$, and $\epsilon_i$ is the relative permittivity of the dielectric material. $E_{2,3}, H_{2,3}$ are the coefficients of E and H field in region 2 and 3. $\Gamma_{1,2}^{TM,TE}, T_{1,2}^{TM,TE}$ are reflection and transmission coefficients respectively. The superscript $TE, TM$ indicates the mode, and the subscript 1, 2 represents the reflection and transmission from upper and lower boundary respectively.

As mentioned above, the total field is the sum of TE and TM modes, and can be represented by [48]

\[ \tilde{E}_x = \frac{jk_x}{\beta^2} \frac{\partial}{\partial z} \tilde{E}_z + \frac{\omega\mu_0 k_y}{\beta^2} \tilde{H}_z \]  
(3.15)

\[ \tilde{E}_y = \frac{jk_y}{\beta^2} \frac{\partial}{\partial z} \tilde{E}_z - \frac{\omega\mu_0 k_x}{\beta^2} \tilde{H}_z \]  
(3.16)

\[ \tilde{H}_x = \frac{jk_x}{\beta^2} \frac{\partial}{\partial z} \tilde{H}_z - \frac{\omega\mu_0 k_y}{\beta^2} \tilde{E}_z \]  
(3.17)

\[ \tilde{H}_y = \frac{jk_y}{\beta^2} \frac{\partial}{\partial z} \tilde{H}_z - \frac{\omega\mu_0 k_x}{\beta^2} \tilde{E}_z \]  
(3.18)

where $\beta^2 = k_x^2 + k_y^2$. The $\tilde{E}_z$ terms refer to the TM mode, and the $\tilde{H}_z$ terms represent the TE mode. For a current source $\vec{J} = \hat{z}\delta(x)$ ($\vec{J} = \hat{z}$), upon enforcing the boundary condition at $z = 0$, using (3.15) to (3.18), we get

\[ E_2(k_x, k_y) = -\frac{k_x}{2\epsilon_1\omega} \cdot A' \]  
(3.19)

\[ H_2(k_x, k_y) = -\frac{k_y}{2k_2} \cdot B' \]  
(3.20)
\[ A' = \frac{\Gamma_{2}^{TM} - 1}{\Gamma_{1}^{TM} \Gamma_{2}^{TM} - 1} \quad \text{and} \quad B' = \frac{\Gamma_{2}^{TE} + 1}{\Gamma_{1}^{TE} \Gamma_{2}^{TE} - 1}. \]  \hspace{1cm} (3.21)

For \( z > t_1 \), the closed form Green’s functions for each field component can then be written as

\[ \tilde{G}_{E_x} = \frac{1}{2} \left[ \frac{Z_0}{k_0} \frac{1}{\epsilon_r} \frac{k_x k_1}{\beta^2} A + k_0 Z_0 \frac{k_y^2}{\beta^2 k_2} B \right] e^{-jk_1z} \]  \hspace{1cm} (3.22)

\[ \tilde{G}_{E_y} = \frac{1}{2} \left[ \frac{Z_0}{k_0} \frac{1}{\epsilon_r} \frac{k_y k_1}{\beta^2} A - k_0 Z_0 \frac{k_x k_y}{\beta^2 k_2} B \right] e^{-jk_1z} \]  \hspace{1cm} (3.23)

\[ \tilde{G}_{E_z} = \frac{1}{2} \left[ \frac{Z_0}{k_0} \frac{1}{\epsilon_r} \frac{k_x}{\beta} A e^{-jk_1z} \right] \]  \hspace{1cm} (3.24)

\[ \tilde{G}_{H_x} = \frac{1}{2} \left[ \frac{k_x k_y k_1}{k_2 \beta^2} B - \frac{1}{\epsilon_r} \frac{k_x k_y}{\beta^2} A \right] e^{-jk_1z} \]  \hspace{1cm} (3.25)

\[ \tilde{G}_{H_y} = \frac{1}{2} \left[ \frac{k_y^2 k_1}{k_2 \beta^2} B + \frac{1}{\epsilon_r} \frac{k_x^2}{\beta^2} A \right] e^{-jk_1z} \]  \hspace{1cm} (3.26)

\[ \tilde{G}_{H_z} = \frac{1}{2} \frac{k_y^2}{k_2} B e^{-jk_1z} \]  \hspace{1cm} (3.27)

where \( A = A' \cdot T_{1}^{TM} \), and \( B = B' \cdot T_{1}^{TE} \). There are still eight unknowns in the above equations to be determined. Specifically the reflection and transmission coefficients \( \Gamma_{1,2}^{TM,TE} \) and \( T_{1,2}^{TM,TE} \). By enforcing boundary conditions at \( z = t_1 \) and \( z = -t_2 \), these coefficients are derived as

\[ \Gamma_{i}^{TM} = e^{-jk_2 t_i}, \Gamma_{i}^{TM} \]  \hspace{1cm} (3.28a)

\[ \Gamma_{i}^{TM} = \frac{k_2 - \epsilon_r k_1}{k_2 + \epsilon_r k_1} \]  \hspace{1cm} (3.28b)

\[ \Gamma_{i}^{TE} = e^{-jk_2 t_i}, \Gamma_{i}^{TE} \]  \hspace{1cm} (3.29a)

\[ \Gamma_{i}^{TE} = \frac{k_2 - k_1}{k_2 + k_1} \]  \hspace{1cm} (3.29b)
\[ T_i^{TM} = \epsilon_r e^{-j(k_2-k_1)t_i}(1 + \Gamma^{TM}) \] (3.30)

\[ T_i^{TE} = e^{-j(k_2-k_1)t_i}(1 + \Gamma^{TE}) \] (3.31)

For \( z < -t_2 \), the Green’s function is found by interchanging \( t_1 \) and \( t_2 \). The spatial domain Green’s function can be found via inverse Fourier Transform using (3.2). The result is then computed using stationary phase method [1][48].

### 3.3 Computed Gain and Patterns

Shown in Fig. 3.3 (a) is a short dipole antenna placed on the surface of a dielectric slab of finite thickness. The radiation pattern of the dipole is computed using the spectral Green’s function derived in the previous section. As expected and illustrated in Fig. 3.3 (b), the presence of the dielectric slab not only detunes the antenna but also reduces the antenna gain by redistributing the radiated power.

![Figure 3.3](image-url)

Figure 3.3: (a) Illustration of a short dipole placed on a dielectric slab of finite thickness. (b) Possible mode inside the slab that traps power.
Fig. 3.4 (a) and (b) plots the peak gain of the dipole above and below the dielectric slab ($\epsilon_r = 10.63$ and $tan\delta = 0$) as a function of thickness. In the plots, the solid and dashed curves represent the peak gain of the dipole in the half space above and below the slab vs thickness. As can be seen in Fig. 3.4 (a), the peak gain is a periodic function with a periodicity of $t = 0.5\lambda_g$. It is also noticed that at thickness $t = 0.25\lambda_g$, the peak gain on both sides experiences significant reduction. In Fig. 3.4 (b), we can
observe that for a thin slab (of thickness $t < 0.25\lambda_g$), the gain below the slab is consistently higher. This is due to the usual lensing [55, 56, 57].

![Graph](image)

Figure 3.5: Peak gain of a short dipole on both half space vs the thickness of the dielectric slab ($\epsilon_r = 10.63$ and $\tan\delta = 0.163$), (a) $0 < \text{thickness} < 5\lambda_g$ (b) $0 < \text{thickness} < 0.5\lambda_g$.

For dielectric media with $\tan\delta \neq 0$, the dipole exhibits similar radiation patterns, but efficiency drops. An example is a short dipole placed on the top surface of a dielectric slab with $\epsilon_r = 10.63$ and $\tan\delta = 0.163$. The peak gain of the dipole above and below the slab is plotted in Fig. 3.5 (a) and (b). In Fig. 3.5, the solid and dashed curves represent the peak gain of the dipole (assuming the antenna is under perfect impedance matching condition) in the space above and below the slab. The peak
gain is again a periodic function with respect to the thickness of the slab, but is lower because \( \tan \delta \neq 0 \). From Fig. 3.4 and Fig. 3.5 it can be surmised that the broadside gain reduction is not only due to material losses, but also due to modes activated within the slab. As such, a substantial amount of power is confined in the near zone and does not radiate. Fig. 3.4 (b) and 3.5 (b) illustrate that for thin slabs (of thickness \( t < 0.25 \lambda_g \)), the gain below the slab is consistently higher due to lensing. Fig. 3.6 gives the peak gain when the dipole is placed at some depth below the surface. This result reinforces the observation that the maximum gain is achieved in the direction opposite to the side of the slab on which the tag is mounted.

### 3.4 Experimental Validation

In this section, a T-slot antenna is designed and optimized for the placement on a dielectric slab.
Figure 3.7: Layout of the T-slot antenna.

Figure 3.8: Impedance of the T-slot antenna mounted on a dielectric slab with $\varepsilon_r = 10$.

Figure 3.9: Power reflection coefficient of the T-Slot antenna in free space and placed on the slab having $\varepsilon_r = 6$ and 10.
3.4.1 Antenna Design

A T-slot antenna is employed to demonstrate the proposed design concepts. The antenna is designed using the simulation software Ansys HFSS. As shown in Fig. 3.7, the T-slot antenna can also be interpreted as an embedded T-match antenna. When placed on a dielectric slab with $\epsilon = 10$, its impedance is plotted in Fig. 3.8, and the associated power reflection coefficient $S_{11}$ is given in Fig. 3.9. As can be seen, the tag antenna has a broad and flat impedance bandwidth. More importantly, it maintains a broad bandwidth when mounted on dielectrics of different permittivities.

![Operational Range of the Tag](image)

**Figure 3.10:** Operational range of the tag (shaded region), (a) designed T-slot, (b) Alien Squiggle.

The operational range of the T-slot tag is computed using the equations in Chapter 2 and plotted in Fig. 3.10 (a), (b). This figure shows that, as compared with
the Squiggle antenna, the T-slot functions on a wider variety of permittivities and thicknesses.

### 3.4.2 Measurement on Dielectric Slabs

![Figure 3.11: T-slot is placed on a dielectric slab.](image)

Antenna position & Pth  
---  
Facing reader & 27 dBm  
Facing opposite direction & 25 dBm  

**Table 3.1: Threshold Power of the T-slot on the Slab**

A fixture was made with a dielectric slab of thickness $t = 10 \text{ mm}$, $\epsilon_r = 10.63$ and $\tan \delta = 0.16$. It was placed 11 ft away from the reader antenna and kept the same height with the reader antenna. As shown in Fig. 3.11, the T-slot tag was placed on the slab facing the reader and on the slab facing the opposite direction of the reader. The threshold power is tabulated in Table. 3.1. This table shows that the tag placed on the back of the rubber sample achieved higher threshold power than that placed on the front side. This is consistent to the analytical result.
3.4.3 Measurement on the Tire

The proposed analysis is used to design tag antennas for automotive tires. Tires are constructed with multiple layers, as shown in Fig. 3.12, having permittivity \( 6 < \varepsilon_r < 13 \), and thickness \( 5 < h < 20 \) mm. For this application, the tag’s performance was quantified by determining the minimum output power needed by the reader for detecting the tag (threshold power). Six positions, labeled P1 to P6, on the sidewall of the tire were used for the placement of the tag. The T-slot and Squiggle tags were examined at each position on the exterior surface. The corresponding threshold power is listed in Table 3.2.

![Figure 3.12: (a) Experiment Setup six positions are labeled on the sidewall of the tire. (b) The T-slot antenna is fabricated and connected to a RFID chip.](image)

Table 3.2 shows that, due to its broad bandwidth, the T-slot has less dependence on the position. Therefore its sensitivity is typically better as compared to the commercial Squiggle tag antenna by Alien. This is consistent with the design approach proposed in Chapter 2. The T-slot was also tested at location P6 but on the interior
Table 3.2: On-Tire Threshold Power Measurement Results

<table>
<thead>
<tr>
<th>Position</th>
<th>P1</th>
<th>P2</th>
<th>P3</th>
<th>P4</th>
<th>P5</th>
<th>P6</th>
</tr>
</thead>
<tbody>
<tr>
<td>T-Slot</td>
<td>29 dBm</td>
<td>25 dBm</td>
<td>23 dBm</td>
<td>22 dBm</td>
<td>21 dBm</td>
<td>21 dBm</td>
</tr>
<tr>
<td>Alien Squiggle</td>
<td>29 dBm</td>
<td>28 dBm</td>
<td>27 dBm</td>
<td>25 dBm</td>
<td>21 dBm</td>
<td>21 dBm</td>
</tr>
</tbody>
</table>

of the sidewall. We found that the threshold power was 2 dB lower as compared to placing it on the exterior surface. This is consistent with the analytical results derived using the spectral Green’s function, and is due to lensing.

3.5 Summary of Chapter 3

In this chapter we used the spectral domain Green’s function to study RFID tag antenna performance on dielectric layers. Our analysis predicts that the performance can be improved by placing the RFID tag on the opposite side of the dielectric from where maximum gain is desired. This was verified experimentally. It was also noted that the slab can support internal fields that degrade the antenna’s efficiency.
Chapter 4: RFID Tag in the Vicinity of a Wire Grating

In this chapter, we investigate the RFID tag antenna’s behavior when the tag is close to a wire grating. This research is inspired by designing RFID tag antennas for automotive tires. Steel plies are embedded in the truck tires to increase their mechanical integrity. Therefore, antennas must be designed or positioned with in-depth consideration of their surrounding environment. In this chapter, a 2D theoretical analysis is performed to study the radiation of a dipole adjacent to a periodic grid of parallel wires. Results indicate that the wire grating acts as a solid PEC plate when the dipole is oriented parallel to the wires, and like free space when the dipole is transverse to the wires. A T-slot antenna is optimized to operate near such a grating. The experimental results indicate that the wire grating can be utilized as a ground plane to increase the gain of the antenna, thereby further extending the read range. The trade-off is that the radiation is primarily broadside to the grating and very weak in the opposite direction.

4.1 Introduction

Recently, major tire manufacturers have shown strong interest in using passive RFID systems for better inventory tracking and identification. A unique challenge of implementing RFID tags with truck tires is the presence of steel plies. As shown in
Figure 4.1, the metallic plies are radically arranged in both of the sidewalls of a tire to reinforce its mechanical integrity. They can be modeled as a wire grating which could impede RF signals from propagation into the tire. Therefore, it is necessary to study the effect of these steel plies on an RFID tag antenna’s performance.

Figure 4.1: Radial metallic plies are embedded in the sidewall of a truck tire.

For tag placement, we have the freedom to choose whether to place it parallel or perpendicular to the steel plies. In this chapter, a theoretical approach is used to study how a short dipole antenna radiates in the presence of a wire grating in both configurations.

4.2 EM Theory of a Periodic Structure

Each steel ply can be modeled as a thin PEC circular cylinder. As shown in Fig. 4.2, within a section on the tire sidewall, steel plies can be modeled locally as an infinite array of circular cylinders. The radius of the steel ply is \( a \), the space between
two plies is $s$, and the periodicity of the array is $d$, $d = 2a + s$. For common truck tires $a \approx 0.75$ mm, $s \approx 1$ mm, and $d \approx 2.5$ mm.

The analysis is performed in 2D. Throughout the analysis, the wire grating is positioned parallel to the $z$ direction and periodic along the $x$ direction. The problem is addressed in TEz (transverse electric to $z$) and TMz (transverse magnetic to $z$) polarizations respectively.

4.2.1 TM Polarization: dipole parallel to the wire grating

For an electric dipole parallel to the wire grating, the 2D problem can be addressed in TMz polarization. The problem is solved in the spectral domain by computing the plane wave scattering from the wire grating at the incidence angle $\theta$. Shown in Fig. 4.3 is a TMz plane wave incident on an infinite periodic structure. The wire grating is oriented along the $z$-axis. Method of Moments (MoM) is used to first compute the plane wave scattering from a thin cylinder [58].

The incident $E$ field is

$$E^i = E^i_{\hat{z}} = E_0 e^{-jk(x\cos\theta-y\sin\theta)\hat{z}}. \quad (4.1)$$
The scattered E field can be computed using the radiation equation [48],

\[
E_s(r) = -j\omega\mu \int \left\{ J(r')G(r, r') + \frac{1}{k^2} \nabla \cdot \{ J(r')G(r, r') \} \right\} dv',
\]  

(4.2)

where \( J(r') \) is the equivalent current density on the surface of the cylinder. Since the incident electric field only has z component, the scattered and the total electric field should also each have only z component which is independent of z variation [59]. Therefore, \( J = J_z \hat{z} \) and \( J_z \) is independent of z variation, so the second term of (4.2) is zero. Because the two dimensional Green’s function is \( G(\rho, \rho') = -\frac{j}{4} H_0^{(2)}(k|\rho - \rho'|) \) [60], and current only distributes on the surface of the conductor, the scattered field can be represented as

\[
E_s^z(\rho) = -\frac{\omega\mu a}{4} \int_0^{2\pi} J_z(\rho') H_0^{(2)}(k|\rho - \rho'|) d\phi',
\]

(4.3)

where \( a \) is the radius of the wire. \( \rho \) and \( \rho' \) are observation and source point respectively. For a thin wire \((a \ll \lambda)\), the current is uniformly distributed on the surface of the cylinder, so the equivalent current density has no \( \phi \) dependence. The scattered field can be represented as

\[
E_s^z(\rho) = -\frac{\omega\mu a}{4} \int_0^{2\pi} H_0^{(2)}(k|\rho - \rho'|) d\phi',
\]

(4.4)
For scattering by a single PEC (perfect electric conductor) cylinder, the equivalent current density $J_z$ is calculated by enforcing the boundary condition on the surface of the cylinder. Here the electric field integral equation (EFIE) is used to compute the scattered field. The boundary condition states that $\hat{n} \times (E^i + E^s) = 0$ so $E_z^i = -E_z^s$ on the surface. By placing the observation point on the surface of the cylinder $\rho = \rho_i$, the electric field integral equation is constructed as

$$E_0 e^{-jk(x_i \cos \theta - y_i \sin \theta)} = \frac{\omega \mu a}{4} J_z \int_0^{2\pi} H_0^{(2)}(k|\rho_i - \rho'|) d\phi',$$  \hspace{1cm} (4.5)

where $\rho_i = x_i \hat{x} + y_i \hat{y}$. Method of Moments (MoM) is used to compute the equivalent current density. Point matching and entire domain basis functions are used. As mentioned above the current is assumed to be uniformly distributed on the surface of the cylinder. Thus, the current density $J_z = \alpha$ for $\rho' = a$ and $0 \leq \phi' < 2\pi$. We obtain the matrix equation

$$[Z_i][\alpha] = [g_i],$$  \hspace{1cm} (4.6)

where $[\alpha]$ is the coefficient of the current density to be determined. The elements $[g_i]$ are

$$[g_i] = E_z^i(\rho_i)$$  \hspace{1cm} (4.7)

and the elements $[Z_i]$ are

$$[Z_i] = \frac{\omega \mu a}{4} \int_0^{2\pi} H_0^{(2)}(k|\rho_i - \rho'|) d\phi'.$$  \hspace{1cm} (4.8)

It is noticed that there is only one unknown in (4.6), so only one testing point is needed. For thin wire approximation, the testing point is chosen to be in the origin, $\rho_i = x_i \hat{x} + y_i \hat{y}$, $x_i = 0, y_i = 0$, in order to force the total electric field to be zero inside
the cylinder. Thus, the equivalent current density can be expressed in closed form,

\[ J_z = \alpha = \frac{g_1}{Z_1} = \frac{2E_0}{\omega \mu \pi a H_0^{(2)}(ka)}. \]  

(4.9)

The scattered E field can be derived by substituting (4.9) into (4.4).

To validate the MoM solution, the eigenfunction solution of the scattered E field from a circular cylinder is used [61]. It is

\[ E_s^z(\rho) = E_0 \sum_{n=-\infty}^{\infty} j^{-n} a_n H_n^{(2)}(k\rho)e^{jn(\phi-\theta)} \quad a_n = \frac{-I_n(ka)}{H_n^{(2)}(ka)} \]  

(4.10)

where \( H_n^{(2)} \) and \( I_n \) are the Hankel function of the section kind and the Bessel function of the first kind respectively. \( \theta \) is the angle of the incident wave, and \( \phi \) is the angle of the observation point. To compare the MoM solution with the eigenfunction solution, the bistatic scattering width \( \sigma_{2-D}/\lambda \) is computed as \( \sigma_{2-D} = 2\pi |E_s^z|^2. \) The result is plotted in Fig. 4.4 for a cylinder of radius \( a = 0.75 \) mm.

In the plot, 101 modes are summed to generate the eigenfunction solution (the eigenfunction solution is converged with 101 modes). The result indicates that the thin wire approximation is sufficient for the UHF RFID band (902-928 MHz). The thin wire approximation starts breaking down when the frequency is beyond 10 GHz. At 10 GHz the wavelength is 3 cm, 40 times the radius of the cylinder. Therefore, it can be concluded that the thin wire approximation can be used for the wire grating for TMz polarization.

For the infinite wire grating the scattered E field can be expressed as

\[ E_z^s(\rho) = -\frac{\omega \mu a}{4} \sum_{n=-\infty}^{\infty} \int_0^{2\pi} J_n(\rho_n') H_0^{(2)}(k|\rho - \rho_n'|)d\phi_n'. \]  

(4.11)

In the equation, the primed parameter indicates the source location and \( n \) represents local coordinates of the \( n^{th} \) cylinder. The unknown quantities are the current densities
Figure 4.4: TMz bistatic scattering vs frequency computed using the eigenfunction (101 modes) and MoM methods. The angle of the incident wave $\theta = 0$ degrees. The angle of the observation point is $\phi = 90$ degrees.

$J_n$ on each cylinder which again is assumed to be uniform around each cylinder. As the boundary condition needs to be satisfied on each cylinder, for such an infinite periodic structure the Floquet condition states that the equivalent current density on each cylinder is [62]

$$J_n = J_0 e^{jn\psi} \quad (4.12)$$

where $J_0$ is the current density on the 0th cylinder and $\psi$ is the phase delay. Given the incident wave in (4.1), $\psi = -dk \cos \theta$ [63], here $d$ is the periodic spacing of the wire grating, $k$ and $\theta$ are the propagation constant and the angle of the incident wave respectively. The solution of scattering from a single cylinder indicates that the thin wire approximation is sufficient for the RFID frequency band. Therefore, it is assumed here that the surface current density also has no $\phi$ dependence.
Eq. (4.11) shows that the infinite summation only works on \( n \), so it can be moved inside the integration. Also in (4.11), \( \rho = x\hat{x} + y\hat{y} \) and \( \rho_n = (a\cos\phi' + nd)\hat{x}' + a\sin\phi'\hat{y}' \).

Therefore, the scattered E field can be expressed as

\[
E_z^s(x, y) = -\frac{\omega \mu a}{4} J_0 \int_0^{2\pi} \sum_{n=-\infty}^{\infty} e^{jn\psi} H_0^{(2)}((k\sqrt{(y - a\sin\phi')^2 + (x - a\cos\phi' - nd)^2})d\phi'.
\]

The Hankel function can be expressed in its plane wave expansion [48]

\[
H_0^{(2)}(k\sqrt{x^2 + y^2}) = \frac{1}{\pi} \int_{-\infty}^{\infty} e^{-jk_y|y|} e^{jk_x x} dk_x
\]

Replacing the Hankel function with its plane wave expansion (4.14), (4.13) becomes

\[
E_z^s(x, y) = -\frac{\omega \mu a}{4\pi} J_0 \int_0^{2\pi} \sum_{n=-\infty}^{\infty} e^{jn\psi} e^{\frac{-jk_y|y|}{k_y}} e^{\frac{jk_x x}{k_y}} e^{-jk_x nd} dk_x d\phi'.
\]

Interchanging the inner integral and summation over \( n \), (4.15) becomes

\[
E_z^s(x, y) = -\frac{\omega \mu a}{4\pi} J_0 \int_0^{2\pi} \sum_{n=-\infty}^{\infty} e^{jn\psi} e^{-jk_x nd} \left[ \int_{k_x = -\infty}^{\infty} \frac{e^{-jk_y|y-a\sin\phi'|+jk_x(x-a\cos\phi')}}{k_y} e^{-jk_x nd} dk_x \right] d\phi'.
\]

An important identity is [63]

\[
\sum_{n=-\infty}^{\infty} e^{jn\psi} e^{-jk_x nd} = \frac{2\pi}{d} \sum_{n=-\infty}^{\infty} \delta(k_x - \frac{2\pi n}{d} - \frac{\psi}{d}).
\]

Substituting (4.17) into (4.16), (4.16) becomes

\[
E_z^s(x, y) = -\frac{\omega \mu a}{2d} J_0 \int_0^{2\pi} e^{\frac{-jk_y y - jk_x x}{k_y}} \sum_{n=-\infty}^{\infty} e^{jn\psi} e^{-jk_x nd} dk_x d\phi'.
\]

The integration only works on the primed parameters, (4.18) can be re-organized as

\[
E_z^s(x, y) = -\frac{\omega \mu a}{2d} J_0 \sum_{n=-\infty}^{\infty} e^{\frac{jk_y y + jk_x x}{k_y}} \frac{1}{k_y} \int_0^{2\pi} e^{\frac{jk_y a \sin\phi' - jk_x a \cos\phi'}{k_y}} d\phi'.
\]
where \( k_{xn} = \frac{2\pi n}{d} + \frac{\psi}{d} = -k \cos \theta + \frac{2\pi n}{d} \),

\[
k_{yn} = \begin{cases} 
\sqrt{k^2 - k_{xn}^2} & k^2 - k_{xn}^2 > 0 \\
-j \sqrt{k^2 - k_{xn}^2} & k^2 - k_{xn}^2 < 0.
\end{cases}
\] (4.20)

Eq. (4.19) is a sum of Floquet modes, so therefore the scattered field is represented as an infinite summation of plane waves. Eq. (4.19) can be expressed as

\[
E_s^z(x, y) = \sum_{n=-\infty}^{\infty} C_n e^{\mp jk_{yn} y + jk_{xn} x} \] (4.21)

where the coefficient of each plane wave is

\[
C_n = -\frac{\omega \mu a}{2d} J_0 \frac{1}{k_{yn}} \int_0^{2\pi} e^{\pm jk_{yn} a \sin \phi' - jk_{xn} a \cos \phi'} d\phi'.
\] (4.22)

It is noticed that there is only one unknown in (4.19), the \(0^{th}\) order current density \(J_0\). It can be computed using the EFIE and MoM as described above. Similar to the single cylinder problem one testing point and entire domain basis function are used. The testing point is at the origin \((x_i = 0 \text{ and } y_i = 0)\) in order to force the total electric field to be zero inside the cylinder, yielding,

\[
E_i^z(0, 0) = E_0 = -E_s^z = \frac{\omega \mu a}{2d} J_0 \int_0^{2\pi} \sum_{n=-\infty}^{\infty} \frac{e^{-jk_{yn} a \sin \phi' - jk_{xn} a \cos \phi'}}{k_{yn}} d\phi'.
\] (4.23)

The last term in (4.23) can be expressed as

\[
= \frac{\omega \mu a}{2d} J_0 \sum_{n=-\infty}^{\infty} \left[ \int_0^{\pi} \frac{e^{-jk_{yn} a \sin \phi' - jk_{xn} a \cos \phi'}}{k_{yn}} d\phi' + \int_{\pi}^{2\pi} \frac{e^{jk_{yn} a \sin \phi' - jk_{xn} a \cos \phi'}}{k_{yn}} d\phi' \right]
\] (4.24)

\[
= \frac{\omega \mu a}{2d} J_0 \sum_{n=-\infty}^{\infty} \int_0^{\pi} \frac{2\pi \cos(ak_{zn} \cos \phi') e^{-jk_{yn} a \sin \phi'}}{k_{yn}} d\phi'.
\]
So the current density can be expressed in closed form

\[ J_0 = -\frac{2dE_0}{\omega \mu a} \sum_{n=-\infty}^{\infty} \frac{1}{2 \kappa y_n} \int_0^{\pi} \cos[ak_{zn} \cos \phi'] e^{-jk_{yn}a \sin \phi'} d\phi'. \] (4.25)

In order to compute \( J_0 \), an infinite summation of integrals needs to be calculated. The number of modes needed to be summed depends on the convergence rate of the integral in the denominator of (4.25). Because there is no closed form solution for this integral, it is computed here using trapezoidal rule of numerical integration. Fig. 4.5 plots the real and imaginary values of the integral \( \frac{2}{\kappa y_n} \int_0^{\pi} \cos[ak_{zn} \cos \phi'] e^{-jk_{yn}a \sin \phi'} d\phi' \) versus mode number. Here \( a = 0.75 \text{ mm}, d = 2.5 \text{ mm}, f = 915 \text{ MHz} \). The angle of the incident wave is \( \theta = 45 \text{ degrees} \).

![Real and Imaginary Values of the Integral](image)

Figure 4.5: Real and imaginary values of the integral \( \frac{2}{\kappa y_n} \int_0^{\pi} \cos[ak_{zn} \cos \phi'] e^{-jk_{yn}a \sin \phi'} d\phi' \) vs mode number \( n \). \( a = 0.75 \text{ mm}, d = 2.5 \text{ mm}, \theta = 45 \text{ degrees} \).
Fig. 4.5 indicates that the convergence rate of the integrand should be very fast. Only a few (3 to 5) modes may be needed to compute the infinite summation.

The reflection and transmission coefficients of the $0^{th}$ order Floquet mode vs frequency are computed using (4.19)-(4.25) and plotted in Fig. 4.6. Here $a = 0.75\ mm$, $d = 2.5\ mm$, and $\theta = 45$ degrees.

Figure 4.6: Reflection and transmission coefficients of the $0^{th}$ order Floquet mode, (a) real part, (b) imaginary part. $a = 0.75\ mm$, $d = 2.5\ mm$, and $\theta = 45$ degrees.

Fig. 4.6 shows that the reflection coefficient is very close to -1 and the transmission coefficient is close to 0 over a large bandwidth, especially at frequencies below 10GHz where the thin wire approximation is valid. To make only the $0^{th}$ order Floquet mode propagate, $k_{yn}$ needs to be real for $n = 0$ and purely imaginary for $n \neq 0$. So, the equation needs to satisfy

$$|k_{xn}| = | - k \cos \theta + \frac{2\pi n}{d} |_{n=1, 0<\theta<\pi/2} > |k|. \quad (4.26)$$
This leads to $2a < d \leq \lambda/2$ (periodicity $d$ needs to be greater than the diameter of the cylinder $2a$). In the RFID frequency band, $\lambda \approx 300$ mm, when $a = 0.75$ mm and $d = 2.5$ mm. As $d < \lambda/2$, only the $0^{th}$ order Floquet mode propagates regardless of the angle of the incident wave. At 915 MHz, using the same equations, the reflection and transmission coefficients versus the angle of the incident wave are plotted in Fig. 4.7.

![Reflection coefficient](image1)

![Transmission coefficient](image2)

Figure 4.7: Real and imaginary values of the coefficients vs the angle of the incident wave. $f = 915$ MHz, $a = 0.75$ mm, and $d = 2.5$ mm. (a) Reflection coefficient, (b) transmission coefficient.

The reflection coefficient is -1 and transmission coefficient is 0 regardless of the angle of incidence. It can be concluded that in the UHF RFID frequency band with wire radius $a = 0.75$ mm and $d = 2.5$ mm, the wire grating acts as a PEC plate for TMz waves.
4.2.2 TE Polarization: dipole transverse to the wire grating

For an electric dipole placed orthogonal to the wire grating the problem is addressed in TEz polarization. As shown in Fig. 4.8, similar to the TMz polarization, the incident field is defined as

$$\mathbf{H}^i = H_z^i \mathbf{\hat{z}} = H_0 e^{-jk(x \cos \theta - y \sin \theta)} \mathbf{\hat{z}}, \quad (4.27)$$

where $\theta$ is the angle of the incident wave. Similar to the TMz solution, the scattered field from a single cylinder is first computed. The scattered H field from a single cylinder can be computed as

$$\mathbf{H}^s(\rho) = \nabla \times \mathbf{A} = \nabla \times \int \mathbf{J}(\rho')G(\rho, \rho')ds'. \quad (4.28)$$

The following vector identity is applied,

$$\nabla \times [\mathbf{J}(\rho')G(\rho, \rho')] = G(\rho, \rho')\nabla \times \mathbf{J}(\rho') - \mathbf{J}(\rho') \times \nabla G(\rho, \rho') \quad (4.29)$$

The first term of (4.29) vanishes as the unprimed $\nabla$ curls the primed parameter. Substituting (4.29) and 2D Green’s function into (4.28), assuming the current density
is only on the surface, the scattered field can be expressed as

\[
\mathbf{H}^s(\rho) = \nabla \times \mathbf{A} = \frac{j}{4} \int_C \mathbf{J}(\rho') \times \nabla H_0^{(2)}(k|\rho - \rho'|) dc'.
\] (4.30)

Here the magnetic field integral equation (MFIE) is used for computing the scattered \( \mathbf{H} \) field. Enforcing the boundary condition on the surface of the cylinder

\[
\hat{n} \times (\mathbf{H}^i + \mathbf{H}^s) = \mathbf{J}
\] (4.31)

As \( \hat{n} \times \mathbf{H}^i = \hat{n} \times H_z^i \hat{z} \), \( \hat{n} \times \hat{z} = -\hat{\phi} \), (4.31) can be written as

\[
-\hat{\phi} H_z^i(\rho) + \hat{n} \times \mathbf{H}^s(\rho) = J_\phi(\rho) \hat{\phi}.
\] (4.32)

As \( \hat{n} = -\hat{z} \times \hat{\phi} \), using vector identity \( (\mathbf{A} \times \mathbf{B}) \times \mathbf{C} = (\mathbf{A} \cdot \mathbf{C})\mathbf{B} - (\mathbf{B} \cdot \mathbf{C})\mathbf{A} \), so we have \( \hat{n} \times \mathbf{H}^s = -(\hat{z} \times \hat{\phi}) \times \mathbf{H}^s = -(\hat{z} \cdot \mathbf{H}^s)\hat{\phi} + (\hat{\phi} \cdot \mathbf{H}^s)\hat{z} \). According to (4.32), the term \( \hat{n} \times \mathbf{H}^s \) should only have a \( \hat{\phi} \) component. Thus, (4.32) becomes

\[
J_\phi(\rho) + \hat{z} \cdot \mathbf{H}^s(\rho) = -H_z^i(\rho).
\] (4.33)

Substituting (4.30) into (4.33), we have

\[
J_\phi(\rho) + \hat{z} \cdot \left[ \frac{j}{4} \int_C \mathbf{J}(\rho') \times \nabla H_0^{(2)}(k|\rho - \rho'|) dc' \right] = -H_z^i(\rho).
\] (4.34)

It may be shown that,

\[
\nabla H_0^{(2)}(k|\rho - \rho'|) = -kH_1^{(2)}(kR)\hat{R}, \quad \mathbf{R} = \rho - \rho'.
\] (4.35)

Eq.(4.34) becomes

\[
J_\phi(\rho) - \hat{z} \cdot \left[ \frac{j}{4} k \int_C \mathbf{J}(\rho') \times H_1^{(2)}(kR)\hat{R} dc' \right] = -H_z^i(\rho).
\] (4.36)

Using a vector identity,

\[
\hat{z} \cdot (J(\rho') \times \hat{R}) = \hat{R} \cdot (\hat{z} \times J(\rho')) = J_\phi(\rho')\hat{R} \cdot (\hat{z} \times \hat{\phi'}) = -\hat{R} \cdot \hat{n}' J_\phi(\rho')
\] (4.37)
Eq. (4.36) becomes
\[ J_\phi(\rho) + \frac{j}{4} k \int_C [\hat{R} \cdot \hat{n}'] J_\phi(\rho') H^{(2)}_1(kR) dc' = -H^z_z(\rho). \] (4.38)

Note that \( \hat{n}' \) is the unit surface normal of the cylinder at the integration point \( \rho' \).

Placing the observation point at \( \rho = \rho_i \) on the surface of the cylinder \( C \), the MFIE is constructed as
\[ J_\phi(\rho_i) + \frac{j}{4} k \int_C [\hat{R}_i \cdot \hat{n}'_i] J_\phi(\rho') H^{(2)}_1(kR_i) dc' = -H^z_z(\rho_i). \] (4.39)

To evaluate the integral in (4.39), a singularity must be dealt with when \( \rho_i \rightarrow \rho' \).

To do so, the integral is split into two parts: \( \Delta C \) (the self-term) and \( C - \Delta C \) (the non-self-term). Thereby, the integral in (4.39) is written as
\[ \frac{j}{4} k \int_C [\hat{R}_i \cdot \hat{n}'_i] J_\phi(\rho') H^{(2)}_1(kR_i) dc' = \frac{j}{4} k \int_{\Delta C} [\hat{R}_i \cdot \hat{n}'_i] J_\phi(\rho') H^{(2)}_1(kR_i) dc' + \frac{j}{4} k \int_{C - \Delta C} [\hat{R}_i \cdot \hat{n}'_i] J_\phi(\rho') H^{(2)}_1(kR_i) dc'. \] (4.40)

To compute the self-term (integration over \( \Delta C \)) in (4.40), a circle (enclosing the observation point) is drawn with radius of \( r \), and the center at \( \rho_i \), as shown in Fig. 4.9.

The source contour \( \Delta C \) is the section of the cylinder surface inside the circle. It is noted that \( \rho_i \) is assumed to be located on the outer side of the contour \( C \) in the evaluation that follows. As \( \Delta C \) is very small, we can assume that the current density is constant (independent of \( \phi' \)) and \( J_\phi(\rho') \approx J_\phi(\rho_i) \) over \( \Delta C \) [60]. Thereby, the self-term in (4.40) can be expressed as
\[ \frac{j}{4} k \int_{\Delta C} [\hat{R}_i \cdot \hat{n}'_i] J_\phi(\rho_i) H^{(2)}_1(kR_i) dc' \approx \frac{j r}{4} k J_\phi(\rho_i) \int_{\Delta \phi} [\hat{R}_i \cdot \hat{n}'_i] H^{(2)}_1(kR_i) d\Delta \phi' \] (4.41)
\[ = \frac{j r}{4} k \Delta \phi J_\phi(\rho_i) H^{(2)}_1(kr), \]
where \( \Delta \phi = 2 \cos^{-1} \frac{r}{2a} \), \( a \) is the radius of the cylinder. For small argument \( (kr \rightarrow 0) \), the Hankel function of the second kind can be expressed as [61]
\[ H^{(2)}_n(kr) = \frac{(kr)^n}{2^n n!} + \frac{j 2^n (n - 1)!}{\pi (kr)^n} \quad n > 0. \] (4.42)
When $r \to 0$, $\Delta \phi = 2 \cos^{-1} \frac{r}{2a} \approx \pi$. Using (4.42), (4.41) can be represented as

$$
\frac{j}{4} k \int_{\Delta C} \left[ \hat{R}_i \cdot \hat{n}' \right] J_\phi(\rho') H_1^{(2)}(kR_i) dc' \approx \frac{j}{4} k \Delta \phi J_\phi(\rho_i) H_1^{(2)}(kr)
$$

$$
\approx J_\phi(\rho_i) \frac{j}{4} k \pi \left( \frac{kr}{2} + \frac{2}{\pi kr} \right) = J_\phi(\rho_i) \left( -\frac{1}{2} + j \frac{k^2 r^2 \pi}{8} \right)
$$

(4.43)
as $r \to 0$, we can write [59, 58]

$$
\frac{j}{4} k \int_{\Delta C} \left[ \hat{R}_i \cdot \hat{n}' \right] J_\phi(\rho') H_1^{(2)}(kR_i) dc' \approx -J_\phi(\rho_i). 
$$

(4.44)
This is the expected principal value of the MFIE contour integral. Substituting (4.44) and (4.40) into (4.39), (4.39) becomes

$$
\frac{J_\phi(\rho_i)}{2} + \frac{j}{4} k \int_{C-\Delta C} \left[ \hat{R}_i \cdot \hat{n}' \right] J_\phi(\rho') H_1^{(2)}(kR_i) dc' = -H_z(\rho_i).
$$

(4.45)
The equivalent current density is computed using the Method of Moments. Similar to TM polarization, point matching and entire domain basis functions are used. For the thin wire approximation, the equivalent electric current density is represented

\[ J_\phi(\rho') = \alpha e^{-j\phi'} + \beta + \gamma e^{j\phi'} \quad \rho' = a \quad \text{and} \quad 0 \leq \phi' < 2\pi. \]  

where \( \alpha, \beta, \) and \( \gamma \) are unknown coefficients. The selection of this basis function is based on the eigenfunction solution of the equivalent current density on a PEC cylinder [61]

\[ J(\rho') = J_\phi \hat{\phi} \quad J_\phi = \frac{2jH_0}{\pi ka} \sum_{n=-\infty}^{\infty} \frac{j^{-n}e^{jn(\phi'-\theta)}}{H'(2n')(ka)}. \]

(4.47)

Where \( \theta \) is the angle of the incident wave. For a small radius \( a \), the \( n = 0 \) term is dominant. However, the \( n = \pm 1 \) terms also radiate, so they cannot be neglected.

Using (4.46), the obtained matrix is

\[ [Z_{i\alpha} \ Z_{i\beta} \ Z_{i\gamma}] [\alpha \ \beta \ \gamma]^T = [g_i] \]  

(4.48)

where

\[ Z_{i\alpha} = \frac{e^{-j\phi_i}}{2} + \frac{jak}{4} \int_{\phi-\Delta\phi}^{\phi+\Delta\phi} [\hat{R}_i \cdot \hat{n}']e^{-j\phi'} H_1^{(2)}(kr_i)d\phi' \]

\[ Z_{i\beta} = \frac{1}{2} + \frac{jak}{4} \int_{\phi-\Delta\phi}^{\phi+\Delta\phi} [\hat{R}_i \cdot \hat{n}'] H_1^{(2)}(kr_i)d\phi' \]

(4.49)

\[ Z_{i\gamma} = \frac{e^{j\phi_i}}{2} + \frac{jak}{4} \int_{\phi-\Delta\phi}^{\phi+\Delta\phi} [\hat{R}_i \cdot \hat{n}'] e^{j\phi'} H_1^{(2)}(kr_i)d\phi', \]

and

\[ [g_i] = -H_z^{(1)}(\rho_i). \]

(4.50)

Trapezoidal rule of integration is used to compute the integrals in (4.49). Three testing points are used to compute the current density. They are \((x_i, y_i) = (0, a), (x_i, y_i) = (a, 0), \) and \((x_i, y_i) = (0, -a)\). The scattering width \( \sigma_{2-D}/\lambda \) of the cylinder is computed using both the eigenfunction and the MoM methods. The results are
compared and plotted in Fig. 4.10. Here \( a = 0.75 \) mm, and the observation angle \( \phi = 180^\circ \).

Figure 4.10: Scattering width \( \sigma_{2-D}/\lambda \) computed using the eigenfunction and MoM methods. \( a = 0.75 \) mm. The angle of the incident wave is \( \theta = 0 \) degrees. The angle of the observation position is \( \phi = 180 \) degrees

The result indicates that the thin wire approximation yields the same result as the eigenfunction solution at the RFID frequency band. The thin wire approximation breaks down when frequency is higher than 30 GHz.

For the infinite wire grating the scattered H field can be expressed as

\[
H^s(\rho) = \frac{ja}{4} \sum_{n=-\infty}^{\infty} \int_0^{2\pi} J_n(\rho'_{rn}) \times \nabla H_0^{(2)}(k|\rho - \rho'_{rn}|) d\phi'.
\]  

(4.51)
Similar to the TM polarization, the boundary condition needs to be satisfied on each cylinder, so the equivalent current density on each cylinder can be expressed as

\[ J_{\phi_n} = J_{\phi_0} e^{j n \psi} \] (4.52)

where \( J_{\phi_0} \) is the current density on the 0th cylinder and \( \psi \) is the phase delay, \( \psi = -dk \cos \theta \) [63].

Enforcing the boundary condition on the 0th cylinder and applying (4.31) to (4.38), the MFIE can be represented as

\[ J_{\phi_0}(\rho) + \frac{ja}{4} k \sum_{n=-\infty}^{\infty} e^{j n \psi} \int_{0}^{2\pi} \left[ \hat{R}_n \cdot \hat{n}'_n \right] J_{\phi_0}(\rho') H^{(2)}_1(kR_n) d\phi' = -H_z^i(\rho). \] (4.53)

Placing the observation point at \( \rho \rightarrow \rho_i \) on the surface of the 0th cylinder and applying (4.40) to (4.45), the MFIE becomes

\[ \frac{J_{\phi_0}(\rho_i)}{2} + \frac{ja}{4} k \int_{\phi_0}^{\phi_0 - \Delta \phi} \left[ \hat{R}_{\rho_0} \cdot \hat{n}'_{\rho_0} \right] J_{\phi_0}(\rho') H^{(2)}_1(kR_{\rho_0}) d\phi' + \sum_{n=-\infty}^{\infty} e^{j n \psi} \int_{0}^{2\pi} \left[ \hat{R}_{in} \cdot \hat{n}'_{n} \right] J_{\phi_0}(\rho') H^{(2)}_1(kR_{in}) d\phi' \]

\[ = -H_z^i(\rho_i). \] (4.54)

MoM is used to compute the current density. The entire domain basis function (4.46) is employed. The three testing points adopted in the single cylinder problem are also used here. The obtained matrix is

\[ [Z_{i\alpha} Z_{i\beta} Z_{i\gamma}] [\alpha \beta \gamma]^T = [g_i] \] (4.55)
where

\[
Z_{i\alpha} = \frac{e^{-j\phi_i}}{2} + \frac{jak}{4} \int_{\phi_0 - \Delta \phi} \left[ \hat{R}_{i0} \cdot \hat{n}_0' \right] e^{-j\phi'} H_1^{(2)}(kR_{i0}) d\phi_0
\]

\[+ \frac{jak}{4} \sum_{n=-\infty}^{\infty} e^{jn\psi} \int_{0}^{2\pi} \left[ \hat{R}_{in} \cdot \hat{n}_n' \right] e^{-j\phi'} H_1^{(2)}(kR_{in}) d\phi'_n. \]

(4.56)

\[
Z_{i\beta} = \frac{1}{2} + \frac{jak}{4} \int_{\phi_0 - \Delta \phi} \left[ \hat{R}_{i0} \cdot \hat{n}_0' \right] H_1^{(2)}(kR_{i0}) d\phi_0
\]

\[+ \frac{jak}{4} \sum_{n=-\infty}^{\infty} e^{jn\psi} \int_{0}^{2\pi} \left[ \hat{R}_{in} \cdot \hat{n}_n' \right] H_1^{(2)}(kR_{in}) d\phi'_n. \]

\[
Z_{i\gamma} = \frac{e^{j\phi_i}}{2} + \frac{jak}{4} \int_{\phi_0 - \Delta \phi} \left[ \hat{R}_{i0} \cdot \hat{n}_0' \right] e^{j\phi'} H_1^{(2)}(kR_{i0}) d\phi_0
\]

\[+ \frac{jak}{4} \sum_{n=-\infty}^{\infty} e^{jn\psi} \int_{0}^{2\pi} \left[ \hat{R}_{in} \cdot \hat{n}_n' \right] e^{j\phi'} H_1^{(2)}(kR_{in}) d\phi'_n. \]

and

\[
[g_i] = -H_z^i(\mathbf{r}_i). \quad (4.57)
\]

In order to compute the equivalent current density, an infinite summation of integrals needs to be calculated. The number of modes needed to be summed depends on the convergence rate of the integrals in (4.56). Because there are no closed form solutions, they are computed here using trapezoidal rule of numerical integration.

The real and imaginary values of the integral \( \frac{jak}{4} k e^{jn\psi} \int_{0}^{2\pi} [\hat{R}_{in} \cdot \hat{n}_n'] H_1^{(2)}(kR_{in}) d\phi'_n \), \( \frac{jak}{4} k e^{jn\psi} \int_{0}^{2\pi} [\hat{R}_{in} \cdot \hat{n}_n'] H_1^{(2)}(kR_{in}) d\phi'_n \), and \( \frac{jak}{4} k e^{jn\psi} \int_{0}^{2\pi} [\hat{R}_{in} \cdot \hat{n}_n'] H_1^{(2)}(kR_{in}) d\phi'_n \) in the element \( Z_{i\alpha}, Z_{i\beta}, \) and \( Z_{i\gamma} \) versus mode number are plotted in Fig. 4.11 respectively.

Here \( a = 0.75 \text{ mm}, d = 2.5 \text{ mm}, f = 915 \text{ MHz}. \) The angle of the incident wave is \( \theta = 45 \) degrees. It is observed that the \( n = 0 \) terms are dominant and the integrals are very small as \( |n| \) increases. This demonstrates that the summations in (4.56) will converge rapidly.
Figure 4.11: Real and imaginary value of the integrals vs mode number a) \( \frac{ja}{4}ke^{j
u\psi}\int_0^{2\pi}[\hat{R}_{in}\cdot\hat{n}_n']H_1^{(2)}(kR_{in})d\phi'_n \) in \( Z_{i\beta} \), b) \( \frac{ja}{4}ke^{j
u\psi}\int_0^{2\pi}[\hat{R}_{in}\cdot\hat{n}_n']e^{-j\nu'H_1^{(2)}(kR_{in})d\phi'_n} \) in \( Z_{i\alpha} \), c) \( \frac{ja}{4}ke^{j
u\psi}\int_0^{2\pi}[\hat{R}_{in}\cdot\hat{n}_n']e^{j\nu'H_1^{(2)}(kR_{in})d\phi'_n} \) in \( Z_{i\gamma} \). Here \( a = 0.75 \) mm, \( d = 2.5 \) mm, \( f = 915 \) MHz, and \( \theta = 45^\circ \).

Similar to the TMz polarization, the scattered H field needs to be represented with a summation of Floquet modes in order to compute the reflection and transmission coefficient. Substituting (4.14) into (4.51), and using identity

\[
\nabla H_0^{(2)}(k|\rho - \rho'|) = -\nabla' H_0^{(2)}(k|\rho - \rho'|),
\]

(4.58)
the scattered H field can be represented as
\[
H^s(x, y) = -\frac{ja}{4\pi} \sum_{n=-\infty}^{\infty} \oint_0^{2\pi} \mathbf{J}_n(x'_n, y'_n) \times \nabla' \left[ \int_{-\infty}^{\infty} \frac{e^{-jk|y-y'_n|}}{k_y} e^{jkz(x-x'_n)} dk_x \right] d\phi'_n. \tag{4.59}
\]
Where \(x'_n = x' + nd\), and \(y'_n = y'\). The current density on each local coordinate can be expressed as
\[
J_n(x'_n, y'_n) = J_{\phi_0}(\phi'_n) \delta'_n = -\sin \phi'_n J_{\phi_0}(\phi'_n) \mathbf{x}' + \cos \phi'_n J_{\phi_0}(\phi'_n) \mathbf{y}'.
\]
Applying (4.52), the scattered field can be written as
\[
H^s(x, y) = -\frac{ja}{4\pi} \sum_{n=-\infty}^{\infty} e^{jn\psi} \oint_0^{2\pi} J_{\phi_0}(\phi'_n)(-\sin \phi'_n \mathbf{x}' + \cos \phi'_n \mathbf{y}')
\]
\[
\times \nabla' \left[ \int_{-\infty}^{\infty} \frac{e^{-jk|y-y'_n|}}{k_y} e^{jkz(x-x'_n)} dk_x \right] d\phi'. \tag{4.60}
\]
For \(y \geq y'\), evaluating the gradient of the integrand in the spectral integral, (4.60) reduces to,
\[
H^s_z(x, y) = -\frac{ja}{4\pi} \sum_{n=-\infty}^{\infty} e^{jn\psi} \oint_0^{2\pi} J_{\phi_0}(\phi'_n)(-\sin \phi'_n \mathbf{y}' + \cos \phi'_n \mathbf{x}')
\]
\[
\times \left[ \int_{-\infty}^{\infty} \frac{e^{-jk|y-y'_n|}}{k_y} e^{jkz(x-x'_n)} dk_x \right] d\phi'. \tag{4.61}
\]
Applying (4.16) to (4.19), (4.61) can be represented as
\[
H^s_z(x, y) = \frac{a}{2d} \sum_{n=-\infty}^{\infty} e^{-jkyny+jkxn} \frac{1}{k_{yn}} \oint_0^{2\pi} J_{\phi_0}(\phi'_n)[-k_{yn} \sin \phi' + k_{xn} \cos \phi']
\]
\[
\times e^{jkyn a \sin \phi' - jkxn a \cos \phi'} d\phi', \tag{4.62}
\]
where \(k_{xn} = -k \cos \theta + \frac{2\pi n}{d}\),
\[
k_{yn} = \begin{cases} 
\frac{\sqrt{k^2 - k_{xn}^2}}{j} & k^2 - k_{xn}^2 > 0 \\
\frac{k^2 - k_{xn}^2}{j} & k^2 - k_{xn}^2 < 0.
\end{cases} \tag{4.63}
\]
The scattered field in (4.62) can be represented as an infinite summation of plane waves
\[
H^s_z(x, y) = \sum_{n=-\infty}^{\infty} C_n e^{jkyn y+jkxn x}. \tag{4.64}
\]
Where

\[ C_n = \begin{cases} \frac{a}{2dkyn} \int_0^{2\pi} J_0(\phi')[-k_{yn} \sin \phi' + k_{xn} \cos \phi']e^{jk_{yn}a \sin \phi' - jk_{xn}a \cos \phi'} d\phi' & y \geq y' \\ \frac{a}{2dkyn} \int_0^{2\pi} J_0(\phi')[k_{yn} \sin \phi' + k_{xn} \cos \phi']e^{-jk_{yn}a \sin \phi' - jk_{xn}a \cos \phi'} d\phi' & y < y' \end{cases} \]

(4.65)

The real and imaginary values of the reflection and transmission coefficients of the 0\(^{th}\) order Floquet mode vs frequency are computed and plotted in Fig. 4.12. Here \(a = 0.75\) mm, \(d = 2.5\) mm, and the incidence angle \(\theta = 45^\circ\).

From Fig. 4.12 we can see that at frequencies below 10GHz (include the RFID frequency band) the reflection and transmission coefficients are very close to 0 and 1 respectively. The result indicates that for TEz polarization the wire grating acts like free space, all the wave penetrates and none is reflected back. According to (4.26), at
the RFID frequency band only the 0\textsuperscript{th} order Floquet mode propagates regardless of the angle of the incident wave.

In conclusion, it may be assumed that the radial steel plies in the sidewall of a tire act as a polarization filter. When the electric field is parallel to the plies, it acts as a PEC surface. When the electric field is transverse it is transparent to the incident wave. These findings may be extended to an electric dipole antenna placed near the plies, as demonstrated in the next section.

4.3 Antenna Design

The broadside gain of a dipole antenna can be increased when positioned parallel to a metallic reflector. We have demonstrated that the wire grating acts as a solid PEC plate when the dipole is placed parallel to the cylinders. So the steel plies embedded in the sidewall can be utilized to increase the gain of a RFID tag antenna and extend the tag’s read range.

A regular truck tire is comprised by over a thousand steel plies in the radial direction and a cylindrical steel belt on its tread region. The radiation pattern of a dipole source in the vicinity of the metallic skeleton of such a tire is first studied using simulation software FEKO Suite 7.0. FEKO offers a wide spectrum of numerical methods, such as Finite Difference Time Domain (FDTD) \[64, 65\]; Finite Element Method (FEM)[66]; Method of Moments (MoM)[67], etc. In this work, Method of Moments solver is used. The simulation model is shown in Fig. 4.13 (a). The tire has more than 1500 steel plies on each sidewall. In this simulation the tire rubber is removed to reduce the computational time. Fig. 4.13 (b) plots the radiation pattern of an electric dipole source in free space, the maximum gain is 1.76 dB. The dipole
source is placed parallel and transverse to the steel plies respectively. The 3D and 2D radiation patterns are plotted in Fig. 4.14 and Fig. 4.15.

Figure 4.13: (a) Structure of the tire in the simulation. (b) Radiation pattern of a dipole source in free space, maximum gain is 1.76 dB.

Fig. 4.14 illustrates that for the dipole perpendicular to the steel plies, a certain amount of energy penetrates through the tire. Antenna still exhibits omni-directional radiation pattern with a broadside gain \((\phi = 90^\circ \, \theta = 90^\circ)\) of 1.27 dB. Fig. 4.15 shows that for the dipole parallel to the steel plies, the steel plies act as a PEC plate with no energy penetrating through. The antenna’s broadside gain \((\phi = 90^\circ \, \theta = 90^\circ)\) is increased to 9.1 dB. These simulation results are consistent with our analytical analysis. The steel plies of the tire can be utilized as a reflector to increase the gain of RFID tag antennas.
A T-slot antenna (introduced in Chapter 3) is optimized to be mounted parallel to the steel plies. This antenna provides broad bandwidth and can be tuned conveniently by changing the length of the slot. An effective way to increase an antenna’s gain and maintain proper impedance matching is to place the antenna horizontally $\lambda/4$ above a large metal reflector. Substrates with high permittivity can be used to reduce the phase velocity, therefore decreasing the distance between the antenna and the reflector. Here a substrate with a thickness of 12 mm is attached to the T-slot antenna. The substrate is made of the same material as the sidewall rubber with $\epsilon_r = 10.62$. The thickness of the sidewall rubber is 10 mm on average. As such, the distance between the antenna and steel plies is about 22 mm, close to the quarter wavelength $\lambda_g/4 = 25$ mm in the material.
4.3.1 Measurements and Results

The T-slot antenna was fabricated and connected to a RFID chip. The performance of the tag mounted on the sidewall of a tire, with and without the rubber substrate, was evaluated and compared with the T-slot tag discussed in Chapter 3. The threshold power was measured. The results are given in Table 4.1. As seen employment of the substrate effectively improved the performance of the tag parallel to the steel plies. More importantly, the parallel tag with the substrate achieved a threshold power 2 dB lower than the perpendicular tag introduced in Chapter 3. The trade-off is that the radiation is primarily broadside to the one side of the tire, and very weak in the opposite direction.
<table>
<thead>
<tr>
<th>Tag configuration</th>
<th>Threshold power</th>
</tr>
</thead>
<tbody>
<tr>
<td>New parallel tag w/out rubber pad</td>
<td>26 dBm</td>
</tr>
<tr>
<td>New parallel tag w/ rubber pad</td>
<td>20 dBm</td>
</tr>
<tr>
<td>T-slot tag in Chapter. 3</td>
<td>22 dBm</td>
</tr>
</tbody>
</table>

Table 4.1: Threshold Power Vs Thickness for a T-slot Tag Mounted on the Sidewall of a Tire with and without a Rubber Pad Substrate.

### 4.4 Summary of Chapter 4

This chapter studied the antenna’s behavior when the tag is close to a wire grating. A theoretical analysis was performed in 2D and the problem was addressed in TM and TE polarization, respectively. Free space Green’s function was replaced with its plane wave spectral expansion. The periodicity constraint was imposed to reduce the solution to a sum of Floquet modes. The theoretical results indicated that the wire grating acts as a solid PEC plate and free space when the antenna is placed parallel and transverse respectively. A T-slot antenna was optimized to validate the theory. The experimental results indicated that the wire grating could be utilized as a reflector to increase antenna gain, thereby further extending the tag’s read range.
Chapter 5: UHF RFID Tag Antennas for Automotive Tire Applications

Two RFID tag antennas are introduced in this chapter. The antennas are designed based on the position of placement. The first half of the chapter discusses the tag embedded in the sidewall of the tire. The designed tag antenna achieves good electrical performance and possesses a robust mechanical structure. A tag antenna operating close to large metallic objects is proposed in the second half of the chapter. The effects of a large conductive plate on a dipole-based antenna’s performance are studied. The experimental results indicate that the designed tag antenna obtained exceptional performance when placed in the vicinity of a metal plate.

5.1 Flexible and Stretchable UHF RFID Tag Antenna for Automotive Tire Sensing

RFID (Radio Frequency IDentification) implementation has diversified due to the increasing demand of automatic inventory tracking and item identification. One of the applications is to integrate RFIDs with rubber tires to verify their history and track the inventory. As opposed to barcodes, RFIDs do not require line-of-sight to read the information. Hence the tag can be embedded inside the tire rubber. Further, by combining RFIDs with sensor technologies, it is possible to monitor the
tire condition including temperature, pressure, stress and vibration. This information can be wirelessly communicated to the driver or tire inspectors.

Existing RFID tags for tires do not operate well in the vicinity of adjacent tires, especially when they are mounted on the inner tire of a dual-wheel situation (placements S3 and S4 in Fig. 5.1). This is due mainly to the frequency shifting of the impedance matching curve of the tag antenna. Tires are composed of multiple layers of rubber which exhibit different dielectric properties and thicknesses along the sidewall as shown in Fig. 5.2. A tag antenna designed for tire surface mount may be detuned when it is embedded in the side wall. Therefore, the tag’s performance will be degraded. There are several prior works on RFID tag antennas. Three RFID tag antenna designs were introduced in [68] for automotive tire applications. However, the authors treated the tire as a homogenous dielectric medium. Therefore, the tag is expected to be severely detuned and fail to operate when embedded in real tires. Bandwidth issues of tag antennas for tires were studied in [69]. However, no applicable design was proposed.

![Figure 5.1: Tires on dual-wheel situation, existing tags cannot be detected when mounted on sidewall S3 and S4 of the inner tire.](image)
Figure 5.2: Cross section of a tire: a tire is composed of multiple layers of rubber, whose thickness changes along the sidewall.

Another challenge in designing tags for tires is the mechanical characteristics of the tag. As vehicles are constantly experiencing different road conditions, tags must be structurally tough, durable, and flexible to endure changes in pressure, stress and temperature that would possibly deform the antenna. At the same time, the presence of an RFID tag should not damage the mechanical strength or balance of the tire and cause safety issues. The structure of the metallic antenna printed on plastic film [70] does not satisfy these requirements. It is therefore of interest to develop a practical solution for various tire conditions.

In this work, a flexible UHF RFID tag antenna is proposed based on the quantitative study of dielectric materials discussed in Chapter 2 and 3. The designed antenna has broad impedance bandwidth, so the performance of the tag is less susceptible to the type of the tire and the location of placement. In addition, a copper wire based RFID antenna is fabricated. The antenna exhibits excellent mechanical strength and flexibility, which favors its use in RFID tire applications.
5.1.1 Tag Antenna Design Approach

A unique challenge of designing antennas for tire tags is the consideration of their mechanical characteristics. RFID tags will be embedded in the sidewall of the tire during the tire manufacturing procedure. Therefore, the antenna of a tire tag should be stretchable and mechanically robust to withstand external forces introduced during the construction process, as well as different road conditions when the tire is in service.

Conventional RFID tags printed on thin plastic film cannot be embedded in the sidewall. The plastic film does not bond with the rubber, thus creating a cavity which damages the mechanical integrity of the tire. Paper-based RFID tag antenna proposed in [37] does not jeopardize the structure of the tire, however, the antenna is fragile, and may not survive the manufacturing procedure.

Our first approach was to print the antenna directly on the sidewall rubber as shown in Fig. 5.3. To prove the concept, Cu lines were printed on the stretched and relaxed rubber samples. The assumption was that Cu printed on the stretched rubber sample would maintain its continuity when the rubber was relaxed. This was an attractive approach. However, the method was proved to be unsuccessful. As shown in Fig. 5.3, many visible cracks appeared even when printing the Cu line on the pre-stretched rubber samples.

RFID tag antennas can be constructed with copper wires to increase their flexibility and mechanical robustness. As opposed to printed antennas, copper wire-based antennas do not need to be attached to substrates (such as plastic films). Furthermore, copper wire antennas can be designed to incorporate flexible structures (such as helix antennas). To prove the concept, a simple meander-line antenna was fabricated with both planar copper tape and a 24 AWG wire as shown in Fig. 5.4. The
Figure 5.3: Cu lines printed directly on the pre-stretched and relaxed rubber samples. The experimental result indicates that when connected to the same RFID tag chip, these two tags achieved same performance.

Figure 5.4: Simple meander line antenna fabricated with planar copper foil and copper wire.

For convenience, planar antennas are designed in the simulation tool. The electrical equivalent radius for the planar antenna can be obtained as [1],

$$a_e = 0.25a$$  \hspace{1cm} (5.1)

where $a$ is the width of the planar copper antenna and $a_e$ is the effective radius of its wire counterpart.
As most RFID chips exhibit high capacitance, tag antennas are designed with an inductive component to compensate for the chip’s capacitance. T-match is a popular method to provide inductance. It includes a center-fed smaller dipole connected to a larger dipole, the two being separated by a small distance as shown in Fig. 5.5 (a). The input impedance at the center is calculated

$$Z_{in} = \frac{2Z_t(1 + \alpha)^2Z_A}{2Z_t + (1 + \alpha)^2Z_A}. \quad (5.2)$$

Here $Z_t = jZ_0\tan\left(\frac{k'l}{2}\right)$ is the transmission line mode impedance; $Z_0$ is the characteristic impedance of two wire transmission line $Z_0 \approx 276\log_{10}\left(\frac{2s}{\sqrt{ww'}}\right)$; $Z_A$ is the input impedance of the larger dipole without presence of the smaller dipole, and $\alpha$ is the current division factor $\alpha = \frac{ln(s) - ln(w/2)}{ln(s) - ln(w'/2)} [1]$.

The equivalent circuit schematic of the T-match network is plotted in Fig. 5.5(b). From Fig. 5.5(b) we can see that the T-match network behaves as an impedance transformer. The impedance of the larger dipole is stepped up by a ratio of $(1 + \alpha)$. Therefore, the input impedance at the center is the result of twice the transmission line impedance in shunt with $(1 + \alpha)^2Z_A$. For designing RFID tag antennas, high
inductance could be achieved by manipulating the dimension of the smaller dipole and the space between two dipoles.

![Figure 5.6: Proposed flexible broadband RFID tag antenna with diamond shaped loops.](image)

The designed tag antenna is shown in Fig. 5.6. The antenna was fabricated with a 24 AWG copper wire. This antenna possesses exceptional electrical and mechanical properties. It is optimized using Ansys HFSS to function on or inside a wide range of tire rubber materials. The power reflection coefficient of the antenna is plotted in Fig. 5.7. As shown in the plot, the antenna has a wide impedance bandwidth of 230 MHz in free space. Importantly, when placed on a rubber sample ($\epsilon_r = 10.62$, $\tan\delta = 0.163$ and thickness = 2 mm), it achieves a bandwidth of 195 MHz. In this design the diamond loop in the center is used to provide sufficient inductance for impedance matching. The diamond shaped loops at the ends improve the antenna’s bandwidth and further reduce its overall size.

This antenna is also mechanically robust. The incorporated diamond shaped loops enable the antenna to endure vertical forces. The meander-line structure introduces flexibility, making the antenna stretchable in the horizontal dimension. It also reduces the overall size of the antenna.
5.1.2 Experimental Results

As shown in Fig. 5.6 the fabricated tag antenna was connected to a RFID chip. The tag was tested and compared with a speedy patch which is a popular commercial RFID tire tag.

Read Range Test

The tags under test were placed on the exterior surfaces of a dual wheel configuration as shown in Fig. 5.1. The read range of the tag placed on each surface was measured. The results are listed in Table 5.1.

Table 5.1 shows that compared with the commercial speedy patch RFID tag, the designed tag achieved much improved read range on all surfaces. Especially when placed on the surfaces of the inner tire (S3 and S4), the designed tag achieved a read range of 4 feet, while the speedy core could not be read.
<table>
<thead>
<tr>
<th>Antenna/Surface</th>
<th>S1</th>
<th>S2</th>
<th>S3</th>
<th>S4</th>
</tr>
</thead>
<tbody>
<tr>
<td>Speedy Patch</td>
<td>7</td>
<td>1</td>
<td>cannot be read</td>
<td>cannot be read</td>
</tr>
<tr>
<td>This work</td>
<td>14+</td>
<td>8</td>
<td>4</td>
<td>4</td>
</tr>
</tbody>
</table>

Table 5.1: Read Range of the RFID Tags on Dual-Wheel (ft)

**Embedded Performance Test**

RFID tags may be embedded in the sidewall of a tire during the tire construction process. A fixture was made to simulate such an environment. As shown in Fig. 5.8, the fixture was comprised by two pieces of foam and six layers of rubber samples. Two different kinds of rubber samples were used. The tag under test was sandwiched in the middle of the layers, and the layers were then squeezed by two pieces of foam. The fixture was placed 5 ft away from the reader antenna, and the threshold power of the tag was measured. The results are listed in Table 5.2.

Figure 5.8: Fixture used to simulate the environment of an embedded tag.
Table 5.2: Threshold Power of the Antenna Embedded in the Fixture

<table>
<thead>
<tr>
<th>Speedy Patch</th>
<th>Squiggle®</th>
<th>This work</th>
</tr>
</thead>
<tbody>
<tr>
<td>28 dBm</td>
<td>32 dBm</td>
<td>19 dBm</td>
</tr>
</tbody>
</table>

Table 5.2 indicates that the designed copper wire-based tag outperformed the commercially available tags. It achieved a much lower threshold power when embedded in the rubber layers. Therefore the designed tag is expected to deliver longer read range when embedded in the sidewall of the tire.

5.2 RFID Tag Antenna for Mounting on Metallic Objects

It is desirable to place RFID tags in the crown area for tire recycling/retreading procedure. As shown in Fig. 5.9, the crown area consists of two parts: the tread region and the carcass. Perhaps, the best location to mount a tag is the interface of these two parts. It is noted that a steel belt is embedded in the carcass 2 mm below the interface. Therefore, it is essential to design a tag antenna functioning close to metallic objects.

Metal mountable RFID tags have been researched extensively. A slim RFID tag antenna design was introduced in [71], an upgraded version using looped bowtie is discussed in [72]. An electrically small tag antenna design was proposed in [73]. Ground planes and shortening pins were used in these designs. Similar techniques can be found in [74, 75]. However, this type of tags are not practical in our application. The structures of these antennas are complicated which increases the cost and fabrication difficulty. Also such tags are generally thick. Thus, they are suitable for mounting
directly on metal objects instead of being sandwiched in between dielectrics. Several low profile and conformal “label-type” antenna designs were proposed in [76, 77, 78]. However, because these antennas were designed based on specific substrates, they may not be applicable to our application.

In this work, we study the antenna’s behavior close to large conductive objects. A low-profile, planar, and conformal tag antenna is designed. The designed antenna does not incorporate a substrate, it utilizes the dielectric material above the steel belts to achieve impedance matching. The experimental results show that the designed antenna achieved exceptional performance when placed 2 mm away from a conductive plate.

5.2.1 Antenna Design Approach and Proposed Tag Antenna

The presence of a large electric conductor has a profound impact on an antenna’s performance. The realized gain of an antenna is determined by the antenna’s directivity, radiation efficiency and impedance matching [1]. When the space between the
antenna and the metal plate \( s \to 0 \), the radiation resistance of the antenna reduces significantly. This severely degrades the radiation efficiency and impedance matching, thus decreasing the realized gain. Taking Alien Squiggle\textsuperscript{®} RFID tag antenna as an example, HFSS simulation results show that at 915 MHz, the impedance of the tag antenna is \( 19 + 193j \) in free space. When this antenna is placed 2 mm above a large metal plate, its impedance changes to \( 0.01 + 136j \). It is also noted that the antenna’s radiation resistance is close to zero over a large frequency band. So this antenna cannot be modified to achieve matched impedance. An effective method to improve the radiation efficiency is to place the horizontal dipole antenna \( \lambda/4 \) above the metal reflector. This method is used in Chapter 4, where high dielectric rubber pads are inserted between the antenna and the reflector. However, in this approach \( s = 22 \) mm, much larger than the required distance.

An antenna’s radiation efficiency can be effectively improved by employing a thin lossy material (\( \tan \delta \neq 0 \)) in between the antenna and the metal reflector as demonstrated in [76, 77, 78]. In our application, a thin (2 mm) dielectric layer exists in between the antenna and the steel belts. The electrical properties of the material is \( \epsilon_r = 10.63 \) and \( \tan \delta = 0.162 \). This dielectric layer can be utilized to increase the antenna’s radiation resistance and improve impedance matching.

Antennas with matched impedance may not achieve high efficiency. Under perfect matching condition, the gain of the Squiggle\textsuperscript{®} antenna and the antenna with circular loops (proposed in Chapter 6) is -45 dB and -20 dB respectively when placed on the grounded dielectric slab. The meander line structure in these antennas introduces more contact with the material that degrades the radiation efficiency.
T-slot antenna proposed in Chapter 3 and 4 delivers sufficient gain when placed on the grounded dielectric slab. This antenna can be analyzed as an embedded T-match antenna [79]. Antenna tuning is achieved by changing its overall length $l_{body}$ and the length of the slot $l_{slot}$. An optimized T-slot antenna is designed using simulation software CST studio. The power reflection coefficient and radiation pattern of this antenna placed on the grounded 2 mm thick dielectric slab are shown in Fig. 5.10. In this design, $l_{body} = 60$ mm and $l_{slot} = 24$ mm.

![Power reflection coefficient and radiation pattern](image)

Figure 5.10: (a) Power reflection coefficient of the T-slot. (b) Radiation pattern of the T-slot. The length of the antenna is 60 mm, the length of the slot is 24 mm.

The plots show that the antenna is tuned at 915 MHz with a gain of -11 dB (under perfect matching condition). It is also noted that this antenna possesses another resonance at higher frequency. Therefore, there exists another design which could also achieve matched impedance. The power reflection coefficient and radiation pattern of another T-slot antenna with the same width but body length of 100 mm and slot length of 80 mm are plotted in Fig. 5.11.
Figure 5.11: (a) Power reflection coefficient of the T-slot. (b) Radiation pattern of the T-slot. The length of the antenna is 100 mm, the length of the slot is 80 mm.

From the plot it can be seen that the large T-slot antenna is also matched at 915 MHz. Importantly, the gain of this antenna is about 6 dB higher than that of the small T-slot. It can be explained by analyzing the current distribution on the antenna. Fig. 5.12 plots the simulated current distribution on the small and large T-slot tag antennas. As can be seen from the plots, current is mainly distributed on

Figure 5.12: Current distribution of the T-slot antenna (a) small T-slot (b) large T-slot.

the edges of both antennas. There are four radiating edges on each antenna. For
the small T-slot antenna \((l_{\text{body}} = 60 \text{ mm})\), the current on the top two edges is 180 degrees out of phase from that on the bottom. For the large T-slot \((l_{\text{body}} = 100 \text{ mm})\) the current is in phase everywhere. The antenna with current flowing in the same direction achieves higher radiation efficiency.

### 5.2.2 Experimental Results

The designed antennas were connected to RFID chips and tested on a fixture. As shown in Fig. 5.13, the fixture is consisted of a 2 mm thick dielectric pad and a large metal plate. The dielectric pad had the same electrical properties as the rubber in the crown region. Threshold power test was used, while the distance between the reader and the fixture was set to 5 ft. Tags under test were: Alien Squiggle®️, the commercial Speedy patch, and the small and large T-slot tags. The results are shown in Table 5.3.

![Figure 5.13: The tags are tested on a fixture which consists of a 2 mm thick dielectric pad and a large metal plate.](image)

Table 5.3 shows that the Squiggle®️ and Speedy patch tags cannot be accessed even with the highest output power (32 dBm). The designed tag with small T-slot
antenna had a threshold power of 30 dBm. The tag with large T-slot antenna achieved exceptional performance, the threshold power was 23 dBm. This antenna may be used for tags embedded in the crown area.

5.3 Summary of Chapter 5

In this chapter, two antennas are designed for automotive tire applications. For the tag embedded in the sidewall, a copper wire-based antenna is designed. The designed antenna incorporates robust mechanical structure and possesses excellent electrical properties. A T-slot antenna is designed/optimized for placement in the crown region of the tire. The design approach of the antenna operating close to metallic objects is discussed. The optimized tag achieved exceptional performance when placed 2 mm away from a large metal plate.
Chapter 6: Textile Based RFID Tag Antenna

With recent interest in wearable technologies, conformal and flexible antenna designs attract more attention. Generally speaking, for wearable applications, antennas are usually designed using the conventional techniques, but fabricated with special materials. Among all available approaches [80, 81, 82], embroidery of metal-coated polymer fibers (E-fiber) is proved to be an effective solution [83]. Prototypes developed using this approach exhibit comparable electrical properties to their copper-based counterparts, while maintaining excellent mechanical performance.

E-fiber based antennas are not limited to body worn applications. They are also applicable to other harsh environment implementations where antennas are required to withstand deformations and changes while maintaining good electrical performance. In this chapter, E-fibers are used to “print” RFID tag antennas for elastic materials. For the first time, herewith, a single E-fiber thread is employed to embroider a stand-alone RFID wire antenna.

6.1 Broadband Textile-Based Passive UHF RFID Tag Antenna for Elastic Material

RFID (Radio Frequency IDentification) technology has been implemented widely for item identification and inventory tracking. Recently there is also strong interest
for integrating RFIDs with sensors to wirelessly monitor conditions of structures or
devices while in service and possibly hostile environments [84, 85]. Generally speak-
ing, the RFID tag/sensor antenna is only designed for a given material to optimize
impedance match and read range. However, a tag’s performance deteriorates signifi-
cantly when mounting location or nearby material changes. Therefore, it is desirable
to design a broad impedance bandwidth RFID tag antenna for operation in a variety
of material structures.

Broadband RFID tag antennas have been considered before, but primarily for
covering multiple bands [86, 25, 87]. Here we focus on a broadband UHF RFID tag
antenna delivering a bandwidth of 263MHz (free space) for the purpose of maintaining
performance over a wide range of materials. Previous UHF RFID antennas have been
designed for operation on metallic objects [88, 89], extreme environments [90], and
metallic/dielectrics [91]. However, there is no solution in the literature for UHF
RFIDs that can operate in different dielectrics, including elastic materials.

Here, we present an RFID tag antenna constructed from conductive textiles for
flexible and stretchable designs [92]. For the first time, herewith, a single E-fiber
thread is employed to embroider a stand-alone RFID wire antenna. This is an attrac-
tive approach since former technologies based on stretchable conductive wires (e.g.,
liquid metal [93], copper traces [94], silver nanowires , etc.) are prone to fatigue
and wear. Using the proposed textile wires, the RFID antenna exhibits excellent
mechanical properties that favor its application in hostile environments.
In this work, the effects of dielectric substrates or superstrates on the RFID tag’s performance are discussed. A textile-based broadband RFID tag antenna is presented. The proposed tag antenna is designed incorporating three aspects: 1) electrical properties: broad bandwidth and electrically small size; 2) mechanical properties: elasticity, flexibility, and mechanical robustness; and 3) fabrication complexity: low profile and easy to mass produce. The designed antenna achieves a wide frequency bandwidth making it functional over a wide range of materials. Various tags are fabricated with copper wire and E-fiber threads. Their performance is evaluated and compared with a commercial RFID tag for tire applications. Measurements indicate that the designed antenna achieves significant improvement as compared to a conventional copper one, without loss of flexibility.

6.1.1 Proposed Tag Antenna

![Design of the proposed RFID tag antenna.](image)

The antenna is designed and optimized based on the three key aspects described in the introduction, namely, 1) electrical properties, 2) mechanical properties, and 3) fabrication complexity. As illustrated in Fig. 6.1, the antenna is composed of three parts: the circular end-loading loops at both ends, the meander-line dipole arms, and
a tuning loop in the center. Design optimization was performed using ANSYS HFSS software.

![Graph showing power reflection coefficient](image)

Figure 6.2: Power reflection coefficient between the RFID chip and antenna. The red curve is in free space and the blue curve is when the tag is mounted on a 4 mm thick material with $\epsilon_r = 10, 13, 16$.

1) Electrical properties: The designed tag antenna is electrically small (87 mm × 19 mm) and broadband. As such, it is able to maintain tuned behavior over a wide range of mounting locations and nearby materials. Fig. 6.2 plots the power reflection coefficient between the RFID chip and antenna. Specifically, in Fig. 6.2, the red curve and the blue curve correspond to the power reflection coefficient of the antenna in free space and when mounted on a 4 mm-thick dielectric with $\epsilon_r = 10, 13, 16$ respectively. As seen, the tag antenna delivers a 3 dB bandwidth of 263 MHz in free space. More importantly, it is tuned to maintain a good impedance match over the US RFID band (902-928 MHz) with sufficient bandwidth (171 MHz $\epsilon_r = 10$ ) to allow variability in the underlying or surrounding material. The impact of dielectric materials placed in the vicinity of the tag antenna and the need for a broadband antenna to mitigate these effects are discussed in the previous chapters and [95, 96].

91
In the proposed antenna design (see Fig. 6.1) the circular loop in the center is used to provide sufficient inductance for impedance matching to the RFID chip. A RFID tag consists of two parts, an antenna and a RFID chip. For maximum power delivery, it is necessary to match the antenna impedance to that of the chip. Therefore, we must minimize the power reflection coefficient given by (2.2). Generally speaking, the RFID chip does not incorporate a matching circuit. Its impedance is highly capacitive and can be modeled as a resistor and a capacitor in parallel as shown in Fig. 6.3(a). Based on transmission line theory, the circular center loop can be viewed as a short stub, and therefore serves as a parallel inductor. The equivalent circuit model of the antenna and the center loop is provided in Fig. 6.3(b). We note that the length of the loop was optimized such that complex conjugate matching can be achieved at the desired frequency.

![Figure 6.3: Equivalent circuit model for (a) the chip impedance, and for (b) the dipole antenna loaded with a short stub.](image)

The meander-line structure was adopted to reduce the tag’s size [79]. As compared with a straight dipole of the same length, the resonance frequency can be significantly reduced using meander-lining, typically at the expense of narrower bandwidth. The circular end-loading wire loops are employed to increase the impedance bandwidth.
of the antenna. The end-loading (or top-loading) of the dipole serves to broaden its bandwidth [97]. Fig. 6.4 (a) compares the designed antenna with and without end-loading. As would be expected, the end-loading loops improve impedance bandwidth. Furthermore, the end-loading reduces the overall size of the antenna. As depicted in Fig. 6.4 (a), the antenna with end-loading is 3 mm shorter than the one without end-loading.

2) Mechanical properties: The tag antenna was designed to be mechanically robust. This implies that its physical integrity and electrical performance remain intact when the tag is subject to external forces. To do so, the shape of the meander-lines was optimized such that the tag’s performance is preserved even if the antenna undergoes deformation. Importantly, the meander line structure enables stretching of the tag. Fig. 6.4(b) compares the performance of the original antenna and the one stretched by 10% (overall length remains the same). As expected, the simulations show that the antenna’s resonance is shifted downwards in frequency after stretching. However, its bandwidth still overlaps the operating frequency band. The end-loading circular loops also allow stretching in any dimension. Fig. 6.4(c) compares the antenna with the original and elongated loops. The elongated loops are of rectangular shape, but maintain the same perimeter.

3) Fabrication complexity: The designed tag antenna incorporates a low profile structure such that fabrication complexity and manufacturing cost are reduced. Previous textile-based RFID tag antennas use densely stitched conductive textile to produce the surface of planar antennas. In this work, the tag antenna only requires a single thread of the conductive textile. This is because end-loading loops were used instead of patches in order to avoid rigid planar surfaces. The adoption of the loops
was inspired by the observation that for end-loading patches, the current is primarily distributed on the edges. Fig. 6.4 (d) compares the antenna when the end-loading is formed of loops and patches. The results show that the antenna with end-loading loops achieves equally good performance as that with patches. When fabricated using single textile threads, the circular loop is more attractive as it avoids sharp corners. That is, rectangular loops are challenging to fabricate. Therefore, circular loops were adopted to avoid sharp corners and increase fabrication accuracy. The meander-line does not require embroidering sharp edges either. As will be shown in the following section, the shape of the antenna could be preserved after cured with the polymer material.

Figure 6.4: Power reflection coefficient of the designed antenna: (a) with vs. without end-loading, (b) non-stretched vs. stretched by 10%, (c) with original vs. elongated end-loading loops, and (d) with end-loading loops vs. patches.
6.1.2 Textile Based Tag Antenna

To realize a flexible and stretchable version of the proposed broadband RFID tag, we used electrically conductive metal-polymer fibers (E-fibers) embedded in elastic polymer. Flexibility and elasticity are important for RFID tags that operate in hostile environments subjecting them to mechanical deformation. These E-fibers have already been validated for some textile antennas and sensors, indicating excellent mechanical strength, flexibility, low DC resistance (about $0.5 - 0.7\Omega/m$), and low loss at RF frequencies. They have been found to have comparable performance to their copper counterparts. Polydimethylsiloxane (PDMS) was employed as the polymer material ($\epsilon_r = 3$, $\tan\delta < 0.02$) because it is flexible and allows for antenna elongation of about 10% of its original size.

![Figure 6.5: Fabrication procedure and resulting flexible single E-fiber constructed RFID tag antenna.](image)

The fabrication procedure may be summarized as follows [98]: (a) convert the antenna CAD file into a digital stitching pattern, or, equivalently, needle path, (b) use a standard sewing machine to embroider the prescribed antenna pattern onto
a polyester fabric (as an intermediate support), (c) solder the RFID chip to the antenna, (d) melt the non-stretchable polyester fabric via heating (melting point ≈ 250°C) without damaging the E-fiber (melting point ≈ 600°C), (e) pour 1-2 mm thick liquid PDMS polymer on the RFID tag, and (f) cure the PDMS polymer by placing it on an elevated temperature (120°C) hot plate. The fabrication procedure and resulting product is the flexible and stretchable E-fiber RFID prototype tag as shown in Fig. 6.5. The prototype was found to withstand repetitive flexing and stretching. The surrounding polymer also preserves the integrity of the E-fiber and protects it from corrosion.

6.1.3 Experimental Results

Four versions of the designed antenna were fabricated and soldered to G2XM RFID chips (impedance: 16-148j): a copper wire antenna and an E-fiber antenna, each with and without being embedded in polymer. Shown in Fig. 6.6 is the copper wire version embedded in polymer. The thin (1-2 mm) and low-permittivity ($\epsilon_r = 3$) polymer was used to support the mechanical integrity of the tag. The tag’s performance was evaluated using read range and threshold power tests (the minimum reader output power required to detect the tag). An Impinj Speedway RFID reader was used to collect data. All measurements were conducted in a large RFID lab.

Read Range Test

In this test the fabricated copper wire tag embedded in polymer and a commercially available RFID tag for tires (the Speedy Core) were tested and compared. Both tags were measured on truck tires. Truck tires are constructed of multiple layers of rubber with different dielectric constants and thicknesses. It is necessary to have a
Figure 6.6: Copper wire version prototype embedded in polymer.

Figure 6.7: (a) The surfaces of the dual-wheel are labeled as S1 to S4. (b) Six positions are labeled as P1 to P6 on the exterior surface.

broadband antenna such that the RFID tag maintains good performance over a wide range of tires.

The measurement setup is shown in Fig. 6.7. The tag under test was mounted at the same position on the outer surfaces S1, S2, S3, and S4 of two truck tires. The space between the two adjacent tires is 15 cm, equal to the separation of dual tires on trucks. The output power of the RFID reader was set to be 30 dBm. The RFID reader antenna was moved away from the tires until the tag cannot be detected. The distance between the reader antenna and the tire on the left is the read range of the tag. Measurement results are shown in Table 6.1.
As shown in Table 6.1, the designed tag has much longer read range than the Speedy Core. More importantly, the designed tag was readable at all positions whereas the Speedy Core was not readable at positions S3 and S4.

**Threshold Power Test**

<table>
<thead>
<tr>
<th>Antenna/Location</th>
<th>P6</th>
<th>P4</th>
</tr>
</thead>
<tbody>
<tr>
<td>Copper Wire</td>
<td>20</td>
<td>21</td>
</tr>
<tr>
<td>Embroider (Single &amp; double thread)</td>
<td>21</td>
<td>22</td>
</tr>
<tr>
<td>Copper wire W/ coating</td>
<td>21</td>
<td>21</td>
</tr>
<tr>
<td>Embroider W/ coating (Single &amp; double thread)</td>
<td>20</td>
<td>21</td>
</tr>
</tbody>
</table>

Table 6.2: Threshold Power Test of the Designed Tag Fabricated with Different Material (dBm)

In this test, all four fabricated tags were evaluated for threshold power sensitivity. The threshold power of a tag is the minimum reader output power required to detect the tag at a fixed range. A lower power indicates a better performing tag. The tags under test were placed in turn on surface S1 of the tire. The reader antenna was fixed 5 feet from the tire. The results are given in Table 6.2. The threshold power is nearly
the same for all four versions of the tag. We may therefore conclude that all of the fabricated tags would outperform the commercial Speedy Core tag.

**Stretch Test**

The performance of the copper wire prototype was tested in an elongated condition. As the tags will be embedded in the sidewall of a tire during the manufacturing process, the tag antenna would suffer a certain amount of deformation along its longitudinal direction. Therefore, it is very important that the tag antenna maintains its electrical properties after the embedding process. Threshold power measurements were carried out and listed in Table 6.3. The designed tag achieved the same good performance after being stretched by nearly 10%.

<table>
<thead>
<tr>
<th>Antenna</th>
<th>Length</th>
<th>$P_{th}$ @ P6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Original</td>
<td>85 mm</td>
<td>20 dBm</td>
</tr>
<tr>
<td>Stretched</td>
<td>93 mm</td>
<td>20 dBm</td>
</tr>
</tbody>
</table>

Table 6.3: Stretch Test of the Copper Wire Tag

**Embedded in the Real Tires**

The fabricated tags were embedded in the sidewall of a real truck tire during the tire built process. Fig. 6.8 shows the X-ray plot of the embedded tag.

The read range of the designed tags and a commercial tag were measured and compared. The maximum read range of the commercial tag is 1.2 m, the maximum read range of the deigned tag is 2.8 m. The designed tag achieved significant improvement in read range.
6.2 Investigation of Broadband Textile Based Antenna Configurations for Passive UHF RFID Tags

Recently, conductive textile materials have been used for RFID tag antenna fabrication due to their lightweight and conformity [99, 100]. The benefits of using textile-based antennas are discussed in 6.1. Generally speaking, textile/embroidered antennas are designed using the same approach of designing copper based antennas, but fabricated with special conductive materials. It will be demonstrated in this chapter that antennas can be re-designed to use least amount of conductive threads to reduce the cost while maintaining the same performance.

Most existing dipole-based embroidered RFID tag antennas use densely stitched conductive textile materials to form a solid surface which resembles a conventional planar antenna fabricated using copper foil. In such a configuration, the density of the stitch and the sewing pattern are critical to the antenna’s performance [100, 101]. This is the most direct approach of antenna fabrication, because most RFID tag antennas incorporate planar structures. However, there are a few drawbacks associated with this configuration. 1) High cost: Conductive textile materials are more expensive than copper. Thus, adoption of densely stitched textiles increases the cost of the hardware. 2) Conformal but not stretchable: Antennas cannot be stretched due to
the solid surface. Thus, they still cannot be used in materials subject to external
forces and deformations. 3) Difficult and time-consuming to fabricate: The density
of the stitch and sewing patterns need to be controlled during the manufacturing
procedure.

A novel method of fabricating dipole-based RFID tag antennas with just a single
thread of conductive material is proposed in 6.1. The material cost is significantly
reduced. In this section, we investigate antennas possessing solid planar surfaces that
cannot be replaced by a single thread of E-fiber. It will be demonstrated that these
antennas can be redesigned such that they can be fabricated with least amount of
conductive threads and obtaining the same electrical performance.

6.2.1 Antenna Design

The T-slot antenna is optimized to demonstrate our design approach. As discussed
in the previous chapters, this antenna exhibits very broad impedance bandwidth. It
has a planar structure, while antenna tuning is closely related to the length of the
body, length of the slot, as well as the width of the body. Therefore, the single thread
fabrication process proposed in 6.1 cannot be employed. As illustrated in 5.2, even
though the antenna has a solid, planar, patch type structure, current is distributed
primarily on the four radiating edges. A redesigned sparse T-slot antenna is shown in
Fig. 6.9. All the parameters of the antenna remain unchanged except that the antenna
only incorporates the radiating edges. It is noticed that the resonate frequency is
shifted to the lower band, because current travels longer distance. The sparse T-slot
antenna is then tuned to operate at the desired frequency by adjusting the length
of the slot. The power reflection coefficient and gain pattern of the sparse and solid
patch type T-slot antennas are compared and plotted in Fig. 6.10. Antennas are placed on a 4 mm thick dielectric slab with $\epsilon_r = 6$.

Figure 6.9: Structure of the sparse T-slot antenna.

![Figure 6.9: Structure of the sparse T-slot antenna.](image)

Figure 6.10: Solid patch vs Sparse. (a) Power reflection coefficient. (b) Antenna pattern.

![Figure 6.10: Solid patch vs Sparse.](image)

As seen in Fig. 6.10 it can be seen, the sparse antenna achieves the same broad bandwidth and even better matching as compared to its solid patch counterpart. The gain of the sparse antenna is the same to that of the solid patch antenna.

The sparse T-slot antenna was fabricated with copper tape and connected to a RFID tag chip as shown in Fig. 6.11. Its performance was measured and compared
with the solid patch T-slot (in Chapter 3) on six positions of the sidewall of a tire.

The results are listed in Table. 6.4.

<table>
<thead>
<tr>
<th>Tag Antenna Type</th>
<th>P1</th>
<th>P2</th>
<th>P3</th>
<th>P4</th>
<th>P5</th>
<th>P6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sparse Antenna</td>
<td>-24</td>
<td>-24</td>
<td>-24</td>
<td>-23</td>
<td>-22</td>
<td>-21</td>
</tr>
<tr>
<td>Solid Patch Antenna</td>
<td>-24</td>
<td>-24</td>
<td>-24</td>
<td>-23</td>
<td>-22</td>
<td>-21</td>
</tr>
</tbody>
</table>

Table 6.4: Threshold Power of the Tags Under Test

Table. 6.4 demonstrates that the tag with sparse T-slot antenna achieved the same performance as that with the solid patch antenna.

6.2.2 Embroidered Antenna

Both the sparse and solid patch antennas were fabricated with E-fibers, as shown in Fig. 6.12. The solid patch T-slot antenna was made with densely stitched double layer E-fiber. The sparse T-slot antenna was fabricated with triple thread E-fiber. The fabricated antennas were connected to RFID chips. They were tested on the sidewall of a tire. The experimental results are listed in Table. 6.5.

Table. 6.4 and Table. 6.5 show that tags with textile based antennas achieved the same good performance as their copper tape counterparts. More importantly,
Figure 6.12: Fabricated textile based T-slot antennas, (a) solid patch T-slot fabricated with double layer E-fiber, (b) sparse T-slot fabricated with triple thread E-fiber.

<table>
<thead>
<tr>
<th>Position</th>
<th>P1</th>
<th>P2</th>
<th>P3</th>
<th>P4</th>
<th>P5</th>
<th>P6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Solid T-slot</td>
<td>24 dBm</td>
<td>24 dBm</td>
<td>24 dBm</td>
<td>23 dBm</td>
<td>23 dBm</td>
<td>22 dBm</td>
</tr>
<tr>
<td>Sparse T-slot</td>
<td>24 dBm</td>
<td>25 dBm</td>
<td>25 dBm</td>
<td>23 dBm</td>
<td>22 dBm</td>
<td>22 dBm</td>
</tr>
</tbody>
</table>

Table 6.5: On-Tire Threshold Power Measurement Results

Table 6.5 shows that the tag with the sparse textile-based antenna achieved similar performance as the tag with embroidered antenna made by densely stitched E-fiber. The new antenna achieved significant surface area and material reduction. Thus, it has great potential for advancing fabric antenna technology.

6.3 Summary of Chapter 6

Different textile-based antenna configurations were investigated in this chapter. A broadband tag antenna designed for elastic materials was presented. The antenna delivers a bandwidth of 265 MHz in free space for the purpose of maintaining tuned behavior over a wide range of materials. The antenna was fabricated with a single thread E-fiber which reduced the fabrication cost, and made the tag less prone to fatigue. It was also demonstrated in this chapter that conventional antennas having
planar solid surface can be redesigned and fabricated with multiple threads of E-fiber. This further reduces the cost and enhances the stretchability of the antenna. A sparse T-slot antenna was designed and tested. The experimental results showed that the antenna made with multiple conductive threads achieved the same good performance as its densely stitched E-fiber counterpart. It is believed that this new type of antenna will help the advancement of fabric antenna technology.
Chapter 7: Contributions and Future Work

This work has explored systematically the effects of dielectrics on the RFID tag antenna’s performance. For the first time, the effects of dielectric media on the antenna’s impedance and gain pattern have been quantitatively analyzed. Novel tag antennas have been designed based on the proposed analysis. Experimental results show that they exhibit excellent performance over a wide range of materials. Different methods and materials have been investigated to fabricate flexible and stretchable antennas for challenging environments. A novel conductive material, E-fiber has been used to construct the tag antennas. New textile-based antenna design approach has been proposed. For the first time, a single/multiple thread E-fiber has been used for antenna fabrication. The fabricated antennas achieved exceptional electrical performance while maintaining robust mechanical structure.

This research focuses on designing RFID tag antennas for automotive tires. However, the proposed analysis and the developed technique may benefit a broad range of applications. With the fast development of the Internet of Things, more devices than ever are connected to the Internet. These devices include a large number of smart tags and sensors that track and monitor a plethora of items and environmental conditions. They will be mounted on different environments. This work provides great insights for designing antennas for these devices.
7.1 Looking Ahead

An attractive future implementation of RFID is to integrate RFID tags with sensor technologies. A RFID tag consists of an ASIC (Application-specific integrated circuit) chip which provides certain data processing capability. It is possible to extend this well-established infrastructure for wireless sensing applications. Existing wireless sensor solutions face the challenge of limited battery life. Replacing the batteries significantly increases the cost of labor and maintenance which outweigh the cost of batteries. An RFID enabled sensor node may not require internal power source for communication, and can depend on the received power from the reader to operate the circuitry. Furthermore, the RFID sensors could be implemented on challenging environments where replacing the batteries is inconvenient and even not possible.

There has been strong interest in developing RFID based passive sensing platforms. Existing battery-free RFID integrated sensor techniques could be summarized into two main categories: 1) analog-based approach, 2) digital-based approach. Analog based approach exploits the dependence of the tag antenna on the materials in the vicinity. It is well known that when the tag is placed at proximity of materials having different electrical properties, the realized gain of the tag antenna will change. This eventually affects the threshold power to operate the tag, the read range and the RSSI (received signal strength indicator). By calibrating the signal level of the RSSI, it is possible to realize functions such as temperature monitoring [102], gas detection [103] etc. This is a low cost approach which does not require any additional electronic hardware and could be developed based on the existing infrastructure. However, the major problem is that RSSI depends strongly on the multipath, polarization mismatch between the tag and the reader antennas, and the manufacture standard. Calibration
needs to be repeated whenever the environment changes. Thereby this approach cannot provide reliable data and is mostly used for low resolution sensing applications. Such as sensors carry only 1-bit sensing info (on or off) \cite{104, 105}. The digital approach is to integrate sensors as well as peripheral circuitry on to the RFID tag chip. The analog sensing info is collected by the on-chip sensor then converted by ADC and eventually stored on the EEPROM (Electrically Erasable Programmable Read-Only Memory) of the tag chip. The sensing info is digitally delivered to the reader using the same communication protocol as the tag broadcasts its serial number. This approach provides high resolution and accuracy. There are several examples of such approach \cite{106, 107}. A challenge for this approach is the power consumption. Although RFID technology does not need power for wireless communication, it requires a certain amount of power to operate the on-chip sensor and ADC. The power budget and processing capability of such sensing platform is discussed in \cite{108}. Furthermore, additional power needs to be provided for data logging function.

At the end of this dissertation, we propose an idea that could be explored for the future work. The goal is to develop an architecture for RFID-based integrated sensing, energy harvesting, and wireless communication. This technology could be developed for automotive tires, but it may benefit a broad range of applications. The sensor info provides real time tire condition and ride and handling information. This information could be stored in the RFID chip and accessed via wireless readers. The whole architecture only involves one tag, one reader and one wireless communication protocol. The proposed system architecture is shown in Fig. 7.1.

The proposed architecture is comprised by three parts: 1) Tag Antenna; 2) Tag chip; 3) Sensors. The tag antenna needs to be designed to provide reliable reading
regardless of the tire construction and condition. The antenna also needs to have robust structure to withstand rubber curing, retreading, and vehicle running/stopping. As opposed to conventional RFID chips, the RFID tag chip in the proposed work needs to receive sensor input from multiple sensors. Optionally, it also processes the sensor data and stores it at the on-chip memory bank. The sensors need to be designed to provide reliable sensing of different tire phenomena, they also need to have robust structure to withstand operating conditions. In order to provide data logging function, additional energy harvesting circuitry needs to be designed and integrated with the RFID tag chip. The harvested energy is used to power data processing and storing.

RFID technology is the key enabler of the Internet of Things. Besides providing a more efficient way of inventory tracking, it could transform the way people interact with objects and how data is being collected and exchanged. Data collected by RFID
tags does not need to be processed by the tag chip, but instead, transferred instantly to the cloud. Companies can utilize this data to make sound strategies accordingly, therefore, increase their profits. As an example, the benefits of adopting RFIDs in the tire industry is illustrated in Fig. 7.2.

Figure 7.2: How tire manufacturers benefit from using RFID technologies.

As shown in Fig. 7.2, the RFID tag is embedded in the tire on the assembly line. After embedded with the tag, the tire goes through a portal (which is also on the assembly line). The information of the tire, such as type and manufacture date is written on the tag chip as well as the cloud computing system. In the warehouses or retailers, whenever a tire is shipped or sold, the employer can access the tag using a hand-held reader or portal. The information of sale/ship date is updated on the tag chip as well as the cloud. After purchasing the tires, customers could check
the conditions of their tires using smart phones or smart phone enabled accessories.
The conditions of the tire will be transmitted to the tire manufacturers. Tire manufacturers can send notifications to the users regarding proper maintenance. During the recycle procedure, the information could be recorded in the tire using the same method.
Bibliography


120


