Propagation and Narrow Cylindrical Antennas for Non-Line-of-Sight Links

by

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Thesis Submitted in Partial Fulfillment of the Requirements for the Degree of Doctor of Philosophy in the School of Engineering Science Faculty of Applied Science

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SIMON FRASER UNIVERSITY
Summer 2020

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Abstract

Marconi’s century-old commercialization of wireless has grown to billions of radio links. In cities there can be thousands of cellular base stations, usually mounted on buildings, to link millions of terminals, and there are many WiFi devices in homes and offices. These links are nearly all non-line-of-sight (NLOS), with signal processing at the terminals striving to cope with the degradation from the propagation and the antennas. The signal processing, antenna design, and the propagation, are now separate disciplines as a result of their expansion. The limitations from the propagation channels and the antennas are often blindly accepted by signal processing. Innovations become most likely when there is an in-depth understanding of each discipline, an increasingly difficult prospect. But no matter how powerful or innovative the electronic signal processing, the propagation and antenna performance remain the biting constraint for communications performance. This motivates a hypothesis: improving the understanding of the bottleneck mechanisms - the propagation and antennas - enables innovation for better link performance. The approach is to select topics in propagation and antennas which bottleneck the link performance.

For NLOS, diffraction is the critical mechanism. The thesis therefore opens with a look at diffraction, in the context of two applications: classical around the corner propagation, where simple arrangements of passive dipoles are demonstrated to drastically improve a diffraction-limited link; and through-forest propagation, where a new model, combining diffraction across the tree tops and direct transmission, is demonstrated to fit the full range of short- to long-distances established from recent experiments.

For the antennas, tubular platforms offer challenges which have not been widely addressed, and yet such platforms are ubiquitous in the form of bicycle frames, drone struts, and masts. Designs are investigated where compactness is a critical requirement: externally-mounted, small narrowband antennas for where the curvature of the cylindrical tube is too small for planar groundplane principles to guide the design; and configurations that deploy the tubular structure as a compact coaxial cylindrical waveguide, to feed slot elements in the cylinder. These are demonstrated to be extremely wideband and have low loss at microwave frequencies.
Keywords: diffraction; propagation around corners; propagation through-forest; small cylindrical antenna; coaxial waveguide antenna.
To my parents, Fattaneh and Lotfollah,
and to my sisters, Maryam and Nina,
for their love, endless support and encouragement.
Our deepest fear is not that we are inadequate.
Our deepest fear is that we are powerful beyond measure.

It is our light, not our darkness.

That most frightens us.

Marianne Williamson
Acknowledgements

First and foremost, I would like to express my special gratitude to my supervisor, Dr. Rodney G. Vaughan who offered me an opportunity to fulfill my dream of completing a PhD. Throughout this journey, he has been a constant source of inspiration. This dissertation would not have been possible without his encouragement, unlimited support and exceptional guidance. I also would like to acknowledge the invaluable guidance provided by my supervisory committee members Dr. Bernhard Rabus of Simon Fraser University and Dr. Daniel Janse Van Rensburg from NSI-MI Technologies, at various stages of this research. I am also thankful to internal examiner Dr. Ash Parameswaran and defense chair Dr. Bonnie Gray for their valuable time and comments.

A special thanks is due to Dr. Derek McNamara from University of Ottawa. He is a pioneer in diffraction and antenna theory. It is a great honor to have such a knowledgeable and eminent person as my external examiner.

I wish to thank the faculty and staff of the School of Engineering Science and Simon Fraser University library for their support during my graduate study. Especially, I want to thank Ms. Raj Pabla, Ms. Mary Palisoc, Mr. Alex Girtonea, Mr. Chao Cheng, Ms. Anise Ladha, and Mr. Donald Taylor.

I am so grateful to all my colleagues in Sierra Wireless Communications Lab for their friendship, advice, and support over these years. Christopher G. Hynes is especially singled out for his helpful comments and advice, from both industrial and academic viewpoints. My gratitude also goes to the past and present members: Ying Chen, Abhijit Bhattacharya, Maral Dehghani, Joe Steyn, Maryam Razmhosseini and Rajveer Singh Brar.

I would like to thank the Natural Sciences and Engineering Research Council of Canada (NSERC) and MITACS for the support for various projects which make up in this dissertation.

Finally, My greatest debt of gratitude is to my family. It is impossible to thank my family for all that they have done to support me over the years. Special thanks to my parents, Fattaneh and Lotfollah, for unconditional love and support. I also want to thank my sisters, Maryam and Nina, for always being a true support and for loving me.
Table of Contents

Approval ii
Abstract iii
Dedication v
Quotation vi
Acknowledgements vii
Table of Contents viii
List of Tables xi
List of Figures xii
List of Acronyms xx

1 Introduction 1
  1.1 Background and Motivation ................................. 1
  1.2 Organization of This Dissertation ........................... 3
  1.3 Works Published/Submitted as Part of this Doctorate Research 5
    1.3.1 Statement of Personal Contributions ........................ 8

2 Diffraction: A Critical Propagation Mechanism 9
  2.1 Introduction and Background .................................. 9
  2.2 Contributions and Organization of the Chapter ................ 10
  2.3 Basic Diffraction ............................................. 10
  2.4 Essential UTD Diffraction .................................... 13
    2.4.1 UTD Diffraction by a Wedge ................................. 13
    2.4.2 Modeling the Half-plane Diffraction Using the UTD .......... 16
  2.5 Comparison of Diffraction with Numerical and Physical Experiments 17
    2.5.1 Simulation Modelling and Results ........................... 18
    2.5.2 Physical Experiments ...................................... 20
2.6 Summary and Conclusion ........................................ 22

3 Propagation Around Corners .................................... 23
  3.1 Introduction and Background ................................ 23
  3.1.1 Challenges ............................................... 24
  3.1.2 Review and Contribution ................................. 25
  3.2 Corner Scattering Analysis .................................. 26
  3.3 Description of the Model Used for Demonstration .......... 30
  3.4 Results and Discussions .................................... 31
  3.4.1 Single Dipole Scatterer near the Corner ............... 31
  3.4.2 Linear Array of Dipole Scatterers ....................... 38
  3.4.3 Three-Dipole Scatterer and its Array ................... 38
  3.4.4 Other Considerations ................................... 40
  3.5 Summary and Conclusion .................................... 41

4 Through-Forest Propagation ..................................... 42
  4.1 Introduction and Background ................................ 42
  4.2 Contributions and Organization of the Chapter .......... 44
  4.3 Basic Radiative Energy Transfer Method .................... 44
  4.4 Simple Transmission Line Model from RET Results for Short-Distance Foliage 53
  4.5 Simplifying Long Distance Through-Foliage Propagation Modelling ........ 55
    4.5.1 Multi-Layer Transmission Line ......................... 57
    4.5.2 Multiple Knife-Edge Diffraction ....................... 59
    4.5.3 Results and discussion .................................. 61
    4.5.4 Model Performance ..................................... 66
  4.6 Summary and Conclusion .................................... 67

5 Antennas on Tubular Platforms ................................. 69
  5.1 Introduction and Background ................................ 69
  5.2 Contribution and Organization of the Chapter ............ 70
  5.3 Antennas on an Electrically Narrow Cylindrical Groundplane .... 71
    5.3.1 Monopole on a Small Cylinder ......................... 71
    5.3.2 Conformal Axial-Direction PIFA on a Small Cylinder .... 76
    5.3.3 A New Conformal PIFA for Small Cylindrical Groundplane Mounting 80
  5.4 Bent Configurations of Waveguide Slot Arrays .......... 82
    5.4.1 Short-End Effect Study .................................. 83
    5.4.2 Circular Waveguide Bend Effect ....................... 87
  5.5 Wideband Antenna using Coaxial Waveguide ............... 90
    5.5.1 Basic Coaxial Waveguide Design Principles .......... 91
    5.5.2 Feeding System ......................................... 95
List of Tables

Table 2.1 Dimensions of the measured baffles in NSI chamber. 21
Table 4.1 RET required parameters for "Common Lime" tree at 1.3GHz [1]. 51
Table 4.2 Transmission line parameters for "Common Lime" tree at 1.3GHz. 54
Table 4.3 RMS Error of the propagation model with various empirical parameters. 66
Table 4.4 RMS Error of propagation model for vegetation. 67
Table 5.1 Dimensions of the CPIFA on a small cylinder. 76
Table 5.2 Simulated −10dB impedance bandwidth for a PIFA on different ground-planes. 77
Table 5.3 Dimensions of the WCPIFA on a small cylinder. 80
Table 5.4 One-slotted coaxial waveguide antenna. $kb = 0.75$ at 1.8GHz and coaxial cable length is 0.22λ. 101
Table 5.5 One X-shaped slotted coaxial waveguide antenna. $kb = 0.75$ at 1.8GHz and coaxial cable length is 0.22λ. 102
Table 5.6 Two-slotted coaxial waveguide antenna. $kb = 0.75$ at 1.8GHz and coaxial cable length is 0.22λ. 103
Table 5.7 Six-slotted coaxial waveguide antenna. Cylinder length $L = 2\lambda$, $kb = 0.65$ and coaxial cable length is 0.22λ. 108
# List of Figures

| Figure 2.1 | The GO field and the total field along the observation screen. | 12 |
| Figure 2.2 | The total electric field over a finite absorbing baffle. | 12 |
| Figure 2.3 | The geometry of wedge diffraction (after [2]). | 14 |
| Figure 2.4 | Single reflection of a smooth surface and a direct signal path [2]. | 16 |
| Figure 2.5 | Diffraction over absorbing baffle: the comparison of UTD and simplest diffraction in 2.2 results. | 17 |
| Figure 2.6 | The total electric field over a PEC baffle for horizontal and vertical polarizations. | 17 |
| Figure 2.7 | Semi-infinite baffle approximated by a finite baffle with the height of 15λ for plane wave illumination in CST simulation for (a) horizontal polarization in z-direction (R = −1) and (b) vertical polarization in y-direction (R = 1) (see [3]). | 19 |
| Figure 2.8 | The simulated and calculated total electric field over a perfectly conducting semi-infinite baffle approximated by a finite baffle with the height of 15λ for plane wave illumination at 8λ away from the baffle for (a) horizontal polarization in z-direction (R = −1) and (b) vertical polarization in y-direction (R = 1) (see [3]). | 19 |
| Figure 2.9 | Finite baffle with a height of 8λ for plane wave illumination in CST simulation for (a) horizontal polarization in z-direction (R = −1) and (b) vertical polarization in y-direction (R = 1) (see [3]). | 20 |
| Figure 2.10 | The simulated total electric field over a perfectly conducting finite baffle with the height of 8λ for plane wave illumination at 8λ way from the baffle for horizontal polarization in z-direction (R = −1) and vertical polarization in y-direction (R = 1) (see [3]). | 20 |
| Figure 2.11 | Experimental setup in an NSI planar scanner chamber for (a) the "semi-infinite" baffle and (b) finite baffle illuminated by a horn antenna, with mild clutter. | 21 |
| Figure 2.12 | The measured total electric field over (a) "semi-infinite" and (b) finite baffle for horn illumination with horizontal polarization. | 22 |
| Figure 3.1 | Propagation around a corner of a building with a fixed dipole scatterer. | 27 |
Figure 3.2  Nomenclature for propagation around a corner - plan view. (a) diffraction parameters, (b) the configuration of the scatterers used for analysis, simulation and measurement. ................................. 27

Figure 3.3  Nomenclature for a two-path GO model (after [2]). ................................. 28

Figure 3.4  Side view of a (a) single half-wavelength dipole scatterer, (b) linear array of dipole scatterers, (c) three-dipole scatterer, and (d) linear array of three-dipole scatterers near the corner. The observation point (not seen in the side view), scattering dipole, and the transmitter are aligned (at the same height up the building) in the plane of incidence. 31

Figure 3.5  Theoretical and simulation comparison with and without a half-wavelength short-circuited dipole scatterer \( (h_R = 2\lambda \text{ and } h_s = \lambda) \). The theoretical model (blue curve) is not expected to be accurate for small \( X \). ................................................................. 32

Figure 3.6  Depiction of the electric field intensity in the \( XY \)-plane at \( Z = 0 \). (Plan view for a building corner): (a) No scatterer (diffraction only), (b) a short-circuit termination on a half-wavelength dipole scatterer, and (c) an array of short-circuited dipole scatterers. ................. 33

Figure 3.7  Electric field intensity in the \( YZ \)-plane at \( X = -2.45\lambda \) (in the shadow region) from the tip of the corner in the \( X \)-direction. (Side view of a building corner, vertical polarization): (a) no scatterer (diffraction only), (b) a short-circuit termination on a half-wavelength dipole scatterer, and (c) an array of short-circuited dipole scatterers. ................. 33

Figure 3.8  Smart corner physical measurements (a) without and (b) without the half wavelength short-circuited dipole scatterer (mounted using a foam stand-off), in an anechoic chamber. The surface is flat with the foil ripples exaggerated by the lighting. ................................. 35

Figure 3.9  Measurement and simulation comparison - with and without a half-wavelength short-circuited dipole scatterer \( (h_R = 2\lambda \text{ and } h_s = 0.8\lambda) \). 35

Figure 3.10  Simulation results comparisons of a short-circuit dipole scatterer with various lengths, \( L_s \). \( (h_R = 2\lambda \text{ and } h_s = \lambda) \). ................................. 36

Figure 3.11  Comparison of the normalized simulated electrical field over the corner \( (XY \)-plane at \( h_R = 2\lambda \)), without and with a "vertical" half-wavelength dipole with a spacing of \( h_s = \lambda \) and \( \theta = 45^\circ \). The transmitter is a dipole antenna with horizontal (perpendicular) polarization (surface reflection coefficient is \( R = 1 \)). The simulated result is also compared to the UTD for validating the simulation model. ............. 37
Figure 3.12 The simulated horizontal (perpendicular) electric field in the XY-plane for a corner (a) without and (b) with a "vertical" half-wavelength dipole scatterer. The transmitter is a dipole antenna with horizontal (perpendicular) polarization (corner reflection coefficient is $R = 1$).

Figure 3.13 Simulation results comparisons- with open-circuited and short-circuited dipole scatterer and array of dipole scatterers. ($h_{RX} = 2\lambda$ and $h_s = \lambda$).

Figure 3.14 Depiction of the electric field intensity in the XY-plane at $Z = 0$.

Figure 3.15 Electric field intensity in the YZ-plane at $X = -2.45\lambda$ (in the shadow region) from the tip of the corner in the X-direction. (Side view of a building corner, vertical polarization): (a) short-circuited three-dipole scatterer, (b) an array of short-circuited three-dipole scatterer.

Figure 3.16 Simulation results comparisons- with open-circuited and short-circuited three-dipole scatterer and array of three-dipole scatterer. ($h_{RX} = 2\lambda$ and $h_S = \lambda$).

Figure 4.1 The received power for different path loss components compared to the measurement of a "Common Lime" tree out of leaf at 1.3GHz given in [4], which is the received power due to the vegetation excess loss apart from the free-space path loss, and the best fit RET curve.

Figure 4.2 (a) Definition of specific intensity emitting from $dA$, (b) Scattering from a homogeneous random vegetation medium (After [5,6]).

Figure 4.3 The RET geometry for mm-wave propagation through vegetation (After [4]).

Figure 4.4 Two components of total specific intensity (After [6]).

Figure 4.5 The total normalized received power and the three individual contributions, i.e., $I_{ri}, I_1$ and $I_2$, in RET equation, i.e., (4.3), compared to the measured data of a 'Common Lime' tree out of leaf at 1.3GHz given in [4].

Figure 4.6 Transmission line model for vegetation illuminated by a plane wave.

Figure 4.7 The normalized received power given by RET and Transmission line model for the "Common Lime" tree (a) in leaf and (b) out of leaf at 1.3GHz.

Figure 4.8 Through-forest propagation model. The transmission line, in parallel and uncoupled with the diffraction path.
Figure 4.9  (a) A depiction of trees between a transmitter and receiver antenna. (b) Two-dimensional multi-layer configuration of in-line trees. The constant thickness of the trees and the free space distance between them are donated by \( d \) and \( d_0 \), respectively. The diffraction and vegetation penetration mechanisms are considered as independent, a simplifying limitation, but their models are connected, again for simplicity, by the number of lossy transmission line sections (trees) being the same as the number of diffraction edges (tree tops).

Figure 4.10  Equivalent transmission line circuit for the two-dimensional multi-layered configuration of in-line trees (after [7]).

Figure 4.11  Geometry for multiple knife-edge diffraction (after [8]).

Figure 4.12  Measurement locations conducted in a typical forest terrain in Denmark [9]. The transmitter position is marked as Tx at the left top in the figure.

Figure 4.13  (a) \( N \) absorbing knife-edge diffraction from the crown of the trees with the same height and spacing, \( r \), (b) two-dimensional \( M \)-layer configuration of in-line trees with \( d \) thickness, \( \varepsilon_r \) dielectric constant and \( r \) spacing.

Figure 4.14  Comparison between the measured path gain in [9] and the propagation model for \( r = 1.5 \text{m} \) (\( N = 1720 \)), \( W_2 = 1 - W_1 = -70 \text{dB} \): (a) \( \varepsilon'' = 0.006 \) and (b) \( \varepsilon'' = 0.01 \). The distance scale is logarithmic. The free space path gain (FSPG), \( W_1P_{D_N} \), and \( W_2P_{MLT_{trans}} \) are also plotted to show the dominant parameters.

Figure 4.15  Comparison between the measured path gain in [9] and the propagation model for \( \varepsilon'' = 0.008 \), \( W_2 = 1 - W_1 = -70 \text{dB} \): (a) \( r = 1 \text{m} \) (\( N = 2580 \)), (b) \( r = 1.5 \text{m} \) (\( N = 1720 \)), (c) \( r = 2 \text{m} \) (\( N = 1290 \)) and (d) \( r = 3 \text{m} \) (\( N = 860 \)). Note that for different \( r \), \( N \) is different, and the distance scale is logarithmic. The free space path gain (FSPG), \( W_1P_{D_N} \), and \( W_2P_{MLT_{trans}} \) are also plotted to show the dominant parameters.

Figure 4.16  Comparison between the measured path gain in [9] and the propagation model for \( \varepsilon'' = 0.008 \), \( r = 1.5 \text{m} \) (\( N = 1720 \)): (a) \( W_2 = 1 - W_1 = -60 \text{dB} \), (b) \( W_2 = 1 - W_1 = -30 \text{dB} \), (c) \( W_2 = 1 - W_1 = -1 \text{dB} \), (d) \( W_2 = 1 - W_1 = -\infty \text{dB} \). The distance scale is logarithmic. The free space path gain (FSPG), \( W_1P_{D_N} \), and \( W_2P_{MLT_{trans}} \) are also plotted to show the dominant parameters.

Figure 5.1  A quarter-wavelength monopole on a small, finite-length cylinder.
Figure 5.2  Calculated $-10\text{dB}$ numerical and theoretical bandwidths for a small monopole with the diameter of $0.1\lambda$ for different cylinder lengths, $L$. 73

Figure 5.3  Measurement setup of a monopole antenna prototype (a) with VNA and (b) in Satimo chamber. 74

Figure 5.4  Simulated and measured $S_{11}$ of a $\lambda/4$ monopole on a small, finite-length cylinder. The cylinder diameter and length is $0.1\lambda$ and $0.26\lambda$. 74

Figure 5.5  Comparison of normalized simulated and measured radiation patterns of the monopole at (a) $\phi = 0^\circ$, (b) $\phi = 90^\circ$ and (c) $\theta = 90^\circ$ cuts; Solid line: measured $\theta$ pol., dash-dot line: measured $\phi$ pol., dashed line: simulated $\theta$ pol. and dotted line: simulated $\phi$ pol. (cannot be seen in this scale). 75

Figure 5.6  (a) Sloping and (b) bending monopoles on a small, finite-length cylinder. (c) Simulated $-10\text{dB}$ impedance bandwidth for the sloping and bending monopoles. 75

Figure 5.7  A CPIFA on a cylinder: (a) CPIFA structure, (b) end view along the cylinder, and (c) 3D view. 77

Figure 5.8  Measurement setup for a CPIFA prototype (a) with VNA and (b) in Satimo chamber. 78

Figure 5.9  Simulated and measured $S_{11}$ of a CPIFA on a small cylinder. The dimensions are given in Table 5.1. 78

Figure 5.10  Comparison of normalized simulated and measured radiation patterns of the CPIFA at (a) $\phi = 0^\circ$, (b) $\phi = 90^\circ$ and (c) $\theta = 90^\circ$ cuts; Solid line: measured $\theta$ pol., dash-dot line: measured $\phi$ pol., dashed line: simulated $\theta$ pol. and dotted line: simulated $\phi$ pol. 79

Figure 5.11  The effect of the (a) cylinder length $L$ and (b) cylinder diameter $D$ on the $-10\text{dB}$ impedance bandwidth. The dimension of the CPIFA is given in Table 5.1. 80

Figure 5.12  WCPIFA on a small cylinder: (a) 3D view, (b) end view along the cylinder. 80

Figure 5.13  Measurement setup for a WCPIFA prototype (a) with VNA and (b) in Satimo chamber. 81

Figure 5.14  Simulated and measured $S_{11}$ of a WCPIFA on a small cylinder. $D = 2b$ is the cylinder diameter. The dimensions are given in Table 5.3. 81
Figure 5.15 Comparison of normalized simulated and measured radiation patterns of the WCPIFA at (a) $\phi = 0^\circ$, (b) $\phi = 90^\circ$ and (c) $\theta = 90^\circ$ cuts; Solid line: measured $\theta$ pol., dash-dot line: measured $\phi$ pol., dashed line: simulated $\theta$ pol. and dotted line: simulated $\phi$ pol. 82

Figure 5.16 (a) One slot on top of the waveguide. (b) Simulated S-parameters in terms of the spacing between the slot element and the short-end termination. 84

Figure 5.17 (a) One slot on top (solid line) and one slot on bottom (dashed line) of the waveguide. (b) Simulated S-parameters in terms of the spacing between the top slot element and the short-end termination. 84

Figure 5.18 (a) Five slots on top of the waveguide. (b) Five slots on top (solid line) / bottom (dashed line) of the waveguide. 85

Figure 5.19 Simulated S-parameters of (a) five slot waveguide array on top of the waveguide, and (b) five slot waveguide array on top/bottom of the waveguide. 85

Figure 5.20 $-6$ dB impedance bandwidth comparisons. 86

Figure 5.21 $-6$ dB impedance bandwidth calculated from Eq.5.10 (solid line), $-6$ dB impedance bandwidth calculated from simulation (markers). 87

Figure 5.22 Left: cross section view, Middle: side view, Right: equivalent circuit, (After [10]). 87

Figure 5.23 Circular bend guided-wavelength in terms of mean radius ($R$). 88

Figure 5.24 Numerical and theoretical reflection coefficient at reference plane, $T$. 89

Figure 5.25 Different configurations of circular bend (with rectangular cross section) and rectangular waveguide. One slot on top (solid line) and one slot on bottom (dashed line) with $0.25\lambda_g$ lengthwise spacing. 89

Figure 5.26 Simulated S-parameters of different configurations of circular bend (with rectangular cross section) and rectangular waveguide. 90

Figure 5.27 (a) Coaxial line geometry: $\epsilon$ is the permittivity of the dielectric, $a$ and $b$ are the inner and outer conductors radius, (b) electric/magnetic field lines and currents for the TEM mode of coaxial line (after [11]). 91

Figure 5.28 Cut-off frequency of the $TE_{11}$ mode of a $50\Omega$ coaxial line with a (lossless) air dielectric insert, $\epsilon_r = 1$. (Here, $k_c = 2/(a + b)$, and $a$ is calculated for $b$ ranging from 1mm to 20 mm). 92
Figure 5.29  (a) Eccentric line (After [12]). (b) Line impedance in terms of eccentricity for dielectric insert air ($\varepsilon_r = 1$).

Figure 5.30  (a) Conductor loss and (b) dielectric loss per meter for a 50Ω coaxial line (TEM mode) in comparison with rectangular waveguide (TE10 mode) for copper ($\sigma = 5.96 \times 10^7$), gold ($\sigma = 4.17 \times 10^7$), stainless steel ($\sigma = 1.45 \times 10^6$) and aluminium ($\sigma = 3.5 \times 10^7$) filled with teflon medium modelled with $\varepsilon_r = 2.1$ and $\tan \delta = 0.0004$.

Figure 5.31  Geometry of the feeding: a 50Ω panel mount N-type connector connected to the coaxial waveguide.

Figure 5.32  Numerical -6dB impedance and −10dB impedance bandwidth of a one-slotted coaxial waveguide with $\alpha = 0^\circ$ located at 0.5λ from the short-end termination in terms of (a) extended coax inner length and (b) eccentricity, see Fig. 5.29b. The length is of about 0.5λ, and the cylinder length is λ.

Figure 5.33  One-slotted coaxial waveguide antenna with cylinder length ($L_{cyl}$) of λ, slot length ($S_l$) of 0.5λ and slot width ($S_w$) of 0.01λ, (b) −6dB and −10dB impedance bandwidth in terms of the angle $\alpha$ with the fixed distance of 0.5λ from the short-end position (top x-axis), −6dB and −10dB impedance bandwidth in terms of the position of the slot from the short-end position with the $\alpha$ angle of 0° (bottom x-axis).

Figure 5.34  (a) Simulated −3dB, −6dB and −10dB impedance bandwidth in terms of the slot position form the waveguide port of one-slotted coaxial waveguide antenna (see Fig. 5.33a), the slot length is fixed at 0.5λ, (b) simulated normalized radiation pattern for ($\phi = 90^\circ$ cut) for different coaxial waveguide lengths of one slotted coaxial waveguide. The distance between the feed and the slot is varied from 0.125λ to 0.5λ. The slot position is fixed at 0.5λ from the short-end with the angle of $\alpha = 0^\circ$.

Figure 5.35  The normalized radiation pattern for a one slotted coaxial waveguide with $\alpha$ angle of 0° located at 0.5λ from the short-end termination for the design frequency ($F_d$), center frequency ($F_0$), lower ($F_L$) and upper ($F_U$) operating frequency. The cylinder length is λ.

Figure 5.36  Effect of varying the number of slots with $\alpha = 0^\circ$ and length of 0.5λ on −6dB and −10dB impedance bandwidth.
Figure 5.37 One-slotted coaxial waveguide antenna with the slot length of $\sim 0.5\lambda$ and $\alpha = 0^\circ$: (a) $d_{\text{short}} = 0.5\lambda$ and $d_{\text{feed}} = 0.125\lambda$, (b) $d_{\text{short}} = 0.125\lambda$ and $d_{\text{feed}} = 0.125\lambda$. (c) Simulated (solid line) and measured (dashed line) S-parameter results.

Figure 5.38 One X-shaped slotted coaxial waveguide antenna with the slot length of $\sim 0.5\lambda$: (a) $U = 10^\circ$, $B_{\text{hh}} > AC = 0.5\lambda$ and $\theta = 0.125\lambda$, (b) $U = 15^\circ$, $B_{\text{hh}} > AC = 0.125\lambda$ and $\theta = 0.125\lambda$. (c) Simulated (solid line) and measured (dashed line) S-parameter results.

Figure 5.39 Two-slotted coaxial waveguide antenna with the slot length of $\sim 0.5\lambda$ and $U = 0^\circ$: (a) $B_{\text{hh}} > AC = 0.5\lambda$ and $\theta = 0.125\lambda$, (b) $B_{\text{hh}} > AC = 0.125\lambda$ and $\theta = 0.125\lambda$, and $B_{\text{hh}} > AC = 0.006\lambda$. (c) Simulated (solid line) and measured (dashed line) S-parameter results.

Figure 5.40 The angle $\theta$ between the beginning and the end of the slot seen from along the cylinder.

Figure 5.41 Slotted coaxial waveguide with (a) two slots (1 slot on top & 1 slot on bottom), (b) four slots (2 slots on top & 2 slots on bottom) and (c) six slots (3 slots on top & 3 slots on bottom). $S_t = 0.44\lambda$, $d_{\text{short}} = 0.5\lambda$, $d_{\text{feed}} = 0.5\lambda$ and $d_{\text{slot}} = 0.5\lambda$ at $f_0 = 1.575\text{GHz}$. (d) Simulated s-parameters comparison. Bottom slots are shown as dotted lines.

Figure 5.42 Effect of varying the number of slots on $-6\text{dB}$ and $-10\text{dB}$ impedance bandwidth (left axis) and directivity (right axis).

Figure 5.43 6-slotted coaxial waveguide (3 slots on top & 3 slots on bottom) cannot be seen in the prototype, but shown as dotted lines in the simulated model) (a) prototype, (b) simulated (solid line) and measured (dashed line) S-parameter results. $d_{\text{short}} = d_{\text{slot}} = d_{\text{feed}} = 0.5\lambda$, $\alpha = 70^\circ$ and the slot length is $0.44\lambda$.

Figure 5.44 Normalized radiation pattern of the prototype with 6 slots (Fig. 5.41c) at the frequency of $1.575\text{GHz}$, (a) $\theta = 90^\circ$ cut, (b) $\phi = 90^\circ$ cut; solid line: simulated $\theta$ pol., dash-dot line: simulated $\phi$ pol., dashed line: measured $\theta$ pol. and dotted line: measured $\phi$ pol.

Figure 5.45 (a) 6-slotted coaxial waveguide antenna (3 slots on top & 3 slots on bottom) with $\alpha = 60^\circ$, $d_{\text{short}} = d_{\text{slot}} = d_{\text{feed}} = 0.5\lambda$ and the slot length is $0.44\lambda$, (b) simulated S-parameter results. The design frequency is $5\text{GHz}$.

Figure 5.46 Normalized radiation pattern of the prototype with 6 slots (Fig. 5.41c) at the frequency of $5\text{GHz}$, (a) $\theta = 90^\circ$ cut, (b) $\phi = 90^\circ$ cut; solid line: simulated $\theta$ pol., dash-dot line: simulated $\phi$ pol.
## List of Acronyms

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>2D</td>
<td>Two-Dimensional</td>
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<tr>
<td>3D</td>
<td>Three-Dimensional</td>
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<tr>
<td>5G</td>
<td>Fifth Generation Mobile Networks</td>
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<tr>
<td>BS</td>
<td>Base Station</td>
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<td>BW</td>
<td>Bandwidth</td>
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<td>CMC</td>
<td>Coupled Mode Cable</td>
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<tr>
<td>CPIFA</td>
<td>Conformal Planar Inverted-F Antenna</td>
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<tr>
<td>FDTD</td>
<td>Finite-Difference-Time-Domain</td>
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<tr>
<td>FEM</td>
<td>Finite Element Method</td>
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<tr>
<td>FIT</td>
<td>Finite Integration Technique</td>
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<td>FWHM</td>
<td>Full Width at Half Maximum</td>
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<td>GO</td>
<td>Geometric Optics</td>
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<tr>
<td>IRS</td>
<td>Intelligent Reflecting Surfaces</td>
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<tr>
<td>ITU-R</td>
<td>International Telecommunication Union Recommendation</td>
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<tr>
<td>LIS</td>
<td>Large Intelligent Surface</td>
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<tr>
<td>MIMO</td>
<td>Multiple-Input and Multiple-Output</td>
</tr>
<tr>
<td>mm-wave</td>
<td>Millimeter-Wave</td>
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<tr>
<td>MOM</td>
<td>Method of Moments</td>
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<tr>
<td>NLOS</td>
<td>Non-Line-of-Sight</td>
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<tr>
<td>OFDM</td>
<td>Orthogonal Frequency-Division Multiplexing</td>
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<tr>
<td>Abbreviation</td>
<td>Description</td>
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<tr>
<td>PEC</td>
<td>Perfectly Matched Layer</td>
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<tr>
<td>PIFA</td>
<td>Planar Inverted-F Antenna</td>
</tr>
<tr>
<td>PIS</td>
<td>Passive Intelligent Surface</td>
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<tr>
<td>PML</td>
<td>Geometrical Theory of Diffraction</td>
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<td>PO</td>
<td>Physical Optics</td>
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<td>RET</td>
<td>Radiative Energy Theory</td>
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<tr>
<td>RIS</td>
<td>Reconfigurable Intelligent Surface</td>
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<tr>
<td>RMC</td>
<td>Radiating Mode Cable</td>
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<tr>
<td>RMSE</td>
<td>Root Mean Square Error</td>
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<tr>
<td>TE</td>
<td>Transverse Electric</td>
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<td>TEM</td>
<td>Transverse Electro-Magnetic</td>
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<td>TM</td>
<td>Transverse Magnetic</td>
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<tr>
<td>UAV</td>
<td>Unmanned Aerial Vehicle</td>
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<tr>
<td>UTD</td>
<td>Uniform Geometrical Theory of Diffraction</td>
</tr>
<tr>
<td>VSWR</td>
<td>Voltage-Standing-Wave-Ratio</td>
</tr>
<tr>
<td>WCPIFA</td>
<td>Wrapped Conformal Planar Inverted-F Antenna</td>
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<tr>
<td>WG</td>
<td>Waveguide</td>
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Chapter 1

Introduction

1.1 Background and Motivation

This section reviews the collection of work undertaken for the doctorate training, and also covers the motivation for the thesis. A hypothesis is laid out in the Abstract: improving our understanding of the bottleneck mechanisms - the propagation and the antennas - enables innovation for better link performance. While there is a certain obviousness to this from a scientific viewpoint, it also lays down a challenge to 'delve and deliver' on new engineering ideas. It stems from a look at the Friis transmission equation, which crops up in the thesis multiple times, along with some observations listed below. The hypothesis is tested by example, and specific topics in propagation and antennas have been selected that are fundamental to communications and sensing processes, and yet still have room for new research. The selection also stemmed from some general observations, as follows.

(i) Wireless communication has been a spectacular and crowning success of electrical engineering, and in fact electrical engineering gained status as its own subject - breaking away from physics - over a hundred years ago, because of the multidisciplinary advances required in wireless communications technology. This is a great area to be involved with!

(ii) Over the last couple of decades, there has been an acceleration in the rate of new techniques being developed and transferred from research to global practice of telecommunications. It is hard for an individual engineer to keep up with the multiple technologies, from micro- and nano-antenna fabrication to deep learning receivers that come together under modern wireless communications.

(iii) This acceleration is expected to continue, and significant new industrial product Standards are now appearing every year to try to keep up with the technology, and maintain international compatibility. It is hard to keep up with just the new Standards.

(iv) Much of the literature about the communications signal processing techniques, although highly cited (perhaps because of the large community working on them), is relatively short-lived because the techniques develop and significantly change their nature before becoming a globally accepted Standard.
The techniques seem to blindly accept the propagation and antenna limitations, and strive to repair the degradations rather than address the cause of the degradations directly. With communications signal processing techniques highly topical and dominant in cited research in communications technology, this is a hard area in which to make an everlasting splash. New communications techniques have always been for countering the channel limitations imposed by physical factors - the communications media in particular, i.e., the propagation, and the antenna systems which interface to it. With adaptive antennas overlapping with the signal processing, the grey area of what is a propagation problem and what is an antenna shortcoming, comes to the fore.

Some clarity is found in the Friis equation, where its extraordinary simplicity highlights the physical factors - the propagation in the form of the path gain term, and the antennas in the form of the transmit and receive gain terms. This small cascade of power gain terms, here with self-evident notation including the collection of efficiencies, \( \eta \),

\[
\frac{P_{Received}}{P_{Transmit}} = G_{Transmit} \cdot G_{Receive} \cdot G_{Path} \cdot \eta, \tag{1.1}
\]

can be viewed as setting limits on the received signal level. Electronic receiver technologies set the limits on the noise power level, and adaptive antennas can have a dominant role in the suppression of interference. The result is limits for the signal-to-noise-plus interference ratio, the primary parameter for digital or analogue communications. It follows that the propagation and antennas are the biting constraint in the most-used communications metrics for a system, the total throughput capacity (correctly detected bits per sec per Hz) or an error rate associated with a given capacity in bits per sec per Hz of uncorrected bits.

The physical factors have been relatively constant in a century of wildly changing and highly impactful technology. Instead of tackling the communications techniques, this thesis selects topics from the underlying transmission phenomena that are the raison d’être for the techniques, viz., the propagation and antennas. In short, new technologies are rampant and will continue to come and go, but the underlying physics of the link will stay the same. This is the motivation for the selection of topics here, with a focus on the propagation and antennas rather than the signal processing.

For propagation, diffraction has been singled out because it does not seem to be as widely understood as its major role in transmission deserves. Propagation around corners, and less obviously, through-forest propagation, both use diffraction. For the antennas, the role of electrically compact antennas on electrically narrow cylindrical mounting platforms and coaxial waveguide antennas are selected as a research and learning vehicle, for its novelty and topicality.

Throughout this thesis, reference is made to results that are from physical measurements, simulations, and analytical (i.e., theoretical) formulations. The physical measurements are sometimes referred to as just "measurements" or "experiments", and the simulations are
sometimes referred to as "numerical experiments", reflecting the use of a commercial simulator tool (mostly CST) for experimenting with designs and parametric studies.

1.2 Organization of This Dissertation

The remainder of this thesis is organized as follows.

**Part I Aspects of Propagation -Diffraction**

Chapter 2 is an introduction to basic diffraction used in this research. In particular, the uniform geometrical theory of diffraction (UTD) is covered because it represents the state of the art for describing diffraction, a dominant mechanism for short-haul propagation where a single building or a tree blocks a propagation path. A simplified expression is found for a rectangular absorbing finite baffle. Numerical and physical experiments are compared with the theoretical expression because this has not been previously investigated. The idea is that it will lead to a wider understanding which will in turn lead to simpler and more accurate models for wireless system designers.

Chapter 3 addresses configurations for improving the diffractive link gain for systems that require non-line-of-sight (NLOS) propagation. Coverage in mobile communications systems requires an acceptable link gain over a maximum service area, with areas of low link gain being problematic. To maximize link gain, multiple-input, multiple-output (MIMO) antennas adapt to the multipath, and there is recent interest in using active surfaces for adaptive reflections within the multipath. This calls for new ideas at the system and component levels. A critical mechanism in urban propagation is corner diffraction which formulates how shadow areas are illuminated, but with low link gain. A simple add-on system for corners, comprising a simple dipole or array, which can significantly improve link gain in the shadow region is described. The concept is demonstrated through analysis, simulation, and physical measurement. Within the context of adaptive reflecting surfaces being considered for future systems, the concept can also be applied as a retrofit for current systems. Because the action is to boost only shadowed areas, the predominant coverage areas are not significantly affected. The demonstration is for a passive, fixed system, but it is also possible to adapt the nature of the coverage in the shadow region with an active system of dipole scatterers with varying terminations.

Chapter 4 of the thesis reviews propagation through vegetation based on radiative energy theory (RET). This is an extremely complicated model and can be very time consuming to calculate, but remains an industry standard model for terrestrial communications in the absence of simpler approaches. It is also used for backscattering in remote sensing. Few wireless practitioners fully understand the RET method. A new approximating model for
forward transmission, based on transmission line techniques, is presented here, providing much simpler expressions for the loss through short distance vegetation. The model is computationally very fast and requires fewer parameters than the RET. By demonstrating that the approach works just as well as the RET model (both are empirical models), it is possible that this much simpler approach will become accepted by practicing wireless system designers for short distance propagation through forest (< 100 m). Transmission through such inhomogeneous mixed media, especially for long distance propagation (> 2500 m), is also complicated by the many different propagation mechanisms and the complexity of the randomness. This means that accurate, purely physics-based analysis is unlikely to be practical (conveniently computed), and similarly, that practical, purely random modeling is unlikely to be accurate. Through-vegetation propagation models, including the standard RET, are not very accurate for long distance propagation in the sense that the uncertainty can be tens of dB, and this seems to be an accepted limitation for vegetation. A simpler propagation model, which maintains or improves accuracy, but keeps a reasonable association with the physics, would be insightful. This chapter includes such a model. It comprises two parallel transmission mechanisms: direct transmission through a succession of trees, which is modeled by a simple linear transmission line; and transmission across the forest top, which is modeled by simplified multiple-edge diffraction. The model is examined using recently-published experiments over a long path-length. It is demonstrated that this two-mechanism, empirical model can provide an accurate fit to the dual-slope profile of through-forest propagation over a long distance which is not possible with the RET model.

**Part II Aspects of Antennas - Antennas on Narrow Cylinders**

The range of different types of antennas is remarkable: from electrically very large reflector antennas that are derived from geometric optics to electrically small elements that stem from an understanding of basic fields behaviour in and near metal and dielectric. When there is a Line-of-Sight (in the Fresnel clearance sense), the directivity, and thereby the aperture size, become critical. Patterns are sometimes referred to as "sophisticated", following "sophisticated signals" in signal theory, when they have high directivity. But for most links, which are mobile or nomadic, there is a reliance on NLOS propagation, and the pattern requirements seldom include high directionality, and so they can be deployed, fortuitously, on an electrically small aperture. The sophistication in the patterns in the NLOS case lies in their relationship to each other, viz., their correlation matrix, in a multiport system. The growing interest in mobile platforms such as bicycles and drones, suggests that antennas on cylindrical structures may have increasing importance. Therefore, this topic has been the vehicle for new antenna ideas and designs.

Chapter 5 presents antennas on cylindrical, or tubular, platforms. First, external antenna designs for mounting on a narrow (electrically small cross-section) conducting cylinder are
addressed. This platform is difficult in the sense that its curvature is too great for using
the usual planar ground plane techniques for design, and the bandwidth performance is
not always intuitive. Parametric studies of new designs as well as conventional ones, and
their physical measurement results are provided. The work is relevant to mounting compact
antennas on structures such as masts, drones, and bicycles. The chapter then addresses
applications that have cylindrical tubular structures where externally mounted antennas
are not suitable. An obvious solution is cutting slots on the cylindrical groundplane to
make it radiate such as waveguide-excited slot antennas. For a bicycle, the idea is to have a
radio mounted internally and radiating from of multiple slots distributed along the frame.
The main drawbacks of waveguide slot arrays is their bandwidth sensitivity to dimensional
variations, and in the case of a bicycle frame, there is variability in the waveguide cross
section, length, and in bends along the waveguide. An investigation is undertaken into
using a bicycle frame as a bent waveguide slot array. The use of slots on the different facets
of a rectangular waveguide are shown to improve the bandwidth and help compensate for
the frame length variability. However, waveguide arrays bandwidth and compactness are
both limited by waveguide cutoff. By using the mode of a coaxial waveguide (TEM mode),
a much more compact structure is possible, but which can also support a larger bandwidth,
at the expense of an increased ohmic loss in the waveguide. Using this basis, a new type of
slotted coaxial waveguide antenna is presented which has unprecedented compactness (cross
section) and bandwidth, and yet remains simple to manufacture. The design of such a
compact array means that the element and array aspects are not independent, and the
design is based on numerical and physical experiments. The bandwidth for different design
parameters is included. Two basic design approaches are treated. The first design has no
constraint on the azimuthal extent of the slots. The second design strives to maintain the
mechanical strength of the waveguide by constraining the azimuthal support of the slot to be
90° or less. With correctly designed slots, the simulated and measured results demonstrate
and verify a wideband design.

Chapter 6 concludes the thesis with a summary of the new contributions and suggestions
for future work directions.

1.3 Works Published/Submitted as Part of this Doctorate

Research

(The thesis author’s name is in boldface. * indicates equal contributions with co-principal
author M. Razmhosseini, and † indicates the presenter of a conference paper.)
Journal:


Conference:


**Professional industry report:**


The following chapters draw heavily on the above papers. Chapter 2 uses conference papers 1, 2, and 11; chapter 3 heavily draws upon journal paper 2 and the conference papers 12 and 13; chapter 4 is from journal paper 1 and the conference papers 7 and 14; and chapter 5 heavily draws upon journal paper 3 and the conference papers 4, 5, 8 and 9.
These publications are not otherwise self-cited in the body of the thesis. The publishers of these papers hold the copyright of the published material, but by their agreement with the authors, the material can be used for a thesis.

1.3.1 Statement of Personal Contributions

It is noted above that the thesis chapters draw heavily on published or submitted papers. These papers have multiple authorship. All papers are co-authored by the doctoral supervisor, Dr. Rodney G. Vaughan. This section clarifies my role in producing these papers relative to other graduate student co-authors.

Journal paper 2, and Conference papers 12 and 13 are co-authored with PhD candidate Mr. Christopher G. Hynes. I am the main contributor of these publications. Mr. Hynes’ role was advisory, including contributing his extensive experience and knowledge to the physical measurements.

Journal paper 3, and Conference papers 9 and 10 are co-first authored with PhD candidate Ms. Maryam Razmhosseini, and we have equal contributions as joint principal authors, including all text and figures. I presented this pair of conference papers. For each of these joint papers (J3, C9, and C10), alongside Ms. Razmhosseini, I contributed to all phases of developing the paper - from planning the content, to the design, simulation, construction and measurement of the antennas, and drafting the paper. Section 5.4 draws on conference paper 9, Section 5.5 draws on Journal paper 3. (The material of Conference paper 10 is not included in this thesis.) Some of the material presented in Section 5.5 is not part of Journal paper 3 (including the structures of Figures 37b, 38b, 39b, and their associated performance plots, and Figures 45 and 46), and these parts, including simulation and measurement results, are solely my own contributions.
Chapter 2

Diffraction: A Critical Propagation Mechanism

2.1 Introduction and Background

Multipath propagation mechanisms and their accurate modeling for NLOS wireless links are always under new scrutiny. This is because better accuracy is always needed for predicting the levels of wanted signals and interference, as smaller cells are deployed for better system capacity. These levels refer to the local means - predicting the detail of the short-term, Rayleigh-like fading in general real-world situations, is still regarded as infeasible due to the vast scale (electrical size) and ever-changing detail (moving scatterers and terminals), of almost every propagation scenario. But with canonical situations such as an ideal building corner without other scattering, it should be possible to formulate the short-term fading accurately. The many complicating theoretical and practical factors (mathematically complex formulations, and estimation inaccuracy from the measurement of non-stationary signals, respectively) even the local mean levels of multipath propagation estimations are fraught with difficulty [13]. Nevertheless, the long-standing need for more accurate and fast propagation prediction for real-world problems [14] remains a powerful driver of research. This need is moving with the times as interference increases and spectral re-use techniques become increasingly sophisticated. Current visions for future systems, such as 5G (so-called fifth generation), are calling for extreme increases in capacity and capacity efficiency over current systems. While much effort is going into intensifying the current techniques – smaller cells, larger scale MIMO, high carrier frequencies, interference suppression, full duplex, and so on, there seems to be a disproportionally small effort going into the basic modelling of the propagation behaviour. Nearly all of the communications signal processing is exactly for countering the problems of the propagation channel (and some is for multiple-user access). An improvement in the accuracy of the propagation modelling, or attempting to control some aspect of the propagation channel, could well better guide the deployment of new communications techniques.
Diffraction is the single-most critical propagation mechanism in most urban wireless links. Much progress has been made over the last 60 years on the physics and its modelling, from the landmark paper of Keller in 1962 [15] to [16,17], and more recent transition zone treatments, e.g., [2,18,19]. Diffraction equations are now used widely for propagation analysis, and they offer a rigorous approach for the idealized free space situation. So the theory of diffraction is becoming more widely understood in scientific circles, but at the same time, antenna and link designers increasingly use numerical solutions for their design craft, with decreasing reference to the underlying field and wave theory. Diffraction theory and its assumptions often get lost in the cook-book formulations, and it is not widely understood by wireless link designers. Very few physical experiments have been reported on diffraction, e.g., [20–22]. This seems to be a major disconnect with engineering reality, given that diffraction is the most prominent propagation mechanism in the great majority of the world’s wireless links. Accurate propagation prediction, even for short distances (such as passing a single building) becomes critical in order to make accurate signal calculations for multi-user MIMO deployment. The basic physics is in place, but there is still a need for engineering models derived from the physics. With MIMO antenna design, a statistical spatial model for the propagation is normally used. The wave physics is well-established to calculate such distributions in idealized situations, but this has not been widely undertaken yet. As cell sizes continue to shrink to serving just a handful of mobile terminals, it seems likely that there will be a place for analytical models rather than statistical models for the small-cell propagation.

2.2 Contributions and Organization of the Chapter

This chapter includes a review of basic single diffraction and its mechanism in short-haul wireless propagation, in section 2.3. It explains how diffraction is used to model the transmission around corners even when there are no physical surface currents on the corner structure. A simple expression is developed for an absorbing finite baffle. Section 2.4 studies the UTD by a wedge and model the half-plane diffraction using UTD. In section 2.5 the analytical results are compared to numerical (simulation) and physical experimental results from laboratory-based measurements. The simulations are from a finite element method (FEM), which can be expected to work reasonably well, whereas a method of moments (MOM) using only metal-born electric currents cannot be expected to fully demonstrate diffraction. Section 2.6 summarizes this chapter.

2.3 Basic Diffraction

Diffraction is the continuity of radiating waves from the illuminated region to the shadowed regions of a physical edge. The effects described by diffraction are not predicted by the purely "black and white" approach from the infinitesimally thin rays of geometric optics
(GO). An instructive view is to consider diffraction to be the difference between GO rays and reality, so in this sense, the diffraction terms perfectly complete the optical terms. Huygens’ principle, an extraordinary breakthrough in the history of understanding propagation, explains how waves can get around obstacles by considering their wavefronts as a continuum of isotropic radiators. Diffraction calculations follow this principle by summing the incident waves’ contributions to calculate the subsequent propagation.

The basic problem of diffraction is solved by considering a thin semi-infinite barrier, illuminated by a plane wave [2]. The situation simplifies nicely by considering the diffraction barrier as perfectly absorbing, i.e., a baffle. As shown in (2.1) the total electric field, \( E_{\text{total}} \), is then the sum of the incident electric fields treated as a GO contribution, \( E_{\text{GO}} \), plus the diffraction contribution, \( E_d \), viz.,

\[
E_{\text{total}}(x, y, z) = E_{\text{GO}}(x, y, z) + E_d(x, y, z)
\]  

(2.1)

Three-dimensional (3D) diffraction is still challenging, so here, two-dimensional scenarios approximate the 3D situation. The notation convention is that the \( z \)-dependence drops out. The field over the baffle is given as an integral over the baffle by using the Fresnel-Kirchhoff integral. The physical optics (PO) approximation over the absorbing baffle, and an incident plane wave, \( E_{\text{inc}} = e^{-jkx} \), where \( k = 2\pi/\lambda \), reduces the field expression as follows

\[
E_{\text{total}}(x, y) = e^{-jkx} \left( 1 - \frac{e^{j\pi/4}}{\sqrt{2}} \text{Fres} \left( \sqrt{\alpha^2 y} \right) \right), x > 0,
\]  

(2.2)

where \( \text{Fres}(x) \) is the Fresnel integral given by

\[
\text{Fres}(x) = 2j\sqrt{\pi}e^{ix} \int_{\sqrt{\pi}}^\infty e^{-\frac{j\pi}{2}u^2} du.
\]  

(2.3)

The \( E_{\text{GO}} \) and \( E_{\text{total}} \), calculated in Matlab, are plotted in Fig.2.1. The \( E_{\text{GO}} \) is discontinuous in shadow boundary, being totally dark in shadow region. \( E_{\text{total}} \) is continuous in shadow boundary \((y = 0)\) indicating that the diffracted field is also discontinuous in the transition region. In the lit region \((y > 0)\) there is ripple in the total field that is caused by the interference from the aperture field contributions, viz., the Huygens sources on the aperture. When \( y \) increases, the diffracted field decreases and a local free space situation results, meaning that \( E_{\text{total}} = E_{\text{GO}} \).
For a finite absorbing baffle illuminated by plan wave incidence, the $E_{total}$ is given by [21]

$$E_{total}(x, y) = e^{-jkx} \left( 1 + \frac{e^{jkx}}{\sqrt{2}} \left( F_{res} \left( \sqrt{\frac{2}{\lambda x}} (-y_L - y) \right) - F_{res} \left( \sqrt{\frac{2}{\lambda x}} (y_L - y) \right) \right) \right),$$

(2.4)

where $y_L$ is the half of the baffle height. The total electric field over a finite absorbing baffle is shown in Fig. 2.2. As expected, at the middle of the baffle, $y = 0$, the Arago spot is seen. The bright spot is due to diffraction, and cannot be explained by GO.
In the next section, the UTD is studied to solve a more complicated problem, i.e.,
diffraction by a wedge that can be absorbing or conducting.

2.4 Essential UTD Diffraction

As noted above, GO is not sufficient to explain the behaviour of the electromagnetic field
in many situations. It is just able to calculate the backscattered field from edges, and
cannot predict a nonzero field in the forward scattering direction, especially in a shadow
region. The development of diffraction goes back to the works of Huygens, Poisson, Fresnel
and Arago, and several others. In more recent history, GO was extended by considering
“diffracted rays” - i.e., originating from Huygen sources - in addition to the usual optical
rays behaviour, viz., reflection, and refraction. The approach is widely attributed to Keller
and called the geometrical theory of diffraction (GTD) [15]. It gives the diffraction rays
emanating from edges and corners and describes the field in the shadow region. However,
this theory fails to explain the fields on or close to the geometric boundaries, i.e., transition
regions, and is only valid for plane wave incidence.

The UTD was developed by Kouyoumjian and Pathak [16] to repair GTD’s deficiencies
by formulating the fields in the transition regions, including having a source different from
a plane wave. This section reviews UTD as a solution for diffraction over a wedge, and then
applies it to the half-plane diffraction configuration.

2.4.1 UTD Diffraction by a Wedge

In the UTD [16], a wedge is parametrized by two faces, its exterior angle, \( n\pi \), and the
physical optics (PO) reflection coefficients of each facet, \( R_1 \) (lit side) and \( R_2 \) (shadow side).
A depiction is shown in Fig.2.3, where the source and receive directions are described by
(distance, angle) parameters, \((s', \phi')\) and \((s, \phi)\), respectively. In the terminology of GO, two
shadow boundaries - one for the incident rays, and one for the reflected rays - divide the
space into three regions. Therefore, Region I contains direct, diffracted and reflected rays,
Region II direct and diffracted rays, and Region III diffracted rays only. So the total field
is given by:

\[
E_{\text{total}}(x, y) = E_{GO}(x, y) + E_d(x, y) \tag{2.5}
\]

where \( E_{GO} \) is the geometrical optics field comprising both the incident and reflected fields
in region I, incident field in region II, and zero in region III.
Figure 2.3: The geometry of wedge diffraction (after [2]).

I. The UTD diffracted field:

The known UTD equations e.g., [2,17], are simplified here and summarized. The scattered (diffracted) electric field, relative to its incident field, \( E_i \), is

\[
E_{UTD} = E_i D (\phi', \phi) A (s) e^{-jk_s}.
\] (2.6)

where \( A (s) \) is a spreading factor, describing how the amplitude of the field varies along the diffracted ray. For plane, cylindrical, and conical wave incidence, \( A (s) = \frac{1}{\sqrt{s}} \), and for spherical waves,

\[
A (s) = \sqrt{\frac{s'}{s(s'+s)}}.
\] (2.7)

The plane wave incidence at the tip of the wedge is \( E_i = E_0 e^{-jk_s} \). It decays by \( \frac{1}{\sqrt{s'}} \) and \( \frac{1}{s'} \) for cylindrical and spherical wave incidence, respectively.

\[
D (\phi', \phi) = -\frac{e^{-i\pi/4}}{2n\sqrt{2\pi k}} (D_1 + D_2 + R_1D_3 + R_2D_4),
\] (2.8)

where \( D_1 \) to \( D_4 \) are the geometric optic contributions between the shadow boundaries,

\[
D_1 = \cot \left( \frac{\pi + (\phi - \phi')}{2n} \right) F (kLa^+ (\phi - \phi')),
\] (2.9)

\[
D_2 = \cot \left( \frac{\pi - (\phi - \phi')}{2n} \right) F (kLa^- (\phi - \phi')),
\] (2.10)
\[ D_3 = \cot \left( \frac{\pi - (\phi + \phi')}{{2n}} \right) F(k\Lambda^-(\phi + \phi')) , \]  
\[ D_3 = \cot \left( \frac{\pi + (\phi + \phi')}{{2n}} \right) F(k\Lambda^+(\phi + \phi')) , \]  
(2.11)

(2.12)

\[ F(x) \] is the transition function, which embeds the Fresnel-like integral, normally calculated using the \textit{erfc} function (below). For \( x > 0 \),

\[ F(x) = 2j\sqrt{x}e^{jx} \int_{\sqrt{x}}^{\infty} e^{-ju^2} \, du \]

\[ = \sqrt{j\pi x}e^{jx} \text{erfc} \left( \sqrt{jx} \right). \]  
(2.13)

\( x \) is zero at the shadow boundaries which results in \( F(x) = 0 \), while for \( x \gg 1 \), far from the shadow boundaries, \( F(x) \approx 1 \). \textit{erfc}(z) is the complementary error function defined by

\[ \text{erfc}(z) = \frac{2}{\sqrt{\pi}} \int_{\sqrt{z}}^{\infty} e^{-t^2} \, dt. \]  
(2.14)

The functions \( a^\pm(\beta) \) are

\[ a^\pm = 2 \cos^2 \left( \frac{2\pi N^\pm - \beta}{2} \right), \]  
(2.15)

and the integers \( N^\pm \) are from the best fit to

\[ N^\pm = \frac{\pm \pi + \beta}{2\pi n}. \]  
(2.16)

\( \Lambda \) is a distance factor which is equal to \( s \) for plane wave incidence. For cylindrical and spherical wave incidence, it is

\[ \Lambda = \frac{ss'}{s + s'}. \]  
(2.17)

II. The geometrical optics field:

The geometrical optics field in (2.5) consists of the incident and reflected field in region I. The reflected field can be obtained from the simplest case of reflection mechanism, i.e., a single reflection from a smooth surface with small grazing and reflection angle, \( \phi \), shown in 2.4. The heights of the transmitter and receiver are represented by \( h_T \) and \( h_R \), respectively. The distance between transmitter and receiver is \( d \). The path length difference between the direct path length \( R_d \) and the indirect path length \( R_i \) given in (2.18) is \( \Delta [2,23] \).
Figure 2.4: Single reflection of a smooth surface and a direct signal path [2].

\[ R_i = \sqrt{(h_T + h_R)^2 + d^2}. \]

(2.18)

For very small grazing angles, \( E_{GO} \) can be estimated as

\[ E_{GO_{h,v}} = E_0 \left( 1 + R_h,v e^{-j\Delta} \right), \]

(2.19)

where \( R_h = 1 \) and \( R_v = -1 \) are the Fresnel horizontal and vertical reflection coefficients. \( E_0 \) is the incident field given by

\[ E_0 = \frac{\lambda}{4\pi} \frac{e^{-jKR_d}}{R_d}. \]

(2.20)

### 2.4.2 Modeling the Half-plane Diffraction Using the UTD

Here, a half-plane (semi-infinite) baffle is modeled using UTD. The wedge angle is parameterized as \( n = 2 \), i.e., the wedge outside angle is \( 2\pi \). The critical 2D diffracted field for plane-wave incidence is attained from (2.6) where the distance factor \( L \) is \( s \), and 2D spreading factor \( A(s) \) is \( \frac{1}{\sqrt{s}} \). The total electric field is given by (2.5). The comparison of UTD and theoretical result in (2.2) for diffraction over a semi-infinite absorbing baffle (\( R = 0 \)) is shown in Fig.2.5. The diffracted field is discontinuous in the transition region, \( y = 0 \), and UTD and the theoretical results match very well in the shadow region and reasonably well in the lit region close to the transition region.
In addition, Fig.(2.6) illustrates the total field calculated by UTD over a perfect electric conductor (PEC) baffle for horizontal and vertical polarizations. As expected, the diffracted field for horizontal polarization \((R = -1)\) decays more quickly.

Figure 2.5: Diffraction over absorbing baffle: the comparison of UTD and simplest diffraction in 2.2 results.

Figure 2.6: The total electric field over a PEC baffle for horizontal and vertical polarizations.

### 2.5 Comparison of Diffraction with Numerical and Physical Experiments

The remainder of this chapter presents new numerical and physical experimental results and comparisons with theoretical derivations. The accuracy of a numerical approach for diffraction depends on the method for computing the fields. The simulations used for this
study are from an FEM approach, which can be expected to work reasonably well, whereas the MOM from just electric current sources cannot be expected to demonstrate diffraction for many situations. Physical measurements provide the ultimate test of accuracy, but they are complex and expensive for most real-world scenarios. Very few experiments have been reported on diffraction e.g., [20], probably due to the complex platform needed to get the spatial field measurements. Here, analytical, numerical and physical experiment results from laboratory measurements are compared. These results give a feel for the accuracy of a commercial solver for diffraction and indicate that a simple empirical model should be suitable for practical propagation modelling.

2.5.1 Simulation Modelling and Results

CST Microwave Studio time domain solver [24] is used for simulating diffraction over a semi-infinite and finite baffles. Perfect absorber is very difficult to simulate in CST-type programs, therefore a perfect electrical conductor (PEC) is used for the semi-infinite and finite baffles. All the simulation boundaries are defined as open boundaries with perfectly matched layer (PML) to accommodate an infinite structure [3]. In time domain solvers, the discrete grid equations of the finite integration technique (FIT) used in CST are identical to the discrete equations of Yee’s finite-difference-time-domain (FDTD) method [25]. However, It was found that the time domain solver leaked electromagnetic energy at the baffle to open boundaries [3,21]. Another solver in CST which supports PML boundaries is the frequency domain solver using FEM method. This solver also performs poorly at open boundaries. But this solver supports the periodic boundary condition to model the symmetric infinite side-boundaries for normally incident plane waves. However, for a semi-infinite baffle, the periodic boundary cannot be used for the bottom boundary, so a PEC boundary is assigned to eliminate the leakage at the bottom boundary. For horizontal polarization ($R = -1$) a perfect magnetic conductor boundary ($H_z = 0$) should be used. For vertical polarization ($R = 1$) a perfect electric conductor ($E_z = 0$) should be used [3]. A PEC boundary at the bottom leads to reflection of the diffracted field in the shadow region, so an electromagnetic absorber material is recommended for above the PEC boundary especially for horizontal polarization simulation(see [3]).

Figure 2.7 illustrates a semi-infinite baffle approximated by a finite baffle with the height of $15\lambda$ simulated in CST with a plane wave illumination for both horizontal and vertical polarization [3]. Figure 2.8 shows the results for a horizontally and vertically polarized plane wave at $8\lambda$ away from the baffle. For both cases, there is a good agreement between UTD and simulations. In horizontal polarization model, a few dB difference between simulation and UTD can be observed due to the reflection, and the total field calculated by UTD and simulation over PEC baffle for horizontal polarization decays more quickly.
Figure 2.7: Semi-infinite baffle approximated by a finite baffle with the height of $15\lambda$ for plane wave illumination in CST simulation for (a) horizontal polarization in z-direction ($R = -1$) and (b) vertical polarization in y-direction ($R = 1$) (see [3]).

Figure 2.8: The simulated and calculated total electric field over a perfectly conducting semi-infinite baffle approximated by a finite baffle with the height of $15\lambda$ for plane wave illumination at $8\lambda$ away from the baffle for (a) horizontal polarization in z-direction ($R = -1$) and (b) vertical polarization in y-direction ($R = 1$) (see [3]).

Figure 2.9 also shows a finite baffle with the height of $8\lambda$ simulated in CST with a plane wave illumination for both horizontal and vertical polarization. The observations are along a parallel line $8\lambda$ away from the baffle. Figure 2.10 shows the results for a horizontally and
vertically polarized plane wave. Same as the semi-infinite baffle, less shadowed is observed for vertical polarization.

Figure 2.9: Finite baffle with a height of $8\lambda$ for plane wave illumination in CST simulation for (a) horizontal polarization in z-direction ($R = -1$) and (b) vertical polarization in y-direction ($R = 1$) (see [3]).

Figure 2.10: The simulated total electric field over a perfectly conducting finite baffle with the height of $8\lambda$ for plane wave illumination at $8\lambda$ way from the baffle for horizontal polarization in z-direction ($R = -1$) and vertical polarization in y-direction ($R = 1$) (see [3]).

2.5.2 Physical Experiments

For physical experiments, an NSI planar scanner [26, 27] mounted behind a baffle, and illuminated by a horn antenna, is used. The mild clutter from other scattering, including absorbers, seems unavoidable. Such a comparison offers a feel for the need (or not) to use diffraction for a cluttered scenario. The set-up is shown in Fig. 2.11 for “semi-infinite”
and finite baffle. The frequency is 12GHz. The baffles are foil-covered cardboard, with the “semi-infinite” one extending to the floor (i.e., the floor represents $y = -\infty$ in the integration leading to (2.2) so the "semi-infinite" baffle is far from infinite of course. Table 2.1 gives the dimensions of the baffles.

![Experimental setup in an NSI planar scanner chamber](image1)

Figure 2.11: Experimental setup in an NSI planar scanner chamber for (a) the "semi-infinite" baffle and (b) finite baffle illuminated by a horn antenna, with mild clutter.

<table>
<thead>
<tr>
<th>Dimension</th>
<th>&quot;Semi-infinite&quot; baffle</th>
<th>Finite baffle</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length ($z$ direction)</td>
<td>56$\lambda$</td>
<td>60$\lambda$</td>
</tr>
<tr>
<td>Height ($y$ direction)</td>
<td>40$\lambda$</td>
<td>8$\lambda$</td>
</tr>
</tbody>
</table>

The planar scanner has its vertical locus centered at $y = 0$ and its spacing (the $x$-direction) set by manually moving the baffle. The distance between horn and probe is 120$\lambda$. For observing the total electric field, the semi-infinite is placed at 8$\lambda$ and 40$\lambda$, and the finite baffle is located at 8$\lambda$ and 30$\lambda$ from the probe. The results are plotted in Fig. 2.12 for both "semi-infinite" and finite baffles with horizontal polarization. The horn illumination is a dominant factor in the experimental results and the total electric field follows the taper of the horn pattern (not shown). Here, the measured results are normalized to the horn illumination i.e., the beamshape of the horn is removed. A close match between the theory and these experiments because of the horn taper and the mild clutter (diffractions, reflections and scattering from other close proximity surfaces) of the experimental situation cannot be expected. But from these experiments, the general form of the total field shows the phenomena of the GO plus diffraction.
Figure 2.12: The measured total electric field over (a) “semi-infinite” and (b) finite baffle for horn illumination with horizontal polarization.

2.6 Summary and Conclusion

The physical measurements show that the non-uniform illumination is a dominant factor. This suggests that a more complicated integration (i.e., including the non-uniform illumination, when it is known and can be modelled appropriately) is required for diffraction calculations in real-world situations. Numerical results seem to work well for the uniform illumination case. For modeling real-world barriers, it is evident that the principles of diffraction can guide the way to simpler and faster-to-solve empirical equations for diffraction situations such as buildings in urban environments. This is recommended for future work because urban propagation coverage calculations remain severely limited in their accuracy and computation time. The applications include backhaul link configuration and design, for example, where there remains a need for fast and accurate mean signal level prediction.
Chapter 3

Propagation Around Corners

3.1 Introduction and Background

The penalty on link gain for NLOS, relative to LOS, is enormous, and it is intuitive, in the sense of focusing on the weakest coverage area, to look for ways to reduce excess attenuation of NLOS propagation. For this reason, future systems can be expected to continue the trend toward smaller cell sizes with an associated higher path gain, which contributes to the link gain. A single-user system is often considered for getting new concepts established. An example is adaptive reflecting surfaces in the NLOS paths, e.g., [28]. For such a single-user system with single-port, fixed-beam antennas, the simplicity of the Friis equation is that the received power (proportional to most communications performance metrics) is proportional to just a few terms - the transmit power, the transmitting and receiving antenna gains, \( (G_T \text{ and } G_R) \), efficiency factors (collected below as \( \eta \)), and the term of interest for this chapter, the path gain, \( G_{Path} \). This is most familiar for the special case of LOS, where the electrical separation distance, \( d/\lambda \), is the only parameter: \( G_{Path}^{(f s)} = (4\pi d/\lambda)^{-2} \). But the Friis equation can be considered as defining the path gain, such as in lossy media or in a multipath environment, as used below. Friis’ choice of placing the frequency in the path gain is often interpreted that lower frequencies mean better link gain, and so the higher frequencies of future systems are seen as a disadvantage on this count. But it is straightforward to show that for fixed physical-size antenna apertures at each end of a link, that the link gain increases with frequency. The mechanism is that, setting the antenna gains from their electrical apertures which are increasing with increasing frequency (the physical aperture size is fixed, so the electrical size is increasing), means that the increasing antenna gains can dominate the decreasing path gain [2]. This assumes that the antenna apertures can be populated with arrays that form uni-directional beams for LOS. For NLOS situations, the arrays are deployed for MIMO communications.

The Friis formulation for the multipath situation can be expressed by the taking the terms as statistically independent:-
\[
\begin{aligned}
P_{Rx} &= P_{Tx}G_{Tx}G_{Rx}G_{Path}\eta, \\
&\text{(fixed single port antennas).} \\
\langle P_{Rx} \rangle &= \langle P_{Tx} \rangle \langle G_{Tx} \rangle \langle G_{Rx} \rangle \langle G_{Path} \rangle \langle \eta \rangle, \\
&\text{(averaging for multipath).}
\end{aligned}
\] (3.1)

The path gain term is the "problem", making the link efficiency (the ratio of receive power to transmit power), for typical NLOS links, to be in the order of \(-160\)dB. This low link efficiency has been disguised by the low energy required for radio signal detection. But for a massive number of links serving a massive number of terminals, the low energy efficiency becomes exposed, and in fact gets worse because of the interference from the other users [29]. From the Friis equation, the antenna gains are the only available solution in the sense that the average path gain depends on the changing physical multipath environment which (traditionally) cannot be changed.

Within the path gain term, the problem remains that higher frequencies do not diffract as well as lower frequencies, intuitive from the higher frequencies being closer to a geometric optics (GO) case. The impact is seen in the decreasing path gain with increasing frequency for shadow areas. For example, the classical Hata model for urban propagation has a frequency dependence of about \(f^{-2.6}\), which is approximately a free space term plus a single diffraction.

### 3.1.1 Challenges

Some control of the NLOS propagation can be considered as directly addressing the problematic path gain term in the Friis equation, rather than accepting it and focusing solely on the the antenna gains. This represents a grand challenge in new-generation wireless systems, and is being addressed using adaptive reflecting surfaces to control reflections in the NLOS propagation, e.g., [30]. Specific challenges for feasible smart surfaces in real-world systems include:

- The need for large areas on buildings for the adaptive reflecting surface. The cost of the surface electronics can be low, but obtaining use of the building area is currently a fraught and extremely costly item.

- The need for different reflection coefficients and directions for different users. Multiuser systems would require independent reflection coefficient control for the users, through their different source and sink locations, mixed polarizations, and variable frequencies of operation (channels and different systems).

- The system must be wideband so that the apertures of the adaptive surfaces can be shared between multiple frequencies.
• The selection and configuration of the adaptive surfaces requires a capacity overhead. For fast changing channels, such as from mobile terminals operating with high frequencies, the channel is changing very quickly and the capacity resource needed for controlling the links in a multiuser system become dominant, defeating the intent of the technology. This is a research topic in MIMO communications where joint optimization (or rather the channel sounding and channel state information interchange required) of the antenna weights of even a single user system can dominate the capacity resource.

• The increasing sophistication of the adaptive surface means that it is starting to function as a base station (BS), and so the benchmark for capacity improvement must be from using of the area resource occupied by the adaptive surfaces for extra BSs (with associated smaller cells and lower transmit power, offset by potentially more power consumption in the electronics).

The treatment of the impact of adaptive surfaces has been from basic information theoretic approaches using simple propagation models. These conceptual treatments are typically for single-user systems, and do not address the above challenges for a multi-user system that can be wideband.

3.1.2 Review and Contribution

This chapter deals with the attenuation of a corner - or rooftop, etc. - of a building and analyzing the path gain. (The ensuing conversion from link gain to communications metrics is omitted here). Such a corner is typically the first onset of NLOS. Here, only one corner is analyzed and a simple modification, here coined a *smart corner*, to improve the path gain relative to diffraction, is presented. The approach averts several of the above challenges because, in its simplest form, it does not require channel state information and adaptivity, and covers multiple users in the sense that it is wideband.

Nearly all radio communications links rely on NLOS propagation which explains a growing interest in influencing the propagation environment in order to increase the quality of service for wireless communications. The recent topic of adaptive surfaces, also referred to as intelligent reflecting surfaces (IRS), has been of significant interest in wireless research, e.g., [31–33]. An IRS comprises a planar array whose elements can change the reflected signal, in phase, amplitude, polarization, or even frequency. Other terms include reconfigurable intelligent surface (RIS) [28,34–36], large intelligent surface (LIS) [37–39], and passive intelligent surface (PIS) [40,41]. No fixed or adaptive corners have been presented within this conceptual theme. However, the approach of a scattering dipole at a diffracting edge draws on a tradition of trying to increase signal coverage, usually over a hill, by using passive (and sometimes active) repeaters. These comprise a pair of back-to-back, directly connected antennas receiving from the incident power direction and illuminating the shadow region. Ex-
amples include four-element patch antenna arrays [42,43], multiple aperture reflectors [44], endfire arrays [45], double flat reflectors [46], large passive reflectors [47], reflectarrays [48], Yagi-Uda antennas [49] and frequency selective surfaces [50,51]. Through-wall passive repeaters have also been used to improve indoor radio coverage [52–54]; active repeaters have been treated, e.g., [55–57]; and there is a large body of work in the communications literature, on communications metrics from using active relays. Recently, some effects of (non-adaptive) engineered surfaces in a wireless environment have been measured [58]. A simpler approach is presented here - the use of the simplest scatterers - a dipole, or some spatial sequence of dipoles (here called an array, or scattering array, but without the signal summation action of a normal array), to improve coverage. For adaptation, if required, the dipole scattering can be modified by changing its load, and this requires a minimally active system. This concept can be related to the use of switched parasitic elements to adaptively change the patterns of antennas, e.g. [59]. But here it is the scattering that is being modified rather than the mean gain of an antenna. This modification is constrained in the case of a single dipole because it has minimum scattering antenna properties and in particular, its transmit pattern is the same as its scattering pattern. The treatment is to use physics-based propagation modelling (as opposed to curve-fit type modelling), and also simulations which assist with visualization. The results are validated by physical measurements for the single dipole case. The concept of the simple dipole scatterer, or configuration of dipoles, at a corner for enhancing the shadow coverage, and all the analyses, are new contributions. The measurements are not trivial - the difficulty of measuring diffraction-like phenomena is recently addressed in [20].

The rest of the chapter is laid out as follows. Section 3.2 reviews the theoretical analysis of propagation around a corner, and section 3.3 lays out the model of a simple physical configuration. Simulation and measurement results are compared with the theory in section 3.4, and the summary and conclusion is given in section 3.5.

3.2 Corner Scattering Analysis

For propagation around a corner of a building (see Fig.3.1), the uniform geometrical theory of diffraction (UTD) [16,60] is used as a benchmark for the simulation and measurement results of diffraction by the corner of a building. The equations, e.g., [2,16,17], are simplified and summarized in section 2.4, to allow the calculations to be readily reproduced.

From the previous chapter, the corner is described by its exterior angle, \( n\pi \), \( (n = 1.5 \) or a right-angled corner) and the reflection coefficients of each facet, \( R_1 \) (lit side) and \( R_2 \) (shadow side). The directions to the corner of the transmitter and receiver are described by (distance, angle) parameters, \((s',\phi')\) and \((s,\phi)\), respectively, see Fig.3.2a.
Figure 3.1: Propagation around a corner of a building with a fixed dipole scatterer.

Figure 3.2: Nomenclature for propagation around a corner - plan view. (a) diffraction parameters, (b) the configuration of the scatterers used for analysis, simulation and measurement.

The received power via a scatterer can be simplified and calculated via the Friis transmission formula, based on GO. A simple Friis model for the propagation from the scatterer
is as follows. The paths are (i) from the transmitter to the scatterer, and (ii) from the scatterer to the receiver (Fig. 3.2b). Both of these paths are two-ray models - a direct ray and a smooth-surface reflection ray (Fig. 3.3). In terms of the model accuracy, the Fresnel zone on a reflecting surface is a larger area than the Fresnel zone for LOS (page 735, [2]). For this analysis, accurate GO analysis cannot be expected without the complete reflecting Fresnel zone being available, so the GO results for where the ray reflection is close to the corner may be inaccurate. Similarly, if the wall is electrically rough, then a rough surface scattering formulation is required, e.g., [2]. When the ray reflection is well away from the corner of a smooth wall, the results can be expected to be accurate. This is especially the case for small grazing angles where the wall surface roughness and permittivity do not strongly influence the reflection coefficient. This is in the direction of interest, i.e., the most shadowed zones. In the small-angle limit, the reflection coefficients become the same as for a perfect electrical conductor, used below in simulations.

Referring to Fig.3.2b, the received power, $P_{Rx}$, is given by

$$P_{Rx} = P_{Tx}G'_{Tx-Rs}G'_{Ts-Rx},$$

(3.2)

where the transmitter-to-scatter gain, following Friis but dropping the efficiency factors, is

$$G'_{Tx-Rs} = G_{Tx}G_{Rs}G_{Tx-Rs},$$

(3.3)

and the scatterer-to-receiver gain is

$$G'_{Ts-Rx} = G_{Ts}G_{Rs}G_{Ts-Rx},$$

(3.4)

where $G_{Tx}$ and $G_{Rx}$ are the gain of the transmitter and receiver. $G_{Ts} = G_{Rs}$ is the scatterer gain. $G_{Tx-Rs}$ and $G_{Ts-Rx}$ are the path gains of the two-ray model from the transmitter to scatterer and from the scatterer to the receiver, respectively. The two-ray path gain between a transmitter and receiver, $G_{T-R}$, is expressed as a complex amplitude, (see Fig. 3.3) is
\[ \sqrt{G_{T-R}} e^{jG_{T-R}} = \left( \frac{\lambda}{4\pi} \right) \left( \frac{1}{d_{los}} + R_{v,h} \frac{e^{j\varphi}}{d_{ref}} \right), \]  

(3.5)

where

\[ \varphi = 2\pi \frac{d_{los} - d_{ref}}{\lambda}, \]  

(3.6)

\[ d_{los} = \sqrt{d_{T-Rx}^2 + (h_{Tx} - h_{Rx})^2}, \]  

(3.7)

\[ d_{ref} = \sqrt{d_{T-Rx}^2 + (h_{Tx} + h_{Rx})^2}. \]  

(3.8)

The smooth-surface Fresnel reflection coefficient for the horizontal polarization is

\[ R_h = \frac{\sin \psi - \sqrt{\epsilon_r - \cos^2 \psi}}{\sin \psi + \sqrt{\epsilon_r - \cos^2 \psi}}, \]  

(3.9)

and for vertical polarization, is

\[ R_v = \frac{\epsilon_r \sin \psi - \sqrt{\epsilon_r - \cos^2 \psi}}{\epsilon_r \sin \psi + \sqrt{\epsilon_r - \cos^2 \psi}}, \]  

(3.10)

where \( \epsilon_r \) is the complex relative dielectric constant of the wall media, and the grazing angle \( \psi \) is

\[ \psi = \tan^{-1} \left( \frac{h_{Tx} + h_{Rx}}{d_{Tx-Rx}} \right). \]  

(3.11)

The received electric field is given by

\[ |E_{Rx}|^2 = \frac{2P_{Rx} \zeta}{Ae}, \]  

(3.12)

where \( \zeta = 120\pi \) and the effective aperture of the receiver antenna is

\[ Ae = \frac{G_{Rx} \lambda^2}{4\pi}. \]  

(3.13)

The two-ray path model is valid for the shadow region for \( X \leq -h_{Rx} \cot(\theta) \), where \( \theta \) is the angle of the scatterer(s) from the X-axis shown in Fig.3.2b. Therefore, the total electric field in this region is

\[ E_{Rx, total} = E_{Rx} + E_{UTD}, \]  

(3.14)

where \( E_{UTD} \) is the diffracted field from the corner in the absence of the scatterer, given by UTD, and \( E_{Rx} \) is calculated by the two-ray model from the transmitter to the scatterer and received by the receiver described in (3.12). As discussed in the context of the Fresnel zone on the reflecting surface, this two-path contribution is calculated in the absence of the corner. (See Fresnel zone discussion above - the reflecting surface needs to occupy the first Fresnel zone, but the corner can truncate this zone).
The received field in the lit region for \( X \geq h_s \cos(\theta) \) is

\[
E_{R\text{total}} = E_{T \rightarrow RX} + E_{T \rightarrow RX} + E_{UTD},
\]

where \( E_{T \rightarrow RX} \) is the calculated by the two-ray model from the transmitter to the receiver. \( E_{T \rightarrow RX} \) is the direct received field from the scatterer with its transmitted power calculated by the two-ray model from the transmitter to the scatterer.

### 3.3 Description of the Model Used for Demonstration

The geometry the smart corner model is recalled from Fig. 3.2b, and this is used for the analysis, simulation and measurement. The corner structure is modest in electrical size for simulation convenience, with a height, width and length of 14\( \lambda \), 6\( \lambda \) and 10\( \lambda \), respectively. A half-wavelength parallel dipole antenna is used as a transmitter, located at \( h_T = y_T = \lambda \).

For a building situation, the polarization is vertical. The "height" (horizontal spacing from the wall) of the observation point (receiving dipole) is \( h_R \) and is moved along the \( X \)-axis.

The observation point, scattering dipole, and the transmitter are aligned (at the same height up the building) in the plane of incidence.

The dipole scatterers are located parallel to the surface, and away from the corner with a spacing \( h_s \) and angle \( \theta \) from the \( X \)-axis. Four models of smart corner configurations are presented: a single dipole scatterer (Fig. 3.4a), dipole linear array (Fig. 3.4b), a three-dipole scatterer (Fig. 3.4c), and three-dipole scatterer array (Fig. 3.4d). The single dipole case is formulated above and is also checked by physical measurement. The other configurations are investigated only with simulation. The dipoles can have different loads, for example, the canonical terminations of open circuit, closed circuit and matched circuit. Adaptive variation with general loads could be achieved using switches or varactors, etc. This is particularly simple in simulation, but in a practical system, it requires active electronics and adaptive controllers, etc.

For simulation, CST Microwave Studio time domain solver [24] is used, with all simulation boundaries open using the so-called Perfectly Matched Layer (PML). Using a plane wave in CST results in some energy leakage from open boundaries [61], and this is why a dipole source is used. Using the dipole source forces the use of a perfectly electrical conductor corner material, otherwise fields will leak through any dielectric materials. So the reflection coefficients of the corner simplify to \( R = -1 \) for vertical (parallel) polarization and \( R = 1 \) for horizontal (perpendicular) polarization. The analyses also indicate (not included here) that the enhancement is less for the horizontal polarization (using magnetic dipoles made from slots); however, the corner attenuation for horizontal polarization is also less than that for the vertical polarization.
Figure 3.4: Side view of a (a) single half-wavelength dipole scatterer, (b) linear array of dipole scatterers, (c) three-dipole scatterer, and (d) linear array of three-dipole scatterers near the corner. The observation point (not seen in the side view), scattering dipole, and the transmitter are aligned (at the same height up the building) in the plane of incidence.

3.4 Results and Discussions

3.4.1 Single Dipole Scatterer near the Corner

Figure 3.4a shows the geometry of the simulation model of a dipole scatterer (half-wavelength, short-circuit or open-circuit load) near the corner. Note this is a side view. The spacing between the dipole and the corner for this example is $h_x = \lambda$ with $\theta = 45^\circ$, and the "height" (horizontal spacing from the wall) of the observation point is $h_{Rx} = 2\lambda$.

The simulation results of the corner diffraction (without scatterer), illuminated by a vertically polarized dipole is plotted in Fig. 3.5, and compared to the UTD results from
Figure 3.5: Theoretical and simulation comparison with and without a half-wavelength short-circuited dipole scatterer \((h_{Rx} = 2\lambda \text{ and } h_s = \lambda)\). The theoretical model (blue curve) is not expected to be accurate for small \(X\).

(2.6). The curves for these are overlapping in the figure. The power for the UTD formulation and the simulations are normalized at 1m from the source dipole.

The simulated result for a half-wavelength dipole scatterer over the corner is also plotted in the same figure, for \(h_{Rx} = 2\lambda\), and this gives a direct comparison with the diffraction-only result. These plots illustrate a dramatic enhancement of \(~10\text{dB}\) in the shadow region close to the corner and \(~20\text{dB}\) further away from the corner.

The two-ray path model is valid for \(-h_{Rx} = -2\lambda\) to \(-h_{Rx} = -6\lambda\) (and further, but not shown) in this example. This is because of the reflecting Fresnel region, as discussed above, the two-ray path model is valid for \(X \leq -h_{Rx} \cot(\theta)\), limited by the reflecting Fresnel region. Away from the corner, the single-ray model holds, and this is plotted for \(X > -1\lambda\). In between, the simulation result is the better choice. The result is normalized to the maximum received field in the lit region given in (3.15). Figures 3.6a and 3.6b show the simulated electrical field intensity for the corner without the scatterer, and with a short-circuit termination on the dipole scatterer. This cut corresponds to the plan view, or looking down from the top of a building corner. While Fig. 3.7a and 3.7b are in the \(YZ\)-plane at \(X = -2.45\lambda\), in the shadow region, from the tip of the corner in the \(X\)-direction, so this is a side view of the vertical corner of a building. These figures offer a visualization of the energy enhancement in the shadow region caused by the dipole scatterer at the corner. As expected from minimum scatter antenna (MSA) principles, the enhancement of the shadow region signal is very strong for the short-circuited dipole. It is less for the open-circuit case which corresponds to a pair of an electrically very short (quarter wavelength) wire scatterers. An MSA is often referred to as "invisible", because its scattering pattern is the same as its transmit pattern. This experiment demonstrates the limitations of such invisibility because
the shadow region fields are clearly illuminated by the combination of the scattering and transmit pattern, i.e., the scatterer is not invisible. The dipole can be also terminated by a matched load, and the result is that no transmit pattern will (re-) radiate, just the scattering pattern. This also illuminates the shadow region but at reduced power compared to the short-circuited case. This case, and other terminations, is a subject for future work. But these presented cases demonstrate the concept of reconfigurable scattering at a corner using a single scattering element.

![Figure 3.6: Depiction of the electric field intensity in the XY-plane at Z = 0. (Plan view for a building corner): (a) No scatterer (diffraction only), (b) a short-circuit termination on a half-wavelength dipole scatterer, and (c) an array of short-circuited dipole scatterers.](image)
Figure 3.7: Electric field intensity in the $YZ$-plane at $X = -2.45\lambda$ (in the shadow region) from the tip of the corner in the $X$-direction. (Side view of a building corner, vertical polarization): (a) no scatterer (diffraction only), (b) a short-circuit termination on a half-wavelength dipole scatterer, and (c) an array of short-circuited dipole scatterers.

Physical Measurements

The set-up is shown in Fig. 3.8 for the smart corner with and without the short-circuited dipole scatterer, in an anechoic chamber. The frequency is $2.45\text{GHz}$. The corner is foil-covered cardboard boxes. (The foil is quite flat with the ripples exaggerated by the lighting in the photos.) The corner size and the position of the transmitter, receiver and the short dipole scatterer are the same as for the simulation model. The receive dipole is moved along the $X$-direction. The transmitter is a standard horn antenna placed in $Y$-axis parallel to the wall, whereas a dipole is used in the simulation. The horn pattern keeps the unwanted directional energy (in the experiment) to a minimum. The results are plotted in Fig. 3.9, comparing with the simulated results, for $h_{RX} = 2\lambda$. The spacing between the dipole and the corner is $h_\delta = 0.8\lambda$ with $\theta = 45^\circ$. The dipole is mounted on a piece of foam taped to the conducting corner.

The measured data is normalized to the received max power in the lit region for both cases; i.e., with and without the short dipole scatterer. The experimental results follow the simulation well and demonstrate the dramatic electric field enhancement by locating the short dipole at the corner.
(a) Without a scatterer.  
(b) With a half wavelength short-circuited dipole scatterer.

Figure 3.8: Smart corner physical measurements (a) without and (b) without the half wavelength short-circuited dipole scatterer (mounted using a foam stand-off), in an anechoic chamber. The surface is flat with the foil ripples exaggerated by the lighting.

Figure 3.9: Measurement and simulation comparison - with and without a half-wavelength short-circuited dipole scatterer ($h_{RX} = 2\lambda$ and $h_s = 0.8\lambda$).
A Simple Wideband Scatterer

The coverage improvement by using a simple short-circuit half-wavelength dipole is shown to be tens of dB above that of diffraction. The simulation results also illustrate that the length of the dipole is not a major factor. Figure 3.10 shows that however a very small dipole scatterer, i.e. $L_s = 0.08\lambda$, does not increase the coverage, a large gain is available in the immediate shadow over a wide range of frequencies.

![Figure 3.10: Simulation results comparisons of a short-circuit dipole scatterer with various lengths, $L_s$. ($h_{Rx} = 2\lambda$ and $h_s = \lambda$).](image)

Perpendicular dipole over the corner

Here, the dipole orientations are all perpendicular to the wall. The UTD diffraction with the surface reflection, $R = 1$, is compared with the simulated corner without a dipole scatterer with horizontal (perpendicular) polarization (see Fig. 3.11). The simulation agrees well with the UTD. The presence of a parallel dipole scatterer does not change the field, as intuitively expected. A perpendicular half-wavelength dipole at the corner can be expected to have some effect. The simulation results in Fig.3.11 illustrate that for this case (with $h_s = \lambda$ and $\theta = 45^\circ$), the field is increased by only a few dB in the shadow region for a receiver "height" of $h_{Rx} = 2\lambda$. So the improvement is not very significant for this case (although for communications aspects, such as coding and modulation, a couple of dB is a large improvement). The reasons for this configuration not being so effective are as follows. Firstly, the dipole scatterer is no longer parallel to the source or receiving polarization. Also, the corner diffraction attenuation for horizontal (perpendicular) polarization is less than that of the vertical (parallel) polarization. In this sense, the diffraction attenuation problem is not as important as for the vertical polarization case. The simulated electrical
Figure 3.11: Comparison of the normalized simulated electrical field over the corner (XY-plane at $h_{Rx} = 2\lambda$), without and with a "vertical" half-wavelength dipole with a spacing of $h_x = \lambda$ and $\theta = 45^\circ$. The transmitter is a dipole antenna with horizontal (perpendicular) polarization (surface reflection coefficient is $R = 1$). The simulated result is also compared to the UTD for validating the simulation model.

(a) Without a passive dipole scatterer.  
(b) With a "vertical" passive $\lambda/2$ dipole ($h_x = \lambda$ and $\theta = 45^\circ$).

Figure 3.12: The simulated horizontal (perpendicular) electric field in the XY-plane for a corner (a) without and (b) with a "vertical" half-wavelength dipole scatterer. The transmitter is a dipole antenna with horizontal (perpendicular) polarization (corner reflection coefficient is $R = 1$).

Field intensities on XY-plane are depicted in Fig. 3.12a, illustrating, in comparison to Fig. 3.12b, the reduced improvement.
3.4.2 Linear Array of Dipole Scatterers

Figure 3.4b illustrated the configuration of dipole scatterers near the corner, and Fig. 3.6c and 3.7c depict the same sequence of results as above for the single dipole scatterer. The main differences are the stronger enhancement in the shadow region, as expected, for the short-circuited case (e.g., almost 30dB at $X = -6\lambda$). These results, presented in Fig. 3.13, demonstrate the concept of further coverage enhancement from using a simple linear array configuration of the simplest scattering elements. Using array techniques (different terminations, and making the terminations dependent in some way) allows significant control of the scattering fields. In this thesis, no attempt has been made to make such an array with differing weights (terminations) on the elements, or to electrically link the elements with circuits, this is an interesting avenue for future work.

![Simulation results comparisons](image)

Figure 3.13: Simulation results comparisons- with open-circuited and short-circuited dipole scatterer and array of dipole scatterers. ($h_{Rx} = 2\lambda$ and $h_s = \lambda$).

3.4.3 Three-Dipole Scatterer and its Array

Another configuration that allows further variation through termination variation is to have coupled dipoles as the scatterer element. An example is shown in Fig. 3.4c (side view). Two extra parallel dipoles (all are in the Z-direction) with a shorter length (same effect as a changed load impedance) are placed beside the dipole at $\theta = 45^\circ$, in a right-angled configuration. The spacing between the shorter and longer dipoles are $0.15\lambda$, which allows strong mutual coupling. This structure relates to a multiple Uda-Yagi configuration. The spacing between the longer dipole and the corner for this example is $h_s = \lambda$ and the "height" (horizontal spacing from the wall) of the observation point is $h_{Rx} = 2\lambda$. 

38
Figure 3.14: Depiction of the electric field intensity in the $XY$-plane at $Z = 0$. (Plan view for a building corner): (a) short-circuited three-dipole scatterer, (b) an array of short-circuited three-dipole scatterer.

Figure 3.15: Electric field intensity in the $YZ$-plane at $X = -2.45\lambda$ (in the shadow region) from the tip of the corner in the $X$-direction. (Side view of a building corner, vertical polarization): (a) short-circuited three-dipole scatterer, (b) and array of short-circuited three-dipole scatterer.

The results, following the same depiction as for the above cases, are in Fig. 3.14a and 3.15a. The array is depicted in Fig. 3.4d. The results for the field intensity are in Fig. 3.14b and 3.15b. The results for these cases are also plotted in Fig. 3.16 for $h_{Rx} = 2\lambda$. There is a lot of information in these plots, but the main point is the demonstration of the concept that the coverage enhancement in the shadow area can be extremely large (e.g., almost 35dB at $X = -6\lambda$, and more than this, farther from the corner). These cases also demonstrate
that the amount of enhancement can be configured by changing the terminations on the scattering elements. The results for other combinations of the terminations are not reported here, for brevity.

Figure 3.16: Simulation results comparisons- with open-circuited and short-circuited three-dipole scatterer and array of three-dipole scatterer. ($h_{Rx} = 2\lambda$ and $h_{S} = \lambda$).

### 3.4.4 Other Considerations

The concept of spectacular enhancement of the shadow region with the smart corner has been demonstrated. The system is simple to deploy, extremely wideband, and extremely effective because it raises the shadow area coverage by tens of decibels. This means that the concept is potentially impactful, but in a real-world environment, other scattering may fill in the shadow region as well.

The question of construction and deployment has not been considered here. For the experiments, a foam "corner" moulding to support the passive dipole elements is used. In practice, such an attachment could be readily adhered to a building wall, either with foam, or mounted on a plastic support. For the higher frequencies of future systems, this could be very uninvasive because the structure would be physically very small. It could be placed above normal human heights to avert human interference.

Another aspect of the results is that the surfaces in the canonical corner model are assumed to be perfectly conducting. A conducting base added to the foam stick-on structure is straightforward. But in practice, concrete-clad corners and glass-corners will behave differently to conducting structures, and the large returns may not be there for such corners, because there may be penetration of the waves through the building. On the other hand, the higher frequencies of future systems are still attenuated/reflected by concrete and glass, and in fact at any frequency, even a rough or lossy surface begins to look like a perfect conductor for small grazing angles - exactly the case here.
3.5 Summary and Conclusion

Future systems for urban communications will feature higher carrier frequencies, larger multiport antenna systems, and perhaps emerging technologies such as seen with the recent interest in adaptive surfaces for changing the nature of reflections in the multiple propagation paths. However, it is also the corners of buildings that offer NLOS coverage through diffraction. The associated shadow regions often dominate the limitations in coverage, with increasing frequencies having decreasing coverage. To counter this problem, a simple concept to improve the link gain in the shadow region has been presented. The concept as a fixed system is demonstrated, showing an improvement of tens of decibels in the severely shadowed region. It is also possible to make the system adaptive by reconfiguring the loads on the dipole scatterers. This can change the nature of the scattering which may lead to better coverage for specific areas of the shadowed multipath environment. Using a matched load, the dipole power is absorbed which can reduce the coverage in the shadow region for the case of the diffracted signal being interference.
Chapter 4

Through-Forest Propagation

4.1 Introduction and Background

Progress in propagation modeling, in particular through vegetation, has not kept pace with wireless deployment. Current propagation models for the various scenarios, are inaccurate in the sense that several (up to tens) of decibels of uncertainty is common. However, as summarized in the previous chapters, much progress has been made on the electromagnetic analysis of canonical situations, such as the diffraction of idealized single [16,17,62] and multiple edges [2,8,18,19,63–65]. A single mechanism, even a critical one such as diffraction, seldom dominates all the coverage zones of a real-world link. There are few measurements of propagation through vegetation, mostly with very high uncertainty, and yet these are the basis of standard models (see below). Ray tracing is a popular technique for multipath propagation modelling, but is best suited for a smooth-surface geometric model, such for a city where the building geometry is either already publicly available or can be readily derived from public databases or map resources [66]. Reflections and diffractions are used, although there is no standard for how many rays (and how many reflections and diffractions), and/or what minimum path gain, etc., is used as a criteria to terminate the ray-tracing process.

The physics of radiowave propagation through vegetation is extremely complicated. The mechanisms include: penetration (a direct wave permeating the dielectric vegetation obstacles); multiple scattering between the wide variety of vegetation components; reflection off the ground, or even a surface wave along the ground; rough surface reflection and scattering; diffraction around and over the vegetation obstacles; and forward scattering along the outside boundaries of the vegetation. (It is assumed here that the propagation around a forest is negligible, which seems reasonable for point-to-point deep within large forest). These mechanisms are all extremely difficult, if not impossible, to model accurately for a real-world situation. The analysis of propagation through/around even a single tree is fraught, and as noted above, physical measurements still form the basis of models, which still have high uncertainty. The analysis for propagation through extended vegetation, such
as just a few trees, or a forest, is similarly fraught, and again the parameters for standard
models (such as RET, specified for short distances only) are empirical.

A standard approach to scattering loss is to characterize the media as containing ran-
domly spaced discrete scatterers from a statistical overlay. In such an approach, the domi-
nant mechanism causing excess propagation loss is the scattering of the energy away from
the point-to-point path \[2,67\]. Empirical models based on measurement results can be in-
cluded as parameters of such a model. The empirical models of \[1,68–73\] are widely used
against specific measurements results. Recent measurements are mainly for the so-called
5G bands, which are higher frequencies than the 1GHz measurements \[9\] of interest for
this work. 5G frequencies have a much lower range because of the decreased path gain of
the higher frequencies. For example, \[74\] reports 28GHz measurements for distances up to
300m, and the reader is also referred to the references therein for samples of recent papers
featuring higher frequency measurements and model-fitting.

There are analytical models \[6\] which can be based on physics models for canonical
(statistical) situations. One approach is based on classical wave theory and uses a fractal-
based scattering overlay, and is reported to work well for "large" propagation distances
\[75–77\] (meaning in this case up to 500m, whereas in this work, "large" is taken to mean
well over 2000m). An earlier, prominent theory is the diffusion process-oriented RET, also
called Transport Theory \[5,78\]. This theory models propagation through a statistically
homogeneous (so the randomness is "simple") continuum of electrically small absorbing
scatterers in free space. The excess loss mechanisms of the RET are the scattering away
from the point-to-point direction and the absorption by the scatterers. No other mechanisms
are included, such as diffraction around a body. RET is equivalent to Boltzmann diffusion, is
mathematically complicated, and it contains a relatively large number of parameters \[5,78\].

In the absence of other classical models (alternative works, such as \[75–77\] are more
recent), the RET model has been assumed as suitable for embedded-in-foliage scenarios
\[4,5,78\], and was adapted for through-vegetation propagation laid out in ITU-R.833 \[1\]. The
scenarios even include a single-tree obstacle - which is a major departure from its founding
physics. In the ITU-R.833, the complicated parameters of the complicated RET formulation
are empirically set based on very few measurements which usually have few, or even single,
observations (i.e, not a respectable statistical ensemble). The end product is a complicated
empirical model which maintains the high uncertainty of the empirical parameters used
for the curve fitting. For species other than those in existing measurements, or for other
radio frequencies and other foliage characteristics, interpolation and extrapolation from the
existing experimental data is used to determine the parameters \[4\].

Despite the empirical approach, the RET-based model does not tend to give very accu-
rate loss prediction even for the specific conditions corresponding to the measurement data.
(It is emphasized that 'accurate' in statistical propagation modeling and prediction, can
still mean an uncertainty of some 10dB, or even several tens of decibels). To improve the
RET model, larger and statistically-planned physical measurement campaigns are required, but new and improved data cannot reduce its complexity. Therefore, there is interest in a simpler propagation model for through-vegetation, that can be empirical and should yield results which are at least as good as the RET model. Published physical experiments are surprisingly sparse (and more would be extremely welcome), but they seem to indicate, in an ensemble sense, reasonably simple mean path gain behaviour.

4.2 Contributions and Organization of the Chapter

In this chapter, vegetation loss based on RET theory is reviewed in section 4.3. In section 4.4 a transmission line model based on the RET results is used as for giving much simpler cook-book formulas for short-distance propagation (less than 100m). Section 4.5 discusses a long-distance through-forest propagation model (over 2000m) which strives to maintain some of the physics by looking at just two mechanisms: penetrative transmission directly through the randomly-layered media of a forest (similar to RET in this sense, although no explicit connection is required to the physics of electrically small scatterers), and along the free-space dominated forest top. Although the equations are known for these propagation mechanisms, a review of them is included below for completeness of this chapter. Applications include terrestrial point-to-point links, for fixed links such as cellular backhaul, where trees seem to get in the way surprisingly often, but in particular for long-distance through-forest communications, such as an Internet-of-Things system requiring device-to-device communications between small terminals and a cellular type network, or other mobile communications links such as cellphones, within a forest. The new contributions here include the use of the very simple transmission line model for short-distance (up to about 200m) behaviour with a very simple diffraction-derived model for long-distance (over 500m) behaviour, and their linear combination for both the dual-slope behaviour and the transition between the slopes to match to experimental data. These experiments are recent [9], describing propagation over a long distance through a forest, using a narrow-band system to achieve the large dynamic range. The RET model can also produce a dual slope phenomena, but its parameters cannot be configured to accurately follow the dual slope behaviour of the long distances of the measurements of [9]. The summary and conclusion is given in section 4.6.

4.3 Basic Radiative Energy Transfer Method

The high-frequency radio wave propagation in the vegetation characterized as a scattering random homogeneous medium is predicted by RET. The large dimension of scattering objects compared to millimeter wavelengths generates strong forward scattering. Their scatter characteristics explained below may be expressed by a transmit pattern forward lobe with an isotropic background. Because of the size of typical vegetation scatterers, the RET is
typically for frequencies above 1GHz [4, 79]. The vegetation medium is taken to have uniform characteristics throughout its volume characterized by following parameters. These parameters depend on tree species, leaf size, foliage density and frequency, and are found in ITU-Rec. 833 [1]. The parameters are:

- $\sigma_a$: The absorption cross section per unit volume ($m^{-1}$);
- $\sigma_d$: The (total) scatter cross section per volume ($m^{-1}$);
- $\beta$: The width of the forward scatter lobe (degree);
- $\alpha$: The ratio of forward scattered power to the total scattered power.

To estimate the parameters, one can fit the RET curves with the experimental results. By formulating the RET equation to estimate the attenuation and scatter of vegetation, relationships between the parameters are obtained. The list of the parameters is obtained from a limited number of measurements and consists of a few species at a number of frequencies. However, for other species and frequencies, these parameters need to be determined through interpolation of frequency and the vegetation characteristics from experiments [1]. To improve this model, larger physical measurements are required for various cases at different frequencies.

Note that the usual simple model for path loss, for example an exponential distance relation, cannot predict the propagation loss through vegetation very well. This is because such simple path loss models do not consider forward scattering caused by random obstacles [80]. The path loss is presumed to be the same at a given transmit-receive distance. However, the received power experiences random variations due to numerous foliage density and species in different seasons and at different frequencies. Thus, the models fail without the exact knowledge of the actual medium. The relation between the simplified path gain $G_{path}$ and distance $d$ in dB is expressed by

$$G_{path} = P_{Rx} - P_{Tx} = G_{Tx} + G_{Rx} + \gamma 10 \log_{10} \left( \frac{\lambda}{4\pi d} \right)$$  \hspace{1cm} (4.1)$$

where $P_{Tx}$ and $P_{Rx}$ are transmitted and received power where transmitter and receiver are located at distance $d$ to each other. $\gamma$ is the path loss exponent depending on the type of environment; i.e., the parameter $\gamma$ equals to 2 in free space conditions. The path gain is plotted in Fig.4.1 for different path loss exponents compared to the physical measurement of a "Common Lime" tree out of leaf at 1.3GHz given in [4], which is the received power due to the vegetation excess loss apart from the free-space path loss and the best fit RET curve to the measured data. Note that the impact of antenna gains; i.e., $G_{Tx}$ and $G_{Rx}$ are neglected in this plot. The measured data is an excessive loss depicting high attenuation rates correspond to short distances and lower attenuation rates associated with long distances [4, 5]. However,
the path gain decreases as distance increases. Note that for large dense region \((\gamma > 2)\), a sharp drop in the receive power is expected.

Figure 4.1: The received power for different path loss components compared to the measurement of a "Common Lime" tree out of leaf at 1.3GHz given in [4], which is the received power due to the vegetation excess loss apart from the free-space path loss, and the best fit RET curve.

Here, the parameters corresponding to the RET model are summarized. The full details of how the equation is derived are in [5, 6].

**Specific Intensity**

The basic RET equation is formulated in terms of the specific intensity \(I(\hat{r}, \hat{s})\), which is proportional to the field squared, e.g., \(|E|^2\). As shown in Fig.4.2a, a flow of energy is considered at a point \(\hat{r}\) in a direction \(\hat{s}\). The specific intensity is given as the power density \(dP\) per unit solid angle instead of the flow of energy. The power flowing through a unit area \(dA\) within a solid angle \(d\Omega\) in direction is given by [5, 6]

\[
dP(\hat{r}, \hat{s}) = I(\hat{r}, \hat{s}) \cos(\theta) dA d\Omega \quad (Watts).
\]  

(4.2)

where \(\theta\) is the angle between the direction of the power emitting from the surface and its unit normal. The unit of the specific intensity is \(W m^{-2} sr^{-1}\). When the differential area \(dA\), is perpendicular to \(\hat{s}\), specific intensity in a random homogenous medium is given by [5, 6]

\[
I(\hat{r}, \hat{s}) = \frac{dP(\hat{r}, \hat{s})}{dAd\Omega}.
\]  

(4.3)

Here, waves are not assigned a polarization and the intensity is scalar. Figure 4.2b demonstrates scattering in a homogeneous vegetation medium.
Figure 4.2: (a) Definition of specific intensity emitting from $dA$, (b) Scattering from a homogeneous random vegetation medium (After [5,6]).

The specific intensity $I(\hat{r}, \hat{s}')$ in direction $\hat{s}'$ incidents a cylindrical volume with unit cross section and length $ds$. The theory is formulated based on the power conservation theorem so the difference between entered intensity $I(\hat{r}, \hat{s}')$ and emanated intensity $I(\hat{r}, \hat{s})$; i.e., $dI(\hat{r}, \hat{s})$, gives the power loss. The emitted specific intensity contains the incident and scattered wave power. The absorption and scattering cross section (total extinction) of the medium decrease the power propagating in the direct path $\hat{s}$. Moreover, the scattered intensities entering from other directions, i.e., $\hat{s}'$ are redirected to $\hat{s}$ direction which boosts the power. Therefore, the basic equation of RET theory is given by [5,6]

$$dI(\hat{r}, \hat{s}) = -(\sigma_d + \sigma_s)I(\hat{r}, \hat{s}) + \frac{1}{4\pi} \iiint P(\gamma)I(\hat{r}, \hat{s}')d\Omega'. \quad (4.4)$$

where $P(\gamma)$ is the (power) scatter function, and $\gamma = \arccos(\hat{s}, \hat{s}')$ is the scatter angle. $P(\gamma)$ is assumed normalized such that

$$\frac{1}{4\pi} \iiint P(\gamma)d\Omega' = 1. \quad (4.5)$$

The scatter function determines how the incident specific intensity from direction $\hat{s}'$ is redirected to the direct path $\hat{s}$. There are some assumptions in this theory, as follows [5]

- The scatter function is modelled as a strongly directive narrow forward lobe with an "isotropic background" - meaning low directivity in other directions, since most forest elements have dimensions larger than mm-wavelengths and produce relatively strong forward scattering. It is conveniently modelled by 2D power pattern as

$$P(\gamma) = a q(\gamma) + (1 - a), \quad (4.6)$$
where \( \alpha \) is the ratio of forward scattered power to the total scattered power and the term \((1 - \alpha)\) implies the so-called isotropic background.

- Experimental evidence indicates that \( q(\gamma) \) can be assumed as a Gaussian forward lobe which is given by

\[
q(\gamma) = \left( \frac{2}{\beta_s} \right)^2 e^{-\left( \frac{\gamma}{\beta_s} \right)^2}
\]  

(4.7)

where \( \beta_s \) is the width of the forward lobe. It is the angle between the points where the magnitude of the power decreases by a factor \( 1/e \). It is related to \( \beta \), i.e., the full width at half maximum (FWHM) of the forward lobe by \( \beta_s = 0.6 \beta \).

- The vegetation is characterized as a homogeneous medium of random scatterers, i.e., \( \alpha_a, \alpha_s, \beta, \alpha \) and \( P(\gamma) \) do not depend on \( \hat{r} \).

Solution for the RET equation

The RET geometry for mm-wave propagation through vegetation is depicted in Fig.4.3.

![Figure 4.3: The RET geometry for mm-wave propagation through vegetation (After [4]).](image)

The theory is applied to homogenous media consist of random discrete scatterers. The vegetation boundary is modelled as two half-spaces, one is the air and the other is the vegetation. The assumptions recommended in ITU-R are as follows [1]:

- The transmit antenna is located in the air half-space far away from the planar interface so the illumination is assumed to be a plane wave incident.

- The plane wave is assumed to be perpendicular to the air-vegetation interface.

- The receive antenna is within the vegetation and its coordinate system is based on the incident plane wave direction. \( \Delta \gamma_R \). In the original literature, the beamwidth of the receiving antenna is the angle where the magnitude of the power decreases by a
factor $1/e$. It is determined by $\Delta \gamma_R = 0.6 \Delta \gamma_{3dB}$, where $\Delta \gamma_{3dB}$ is the 3dB beamwidth of the receiving antenna [5].

Two essential components of the RET equation; the total extinction, i.e., $\gamma_a + \gamma_s$, and the phase function $P(\gamma)$, are attained from a single scatterer and then the RET equation is solved subject to appropriate boundary conditions. The solution is explained in [5] and the following equations are from [5]. In the case of normal incidence, because of the symmetry about the $z$ direction, the specific intensity $I(z, \theta)$ changes with the vegetation depth $z$, and $\theta$ (i.e., the elevation angle) only, and it is independent of $\phi$ (i.e., the azimuth angle) . Solving the (4.4) provides the solution $I(z, \theta)$ which is expressed by the sum of two components as given below (see Fig.4.4)

$$I(z, \theta) = I_{r_1}(z, \theta) + I_d(z, \theta) = I_{r_1}(z, \theta) + I_1(z, \theta) + I_2(z, \theta). \quad (4.8)$$

**Figure 4.4:** Two components of total specific intensity (After [6]).

$I_{r_1}(z, \theta)$ is the coherent component reducing incident intensity in the direction $\hat{s}$ due to absorptions and scatters at short distances. $I_d(z, \theta)$ is the incoherent component. It is the diffuse intensity which dominates at large distances. The diffuse intensity includes two parts: $I_1(z, \theta)$ and $I_2(z, \theta)$ which are determined by scattering into the forward lobe of the phase function, and scattering into the isotropic background, respectively. Finally, the RET equation is solved subject to appropriate boundary conditions and the resulting equation is given by [5,6].

$$\frac{P_R}{P_{max}} = I_{r_1} + I_1 + I_2, \quad (4.9)$$

where

$$I_{r_1} = e^{-\tau}, \quad (4.10)$$
\[ I_1 = \frac{\Delta\gamma_R^2}{4} \left\{ e^{-\hat{\tau}} - e^{-\tau} \right\} \bar{q}_m + \sum_{m=1}^{M} \frac{1}{m!} (\alpha W \tau)^m \left[ \bar{q}_m - \bar{q}_M \right] \}, \quad (4.11) \]

and

\[ I_2 = \frac{\Delta\gamma_R^2}{2} \left\{ -e^{-\tau} + \sum_{k=N+1}^{M} \left[ A_k e^{-\frac{\hat{\tau}}{s_k}} \frac{1}{s_k} \right] \right\}. \quad (4.12) \]

\( P_R \) is the received power by the receiving antenna and \( P_{max} \) is the received power by the receiving antenna placed at the air-vegetation interface \((z = 0)\). \( m \) is the order of the term \( I_1 \) with the maximum order \( M \). Higher values of \( m \) give a more accurate solution for \( I_1 \), though, \( I_1 \) will not change for \( m > 10 \). The approximation used to solve \( I_1 \) is valid as long as the forward lobe of the phase function is narrow \((\beta_s \ll \pi)\). The distance into the vegetation is called optical thickness or optical density which is a unit-less quantity and can be given by

\[ \tau = \int_0^z \sigma_T(z) zdz, \quad (4.13) \]

where \( z \) is the distance into the medium. Recall that the total extinction, \( \sigma_T = \sigma_a + \sigma_s \) for a homogeneous medium of random scatterers is the same at every points of the medium so

\[ \tau = (\sigma_a + \sigma_s)z = \sigma_T z. \quad (4.14) \]

\( W \) is the single-scattering albedo expressed as

\[ W = \frac{\sigma_s}{\sigma_a + \sigma_s}, \quad (4.15) \]

and this is different from \( \hat{W} \) which is the reduced albedo. It is reduced by multiplying \( \sigma_s \) by \((1 - a)\). This parameter contributes to term \( I_2 \) caused by scattering in the isotropic background. Hence

\[ \hat{W} = \frac{(1 - a)\sigma_s}{\sigma_a + (1 - a)\sigma_s} = \frac{(1 - a)W}{1 - aW}. \quad (4.16) \]

\( \hat{\tau} \) also differs from optical density \( \tau \) and is defined as

\[ \hat{\tau} = (\sigma_a + (1 - a)\sigma_s)z = 1 - aW. \quad (4.17) \]

The attenuation coefficients \( s_k \) are determined by solving the following equation for \( s \)

\[ \frac{\hat{W}}{2} \sum_{n=0}^{N} \frac{P_n}{1 - \frac{\mu_n}{s}} = 1, \quad (4.18) \]
Table 4.1: RET required parameters for "Common Lime" tree at 1.3GHz [1].

<table>
<thead>
<tr>
<th>Leaf condition</th>
<th>$\alpha$</th>
<th>$\beta$</th>
<th>$W$</th>
<th>$\sigma_e (m^{-1})$</th>
</tr>
</thead>
<tbody>
<tr>
<td>In leaf</td>
<td>0.9</td>
<td>76</td>
<td>0.95</td>
<td>0.22</td>
</tr>
<tr>
<td>Out of leaf</td>
<td>0.95</td>
<td>50</td>
<td>0.95</td>
<td>0.591</td>
</tr>
</tbody>
</table>

where $N$ is an odd integer between 11 and 21, and

$$
\begin{align*}
P_n &= \sin^2 \left( \frac{\pi}{2N} \right) \quad \text{for} \quad n = 0, N, \\
\left\{
\begin{array}{ll}
P_n &= \sin \left( \frac{\pi}{N} \right) \sin \left( \frac{n\pi}{N} \right) & \text{for} \quad n = 1, \ldots, N - 1.
\end{array}
\right.
\end{align*}

(4.19)

$$

$\mu_n$ is given by

$$
\mu_n = -\cos \left( \frac{n\pi}{N} \right)
$$

(4.20)

Equation (4.18) has $N + 1$ solutions, so that

$$
\frac{s_0, \ldots, s_{N - 1}}{2} = -\frac{s_{N + 1}, \ldots, s_N}{2}
$$

(4.21)

Therefore, the amplitude factors $A_k$ are determined only for $k$ from $\frac{N + 1}{2}$ to $N$ by solving the below system of $\frac{N + 1}{2}$ equations

$$
\sum_{k = \frac{N}{2} + 1}^{N} \frac{A_k}{\mu_n} = \frac{1}{\sigma_n}, \quad n = \frac{N + 1}{2}, \ldots, N,
$$

(4.22)

where

$$
\begin{align*}
\sigma_n &= 0 \quad \text{for} \quad n \neq N, \\
\sigma_n &= 1 \quad \text{for} \quad n = N,
\end{align*}
$$

(4.23)

and

$$
q_m = \frac{4}{\Delta\gamma_R^2 + m \beta_s^2}
$$

(4.24)

For example, the RET best curve for measured data and the effect of $I_{i1}$, $I_1$ and $I_2$ contributions on the total term given in (4.3) are plotted in Fig.4.5. The plots are evaluated by using the required parameters for the "Common Lime" tree out of leaf at 1.3GHz given in [1] and summarized in Table 4.1. The beamwidth of the receiving antenna is assumed to be $\Delta\gamma_R = 0.6 \times 1.2^\circ = 0.7^\circ$. 

51
Figure 4.5: The total normalized received power and the three individual contributions, i.e., $I_{ri}$, $I_1$ and $I_2$, in RET equation, i.e., (4.3), compared to the measured data of a "Common Lime" tree out of leaf at 1.3GHz given in [4].

There are some observations that can be made from the above results [4,5]

- At small distances into the vegetation, the attenuation is obtained only from the coherent component $I_{ri}$ in the direction of incident plane wave due to absorption and scattering. The attenuation rate is $\tau/d = \sigma_a + \sigma_s = \sigma_T$.

- The incoherent components $I_1$ and $I_2$ are generated by the scattering of the coherent component $I_{ri}$. At the air-vegetation interface ($d = 0$), no scattering of the coherent components has happened so $I_1 = I_2 = 0$. $I_1$ and $I_2$ increase as $d$ increases and reaches a maximum and then decreases due to the exponential term at large distances.

- In the range of intermediate distances, the forward scattered component $I_1$ dominates, where $a$ is close to unity and the medium is strongly scattering, i.e., $\sigma_s \gg \sigma_a$. Thus $W$ is close to unity. The attenuation rate $\tau/d = \sigma_a + (1-a)\sigma_s$ is smaller than that of the coherent component. Thus, the rate is determined mainly by absorption as well as scattering into the isotropic background only. At large distances, the received power is determined by the contribution of $I_2$. Note that the intensity $I_2$ decreases essentially due to isotropic scattering with smaller attenuation rate. The reduced albedo $\hat{W}$ contributing to $I_2$ is close to unity because of the scattering medium.

Hence, the analytical results predict the strong received power attenuation due to the coherent (direct path) component at short distances. At large distances, less attenuation is expected because of the incoherent (multiple-scattered) component. Therefore, the RET model can produce a dual slope phenomena. RET theory models the foliage as a random
medium composed of discrete scatterers which is used widely for the vegetation loss prediction in the microwave and millimeter-wave bands. The parameters of the model are determined from a limited number of measured data via best-fit curve fitting. Therefore, this theory is validated for a specific type of trees at a few frequencies. Also, the measurements are limited to the configuration and density of the foliage in specific season and measurement site geometry. This model is mathematically complex. Thus there is a need for a faster and simpler model with fewer parameters. In section 4.4, a simple transmission line model with fewer parameters is introduced for short-distance foliage. Section 4.5 presents a two-mechanism model can provide an accurate fit to the dual-slope profile of through forest propagation over a long distance which is not possible with the RET model.

4.4 Simple Transmission Line Model from RET Results for Short-Distance Foliage

To simplify the complex RET solution, the propagation loss calculated by RET equations is modelled using transmission line model. This model is simple and allows quick calculation of the propagation loss compared to RET equations. In this section, a simple transmission line model for propagation through short distance vegetation (<100m) using the calculated loss from RET equations is presented. For this purpose, fewer parameters are presented and tabulated.

In RET geometry, the transmit antenna is located in the air half-space far away from the planar interface so the illumination is assumed to be a plane wave incidence, while the receive antenna is a highly directive antenna having a narrow beamwidth and is placed in the vegetation region. Propagation losses predicted by RET show that the vegetation can be modeled as a three-layer transmission line model illuminated by a plane wave shown in Fig.4.6. It is presumed that the vegetation medium comprises two different characteristics regions, i.e., $\gamma_i$ and $\eta_i$, with various thicknesses $d_i$ for region $i$, where $i \in 1, 2$, separated at $d = d_1$. The propagation constant is

$$\gamma_i = j\omega\sqrt{\mu\epsilon}, \quad (4.25)$$

where $\mu$ and $\epsilon$ are respectively the permeability and the permittivity of the dielectric, and $\omega$ is angular frequency. The characteristic impedance of the region is given by $\eta_i = j\mu\omega/\gamma_i$.

The relation between the transmitted field and the transmission factor for the propagation in each region is given by

$$E_{TL_i} \approx e^{-\gamma_i d} \quad \text{for region I}(d < d_1),$$

$$E_{TL_2} \approx e^{-\gamma_2 d} \quad \text{for region II}(d > d_1). \quad (4.26)$$

53
Therefore, the normalized transmitted power for region \( i \in 1, 2 \) is

\[
P_{TL_i} = \frac{1}{2\eta_i} |E_{TL_i}|^2 \approx e^{-2\gamma_i d}.
\]  
(4.27)

Figure 4.6: Transmission line model for vegetation illuminated by a plane wave.

Figure 4.7a and 4.7b illustrates the normalized received power predicted by RET equations and the transmission line model derived from the RET curves for the "Common Lime" tree in leaf and out of leaf at 1.3GHz. The RET curve is plotted using the parameters given in [1] and shown in Table 4.1. Then, the transmission line parameters, i.e., \( \gamma_i \) for region \( i \in 1, 2 \), and the breakpoint between two regions are derived from RET curves, and tabulated in Table 4.2.

Figure 4.7: The normalized received power given by RET and Transmission line model for the "Common Lime" tree (a) in leaf and (b) out of leaf at 1.3GHz.

<table>
<thead>
<tr>
<th>Leaf condition</th>
<th>( 2\gamma_1 )</th>
<th>( 2\gamma_2 )</th>
<th>Break Point (( d(m) ), power (dB))</th>
</tr>
</thead>
<tbody>
<tr>
<td>In leaf</td>
<td>0.22</td>
<td>0.024</td>
<td>(47, -44.91)</td>
</tr>
<tr>
<td>Out of leaf</td>
<td>0.59</td>
<td>0.039</td>
<td>(18, -46.12)</td>
</tr>
</tbody>
</table>

Table 4.2: Transmission line parameters for "Common Lime" tree at 1.3GHz.
Transmission through such inhomogeneous mixed media is complicated by the many different propagation mechanisms and the complexity of the randomness. Through-vegetation propagation models, including the standard RET, are not very accurate in the sense that the uncertainty can be tens of decibels, and this seems to be an accepted limitation for vegetation. A simpler propagation model, which maintains or improves accuracy, but keeps a reasonable association with the physics, would be insightful, especially for long distance propagation (> 2000m). Next section discusses such a model. It comprises two parallel transmission mechanisms: direct transmission through a succession of trees, which is modeled by a simple linear transmission line; and transmission across the forest top, which is modeled by simplified multiple-edge diffraction. The model is examined using recently-published experiments over a long path-length. It is demonstrated that this two-mechanism model can provide an accurate fit to the dual-slope profile of through-forest propagation over a long distance which is not possible with the RET model.

4.5 Simplifying Long Distance Through-Foliage Propagation Modelling

The Friis transmission equation is recalled (here with the efficiency factor) as being able to define the path gain [2],

\[
P_{Rx} \cdot G_{Rx} \cdot G_{path} \cdot \eta
\]

where \( P_{Rx} \) and \( P_{Tx} \) are the usual received and transmitted powers, respectively, \( G_{Tx} \) and \( G_{Rx} \) are the transmit and receive antenna effective gains which must be well-defined, and the various efficiency factors (in particular the polarization efficiency) are collected in \( \eta \). In multipath, these factors are recalled as being taken as statistically independent, and they combine with (4.28) to define \( G_{path} \).

For free space media, the path gain corresponds to the spherical spreading. For the Friis approach, the path gain has an inverse frequency dependence stemming from considering an electrically-dimensional (in square wavelengths) receiving aperture, which is spaced by distance \( d \) meters between the phase centers of the antennas, \( G_{path} = G_{(FS)}_{path} = (4\pi d/\lambda)^{-2} \). (This frequency dependence is normally omitted in acoustics and optics formulations).

In multipath media which couples the polarizations randomly, the (mean) polarization efficiency becomes a half, and relatedly, for any single port antenna in full-sphere, homogeneous, uncorrelated multipath, the (mean) antenna gain reduces to half of its radiation efficiency [2]. In this way, for propagation through vegetation, the antenna gain depends on the illuminating wave scenario, and this alone can introduce several tens of decibels of variation (as a constant offset value) in the path gain estimate. This variation is from choosing an antenna gain to be the radiation efficiency over two at one extreme (the dense multipath likely in a dense forest), and the usual maximum co-polarized directional gain at
the other extreme (free space), or something in between depending on the multipath model and the pattern of the antenna.

With definitions for the antenna gains in place (and associated offsets in the path gain estimate), the measured received power gives the path gain through vegetation. It can be extremely variable for a given transmit-receive distance. A single narrow-band measurement is subject to the Rayleigh-like fading, with a variation at least 30dB. Averaging over the Rayleigh-like fading gives an estimate of the local mean path gain, although good accuracy of the mean estimate requires a large number of local spatial measurements (or spaced apart in frequency using a wideband measurement) [2].

Different experimental approaches treat these offset-producing factors differently (if at all), so the slopes of the path gain with distance become the more interesting feature for analysis and modeling. The local variations in the mean estimate are then considered to be the numerous propagation media factors: foliage densities and sizes, species, seasons, and of course, different propagation frequencies [81].

It has been observed experimentally [5,78] that a signal is strongly attenuated at short distances due to absorption and scattering by leaves and branches, while at larger distances (in this context meaning about 100m), in-line scattering becomes a dominant mechanism of propagation resulting in a lower attenuation rate. This foliage path loss is also been referred to as a dual-slope propagation loss, e.g. [77] on a log-linear scale. These experimental observations suggest that a simpler model than the RET approach, comprising just two propagation mechanisms, should be feasible. As explained above, separate transmission line models was discussed in the context of dual slope path gain using the variables as the pair of slopes and the breakpoint separating them, for a given frequency.

The long distance (up to 2580m) experiments also show a dual slope in log-log scale [9]. The slope of the short-distance transmission can be somewhat less than free space, and [9] fits this with a two-ray model in free space. The RET model (not considered in [9]) can also fit to such behaviour. Here, a compromise by using a transmission line model is taken for the short-distance. This is much simpler than the RET model, and may be closer to the physics of the dense scattering of a forest than the pair of free-space rays used in [9]. In the (linear) transmission line model [7,82], the wave is guided linearly, and so the attenuation mechanism is, instead of spherical spreading and rescattering, only from passing through the lossy sections of the line. These transmission line sections can be spaced akin to an effective tree spacing. The lossy sections of the linear transmission line are here very low loss (all transmission line sections have a free space real permittivity, and the tree sections have a small loss added, see below), and so the reflections in the transmission line are negligible. In this way, the extra complexity of cylindrical or spherical transmission lines is not needed. A limitation of this model is that physically reflected power may not be well modelled. It is reasonable to attribute the long-distance propagation to diffraction over the top of the trees. For this, multiple knife-edge diffraction is used [8,63,64]. The output (receive power)
is a weighted sum of these two mechanisms (see Fig. 4.8). The weighting coefficients, $W_1 < 1$ and $W_2 = 1 - W_1$, must be empirical. This model has a minimum of parameters, and retains a reasonable attachment to the physics.

The two propagation mechanisms are treated as independent, so that the diffraction path is not continually fed by the scattering through the trees. This is unlikely to be the case in practice, which is a limitation of the model, but the independence keeps the model as simple as possible. Time domain measurement (requiring wide bandwidth, which means that the required large dynamic range of the measurement system must have some narrow bandwidth-resolving capability such as OFDM) may resolve this issue, but no long-range wideband measurements are publicly available.

Although the two propagation mechanisms are considered independent (e.g. the transmission line lossy sections do not continually feed the diffraction links), there is a form of dependence between the models of the two propagation mechanisms - the number of transmission line sections is taken as the same as the number of diffraction edges. While this is not strictly necessary for the dual-mechanism model to work, each transmission line section can be viewed as relating a tree with a transmission loss to the same tree having an edge diffraction. So the more transmission line sections, the more diffraction edges. This tends to change the offsets of the various contributions (as do the many other factors, discussed above) but it is the slopes that are of particular interest.

![Figure 4.8: Through-forest propagation model. The transmission line, in parallel and uncoupled with the diffraction path.](image)

### 4.5.1 Multi-Layer Transmission Line

The transmit antenna is typically placed away from the trees so the illumination is essentially a plane wave, as illustrated in Fig. 4.9a. Intuitively, a tree can be modeled as a two-dimensional block. For simplicity, the trees are modeled as regularly spaced and with constant thickness, see Fig. 4.9b. Randomness in the spacing and thickness provides only a second-order effect. The transmission line can be configured in many ways by interchanging between the spacing, thickness and loss of each section (tree), but a choice of fixing both the spacing and the thickness allows fewer parameters and of course simplicity.
Figure 4.9: (a) A depiction of trees between a transmitter and receiver antenna. (b) Two-dimensional multi-layer configuration of in-line trees. The constant thickness of the trees and the free space distance between them are donated by \( d \) and \( d_0 \), respectively. The diffraction and vegetation penetration mechanisms are considered as independent, a simplifying limitation, but their models are connected, again for simplicity, by the number of lossy transmission line sections (trees) being the same as the number of diffraction edges (tree tops).

The transmission line relations \([7, 82]\) are now briefly summarized for the vegetation modeling. The vegetation has a complex (i.e., lossy) propagation constant for each tree section

\[
\gamma_d = j\omega\sqrt{\mu\epsilon},
\]

where \( \mu = \mu_0\mu_d \) and \( \epsilon = \epsilon_0\epsilon_d \) are respectively the permeability and the permittivity of the dielectric (tree). The relative permeability is \( \mu_d = 1 \) and the complex relative permittivity is \( \epsilon_d = \epsilon'_d - j\epsilon''_d \). \( \mu_0 \) and \( \epsilon_0 \) are for free space, and \( \omega \) is angular frequency. The characteristic impedance of the dielectric is \( \eta_d = j\mu_0\omega/\gamma_d \). The constant thickness of the trees and the free space distance between them are denoted by \( d \) and \( d_0 \), respectively. The transmission line model for such in-line trees is illustrated in Fig. 4.10. For the nature and goal of this transmission line model, no attempt was made to differentiate detail such as trunks and leaves, or foliage density for a tree, whereas some curve fitting approaches rely on the trunk size, for example. The transmission coefficient is

\[
T = \left( \frac{2\eta_d}{\eta_d + \eta_0} \frac{2\eta_0}{\eta_d + \eta_0} \ldots \frac{2\eta_0}{\eta_d + \eta_0} \right) \times D^{-1} e^{-\gamma_d d - \gamma_0 d_0 - \ldots - \gamma_d d},
\]

(4.30)
where

\[
D = \left[ 1 - \frac{(\eta_0 - \eta_d)}{\eta_0 + \eta_d} \right] \left[ \left( \frac{Z_{in2} - \eta_d}{Z_{in2} + \eta_d} \right) e^{-2\gamma_d d} \right] \\
\times \left[ 1 - \frac{(\eta_d - \eta_0)}{\eta_d + \eta_0} \right] \left[ \left( \frac{Z_{in3} - \eta_0}{Z_{in3} + \eta_0} \right) e^{-2\gamma_0 d_0} \right] \ldots \tag{4.31}
\]

and the input impedance of the \( m \)th layer is, using the lengths of the free-space and tree thickness (see Fig.4.10),

\[
Z_{inm} = \frac{\eta_d Z_{in_{m+1}} + \eta_m \tanh \gamma_m d_m}{\eta_m + Z_{in_{m+1}} \tanh \gamma_m d_m}, \tag{4.32}
\]

The incident power relates to the incident electric field as

\[
P_{inc} = \frac{|E_0|^2}{2\eta_0}, \tag{4.33}
\]

with \( \eta_0 = \sqrt{\mu_0/\epsilon_0} = 120\pi \ \Omega \). The electric field can be calculated from the experimental configuration. Finally, the power transmitted through the \( M \) layers is

\[
P_{MLT_{trans}} = \frac{|E_0|^2 |T|^2}{2\eta_0}. \tag{4.34}
\]

Figure 4.10: Equivalent transmission line circuit for the two-dimensional multi-layered configuration of in-line trees (after [7]).

4.5.2 Multiple Knife-Edge Diffraction

A generalized formulation for propagation over various configurations of inhomogeneous terrain is given in [83]. This includes a two-dimensional propagation over multiple rounded obstacles in the form of a residue series, and the knife-edge case when the radius of the curvature decreases to zero. This series was transformed into a multiple integral by Vogler [8,64] for perfectly absorbing edges. This model is used as a reference for comparison of other
techniques such as the UTD solution for multiple-edge transition zone diffraction \[2, 18, 19\]. The multiple-edge diffraction integral is a very useful tool and the key aspects are briefly summarized here.

Figure 4.11: Geometry for multiple knife-edge diffraction (after [8]).

The geometry associated with the Vogler multiple knife-edge diffraction is depicted in Fig. 4.11. The height of the edges above some reference plane are \( h_i \), \( i = 1, 2, ..., N \), and the height of the transmitter and receiver are \( h_0 \) and \( h_{N+1} \). The separation distances between knife-edges are \( A_i \), \( i = 1, 2, ..., N + 1 \). The diffraction angles are \( \theta_i \), \( i = 1, 2, ..., N \), obtained from the edge heights and the separation distances. Vogler expresses the excess diffraction loss, which is the attenuation of field strength relative to free space \[8, 64\],

\[
A_N = (1/2^N)C_N e^{\sigma_N} (2/\pi^{1/2})^N I_N, \tag{4.35}
\]

where

\[
\sigma_N = \beta_1^2 + ... + \beta_N^2, \tag{4.36}
\]

(note that \( \sigma, \alpha \) and \( \beta \) are used differently in this subsection, following convention)

\[
C_N = \begin{cases} 
1, & N = 1 \\
\left[ \frac{r_2 r_3 ... r_N r_T}{(r_1 + r_2)(r_2 + r_3)...(r_N + r_{N+1})} \right]^{1/2} & N \geq 2
\end{cases} \tag{4.37}
\]

and

\[
r_T = r_1 + r_2 + ... + r_{N+1}. \tag{4.38}
\]

\[
I_N = \int_{\beta_1}^{\infty} ... \int_{\beta_N}^{\infty} e^{2f - (x_1^2 + ... + x_N^2)} \, dx_1 ... dx_N \tag{4.39}
\]

with

\[
f = \begin{cases} 
0, & N = 1 \\
\sum_{m=1}^{N-1} a_m (x_m - \beta_m)(x_{m+1} - \beta_{m+1}), & N \geq 2
\end{cases} \tag{4.40}
\]
\[ \alpha_m = \left[ \frac{r_mr_{m+2}}{(r_m + r_{m+1})(r_{m+1} + r_{m+2})} \right]^{1/2}, \]

\[ m = 1, \ldots, N - 1 \]

\[ \beta_m = \theta_m \left[ \frac{jk(r_mr_{m+2})}{2(r_m + r_{m+1})} \right]^{1/2}, \]

\[ m = 1, \ldots, N \]

To reduce the computation time, the multiple-edge integral in (4.39) is transformed into the series of terms involving the repeated integral of the error function, which is reported in [8] for up to 10 knife-edges, after which the general accuracy falls away. A special case (along the equi-spaced and same-height tree tops) can be used, which has a simpler, accurate solution. Experiments with statistical perturbations from this arrangement, using the UTD, have been investigated in [84].

### 4.5.3 Results and discussion

#### Model Comparison

In this section, the model is examined against results from an extraordinary (for their distance, accuracy, and dynamic range from using an extremely narrowband) set of through-forest measurements conducted in a typical forest terrain in Denmark at 917.5MHz, recently published [9]. The trees are predominantly fir (pine), oak, and beech. The foliage density changes significantly between summer and winter because of the deciduous trees. The measurements were taken in summer. The forest is reported to be a mix of coniferous and deciduous trees in [9]. The transmitter and receiver have the same height of 1.5m with the transmit power 40dBm, and there are 71 distances along the path. The furthest measurement position has a straight line distance to the transmitter of 2580m. The forest map and the measurement locations [9] are shown in Fig. 4.12, and the results of the measurement are included in the figures further below. In these measurements, the short-term, Rayleigh-like fading which is from the vegetation scattering, is averaged at each measurement point, using a circular locus of about 3.3 wavelengths. (In dense multipath, this finite measurement aperture gives a one-standard deviation spread of just over a decibel, [2]). In some recent papers, the Rayleigh-like fading is not removed from the propagation measurement at each range point, and it tends to dominate the path gain variations because of its distribution spanning several tens of decibels for a narrow-band signal.
Figure 4.12: Measurement locations conducted in a typical forest terrain in Denmark [9]. The transmitter position is marked as Tx at the left top in the figure.

**Propagation Model**

The transmission line, in parallel and uncoupled with the diffraction path, is depicted in Fig.4.8. By considering trees as absorbing baffles with equal heights, and spacing $r$, (see Fig. 4.13), the multiple knife-edge diffraction attenuation, $A_N$, becomes frequency independent and simplifies to an exact solution given in [63]; i.e.,

$$A_N = \frac{1}{N + 1},$$  \hspace{1cm} (4.43)

where $N = d_T / r$ is the number of baffles between the transmitter and receiver. So for $r = 1$ m, for example, and $d_T = 2580$ m, $N = 2850$ is obtained (see caption of Fig.4.15 for other choices of $r$), and the offset (Fig.4.15a) is $-68$ dB. The diffracted power is calculated from

$$P_{DN} = G_{path}^{(FS)} \cdot (A_N)^2.$$  \hspace{1cm} (4.44)

In summary, the received power through the transmission line is

$$P_{MLT_{trans}} = \frac{|E_0|^2 |T|^2}{2\pi_0},$$

using a frequency of 917.5 MHz and an $M$-layer transmission line ($N$ trees = $M/2$) with spacing $r$ and the tree thickness $d = r/4$. These choices are almost arbitrary because the thickness, spacing and the value of the lossy permittivity can be interchanged for the same effect. The real part of the relative permittivity is taken as one (i.e. same as free space), because only the loss behaviour is of interest, rather than the scattering detail. The total received power is just

$$P_{R_{total}} = W_1 P_{DN} + W_2 P_{MLT_{trans}}.$$  \hspace{1cm} (4.45)

where $W_2 = 1 - W_1$.  

62
Figure 4.13: (a) $N$ absorbing knife-edge diffraction from the crown of the trees with the same height and spacing, $r$, (b) two-dimensional $M$-layer configuration of in-line trees with $d$ thickness, $\varepsilon_r$ dielectric constant and $r$ spacing.

**Empirical parameters**

The propagation model includes three empirical parameters; i.e., $r$, $\varepsilon''$ and $W_1$, found by fitting the experimental data. This section illustrates, using figures, the effect of these parameters on the dual-slope behaviour, with a simple numerical study. The figures below for the path gain have a large dynamic range in order to emphasize the macro-behaviour. The detail, within the dynamic range, is covered by evaluating the uncertainty in the form of a dB root mean square error (RMSE).

Figure 4.14: Comparison between the measured path gain in [9] and the propagation model for $r = 1.5$ m ($N = 1720$), $W_2 = 1 - W_1 = -70$ dB: (a) $\varepsilon'' = 0.006$ and (b) $\varepsilon'' = 0.01$. The distance scale is logarithmic. The free space path gain (FSPG), $W_1 P_{DN}$, and $W_2 P_{MULT_{trans}}$ are also plotted to show the dominant parameters.
First, the effect of $\varepsilon''$, is plotted in Fig. 4.14 for $\varepsilon'' = 0.006$ and 0.01. The other parameters are fixed to $r = 1.5m$ ($N = 1720$) and $W_2 = 1 - W_1 = -70dB$. The short-distance slope of the model, from 15 to around 200m, as well as the transition between two slopes from 200 to 500m change, are shown in the figure. In particular, it indicates how, in this model, changing $\varepsilon''$, while the trees’ thickness-to-spacing ratio is fixed, or vice versa, affects the short-distance slope.

![Figure 4.14](image1)

Figure 4.15: Comparison between the measured path gain in [9] and the propagation model for $\varepsilon'' = 0.008$, $W_2 = 1 - W_1 = -70dB$: (a) $r = 1m$ ($N = 2580$), (b) $r = 1.5m$ ($N = 1720$), (c) $r = 2m$ ($N = 1290$) and (d) $r = 3m$ ($N = 860$). Note that for different $r$, $N$ is different, and the distance scale is logarithmic. The free space path gain (FSPG), $W_1P_{D_{N}}$, and $W_2P_{MLT_{trans}}$ are also plotted to show the dominant parameters.

![Figure 4.15](image2)

The effect of the spacing $r$ is shown in Fig. 4.15 for $r = 1, 1.5, 2$ and 3m. The other parameters are fixed to $\varepsilon'' = 0.008$ and $W_1 = 0.9999999$ (so the transmission line contribution offset is $W_2 = 1 - W_1 = -70dB$). The results indicate that the long-distance slope of the model which follows the diffracted power from around 500m to 2580m depends on the spacing $r$.  

64
Alternatively $r$ can be fixed and $\epsilon''$ can be changed, etc, see below.) For $r = 1$m, the model fits the measurement nicely, giving the $N$.

Figure 4.16: Comparison between the measured path gain in [9] and the propagation model for $\epsilon'' = 0.008, r = 1.5$m ($N = 1720$): (a) $W_2 = 1 - W_1 = -60$dB, (b) $W_2 = 1 - W_1 = -30$dB, (c) $W_2 = 1 - W_1 = -1$dB, (d) $W_2 = 1 - W_1 = -\infty$dB. The distance scale is logarithmic. The free space path gain (FSPG), $W_1P_{DN}$, and $W_2P_{MULT TRANS}$ are also plotted to show the dominant parameters.

Finally, Fig. 4.16 illustrates the weighting $W_1$ changing from 0.9 to 1 (i.e., $W_2$ changes from $-1$dB to $-\infty$dB) with $r = 1.5$m and $\epsilon'' = 0.008$. It shows how the diffraction is the more dominant mechanism for the long distance. But if only the diffraction mechanism is considered (i.e., when $W_1 = 1$), then the measured data from 15 to 500 meters cannot be fitted. By choosing a best value for $W_1$, the first slope from 15 to around 200 meters, as well as the transition between the two slopes between 200 and 500 meters, can also be fitted to the experimental data.
Table 4.3: RMS Error of the propagation model with various empirical parameters.

<table>
<thead>
<tr>
<th>Fig.</th>
<th>( r ) (m)</th>
<th>( N = M / 2 )</th>
<th>( \epsilon' )</th>
<th>( W_2 = 1 - W_1 ) (dB)</th>
<th>RMSE (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.14a</td>
<td>1.5</td>
<td>1720</td>
<td>0.006</td>
<td>−70</td>
<td>6.9</td>
</tr>
<tr>
<td>4.14b</td>
<td>1.5</td>
<td>1720</td>
<td>0.01</td>
<td>−70</td>
<td>6.8</td>
</tr>
<tr>
<td>4.15a</td>
<td>1</td>
<td>2580</td>
<td>0.008</td>
<td>−70</td>
<td>4.6</td>
</tr>
<tr>
<td>4.15b</td>
<td>1.5</td>
<td>1720</td>
<td>0.008</td>
<td>−70</td>
<td>4.9</td>
</tr>
<tr>
<td>4.15c</td>
<td>2</td>
<td>1290</td>
<td>0.008</td>
<td>−70</td>
<td>5.8</td>
</tr>
<tr>
<td>4.15d</td>
<td>3</td>
<td>860</td>
<td>0.008</td>
<td>−70</td>
<td>7.5</td>
</tr>
<tr>
<td>4.16a</td>
<td>1.5</td>
<td>1720</td>
<td>0.008</td>
<td>−60</td>
<td>8.3</td>
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<td>4.16b</td>
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<td>29.8</td>
</tr>
<tr>
<td>4.16c</td>
<td>1.5</td>
<td>1720</td>
<td>0.008</td>
<td>−1</td>
<td>45.6</td>
</tr>
<tr>
<td>4.16d</td>
<td>1.5</td>
<td>1720</td>
<td>0.008</td>
<td>−∞</td>
<td>20.3</td>
</tr>
</tbody>
</table>

4.5.4 Model Performance

Here, an overview of the performance of the model is given by computing RMSE in dB which is calculated between the measured and predicted path gain (reciprocal of the path loss, or rather negative in dB, as used in the equation below [85]).

\[
RMSE = \sqrt{\frac{\sum_{i=1}^{n} \left( G_{path_{\text{meas}}}^{(dB)} - G_{path_{\text{pred}}}^{(dB)} \right)^2}{n}} \text{ [dB]},
\]

(4.46)

where \( n \) is the total number of measurements. The RMSE using the various empirical parameters of \( r \), \( \epsilon'' \) and \( W_1 \) are tabulated in Table 4.3.

The RMSE of the best fit to the experiments of [9] is 4.6 which is shown in Fig. 4.15a for \( \epsilon'' = 0.008 \), \( r = 1 \)m and \( W_2 = 1 - W_1 = -70 \)dB. This model is compared here using the RMSE of the following models. The results are summarized in Table 4.4.

The first model for comparison is a three-parameter (for a fixed frequency) empirical model proposed by Tewari et. al in [86], where the path gain in dB is given by

\[
G_{\text{path}} = -PL_{\text{Tewari}} = 27.57 - 20 \log_{10}(f) + 20 \log_{10}\left( \frac{Ae^{-\alpha_2\text{dist}}}{\text{dist}} + \frac{B}{\text{dist}^2} \right) \text{ [dB]},
\]

(4.47)

where \( f \) is the frequency in MHz, \( \text{dist} \) is the propagation distance in meters. \( A \) and \( B \) are constants found empirically, and \( \alpha_2 \) is the constant describing the rate of the attenuation in dB/m [86]. This model is claimed to be been verified [86] for discrete frequencies between 50 to 800MHz. For comparison with the measured data at 917.5MHz, the empirical values for 800MHz have been used [9]. The RMSE of the Tewari model with the measured data for the total distance, as well as for the diffraction-dominated distances from 200m to 2580m, are given in Table 4.4.
Table 4.4: RMS Error of propagation model for vegetation.

<table>
<thead>
<tr>
<th>Model</th>
<th>RMSE (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>The propagation model in Fig. 4.15a</td>
<td>4.6</td>
</tr>
<tr>
<td>Tewari∗ [86] for all measured data</td>
<td>10.2 [9]</td>
</tr>
<tr>
<td>Tewari∗ [86] from 200m</td>
<td>6.7 [9]</td>
</tr>
<tr>
<td>Two–Ray+2×ITU-R P.2108-0 [9] from 200m (Same height Tx and Rx)</td>
<td>7.6</td>
</tr>
<tr>
<td>ITU-R P.1546+2×ITU-R P.2108-0 [9] from 200m (Same height Tx and Rx)</td>
<td>11.2</td>
</tr>
<tr>
<td>RET model</td>
<td>13</td>
</tr>
<tr>
<td>Polynomial regression (n=1)</td>
<td>18.5</td>
</tr>
<tr>
<td>Polynomial regression (n=2)</td>
<td>7.9</td>
</tr>
<tr>
<td>Polynomial regression (n=3)</td>
<td>4.5</td>
</tr>
<tr>
<td>Polynomial regression (n=4)</td>
<td>4.33</td>
</tr>
<tr>
<td>Polynomial regression (n=5)</td>
<td>4.1</td>
</tr>
<tr>
<td>Polynomial regression (n=6)</td>
<td>3.5</td>
</tr>
</tbody>
</table>

The RMSE of the two models presented in [9] for distance from 200m to 2580m, are also included in Table 4.4. Finally, the measured data is modeled using linear polynomial regression function, i.e. purely as a curve fit, with no reference to propagation mechanisms, given by

\[ y = \beta_0 + \beta_1 x + \beta_2 x^2 + \beta_3 x^3 + \cdots + \beta_n x^n \] (4.48)

for \( n \in \{1, 2, 3, 4, 5, 6\} \). (See Table 4.4).

This two-mechanism model, with three empirical parameters for a fixed frequency, has a smaller uncertainty (RMSE = 4.6) compared to the other models and its uncertainty is close to the 3rd degree polynomial function. Compared to the various models treated in [9], the model gives a better fit for the distance from 15m to the furthest point at over 2580m, whereas the model in [9] is from 200m. The four-parameter RET model (proposed for only short distances), for a fixed frequency, has RMSE = 13dB. An explicit frequency dependence for the model has not been included, however the free space component in the diffraction contribution, and the transmission line equations, contain the frequency dependence. Without similar, long-distance measurements being available at other frequencies, a frequency dependence for this model is not verified. Some measurements well spread around 1GHz for long distances is given in [87], but these are over undulating terrain using a vehicular antenna (the measurements are used here, are truly through-forest with a personal antenna system) where other mechanisms, such as terrain-shadowing, are at play.

### 4.6 Summary and Conclusion

Propagation through a forest features dual-slope behaviour in the log-log path gain. For NLOS short distances (less than a couple of hundred meters), there is a free-space-like
slope with a large offset of about −60 dB; and for long distances (more than about five hundred meters), there is a steeper slope with another offset. This dual-slope behaviour suggests two principal propagation mechanisms. For the short-distance, it is intuitive that there is direct transmission through the foliage. This can be modeled by the RET, but the simplest model for this is a linear, low-attenuation transmission line where a succession of lossy sections are separated by lossless sections. There is an interchangeable choice of parameters for configuring such a transmission line model, but the loss is only from the lossy sections instead of also from spherical or cylindrical spreading. This transmission line behaviour is shown here to be able to match the experimental short-distance propagation. It has the advantage over the standard short-distance RET model of being simpler and having fewer fitting parameters. This transmission line (or the RET model) cannot also provide a match for the long-distance propagation, nor the transition region between the dual slopes. This is because even a linear transmission line gain falls away very quickly with longer distance, which is contrary to the experimental facts. The simplest and most reasonable physical mechanism is a parallel transmission path over the tree tops. A model for this is multiple knife-edge diffraction, where the trees are the knife-edges. This model can be further reduced by modeling the tree tops as being constant height and spacing. This model is therefore extremely simple, has minimal parameters, and yet retains a reasonable attachment to the physics. Such a diffraction wave can match the long-distance slope of through-forest propagation. A weighted combination (i.e., parallel paths) of the two modeled mechanisms accurately matches (RMSE as low as 5 dB) the whole range of the experimental measurement - the dual slopes, their offsets, and the transition region between the short- and long-distances.
Chapter 5

Antennas on Tubular Platforms

5.1 Introduction and Background

Wideband, highly efficient and compact antennas with low-cost fabrication are in demand in wireless and mobile communication systems. Some applications call for antennas on, or within, an electrically small curved platform such as unmanned aerial vehicles (UAVs), drones, bicycle frames, and cylindrical radio mast or strut.

An electrically small antenna typically fits within a radian-sphere, i.e., with radius \( b = 1/k \) where \( k = 2\pi/\lambda \) is the wavenumber [88, 89]. For cylindrical structures, the cross section (i.e., not the cylinder length) follows a similar guideline. An electrically small cross section means that, for antenna design purposes, the surface cannot be considered as having the convenient properties of a planar groundplane. Specifically, a threshold size for a cylindrical groundplane to appear planar, for monopole impedance, is a diameter of about two wavelengths, i.e., \( kb \sim 6 \) [90]. Design solutions for non-planar configurations are found most conveniently from simulation because accurate analytical and transmission line models are challenging to formulate. These use radio services including wireless communications, navigation, and sensing such as radar. It is often of interest to use the same antenna for such multiple functions. This compounds the wide bandwidth requirement which is already a challenge because of the small antenna system. While the bandwidth is the priority, for most mobile communications terminals, the radiation pattern shape can be of lower priority because of the diversity of directional energy transfer via multipath.

Previous works on low directivity antennas on a curved surface include the monopole work of Henderson [90], the loop [91] and the PIFA antennas [92]. For applications on metallic surfaces, surface slot elements have the advantage of being flush, as opposed to externally mounted elements, and can use the cylinder as the feed line for the elements.

Slotted waveguide antennas with the TE mode (the usual hollow waveguide used for slot arrays) are not suitable for compact antenna design. This is because their bandwidth and compactness (cross section) are both limited by the mode cutoff rate [23, 93, 94] with the size of a waveguide cross section \( kb > 1 \). Therefore, slotted coaxial waveguide (in TEM mode) is
a good candidate for a highly practical radiating structure [95]. By using the TEM mode of coaxial waveguide, a more compact structure is possible which supports a wider bandwidth, compared to hollow waveguide. The cost is a higher ohmic loss than for hollow waveguide, but the extra metallic conductor loss is not much for microwave frequencies. Slotted coaxial waveguide antennas have already been reported for various applications in cellular, mobile, radio and satellite communication, on the basis of their high efficiency (relative to other design options for the application), low cost and easy fabrication [96,97]. An infinitely-long coaxial antenna concept with a single slot was investigated in [98,99], and a coaxial antenna with circumferential slots has been proposed for mobile terminals [14]. A traditional use of slot radiators on coax is "leaky feeders", often deployed along tunnels [100,101], but also reported for position detection [102,103], and for 2G and 3G wireless communications [104] in situations where radiation from a conventional discrete antenna is not feasible. Leaky feeder structures are not suitable for electrically compact radiators, they are instead for a large (specifically, long) aperture. A recent text [105] categorizes slotted coaxial waveguides into the radiating mode cable (RMC) and the coupled mode cable (CMC). In a CMC, the spacing of the apertures (slots) is smaller than the operating wavelength and the cable can work within a wide frequency range. For RMCs, the slots are typically configured with spacings comparable with the operational wavelength [104]. Some work has been undertaken for coaxial waveguide slot arrays. For example, Iigusa [106] proposed a slot array antenna on a coaxial cylinder with circular polarization (CP). The maximum gain is 6.3dB and the radiation efficiency 73%, but because the pattern (directive CP) was the design priority, the bandwidth is very narrow. A single- and a dual-CP omnidirectional antenna using a slot array on coaxial cylinder with dielectric insert ($\varepsilon_r = 2.1$) was recently proposed [107–109]. This has 32 small slots (length $0.11\lambda_g$) and a $-10$dB impedance bandwidth of $\sim 16.4\%$. These cylindrical antennas have a large electrical cross section, with $kb \sim 2.5$.

### 5.2 Contribution and Organization of the Chapter

The contribution of this chapter is an antenna design on a small cylinder such as a bicycle frame. The numerical and physical measurements show that carbon fiber which is widely used as a bicycle frame works like metal, so an internal antenna, radiating through the frame material, is not feasible. This prompts the use of antennas located on the outside surface of the frame. These are discussed in section 5.3, and include a bent monopole, a small Conformal PIFA (CPIFA) and new wrapped conformal PIFA (WCPIFA). The WCPIFA has $-10$dB impedance bandwidth of $\sim 30\%$ and a simple structure.

External antennas can be broken easily. But if the whole frame radiates, a more robust system is obtained. Section 5.4 presents bent waveguide-excited slot antennas (in TE mode). The idea is to have a radio mounted internally and radiating from multiple slots distributed along the cylinder. However, the hollow slotted waveguide is not suitable for a compact
(cross section) cylinder platform. This is because their bandwidth and compactness (cross section) are both limited by the mode cutoff rate with the size of a waveguide cross section. A solution is to use a coaxial type TEM mode.

New designs of slotted coaxial antennas which are compact and wideband are presented in section 5.5, comprising compact (cross section) slotted coaxial waveguide antennas with both short length of a cylinder, \( L < \lambda \), and the long length of cylinder (for bicycle frames), \( L = 2\lambda \). Section 5.5.4 presents slotted coaxial waveguide antennas with the cross section of \( kb = 0.75 \) and the prototypes are for a frequency around 1.8GHz. Two different length of antennas, i.e., \( L = 0.25\lambda \) and \( 0.625\lambda \) are studied in section. The bandwidth performance of the design with two slot on different sides of the waveguide is remarkable. The −10dB bandwidth of \( \sim 50\% \) is obtained for the cylinder length of \( L = 0.625\lambda \), while the cylinder length of \( L = 0.25\lambda \) gives −10dB bandwidth of \( \sim 40\% \). In comparison, a wide bandwidth open-slot antenna with slot length \( 0.8\lambda \) (uses the second resonance for high bandwidth) in a ground plane of \( 1.6\lambda \times 1.6\lambda \), has −10dB bandwidth of 25\% \([110]\). Section 5.5.5 features longer structures for the bicycle application, and the fabricated example has 6 slots with \( kb = 0.65 \) for a frequency of 1.575GHz. A longer antenna can have pattern directionality advantage, but the bandwidth tends to decrease because of the long-line effect \([111]\), see below). The −10dB impedance bandwidth is \( \sim 17\% \). The summary and conclusion is given in section 5.6.

5.3 Antennas on an Electrically Narrow Cylindrical Ground-plane

5.3.1 Monopole on a Small Cylinder

A benchmark candidate antenna mounting on a small cylinder is an outward-pointing quarter wavelength monopole. This antenna is one of the simplest possible structures and the most effective small antenna. The impedance of a quarter wavelength monopole on an electrically small cylinder in CST simulation is solved, shown in Fig.5.1. A thin \( \lambda/4 \) monopole of diameter 0.0004\( \lambda \) is mounted on a cylinder with the diameter and length of 0.1\( \lambda \) and 0.26\( \lambda \), respectively. The cylinder is PEC with the thickness of 0.005\( \lambda \). Simulation results indicate a −10dB impedance bandwidth of \( \sim 12\% \) for a nominal frequency of 1.575GHz. Simulation results also show that the −10dB impedance bandwidth of the monopole on a larger diameter of the cylinder; i.e., \( 2\lambda \) is almost 15.5\%, while the same monopole mounted on a large planar groundplane has a bandwidth of \( \sim 13\% \).
The bandwidth calculation for a monopole on a small cylinder has not been treated before. Here, the bandwidth is also derived from theoretical and numerical methods. The theoretical bandwidth is related to the quality factor $Q$, which is traditionally defined as the ratio of the stored reactive energy to the dissipated energy in a circuit (or radiated and ohmic losses in an antenna) $[112,113]$. It is a good theoretical metric to evaluate electrically small antenna fundamental limits. At a single local resonance, $Q$ can be expressed in terms of the slope of the input reactance of a lossless circuit as $[112,113]$

$$Q = \frac{\omega}{2R_{in}} \frac{\partial X}{\partial \omega}. \quad (5.1)$$

Since the slope of the reactance can be negative in a lossy situation (in particular, radiation), it is called the signed $Q$. To have a positive $Q$, consistent with its definition, the unsigned $Q$ uses the absolute value of the derivative. The unsigned $Q$ is also approximated with the slope of the full impedance $[112,113]$, and although in an ideal model, the resistance slope is zero at resonance, the full derivative tends to give more stable results from measured or simulated data,

$$Q = \frac{\omega}{2R_{in}} |\frac{\partial Z}{\partial \omega}|. \quad (5.2)$$

$Q$ can be approximated from other expressions of the input impedance, such as the matched voltage-standing-wave-ratio (VSWR) bandwidth as $[112,113]$

$$Q = \frac{2\sqrt{\beta}}{FBW_v(\omega_0)}, \quad (5.3)$$

where $FBW_v$ is the fractional matched VSWR bandwidth at the resonate frequency $\omega_0$, determined for any arbitrary $VSWR = s$, i.e., $s = 2$ for $s_{11} = -10$dB, and

$$\sqrt{\beta} = \frac{s-1}{2\sqrt{s}} \leq 1. \quad (5.4)$$
These above expressions are useful in the sense that they allow a theoretical bandwidth to be expressed, without actually having to match the antenna port. Alternatively stated, the theoretical bandwidth assumes an ideal (lossless) matching circuit if it is to be translated into a matched-port impedance bandwidth. This separates the antenna’s bandwidth, as defined at the antenna port without an explicit matching network (good for theorists and simulations), from the practical impedance bandwidth (good for practitioners) available from using an ideal matching circuit. $Q$ has its definitions pointing to a strictly narrowband parameter for a strictly single resonant circuit, but has moved in practice to cover wideband interpretations.

The relative impedance bandwidth is also calculated from numerical results, viz.,

$$BW = 100\frac{f_H - f_L}{f_C} = 200\frac{f_H - f_L}{f_H + f_L},$$

where $f_H$, $f_L$ and $f_C$ denote higher cut-off frequency, lower cut-off frequency and center frequency, respectively. Figure 5.2 illustrates a \(-10\text{dB}\) fractional and relative bandwidth for the small diameter cylinder; i.e., \(0.1\lambda\). The results depicts that the cylinder length is a minor factor.

![Figure 5.2: Calculated \(-10\text{dB}\) numerical and theoretical bandwidths for a small monopole with the diameter of \(0.1\lambda\) for different cylinder lengths, $L$.](image)

A quarter-wavelength monopole is also built on a small, finite-length aluminum tube at frequency of \(\sim 1\text{GHz}\). The cylinder diameter and length is \(0.1\lambda\) and \(0.26\lambda\), respectively. The fabricated prototype is measured, and the results are compared with the simulations. The measurement setup is shown in Fig. 5.3. The results from the physical measurements and numerical simulation of the prototype are compared below. It can be observed from Fig. 5.4 that the measured impedance bandwidth is larger, probably due to the ohmic loss in

73
the external coaxial feed cable. Equation (5.6) shows that increasing the input impedance will decrease the quality factor, and increases the bandwidth. Some pattern information is depicted in Fig.5.5. The results indicate that the structure looks like a dipole antenna in free space. So the small, finite length PEC cylinder as a groundplane has different effects compared to a finite planar groundplane on a monopole [2,113,114].

\[ Q = \frac{\omega}{2R_{in}} \left| \frac{\partial Z}{\partial \omega} \right| = \frac{\omega}{2(R_{rad} + R_{\Omega})} \left| \frac{\partial Z}{\partial \omega} \right|. \quad (5.6) \]

Figure 5.3: Measurement setup of a monopole antenna prototype (a) with VNA and (b) in Satimo chamber.

Figure 5.4: Simulated and measured $S_{11}$ of a $\lambda/4$ monopole on a small, finite-length cylinder. The cylinder diameter and length is 0.1$\lambda$ and 0.26$\lambda$. 

74
Figure 5.5: Comparison of normalized simulated and measured radiation patterns of the monopole at (a) $\phi = 0^\circ$, (b) $\phi = 90^\circ$ and (c) $\theta = 90^\circ$ cuts; Solid line: measured $\theta$ pol., dash-dot line: measured $\phi$ pol., dashed line: simulated $\theta$ pol. and dotted line: simulated $\phi$ pol. (cannot be seen in this scale).

The high profile of the monopole is a problem. An alternative is a sloping or bending monopole antenna as shown in Fig.5.6a and 5.6b. However, simulation results shown in Fig.5.6c illustrate that $-10$ dB impedance bandwidth decreases sharply for an increasing slope (decreasing profile) of the monopole along the direction of the cylinder axis.

Figure 5.6: (a) Sloping and (b) bending monopoles on a small, finite-length cylinder. (c) Simulated $-10$ dB impedance bandwidth for the sloping and bending monopoles.
5.3.2 Conformal Axial-Direction PIFA on a Small Cylinder

The planar inverted-F antenna (PIFA) is a good candidate as a low profile antenna on a small curved platform such as bicycle frame PIFA has become the most popular antenna for communication devices due to its small size and low profile. It offers an excellent trade-off between compactness and bandwidth, e.g., [2] which gives it an advantage over the conventional patch antenna. It is low cost, easily fabricated and provides a higher impedance bandwidth than patch antennas. The PIFA and its derivatives have been presented for various applications, such as WLAN and satellite communications, e.g., [115], cellular communications e.g., [116], GPS, e.g., [117], and Bluetooth, e.g., [118]. PIFA antennas can readily be integrated into small handheld wireless devices or on parts of vehicular platforms. The performance of a conventional PIFA mounted on various sizes of planar groundplane has been reported in e.g., [119], providing a very useful guideline for designing a PIFA on a small planar groundplane. In [92], a conformal PIFA antenna array mounted on a 0.34λ diameter cylinder has been presented (with the bandwidth of ~ 8% at 2.45GHz), however, there is little reported on electrically smaller cylindrical structures in the literature. The contribution of this section is a design for a small CPIFA on a small diameter cylinder (~ 0.14λ) such as found on bicycle frames at GPS frequencies.

Antenna Structure

Figure 5.7 illustrates the CPIFA structure of interest. The CPIFA is fed by a coaxial cable, and the probe position is determined such that the input impedance is matched to 50Ω at resonance. Copper sheet and PEC are solved to check the sensitivity of conductivity in a compact structure. The thickness of the conductor is 0.005λ for both the CPIFA and the cylinder. The dimensions of a flat PIFA on a large groundplane are calculated as follows, e.g., [2,114]

\[ F_r = \frac{c}{4(L + W - W_s)} \]

where \( F_r \) is the resonant frequency, \( c \) is the speed of light, and \( W_s \) is the width of the shorting pin. For good radiation efficiency, \( W_s \) is chosen to be equal to the width of the PIFA, \( W \), and the length of the PIFA is \( L = \lambda/4 \). These dimensions are all perturbed in numerical experiments for a hand-optimized design which is matched to 50Ω. The details of the sizes for the CPIFA are summarized in Table 5.1.

Table 5.1: Dimensions of the CPIFA on a small cylinder.

<table>
<thead>
<tr>
<th>( W_{\text{CPIFA}} )</th>
<th>( L_{\text{CPIFA}} )</th>
<th>( H_{\text{CPIFA}} )</th>
<th>( L_{\text{cylinder}} )</th>
<th>( D_{\text{cylinder}} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.12λ</td>
<td>0.22λ</td>
<td>0.04λ</td>
<td>2.62λ</td>
<td>0.14λ</td>
</tr>
</tbody>
</table>
Figure 5.7: A CPIFA on a cylinder: (a) CPIFA structure, (b) end view along the cylinder, and (c) 3D view.

Numerical and Measurement Results

The simulations suggest that the design has a good performance with a $-10\text{dB}$ impedance bandwidth of 16% for a nominal frequency of 1.575GHz. The design gives a higher bandwidth compared to the same dimension PIFA on a large planar groundplane, indicating how the cylindrical platform is an important part of the radiating structure. In fact, the simulations show that the bandwidth decreases to $\sim 5\%$ when PIFA is mounted on a planar surface (see Table 5.2).

Table 5.2: Simulated $-10\text{dB}$ impedance bandwidth for a PIFA on different groundplanes.

<table>
<thead>
<tr>
<th>Groundplane</th>
<th>BW (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Small cylinder ($D_{cylinder} = 0.14\lambda$)</td>
<td>16</td>
</tr>
<tr>
<td>Large cylinder ($D_{cylinder} &gt; \lambda$)</td>
<td>5</td>
</tr>
<tr>
<td>Small groundplane (the same size as the PIFA)</td>
<td>0</td>
</tr>
<tr>
<td>Large groundplane</td>
<td>5</td>
</tr>
</tbody>
</table>

Based on the dimensions derived from the simulation study, a prototype of the designs is constructed and measured at GPS frequency. The measurement setup is shown in Fig. 5.8. The measurements and numerical results are compared below.
Figure 5.8: Measurement setup for a CPIFA prototype (a) with VNA and (b) in Satimo chamber.

Figure 5.9: Simulated and measured $S_{11}$ of a CPIFA on a small cylinder. The dimensions are given in Table 5.1.

As explained before, the measured impedance bandwidth is larger than simulated bandwidth (see Fig. 5.9) probably because of the ohmic loss in the external coaxial feed cable. The normalized radiation patterns of the antenna at the centre frequency from simulation and measurement are shown in Fig. 5.10. The gain is $\sim 5$dB, and not significantly affected by the copper conductivity. The patterns are not a major concern for compact antennas here, and the goal is instead to simply get power radiated. However, the patterns are nevertheless of interest in order to estimate the distributed gains for specific propagation scenarios for multipath situations. The CPIFA patterns and their complexity compared to the compact structure of the PIFA reflects the impact of the cylindrical platform. It is emphasized that
these results are for a finite length cylindrical rod. A more complex structure such as a complete UAV or bicycle, etc., will change these patterns.

![Diagram](image)

Figure 5.10: Comparison of normalized simulated and measured radiation patterns of the CPIFA at (a) $\phi = 0^\circ$, (b) $\phi = 90^\circ$ and (c) $\theta = 90^\circ$ cuts; Solid line: measured $\theta$ pol., dash-dot line: measured $\phi$ pol., dashed line: simulated $\theta$ pol. and dotted line: simulated $\phi$ pol.

**Other Considerations**

There is little analysis available on electrically smaller cylindrical structures and no parametric study has been previously reported. Here, a parametric study of a CPIFA is presented including the impact of cylinder diameter and length on the antenna’s bandwidth since the bandwidth is very fundamental in designing such small antenna structures. This study gives a very useful guideline for designing a CPIFA on a small cylinder. Therefore, the effect of the cylinder length and diameter are evaluated. The $-10\,\text{dB}$ impedance bandwidth are compared for different cylinder diameters in Fig. 5.11a, while Fig. 5.11b shows the effect of the cylinder length on the $-10\,\text{dB}$ impedance bandwidth. The dimension of the CPIFA is given in Table 5.1.

The following observations can be made from the above results:

- The numerical results indicate how the cylindrical platform is an important part of the radiating structure.

- In particular, the bandwidth is very sensitive to the cylinder length for the smaller cylinder diameter (e.g., varying between 11% and 20% for $D \sim 0.15\lambda$) when the length of the cylinder is less than about $L = 2\lambda$. For longer length, the bandwidth settles at about 16%.

- Bandwidth decreases by increasing the diameter of the cylinder and converges to $\sim 5\%$ for the cylinder diameter $D > \lambda$. 

79
Figure 5.11: The effect of the (a) cylinder length L and (b) cylinder diameter D on the −10dB impedance bandwidth. The dimension of the CPIFA is given in Table 5.1.

5.3.3 A New Conformal PIFA for Small Cylindrical Groundplane Mounting

It is fed at its end for construction simplicity, by coax. All conductors are 0.005λ thick, with copper for the WCPIFA and PEC for the cylinder. The dimensions in wavelengths are given in Table 5.3.

![Figure 5.12: WCPIFA on a small cylinder: (a) 3D view, (b) end view along the cylinder.](image)

Table 5.3: Dimensions of the WCPIFA on a small cylinder.

<table>
<thead>
<tr>
<th>$W_{WCPIFA}$</th>
<th>$d_{WCPIFA}$</th>
<th>$H_{WCPIFA}$</th>
<th>$L_{cylinder}$</th>
<th>$D_{cylinder} = 2b$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.14λ</td>
<td>0.22λ</td>
<td>0.08λ</td>
<td>2.62λ</td>
<td>0.27λ</td>
</tr>
</tbody>
</table>

The design has good bandwidth and a simple structure. The measurement setups are illustrated in Fig.5.13. The measurement and numerical results given in Fig.5.14 show a
−10 dB impedance bandwidth of ∼ 30%. The design gives a higher bandwidth compared to the same size PIFA on a large planar groundplane. Simulations show that the bandwidth decreases to ∼ 3% when the PIFA is mounted on a flat surface. It indicates how the cylindrical platform is an important part of the radiating structure.

![Measurement setup for a WCPIFA prototype (a) with VNA and (b) in Satimo chamber.](image)

**Figure 5.13:** Measurement setup for a WCPIFA prototype (a) with VNA and (b) in Satimo chamber.

![Simulated and measured $S_{11}$ of a WCPIFA on a small cylinder. $D = 2b$ is the cylinder diameter. The dimensions are given in Table 5.3.](image)

**Figure 5.14:** Simulated and measured $S_{11}$ of a WCPIFA on a small cylinder. $D = 2b$ is the cylinder diameter. The dimensions are given in Table 5.3.

The normalized radiation patterns of the antenna at the center frequency ($D/\lambda = 2.62$) from simulation and measurement are also compared in Fig.5.15. The gain is about 6dB. Although the patterns are not a major factor associated with this design, they give some
useful information. At $\phi = 0^\circ$ cut, the pattern is asymmetric, confirming that the excitation is not symmetric. Again, the patterns indicate that the antenna is not a dipole which has only one polarization.

![Radiation Patterns](image)

Figure 5.15: Comparison of normalized simulated and measured radiation patterns of the WCPIFA at (a) $\phi = 0^\circ$, (b) $\phi = 90^\circ$ and (c) $\theta = 90^\circ$ cuts; Solid line: measured $\theta$ pol., dash-dot line: measured $\phi$ pol., dashed line: simulated $\theta$ pol. and dotted line: simulated $\phi$ pol.

## 5.4 Bent Configurations of Waveguide Slot Arrays

In this section, the possibilities of using a bicycle frame as a bent waveguide slot array is presented. For simplicity, a rectangular waveguide, including doublets (see below), and variations of length of the waveguide is used. The frequency range resulting from the waveguide cross-section is not covered here, apart from noting that a typical bicycle frame size has a cut-off of about 5GHz, and for lower frequency operation, a coaxial type TEM mode is required discussed in the next section. Many researches on waveguide slot arrays have been undertaken for improved bandwidth e.g., [76,111,120,121]. These techniques include reducing the waveguide wall thickness, dividing the array into subarrays, and reducing the waveguide cross section by using ridged waveguide. As noted, the waveguide feed bandwidth of any waveguide or cavity slot array is the major problem for bandwidth. Using multiple waveguide sides includes placing longitudinal slots on opposing broad-walls, sometimes called a "waveguide doublet". This can enhance the bandwidth when the slots are positioned appropriately [122]. The pattern is no longer from a classical single-sided array and so the beam no longer has a single dominant lobe. Coetzee [123] has shown an equivalence between an array of slot doublets and a single-slot array in a half-height waveguide. A design producing two pattern minima in specific directions to avoid unwanted interference was presented [124], and designs for horizontally polarized, omnidirectional waveguide doublet slot array antennas have been investigated in [125–128]. Finally, work has been presented on annular waveguide slot antennas with rectangular cross section [129,130], mo-
tivated by seeking linear and circular polarization configurations. This section focuses on the effect of bends in the waveguide and the unknown waveguide length on the bandwidth. The complexity in modelling the waveguide array when the elements are potentially closely spaced and on multiple faces, and the waveguide bend, means that a simulation-based approach is required. In fact a design cannot be completed well without simulation because the waveguide-element interactions mean that the element design and the array design cannot be separated.

5.4.1 Short-End Effect Study

The waveguide used here is a standard form, WR159, with \( a = 40.38 \text{mm}, \ b = 20.19 \text{mm}, \) and wall thickness 2mm. This size is reasonably compatible with a cross section of a bicycle frame. The simulation results are from CST Microwave Studio [24]. The slots are designed for center frequency of 5GHz, and are a half wavelength long. The waveguide is excited with the CST waveguide port, and the slot positions are offset and alternating around the centre line in the usual way for the 'plus and minus' co-phased excitation. The unexcited end of the waveguide is a short-circuit. The minimum spacing between the slot closest to the short circuit is \( 3/4 \lambda_g, \) where [23,93],

\[
\lambda_g = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{2a}\right)^2}}.
\] (5.8)

The length of the waveguide from waveguide (WG) port excitation is not a critical parameter here. This is because the waveguide port is matched and so the waveguide length at the feed end has no effect on the impedance bandwidth. In practice, the waveguide excitation is seldom well matched across a wide bandwidth, and so this length can in fact have an impact in real-world designs. The distance between the elements and the short-circuit termination is critical, motivating the research reported here. The associated bandwidth reduction of an increased waveguide length is sometimes called the "long line effect". The impact of the short-circuit position on the bandwidth has been previously studied, e.g., [131]. It is noted that this length plays no direct role in the radiation because the array (containing the elements) aperture is not changed. (But there is an indirect role, from the structure). A goal is to make the design as independent as possible of the position of the short-circuit. To study the basic "long line" effect, the spacing from a single slot element to the short-end position is increased from \( 0.75\lambda_g, \) by \( n\lambda_g, \) with \( n \) varying from 0 to 1. The waveguide situation is shown in Fig. 5.16a, and the simulated S-parameter results are in Fig. 5.16b.
Figure 5.16: (a) One slot on top of the waveguide. (b) Simulated S-parameters in terms of the spacing between the slot element and the short-end termination.

At $n = 0.25$ and $0.75$, the impedance bandwidth collapses since the spacing from the slot element to the short-end becomes $\lambda_g$ and $1.5\lambda_g$ respectively. This illustrates the vulnerability of a waveguide slot array to the location of the short circuit termination. At the other lengths of interest ($n = 0$, $0.5$, and $1$) the changing shape of the return loss is apparent in the figure.

In order to get the impedance bandwidth less dependent of the short-end position, another slot is placed at the bottom of the waveguide, with an offset from the first slot, of $0.25\lambda_g$, as depicted in Fig. 5.17a, and with the S-parameter results in Fig. 5.17b.

Figure 5.17: (a) One slot on top (solid line) and one slot on bottom (dashed line) of the waveguide. (b) Simulated S-parameters in terms of the spacing between the top slot element and the short-end termination.

These show that the vulnerability of the impedance bandwidth to the position of the short-end is fixed by using a waveguide doublet, and a good bandwidth is now obtained for
the troublesome lengths of \( n = 0.25 \) and 0.75. The improvement is particularly significant for modest return losses of less than \(-6\) dB, typical for mobile communications.

It is evident from Fig. 5.17b that the impedance bandwidth is improved by having the extra slot. The reason for this is simply that the second slot is on a different side of the waveguide. The radiation from these two slots are quasi-independent, in the sense that they do not comprise a resonant array as with a single-sided slot array.

The numerical experiment can be repeated for (say) a five slot waveguide array, as shown in Fig. 5.18a and Fig. 5.18b, with the S-parameter results Fig. 5.19a (five slots on top) and Fig. 5.19b (five slots on top and also on the bottom). The trends are quite similar to the one- and two-slot cases above.

![Figure 5.18: (a) Five slots on top of the waveguide. (b) Five slots on top (solid line) / bottom (dashed line) of the waveguide.](image)

![Figure 5.19: Simulated S-parameters of (a) five slot waveguide array on top of the waveguide, and (b) five slot waveguide array on top/bottom of the waveguide.](image)

The \(-6\) dB impedance bandwidths of the above configurations are summarized in Fig. 5.20.
It is seen that increasing the number of slots on one side of the waveguide decreases
the bandwidth, following the model of (5.9) and (5.10). The addition of the slot(s) on
the opposite side of the waveguide helps to improve the bandwidth, summarized in Fig.
5.21. In this figure, the modelled single slot bandwidth from the following equations given
in [111,132] for a given VSWR is

\[ VSWR = 1 + \frac{2}{\alpha^2} + \frac{2}{\alpha} \sqrt{1 + \frac{1}{\alpha^2}}, \]  
(5.9)

\[ \alpha = \frac{\pi NB}{300} \left[ 1 + \frac{(\pi NB)^2}{2 \times 10^4} \right], \quad B \text{ in Percent.} \]  
(5.10)

where \( N \) is the number of slots and \( B \) is the percentage impedance bandwidth of a slotted
waveguide array in which the single-side slot spacing is \( \frac{\lambda_g}{2} \). For the cases shown (2 slots,
5 slots and 10 slots), the decrease in the bandwidth follows the model’s curve very nicely,
apart from a bandwidth offset which makes the model look optimistic. For this preliminary
study, there is no attempt to optimize the bandwidth by further varying the design of the
slots.
5.4.2 Circular Waveguide Bend Effect

Here, the effect of an E-plane circular bend on the bandwidth is studied. Figure 5.22 illustrates an E-plane junction of a rectangular waveguide and a uniform circular bend of rectangular cross section.

An equivalent circuit at the reference plane, $T$, for $2b/\lambda_g < 1$, is given by [10], in which

$$\frac{Z'_0}{Z_0} = 1 + \frac{1}{12} \left( \frac{b}{R} \right)^2 \left[ \frac{1}{2} - \frac{1}{5} \left( \frac{2\pi b}{\lambda_g} \right)^2 \right],$$

(5.11)

$$\frac{X}{Z_0} = \frac{32}{\pi^2} \left( \frac{2\pi b}{\lambda_g} \right)^2 \left( \frac{b}{R} \right)^2 \sum_{n=1,3,...}^{\infty} \frac{1}{n^2} \sqrt{1 - \left( \frac{2b}{n\lambda_g} \right)^2},$$

(5.12)
where $\lambda_g$ is given by Eq. 5.8 for rectangular waveguide. The circular bend guided-wavelength can be modelled by [10],

$$\lambda_{g,\phi} = \lambda_g \left\{ 1 - \frac{1}{12} \left( \frac{b}{R} \right)^2 \left[ -\frac{1}{2} + \frac{1}{5} \left( \frac{2\pi b}{\lambda_g} \right)^2 \ldots \right] \right\},$$  \hspace{1cm} (5.13)

where the mean radius $R$ is $R_0 + b/2$. So the minimum radius is for when $R_0 = 0$ and $R = b/2$. It turns out that the effect of the bend, from this model, is negligible. The details are as follows. Figure 5.23 shows $\lambda_{g,\phi}$ in terms of $R$. $\lambda_{g,\phi}$ is varying between 0.0926 and 0.0897 for $R = b/2$, to a very large value. $\lambda_{g,\phi}$ becomes equal to $\lambda_g$ for a large value of $R$, corresponding to a very gentle bend.

![Figure 5.23: Circular bend guided-wavelength in terms of mean radius ($R$).](image)

In order to see the effect of a circular bend on the impedance match, the first order theoretical reflection coefficient is calculated from $\Gamma = \frac{Z_0' - jX - Z_0}{Z_0' - jX + Z_0}$ and is compared with the modelled one, in Fig. 5.24 at reference plane $T$. The results show that the first order approximation is good enough for practical design where the results better than $-6$dB are of interest. For simulation of the reflection coefficient, both sides of the waveguide are terminated with a matched load. The difference between the model and the numerical experiments is significant (tens of decibels). The model is good enough because both results are very low reflections.
To complete the study, the effect of multiple circular bends in a triangular-shaped bent waveguide is used to model a bicycle frame. Several different configurations are simulated as shown in Fig. 5.25, with the S-parameters in Fig. 5.26.

![Different configurations of circular bend (with rectangular cross section) and rectangular waveguide. One slot on top (solid line) and one slot on bottom (dashed line) with 0.25\(\lambda_H\) lengthwise spacing.]

Figure 5.25: Different configurations of circular bend (with rectangular cross section) and rectangular waveguide. One slot on top (solid line) and one slot on bottom (dashed line) with 0.25\(\lambda_H\) lengthwise spacing.
Figure 5.26: Simulated S-parameters of different configurations of circular bend (with rectangular cross section) and rectangular waveguide.

For all of these configurations, one slot is placed on top and another slot is placed at the bottom of the waveguide, with a length offset from the first slot, of $0.25\lambda$. The $-6\text{dB}$ impedance bandwidth is 7.2% for the one waveguide and one bend configuration which is the shortest configuration, and decreases to 5.5% for the longest structure with three straight waveguides and three bends confirming that the "long line" is the dominant effect in these structures and that the impedance bandwidth is less dependent of the short-end position. The result of the circular bend waveguide compared to a straight rectangular waveguide (not shown here) verifies that the circular waveguide bend does not change the $-6\text{dB}$ impedance bandwidth (less than 0.4%).

Here, a preliminary study of using a bicycle frame for a waveguide slotted antenna is presented. The cross section of a typical frame limits the hollow waveguide frequencies to above about 5GHz. WR159 rectangular waveguide is used for this study because it corresponds approximately to the cross section of a frame. The variations of cross section of frames is not addressed, but these incur a range of cut-off frequencies for practical frames. The variability of the distance to a short-circuit termination (or some other termination, depending on the frame design) of a waveguide array, severely compromises the bandwidth of a standard (one-sided) waveguide slot array. This can be compensated by using multiple faces of the waveguide with the slots spaced appropriately. The next section presents new designs of slotted coaxial antennas which are compact and wideband.

5.5 Wideband Antenna using Coaxial Waveguide

In this section, by using the mode of a coaxial waveguide (TEM mode), a much more compact structure is possible, but which can also support a larger bandwidth, at the expense of
an increased ohmic loss in the waveguide. Using this basis, a new type of slotted coaxial waveguide antenna is presented which has unprecedented compactness (cross section) and bandwidth, and yet remains simple to manufacture. The design of such a compact array means that the element and array aspects are not independent, and the design is based on numerical and physical experiments. The bandwidth against different design parameters are included. Two basic design approaches are treated. The “first” design, with no constraint on the extent of the slots, and the “second” design where the mechanical strength of the waveguide is partially maintained by constraining the azimuthal support of the slot to be 90° or less.

5.5.1 Basic Coaxial Waveguide Design Principles

Coaxial line impedance

The coaxial line nomenclature and indicative electric field lines of TEM mode are shown in Fig. 5.27a and Fig. 5.27b, respectively.

![Coaxial line geometry](image)

**Figure 5.27:** (a) Coaxial line geometry: $\varepsilon$ is the permittivity of the dielectric, $a$ and $b$ are the inner and outer conductors radius, (b) electric/magnetic field lines and currents for the TEM mode of coaxial line (after [11]).

The characteristic impedance of the coaxial line is [11]

$$Z_0 = \sqrt{\frac{\mu \ln b/a}{\varepsilon}}$$

where the usual notation for the permittivity is $\varepsilon = \varepsilon_0 \varepsilon_r$ with $\varepsilon_r = \varepsilon'_r - j \varepsilon''_r$ and electrical loss tangent, $\tan \delta = \varepsilon''_r / \varepsilon'_r$; and similarly for the permeability, $\mu = \mu_0 \mu_r$, of the dielectric insert. Magnetic materials or magnetic loss are not considered here, so $\mu_r = 1$. The coaxial line can support TE and TM modes as well, and these higher modes are usually under cut off.

Preventing propagation of higher order modes sets an upper limit on the size of a coaxial cable or, equivalently, an upper limit on the frequency of operation for a given cable. The
cut-off frequency before the second mode ($TE_{11}$) starts to propagate is [11],

\[ f_c = \frac{ck_c}{2\pi \sqrt{\varepsilon_r}}, \quad (5.15) \]

where, \( k_c = \frac{2}{a+b} \) is the cut-off wave number. In practice, a 5% safety margin is used for this cut-off frequency [11]. For design, Fig. 5.28 illustrates the cross section size for the cut-off frequency of the $TE_{11}$ of a $50\Omega$ coaxial line with dielectric insert air, i.e., \( \varepsilon_r = 1 \), where \( b/a \) is calculated by (5.14). This is the highest frequency before the $TE_{11}$ waveguide mode starts to propagate.

![Figure 5.28: Cut-off frequency of the $TE_{11}$ mode of a $50\Omega$ coaxial line with a (lossless) air dielectric insert, $\varepsilon_r = 1$. (Here, $k_c = 2/(a + b)$, and $a$ is calculated for $b$ ranging from 1 mm to 20 mm).](image)

**Eccentric coaxial line**

The eccentricity of the coax structure is important for practical applications where the centre conductor cannot be accurately placed. This happens when post-fitting a centre conductor to an existing cylindrical structure or when there is a bend in the structure, such as in a bicycle frame. Figure 5.29a shows the geometry of a coaxial line with off-center inner conductor. The characteristic impedance of the coaxial line is known [12],

\[ Z_0 = \left( \frac{60}{\varepsilon_r^{1/2}} \right) \cosh^{-1} \left( \frac{1}{2} \left( \left( \frac{b}{a} \right) + \left( \frac{a}{b} \right) - \left( \frac{c^2}{ab} \right) \right) \right), \quad (5.16) \]

and this is plotted against the eccentricity in Fig. 5.29b for a coaxial line with dielectric insert air, i.e., \( \varepsilon_r = 1 \). For design, this gives insight into the accuracy required of the coaxial structure, allowing distributed reflection coefficients to be estimated from the geometry.
It is evident that quite a large tolerance of eccentricity is possible, with centre conductor offsets of up to 40% of the outer conductor radius having a only a small impact.

Figure 5.29: (a) Eccentric line (After [12]). (b) Line impedance in terms of eccentricity for dielectric insert air ($\epsilon_r = 1$).

Coaxial line ohmic loss

Attenuation in waveguide is from dielectric $\alpha_d$ (Np/m) and conductor $\alpha_c$ (Np/m) losses, and these are also well known. The attenuation per length due to dielectric loss for TEM mode is [11]

$$\alpha_{d_{TEM}} = \frac{k}{2} \tan \delta \left( \frac{1}{a} + \frac{1}{b} \right) \text{ (Np/m)},$$

where $k = \omega \sqrt{\mu \epsilon}$. The conductor-loss due to a perfectly smooth conductor for the TEM mode is [11],

$$\alpha_{c_{TEM}} = \frac{R_s}{2 \eta \ln \left( \frac{b}{a} \right)} \left( \frac{1}{a} + \frac{1}{b} \right) \text{ (Np/m)},$$

where $R_s = \sqrt{\frac{\omega \mu_0 \mu_r}{2 \sigma}}$ is the surface resistance, and $\eta = \sqrt{\frac{\mu_0 \mu_r}{\epsilon_0 \epsilon_r}}$ is the intrinsic impedance.

The loss due to the dielectric and conductor, in dB/m, of a 50$\Omega$ coaxial line with a teflon dielectric, are shown in Fig. 5.30a and Fig. 5.30b. The dimensions of the coaxial line (i.e., the inner conductor radius $a$ and the outer conductor radius $b$) are varying in terms of frequency so that the real part of the characteristic impedance (governed by the ratio $b/a$), $Z_0$, is kept to $R_0 = 50\Omega$. Conductor loss increases with the square root of frequency, while dielectric loss increases linearly. For air dielectrics, only the conductor loss is of interest.
Figure 5.30: (a) Conductor loss and (b) dielectric loss per meter for a 50Ω coaxial line (TEM mode) in comparison with rectangular waveguide (TE10 mode) for copper ($\sigma = 5.96 \times 10^7$), gold ($\sigma = 4.17 \times 10^7$), stainless steel ($\sigma = 1.45 \times 10^6$) and aluminium ($\sigma = 3.5 \times 10^7$) filled with teflon medium modelled with $\epsilon_r = 2.1$ and $\tan \delta = 0.0004$.

It is of interest to compare these losses with that of hollow waveguide - the lowest-loss metallic guiding structure. The conductor loss for TE$_{10}$ mode rectangular waveguide is

$$\alpha_{cTE_{10}} = \frac{R_s (2b \pi^2 + a^2 k^2)}{a^3 b \beta k \eta} \text{ (Np/m)}, \quad (5.19)$$

where propagation constant $\beta = \sqrt{k^2 - k_c^2}$ is real with $k_c = \pi / a$ the cut-off wavenumber for TE$_{10}$ rectangular waveguide. (Note the clash of notation conventions, this $k_c$ is different to the geometric parameter $k_c$ used above.) The dielectric loss associated with TE$_{10}$ mode is given by (cf, equation 5.17)

$$\alpha_{dTE_{10}} = \frac{k^2 \tan \delta}{2 \beta} \text{ (Np/m)}. \quad (5.20)$$

These losses are included in Fig.5.30a and 5.30b for coaxial line (TEM mode) and the TE$_{10}$ mode rectangular waveguide (because this is the usual choice for slot arrays) with the commonly-used different metals: copper, aluminium, and stainless steel. The conductor loss is sensitive to the material, with the stainless steel loss demonstrating the jump from the good conductors. This loss has particular relevance to longer waveguide arrays. For example, for a frequency of 60GHz, a 1 meter long antenna (i.e, a length of almost 200 free space wavelengths), as could be encountered in highly directive mmwave antennas for SAR, would have feed-end-to-feed-end waveguide ohmic loss of 4dB, or 60%, using aluminum. But for the small electrical lengths of interest here, the loss is small. Also, there is not much difference in the loss between the waveguide types. The dielectric loss is also compared, in the lower figure. For a low-loss material such as teflon, this is still significant compared to...
the conductor losses. However, teflon is expensive, and the above method allows other more practical materials to be investigated.

5.5.2 Feeding System

Waveguide antenna structures are conveniently simulated using a waveguide feed which is an idealized (wideband), pure mode excitation. It is insightful to separate the feed from the antenna structure because the bandwidth limitation of the antenna structure can be separated from the bandwidth limitation of the feed. A practical feed can constrain the bandwidth to be less than that of a wideband antenna structure, e.g., [133], so understanding the feed performance is critical for realizing wideband antennas.

The coaxial connector used in the prototypes is a standard panel-mount N-type. A standard connector is unlikely to be the same size as the coaxial waveguide antenna which is designed for a specific frequency, so a transition is required. A tapered coaxial section can be used, but this is expensive. A "lumped" section, as shown in Fig. 5.31, is more practical. The antenna designs presented below do not have a solid dielectric material inside the coaxial cylinder, making the eccentricity a practical problem for the applications such as a post-fit within a bicycle frame. The coaxial eccentricity of the transition (already looked at for the waveguide, above), and the length of the extended coaxial inner conductor, shown in the figure, are important parameters to vary for realizing a wideband feed.

![Figure 5.31: Geometry of the feeding: a 50Ω panel mount N-type connector connected to the coaxial waveguide.](image)

For the choice of operating frequencies, the inner radius of the N-type connector is smaller than the inner of the coaxial waveguide. Fig. 5.32b and 5.32a show parametric simulation results for eccentricity and lengths of the extended coaxial inner transition. The impedance bandwidth decreases with increasing eccentricity as expected, but as with the waveguide study above, there is good tolerance to eccentricity. An electrical length of about 0.03λ is a good solution, albeit rather sensitive, and this length is used in the experimental feeds and the simulation results, below.
Slotted Coaxial Waveguide Antenna Design Concept

This section details the design of three examples of novel slotted waveguide antennas for wide bandwidth. For a given cylindrical outer radius \( b \) of \( 0.12\lambda \) with dielectric insert of air; the inner radius \( a \) is determined to be about \( 0.05\lambda \) in order to obtain 50 characteristic line impedance at the frequency of \( 1.8\text{GHz} \). The length of the cylinder is about \( \lambda \) with wall thickness of \( 0.006\lambda \). The slot length is about \( 0.5\lambda \) similar to the traditional slotted hollow waveguide antenna, and the slot width of \( 0.01\lambda \). Broadening the slot width can improve the impedance bandwidth, and simulations (not included here) show that continuing to broaden the slot width to \( 0.08\lambda \) continues to improve the bandwidth. However, the slot width is kept thin in the designs for structural stability of the air-dielectric designs.

Slot antenna position and orientation

The slots are cut so that the current along the waveguide (see Fig. 5.27b) is interrupted, creating an effective voltage excitation across the slot, which radiates. In a simple equivalent circuit, the radiation loss from the slot is represented by the resistive loss in a series impedance [134]. The angle of the slot, \( \alpha \), and its position, are the important parameters (see Fig. 5.33a). Figure 5.33b shows how the \(-6\text{dB}\) and \(-10\text{dB}\) impedance bandwidths are impacted by different positions of the slot from the shorted-end position (Fig. 5.33b left) and \( \alpha \) angle (Fig. 5.33b right). The maximum \(-6\text{dB}\) impedance bandwidth of \( \sim 50\% \) and the maximum \(-10\text{dB}\) impedance bandwidth of \( \sim 10\% \), are obtained for a slot position of \( 0.5\lambda \) from the short-end termination. They reduce to zero when the slot is at an odd multiple factor of \( 0.25\lambda \) from the short-end position, and, also expected, increasing \( \alpha \) from \( 0^\circ \) to \( 80^\circ \).
collapses the bandwidth. The standing wave distribution can be obtained by either short or open-end termination. Using an open-end termination results in different slot positions of course, and the fringing fields vary with frequency. All the results are for the shorted termination since it is easier to model across a wide bandwidth.

![Diagram of a one-slotted coaxial waveguide antenna with cylinder length \( L_{\text{cyl}} \) of \( \lambda \), slot length \( S_l \) of 0.5\( \lambda \) and slot width \( S_w \) of 0.01\( \lambda \).](image)

![Graph showing impedance bandwidth in terms of the position of the slot from the short-end position with the \( \alpha \) angle of 0° (bottom x-axis).](image)

Figure 5.33: One-slotted coaxial waveguide antenna with cylinder length \( L_{\text{cyl}} \) of \( \lambda \), slot length \( S_l \) of 0.5\( \lambda \) and slot width \( S_w \) of 0.01\( \lambda \). (b) -6dB and -10dB impedance bandwidth in terms of the position of the slot from the short-end position with the \( \alpha \) angle of 0° (bottom x-axis).

Coaxial waveguide length effect

The length of the waveguide feed does not have a strong effect on the impedance bandwidth in long antennas [94] because the fields are mostly governed by the internal waveguide structure. But for short antennas, the outside of the structure becomes important. Figure 5.34a shows how the position of the feeding (N-type connector) affects the -3dB, -6dB and -10dB bandwidths. It can be seen that a distance of 0.125\( \lambda \) between the slot and the N-type feeding position gives good bandwidth. The length of the slots is quite long relative to the cylinder circumference, and their proximity to the waveguide port can disturb the mode purity. As a result, the impedance matching varies with this proximity. This distance also impacts the radiation pattern, as illustrated for the \( \phi = 90^\circ \) cut of single-slotted waveguide antenna in Fig. 5.34b. For the applications, the pattern is not of priority but its form is still of interest. For this simulation there was no feed cable attached to the antenna, and for pattern-critical applications, the cable presence may well change the pattern and impedance behaviour. In the measurements below, a cable was present of course.
Figure 5.34: (a) Simulated $3\text{dB}$, $6\text{dB}$ and $10\text{dB}$ impedance bandwidth in terms of the slot position form the waveguide port of one-slotted coaxial waveguide antenna (see Fig. 5.33a), the slot length is fixed at $0.5\lambda$, (b) simulated normalized radiation pattern for ($\phi = 90^\circ$ cut) for different coaxial waveguide lengths of one slotted coaxial waveguide. The distance between the feed and the slot is varied from $0.125\lambda$ to $0.5\lambda$. The slot position is fixed at $0.5\lambda$ from the short-end with the angle of $\alpha = 0^\circ$.

**Radiation patterns over the operating frequency band**

The normalized radiation pattern for a one slotted coaxial waveguide with $\alpha$ angle of $0^\circ$ located at $0.5\lambda$ from the short-end termination is plotted in Fig.5.35 for the design frequency, center frequency, lower and upper operating frequency. The cylinder length is $\lambda$. Since this is a low gain antenna (about 3 to 4dB), the patterns are expected to change over the operating frequency band.

Figure 5.35: The normalized radiation pattern for a one slotted coaxial waveguide with $\alpha$ angle of $0^\circ$ located at $0.5\lambda$ from the short-end termination for the design frequency ($F_D$), center frequency ($F_0$), lower ($F_L$) and upper ($F_U$) operating frequency. The cylinder length is $\lambda$. 

98
Effect of the number of slots

In a slotted waveguide antenna, the length can be expressed as the number of array elements. Hamadallah’s classic paper [111] models how increasing the number of array elements on one side of the waveguide decreases the impedance bandwidth because of the long-line effect. This model assumes that the array (i.e. waveguide fields) and the element effects, are reasonably decoupled. The numerical results confirming this trend are plotted in Fig. 5.36 for a coaxial waveguide with slot elements with $\alpha = 0^\circ$, from the basic design of the previous section. Note that the slots are on one side only of the waveguide, as in Hamadallah’s model.

![Figure 5.36: Effect of varying the number of slots with $\alpha = 0^\circ$ and length of $0.5\lambda$ on $-6$dB and $-10$dB impedance bandwidth.](image)

5.5.4 Slotted Coaxial Waveguide Antenna Design I

Based on the findings from parametric studies above, three antennas are now presented, including measurements. These are for a single azimuthal slot, a dual azimuthal slot (one slot each side of the coaxial waveguide) and dual intersecting slots with angles offset from azimuth. These demonstrate the development of higher bandwidth with increasing complexity. Fabrication inaccuracy and the presence of a feed cable cause some minor differences compared to the basic simulation structure. In this section, three new compact (cross section) and wideband slotted coaxial configurations are presented with two different cylinder lengths; i.e., $L = 0.625\lambda$ and $0.25\lambda$. All the designs have been fabricated. The outer and the inner conductor of the prototypes are hollow copper tubes and the slots are cut using laser cutter. The antennas are fed by a N-type connector as discussed above. Figure 5.37 shows the simulation model and the prototype of the first design with two different cylinder lengths: Fig. 5.37a illustrates a single slot with $\alpha = 0^\circ$ at $d_{\text{short}} = 0.5\lambda$ from the short-end position. Cylinder length is $L = 0.625\lambda$, and the distance between the slot and the feed-end is $d_{\text{feed}} = 0.125\lambda$. While Fig. 5.37b depicts one slotted coaxial waveguide with the
cylinder length of $L = 0.25\lambda$, $d_{\text{short}} = 0.125\lambda$ and $d_{\text{feed}} = 0.125\lambda$. The simulated and measured impedances are shown in Fig. 5.37c. The simulated and measured $-6\text{dB}$ and $-10\text{dB}$ impedance bandwidths, antennas directivity, radiation efficiency and total radiation efficiency are summarized in Table 5.4. As expected the smaller antenna maintains the smaller bandwidth. Moreover, the mismatch loss with the N-type connector feeding is less than than with the waveguide port excitation resulting the better total efficiency achieved using the N-type connector feeding. It also shows that there are multiple modes present with N-type feeding, quite interesting for general feed design.

Figure 5.37: One-slotted coaxial waveguide antenna with the slot length of $\sim 0.5\lambda$ and $\alpha = 0^\circ$: (a) $d_{\text{short}} = 0.5\lambda$ and $d_{\text{feed}} = 0.125\lambda$, (b) $d_{\text{short}} = 0.125\lambda$ and $d_{\text{feed}} = 0.125\lambda$. (c) Simulated (solid line) and measured (dashed line) $S$-parameter results.
Table 5.4: One-slotted coaxial waveguide antenna. $kb = 0.75$ at 1.8GHz and coaxial cable length is $0.22\lambda$.

<table>
<thead>
<tr>
<th>Design</th>
<th>Cylinder Length</th>
<th>Simulated (Measured) $-6\text{dB}$ Imp. BW (%)</th>
<th>Simulated (Measured) $-10\text{dB}$ Imp. BW (%)</th>
<th>Directivity (dBi)</th>
<th>Radiation Efficiency (dB)</th>
<th>Total Efficiency (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>$L = 0.625\lambda$</td>
<td>$\sim 40$ ($\sim 36$)</td>
<td>$\sim 27$ ($\sim 22$)</td>
<td>3.4</td>
<td>$-0.01$</td>
<td>$-0.12$</td>
</tr>
<tr>
<td>II</td>
<td>$L = 0.25\lambda$</td>
<td>$\sim 20$ ($\sim 22$)</td>
<td>$\sim 13$ ($\sim 13$)</td>
<td>5.4</td>
<td>$-0.034$</td>
<td>$-0.15$</td>
</tr>
</tbody>
</table>

Figure 5.38: One X-shaped slotted coaxial waveguide antenna with the slot length of $\sim 0.5\lambda$: (a) $\alpha = 10^\circ$, $d_{\text{short}} = 0.5\lambda$ and $d_{\text{feed}} = 0.125\lambda$, (b) $\alpha = 15^\circ$, $d_{\text{short}} = 0.125\lambda$ and $d_{\text{feed}} = 0.125\lambda$. (c) Simulated (solid line) and measured (dashed line) $S$-parameter results.

The parametric study also gives rise to the two-slot ("X-shaped slot") designs of Fig. 5.38, with $\alpha = 10^\circ$ for the longer antenna shown in Fig. 5.38a and $\alpha = 15^\circ$ for the shorter antenna in Fig. 5.38b. The simulated and measured $S$-parameters are illustrated in Fig. 5.38c. The
simulated and measured −6dB and −10dB impedance bandwidths, antennas’ directivity, radiation efficiency and total radiation efficiency are summarized in Table 5.5.

Table 5.5: One X-shaped slotted coaxial waveguide antenna. $kb = 0.75$ at 1.8GHz and coaxial cable length is $0.22\lambda$.

<table>
<thead>
<tr>
<th>Design</th>
<th>Cylinder Length</th>
<th>Simulated (Measured) −6dB Imp. BW (%)</th>
<th>Simulated and (Measured) −10dB Imp. BW (%)</th>
<th>Directivity (dBi)</th>
<th>Radiation Efficiency (dB)</th>
<th>Total Efficiency (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>I</td>
<td>$L = 0.625\lambda$</td>
<td>$~ 49$ (~ 50)</td>
<td>$~ 37$ (~ 37)</td>
<td>3.42</td>
<td>−0.01</td>
<td>−0.29</td>
</tr>
<tr>
<td>II</td>
<td>$L = 0.25\lambda$</td>
<td>$~ 29$ (~ 20)</td>
<td>$~ 14$ (~ 11)</td>
<td>5.4</td>
<td>−0.029</td>
<td>−0.64</td>
</tr>
</tbody>
</table>

Figure 5.39: Two-slotted coaxial waveguide antenna with the slot length of $\sim 0.5\lambda$ and $\alpha = 0^\circ$:
(a) $d_{\text{shortslot}} = 0.5\lambda$ and $d_{\text{feedslot}} = 0.125\lambda$, and $d_{\text{slot}} = 0.125\lambda$, (b) $d_{\text{shortslot}} = 0.125\lambda$, and $d_{\text{feedslot}} = 0.125\lambda$, and $d_{\text{slot}} = 0.006\lambda$. (c) Simulated (solid line) and measured (dashed line) S-parameter results.
Placing another slot on the opposite side of the waveguide can improve the bandwidth [122]. The second slot also has $\alpha = 0^\circ$, and a spacing of $d_{slot} = 0.125\lambda$ for the longer cylinder with the cylinder length of $L = 0.625\lambda$ (Fig. 5.39a), and $d_{slot} = 0.006\lambda$ for the antenna with the cylinder length of $L = 0.25\lambda$ (Fig. 5.39b). The simulated and the measured impedance bandwidth are compared in Fig.5.39c. The simulated and measured results are summarized in Table 5.6 showing a remarkable impedance bandwidth.

Table 5.6: Two-slotted coaxial waveguide antenna. $kb = 0.75$ at 1.8GHz and coaxial cable length is $0.22\lambda$.

<table>
<thead>
<tr>
<th>Design</th>
<th>Cylinder Length</th>
<th>Simulated (Measured) $-6\text{dB}$ Imp. BW (%)</th>
<th>Simulated (Measured) $-10\text{dB}$ Imp. BW (%)</th>
<th>Directivity (dBi)</th>
<th>Radiation Efficiency (dB)</th>
<th>Total Efficiency (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>$L = 0.025\lambda$</td>
<td>$\sim 66$ ($\sim 66$)</td>
<td>$\sim 57$ ($\sim 53$)</td>
<td>3.94</td>
<td>$-0.02$</td>
<td>$-0.25$</td>
</tr>
<tr>
<td>II</td>
<td>$L = 0.25\lambda$</td>
<td>$\sim 57$ ($\sim 57$)</td>
<td>$\sim 45$ ($\sim 33$)</td>
<td>3.5</td>
<td>$-0.029$</td>
<td>$-0.32$</td>
</tr>
</tbody>
</table>

### 5.5.5 Slotted Coaxial Waveguide Antenna Design II

In the examples above, remarkable bandwidth could be achieved, but the disadvantage is that the mechanical strength of the antenna is compromised because of the long slot length relative to its circumferential support. A way around this is to use a rigid dielectric filler, or an internal or external dielectric shell section. Another way is to constrain the azimuthal extent of the slots. This Section presents such a design and includes small-array cases for distributing the radiation area. A motivation of such a distribution is to provide a diversity of radiators for situations where one or more of the slot elements may be blocked, such as with a bicycle frame where equipment is attached to the frame or where the rider effectively blocks individual elements. The azimuthal angle between the beginning and the end of the slot around the cylinder axis, $\theta$, is here limited to a maximum of $\theta = 90^\circ$, see Fig.5.40. This choice is somewhat arbitrary but mechanical studies (not included here) indicate that this is a reasonable choice. This constraint requires the angle $\alpha$ to be about $70^\circ$.

Figure 5.40: The angle $\theta$ between the beginning and the end of the slot seen from along the cylinder.
The cylindrical structure has outer radius $b = 0.1\lambda$, inner radius $a = 0.04\lambda$ and wall thickness of $0.005\lambda$ (see Fig. 1). From Fig. 5.33b, the bandwidth for $\alpha = 70^\circ$ has collapsed. But placing another slot on the other side of the cylinder as depicted in Fig. 5.41a can restore some bandwidth: the $-6\text{dB}$ and $-10\text{dB}$ bandwidths become 7\% and 4\% respectively. Reducing the slot length to $0.44\lambda$, which corresponds to the constraint ($\theta < 90^\circ$), $\theta$ becomes $82^\circ$. The length of the cylinder for this configuration is $L = \lambda$.

Figure 5.41: Slotted coaxial waveguide with (a) two slots (1 slot on top & 1 slot on bottom), (b) four slots (2 slots on top & 2 slots on bottom) and (c) six slots (3 slots on top & 3 slots on bottom). $S_{11} = 0.44\lambda$, $d_{\text{short}} = 0.5\lambda$, $d_{\text{feed}} = 0.5\lambda$ and $d_{\text{slot}} = 0.5\lambda$ at $f_0 = 1.575\text{GHz}$ (d) simulated s-parameters comparison. Bottom slots are shown as dotted lines.
The bandwidth of this dual-slot arrangement is still too low. By adding a second pair rotated to $-\alpha$ and placed at $\lambda/2$ from the first pair, as shown in Fig. 5.41b, the $-6\,\text{dB}$ and $-10\,\text{dB}$ impedance bandwidths increase to about 16% and 10%. So this is an two-sided array configuration, and the cylinder length is $L = 1.5\lambda$. Hand-optimizing readily leads to $-6\,\text{dB}$ and $-10\,\text{dB}$ bandwidths of 20% and 17% for a six-slotted coaxial antenna with the cylinder length of $L = 2\lambda$ (see Fig. 5.41c) and has three slot pairs cut on the cylinder.

From the parametric studies, decreasing $\alpha$ increases the bandwidth. For structures strengthened by a thin external dielectric shell, the constraint can be removed, and allowing $\alpha < 70^\circ$ (i.e., $\theta > 90^\circ$), then the $-6\,\text{dB}$ bandwidth can go to 34% for $\alpha = 60^\circ$ for the design with 6 slots. (This is the maximum angle that prevents the top and bottom slots from overlapping.) Figure 5.41d illustrates the s-parameters of the above designs showing that increasing the number of slots is initially increasing the impedance bandwidth. This is contrary to the long line effect where the array aspect (waveguide fields) is not strongly affected by the slot elements. But here the presence of slots interferes more with the fields within the compact coaxial waveguide. In Fig.5.42 (left y-axis), the $-6\,\text{dB}$ and $-10\,\text{dB}$ impedance bandwidth is against the number of slots varying from 2 to 28. For example, each has a maximum of about 22% for 8 to 14 slots and reduces to about 18% for 28 slots. This is also a remarkable bandwidth for such a large array, but it must be remembered that the pattern is not a priority here whereas most arrays are for, and prioritize, the pattern directivity. Here as shown in Fig. 5.42 (right y-axis), the peak directivity of about 9dBi for the 28-slot antenna, is in two directions in the azimuth plane, see the patterns for 6-slot version below.

![Figure 5.42: Effect of varying the number of slots on $-6\,\text{dB}$ and $-10\,\text{dB}$ impedance bandwidth (left axis) and directivity (right axis).](image)

The design with six slots on a $2\lambda$ cylinder length (about the length of bicycle frame section at the frequencies considered) is fabricated as depicted in Fig. 5.43a, and its s-parameter results are in Fig. 5.43b.
The normalized pattern cuts ($\theta = 90^\circ$, $\phi = 90^\circ$) at the design frequency 1.57GHz are in Fig. 5.44a and 5.44b, respectively. The pattern is linearly polarized with dipole pattern shape in the plane around the cylindrical aperture. Therefore, it suggests to utilize this design in higher frequency, e.g., 5GHz WiFi band as shown in Fig. 5.45a. With no constraint on the extent of the slot; i.e., $\alpha = 60^\circ$, and the slot width of 0.1$\lambda$, the simulated impedance bandwidth in Fig. 5.45b shows the $-6$dB and $-10$dB impedance bandwidth of $\sim 27\%$ and $\sim 21\%$. Figure 5.46a and 5.46b also depict the normalized pattern at $\theta = 90^\circ$ and $\phi = 90^\circ$ cut at the design frequency of 5GHz. Again, it shows that the pattern is linearly polarized with dipole pattern shape in the plane around the cylindrical aperture. Table 5.7 summarize the simulated and measured $-6$dB and $-10$dB impedance bandwidths, antennas’ directivity, radiation efficiency and total radiation efficiency of the six-slotted coaxial antenna at 1.57GHz and 5GHz.
Figure 5.44: Normalized radiation pattern of the prototype with 6 slots (Fig. 5.41c) at the frequency of 1.575GHz, (a) $\theta = 90^\circ$ cut, (b) $\phi = 90^\circ$ cut; solid line: simulated $\theta$ pol., dash-dot line: simulated $\phi$ pol., dashed line: measured $\theta$ pol. and dotted line: measured $\phi$ pol.

Figure 5.45: (a) 6-slotted coaxial waveguide antenna (3 slots on top & 3 slots on bottom) with $\alpha = 60^\circ$, $d_{\text{short}} = d_{\text{slot}} = d_{\text{feed}} = 0.5\lambda$ and the slot length is $0.44\lambda$, (b) simulated S-parameter results. The design frequency is 5GHz.
(a)

(b)

![Normalized radiation pattern of the prototype with 6 slots](image)

Figure 5.46: Normalized radiation pattern of the prototype with 6 slots (Fig. 5.41c) at the frequency of 5GHz, (a) $\theta = 90^\circ$ cut, (b) $\phi = 90^\circ$ cut; solid line: simulated $\theta$ pol., dash-dot line: simulated $\phi$ pol.

Table 5.7: Six-slotted coaxial waveguide antenna. Cylinder length $L = 2\lambda$, $kb = 0.65$ and coaxial cable length is $0.22\lambda$.

<table>
<thead>
<tr>
<th>Design</th>
<th>Design Frequency (GHz)</th>
<th>$\alpha$</th>
<th>Simulated (Measured)</th>
<th>Simulated (Measured)</th>
<th>Directivity (dBi)</th>
<th>Radiation Efficiency (dB)</th>
<th>Total Efficiency (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>1.575</td>
<td>70$^\circ$</td>
<td>~20 (~20)</td>
<td>~17 (~17)</td>
<td>5.7</td>
<td>-0.03</td>
<td>-0.19</td>
</tr>
<tr>
<td>II</td>
<td>5</td>
<td>60$^\circ$</td>
<td>~27 (no measurement)</td>
<td>~21 (no measurement)</td>
<td>4.36</td>
<td>-0.05</td>
<td>-0.13</td>
</tr>
</tbody>
</table>

5.6 Summary and Conclusion

Three new possibilities for antenna designs on a tubular platform were presented in this Chapter. First, a new wrapped conformal PIFA (WCPIFA) antenna mounted on a small cylindrical platform ($kb \sim 0.8$) was presented. Finding design solutions for this type of configuration require simulation since accurate analytical and transmission line models are too challenging to formulate. The numerical and measurement results indicated how the cylindrical platform is an important part of the radiating structure. The antenna has a $-10$dB impedance bandwidth of $\sim 30\%$ at the frequency of 1.575GHz. This chapter also investigates bent waveguide-excited slot antennas (in TE mode). The drawbacks of waveguide slot arrays, and how to get round them, are discussed. These include their bandwidth sensitivity to dimensional variations, and in some cases such as a bicycle frame, there is variability in the waveguide cross section, length, and there are bends along the waveguide. The use of slots on different facets of a rectangular waveguide are shown to improve the bandwidth and help compensate for the frame length variability. Finally, this chapter presents and demonstrates new antenna structures which can have wide bandwidth for their electrical size. The basis is a coaxial TEM mode exciting slot radiators in the outer conductor. The location
and size of the slot elements is particularly important, and excellent bandwidth is possible when these are designed correctly. The elements can be extended to an array for higher directivity. The first set of designs have just two slots on opposite sides of the waveguide. When the slots are aligned with the circumference, i.e., $\alpha = 0^\circ$, and with no constraint on the angular extent of the slots, and the cylinder length of $L = 2\lambda$, the $-10\text{dB}$ impedance bandwidth of $\sim 50\%$ is obtained. By decreasing the length of the cylinder to $L = 0.25\lambda$, the impedance bandwidth decreases only by $\sim 10\%$ so there is still a remarkable bandwidth for such antenna size. The second set of designs have the azimuthal slot extent constrained to be less than $90^\circ$ for bicycle frames. For the 6 slot example, the $-10\text{dB}$ impedance bandwidth is $\sim 17\%$ at the frequency of 1.575GHz, and without constraint on the angular extent of the slots, the impedance bandwidth increases to $\sim 21\%$ at the frequency of 5GHz. The feed design of the antennas is included, and is part of the simulations and prototypes confirming that the feed is not limiting the bandwidth of the antennas.
Chapter 6

Thesis Contributions and Future Directions

The theme of this thesis has been exploring the propagation and small cylindrical antennas for NLOS. The work in this thesis is organized into two parts - Part I (Chapters 2, 3 and 4) discusses diffraction as NLOS of propagation mechanism, propagation around corners, and propagation through-forest, respectively. Part II (Chapters 5 and 6) investigated electrically small cylindrical antennas for NLOS. Several new contributions are presented. This chapter concludes the thesis with a summary of those contributions, and suggests future extensions.

Chapter 2 presents an introduction to diffraction principles needed for this research. It covered the uniform geometrical theory of diffraction (UTD) for short-haul propagation where a single building or tree block in the propagation path. Semi-infinite and finite baffles are studied using theoretical, numerical and physical measurement results. The idea is that a wider understanding will lead to simpler and more accurate models for wireless system designers. Future directions for this section can include the following:

- Measurement of two or three baffles;
- Comparing the physical measurement results with UTD multiple-edge transition zone diffraction.

Chapter 3 addresses improving the link gain between antennas in designing for efficient communications. Maximizing the link gain between non-line-of-sight (NLOS) antennas is a goal in multiple-input and multiple-output (MIMO) communications. There is recent interest in also controlling reflections in the propagation paths, calling for new ideas at the system and component levels. The critical mechanism in the NLOS propagation is diffraction. A simple add-on system for corners, comprising a simple dipole or array is presented. This significantly improves the coverage in the shadow area. The analysis, simulations, and physical measurements are also presented to demonstrate the concept. The use of adaptive surfaces for improved link gain is being considered for future systems, but the technique can be applied as a wideband, fixed retrofit for current systems as well, or as an adaptive
system for adaptive control of the scattering propagation. Future directions for this section can include the following:

- Outdoor measurement for 5G application;
- Using systems comprising antenna arrays and curved duct along the corner.

Chapter 4 of the thesis reviews propagation through vegetation based on radiative energy theory (RET). This is an extremely complicated model and is very time consuming to calculate, but remains an industry standard model for terrestrial communications in the absence of simpler approaches. It is also used for backscattering in remote sensing. Few wireless practitioners fully understand the RET method. A new approximating model for forward transmission, based on transmission line techniques, is presented here, providing much simpler expressions for the loss through short distance vegetation. The model is computationally very fast and requires fewer parameters than the RET. By demonstrating that the approach works just as well as the RET model (in fact, both are empirical models), it is possible that this much simpler method will become widely accepted by practicing wireless system designers for short distance propagation through forest (< 100 m). Transmission through such inhomogeneous mixed media especially for long distance propagation (> 2500 m) is also complicated by the many different propagation mechanisms and the complexity of the randomness. This means that accurate, purely physics-based analysis is unlikely to be practical (conveniently computed), and similarly, that practical, purely random modeling is unlikely to be accurate. Through-vegetation propagation models, including the standard RET, are not very accurate for long distance propagation in the sense that the uncertainty can be tens of decibels, and this seems to be an accepted limitation for vegetation. A simpler propagation model, which maintains or improves accuracy, but keeps a reasonable association with the physics, would be insightful. This chapter discusses such a model. It comprises two parallel transmission mechanisms: direct transmission through a succession of trees, which is modeled by a simple linear transmission line; and transmission across the forest top, which is modeled by simplified multiple-edge diffraction. The model is examined using recently-published experiments over a long path-length. It is demonstrated that this two-mechanism model can provide an accurate fit to the dual-slope profile of through forest propagation over a long distance which is not possible with the RET model.

Chapter 5 presents antenna design for tubular platforms. First, antenna designs for the case of mounting on an electrically small conducting cylinder are studied using numerical and physical experiments. The work is relevant to mounting compact antennas on structures such as masts, drones, and bicycles. Then the chapter addresses waveguide arrays comprising slot radiators cut into hollow, single mode, standing wave structures. Their impedance bandwidth and compactness are limited by, for example, the waveguide cut-off. By using the TEM mode of a coaxial waveguide, a more compact structure with a larger bandwidth is possible at the expense of an increased ohmic loss in the waveguide. Using this basis, a
new type of slotted coaxial waveguide antenna is presented. A compact waveguide array means that the element and array aspects are not independent, and with no accurate transmission line models for the structure, the design is from parametric study by simulation. Two basic designs are demonstrated: the first with no constraint on the azimuthal extent of the slots, and the second where some mechanical strength of the waveguide is maintained by constraining the azimuthal extent of the slots to be 90° or less. The parametric studies provide insight for realizing the high bandwidth, and measurements of prototypes confirm the performance. These antennas are useful for relatively thin cylindrical structures such as found on bicycles, drones, and masts. There are many aspects to the designs presented in this chapter that pave ways for interesting new research avenues:

- Conducting measurement with the bicycle frame;
- Coaxial waveguide antennas on a ground plane, and effect of the groundplane on the pattern and bandwidth of the coaxial waveguide antenna;
- Varying cross section shapes of the coaxial waveguide antenna;
- An array configuration of the coaxial waveguide antenna;
- Exciting different polarizations.
Bibliography


