Characterisation of Multiple Antennas and Channel for Small Mobile Terminals

by

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TO MY FAMILY
Abstract

MIMO (Multiple Input Multiple Output) technology has been regarded as a practical approach to increase the wireless channel capacity and reliability. In this technology, the performance of multiple antennas has a significant effect on the MIMO channel capacity. Currently, the MIMO channel capacity is mostly evaluated without accounting for the practical aspects of multiple antennas. Though multiple antennas have been considered in a number of studies, only dipole or monopole antennas were used to evaluate the channel capacity. Practically, multiple dipoles and monopoles are no longer used for small mobile terminals because of their high profile and poor isolation performance. Implementing multiple antennas on a small mobile terminal remains a major engineering challenge.

The objectives of this thesis are to design practical antenna arrays for small mobile terminals running MIMO applications, and to obtain the corresponding channel characteristics in a realistic environment. The research is conducted in the following areas.

A novel PIFA (Planar Inverted-F Antenna) design and its array are proposed and studied. The single PIFA with a small ground plane constitutes a stand-alone structure. The ground plane, as small as the antenna, is located between the PIFA and the PCB (Printed Circuit Board) so that the PCB is no longer acting as a ground plane for the PIFA. The two PIFAs mounted on the PCB do not share the same ground plane. Consequently, the isolation performance between the two modified PIFAs is significantly improved. The characteristics of both single-element and dual-element PIFA in the 5.2GHz and 2.5GHz frequency bands are evaluated in simulation and measurement.

Following the study of multiple antennas, the concept of antenna diversity technology was introduced into the Galileo navigation system for the first time in the GAC (Galileo
Advanced Concept) project funded by GJU (Galileo Joint Undertaking). A statistical model was developed to analyse antenna diversity performance. The key parameters such as angle of arrival and cross-polar ratio of this model are specified for the Galileo system. The antenna diversity technique studied could be deployed for a small Galileo terminal.

A realistic MIMO channel model is established based on a ray tracing simulator, i.e. Wireless InSite. Practical aspects such as the antenna configurations and orientations, radiation patterns, and specific environments are included in the model to evaluate the MIMO channel capacity. The capacity of a MIMO system employing four ideal dipole antennas in an indoor environment is investigated numerically by using the realistic MIMO channel model above. The results agree with those of the IEEE 802.11 MIMO model. Furthermore, the model is used to study the channel capacity of a dual-element modified PIFA array on a mobile terminal. It has been demonstrated that this PIFA array is suitable for practical MIMO applications.
Acknowledgments

Over the past three years of my PhD studies, I have been lucky to study and work with some of the most renowned researchers in the Department of Electronic Engineering at Queen Mary, University of London. When thinking back of my first day in London and the achievement of today’s work, I am beholden to a long list of research staffs, friends and my parents. First of all, I express my deep and sincere thanks to my supervisor, Professor Xiaodong Chen for his guidance, support and encouragement over the past three years. His office door has always been opened, both literally and metaphorically, and I could always talk to him no matter how busy he was. I can only feel flattered for the confidence he has always shown in me. Fortunately, I have benefited from his extraordinary motivation, great intuition and technical insight. I just hope my thinking and working attitudes have been shaped according to such outstanding qualities. I would also like to express my warm and sincere thanks to Professor Clive G. Parini for his valuable suggestions and encouragement during the course of my PhD study.

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<tr>
<td>3D</td>
<td>Three-Dimensional</td>
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<tr>
<td>3G</td>
<td>Third Generation</td>
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<td>3GPP</td>
<td>Third Generation Partnership Project</td>
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<td>4G</td>
<td>Fourth-Generation</td>
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<tr>
<td>A-GPS</td>
<td>Assisted-GPS</td>
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<tr>
<td>AOA</td>
<td>Angle of Arrival</td>
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<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
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<tr>
<td>BLAST</td>
<td>Bell Labs Layered Space Time</td>
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<td>BT</td>
<td>Base Station</td>
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<tr>
<td>CA</td>
<td>Circular Array</td>
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<tr>
<td>CDF</td>
<td>Cumulative Distribution Function</td>
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<tr>
<td>DECT</td>
<td>Digital Enhanced Cordless Telecommunications</td>
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<tr>
<td>DF</td>
<td>Degradation Factor</td>
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<tr>
<td>DLR</td>
<td>German Aerospace Centre</td>
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<tr>
<td>DR</td>
<td>Dielectric Resonator</td>
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<td>EU</td>
<td>European Commission</td>
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<tr>
<td>EVD</td>
<td>Eigen Value Decomposition</td>
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<td>GAC</td>
<td>Galileo Advanced Concept</td>
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<td>Abbreviation</td>
<td>Description</td>
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<td>GJU</td>
<td>Galileo Joint Undertaking</td>
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<td>GO</td>
<td>Geometrical Optics</td>
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<td>GPS</td>
<td>Global Positioning System</td>
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<td>GSM</td>
<td>Global System for Mobile Communications</td>
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<td>GTD</td>
<td>Geometrical Theory of Diffraction</td>
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<tr>
<td>IEEE</td>
<td>Institute of Electrical and Electronics Engineers</td>
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<td>IFA</td>
<td>Inverted-F Antenna</td>
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<td>iid</td>
<td>independent and identically distributed</td>
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<td>ILA</td>
<td>Inverted-L Antenna</td>
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<td>IP</td>
<td>Internet Protocol</td>
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<td>IST</td>
<td>Information Society Technologies</td>
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<tr>
<td>LTE</td>
<td>Long Term Evolution</td>
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<tr>
<td>MIMO</td>
<td>Multiple Input Multiple Output</td>
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<tr>
<td>MT</td>
<td>Mobile Terminal</td>
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<tr>
<td>NLOS</td>
<td>Non-line-of-sight</td>
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<td>NMHA</td>
<td>Normal Mode Helix Antenna</td>
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<tr>
<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplexing</td>
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<tr>
<td>PCS</td>
<td>Personal Communication Services</td>
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<td>PDA</td>
<td>Personal Digital Assistant</td>
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<td>PDC</td>
<td>Personal Digital Communications</td>
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<td>PIFA</td>
<td>Planar Inverted-F Antenna</td>
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<td>QMUL</td>
<td>Queen Mary, University of London</td>
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<td>RF</td>
<td>Radio Frequency</td>
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<td>RHCP</td>
<td>Right Hand Circular Polarisation</td>
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<td>RT</td>
<td>Ray Tracing</td>
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<td>Rx</td>
<td>Receiver</td>
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<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<tr>
<td>SBR</td>
<td>Shooting and Bouncing Ray</td>
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<tr>
<td>SC</td>
<td>Selection Combiner</td>
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<tr>
<td>SIMO</td>
<td>Single Input Multiple Output</td>
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<tr>
<td>SISO</td>
<td>Single Input Single Output</td>
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<tr>
<td>SNR</td>
<td>Signal-to-noise ratio</td>
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<td>SVD</td>
<td>Singular Value Decomposition</td>
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<tr>
<td>Tx</td>
<td>Transmitter</td>
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<tr>
<td>ULA</td>
<td>Uniform Linear Array</td>
</tr>
<tr>
<td>UTD</td>
<td>Uniform Theory of Diffraction</td>
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<tr>
<td>WiMAX</td>
<td>Worldwide Interoperability for Microwave Access</td>
</tr>
<tr>
<td>WLAN</td>
<td>Wireless Local Area Network</td>
</tr>
<tr>
<td>WP</td>
<td>Work Package</td>
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<tr>
<td>XPR</td>
<td>Cross-polar ratio</td>
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Chapter 1

Introduction

The progress of communication technologies over the last century has been spectacular. In particular, the communication revolution that began with the invention of telephone by Bell continued with the invention of radio by Marconi. It was elevated to a whole new level by the introduction of other new technologies such as satellite and cellular networks. This revolution continues today with the widespread use of the Internet for data transfer and electronic correspondence as well as voice and video transmission. Recent technological advancements in science and engineering have blurred the traditional boundaries between previously different services such as the long distance telephone, radio and television broadcasting, and data transmission through the Internet. The number of Internet radio and television stations have increased significantly and continue to increase at a fast pace. Voice over IP technology has been successfully exploited by small corporations to provide a cheaper alternative to traditional long distance phone services offered by established providers. These technological advancements and the growth in consumer’s demands have tremendously increased the need for communication with high data rate, more reliability, power efficient and cheaper wireless services. However, the radio spectrum available for new wireless systems is scarce and expensive, and therefore it is crucial to increase the channel capacity and reliability of current wireless systems
Chapter 1. Introduction

without additional spectrum consumption.

1.1 Background of MIMO systems

In conventional wireless communication systems, only one antenna is used at the transmitter and one at the receiver. This is defined as a SISO (Single Input Single Output) system. In 1948, Shannon worked on the fundamental capacity limit of this system [1]. Having a channel bandwidth $W$ and SNR (Signal-to-noise ratio) over this bandwidth, Shannon stated that the maximum capacity $C$ of the SISO system is

$$C = W \log_2(1 + \text{SNR}) \quad \text{bits/s} \quad (1.1)$$

It is noted from the above equation (1.1) that the channel capacity can only be increased by an increase in bandwidth or signal power. On the one hand it is very expensive to occupy additional spectrum, and on the other hand, the signal power cannot be readily increased as the communication system is interference-limited.

Many coding techniques have been developed over the years; from the initial work of Hamming [2], through to Low Density Check codes [3] and Turbo codes [4], and they are able to achieve a channel capacity close to the Shannon limit. Despite this, no coding scheme could overcome the channel capacity limitation of a traditional SISO system.

During last decade, the Shannon capacity limit has been expanded by introducing multiple antennas at both the receiver and transmitter. The use of multiple antennas at both ends of a wireless communication system is called a MIMO (Multiple Input Multiple Output) system. Without increasing the bandwidth or signal power, the channel capacity of a MIMO system can be increased linearly with the number of the antennas under ideal conditions.

In 1987, Winters first reported the capacity enhancement of the multi-antenna fading
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channels applying multiple antennas at both the transmitter and receiver [5]. Lately, the potential capacity of the MIMO system was theoretically demonstrated by Foschini and Gans [6, 7], and Telatar [8]. Since then, MIMO systems have attracted a considerable interest in both the academic and industry worlds, and tremendous efforts have been put into the research and development of MIMO systems [9–16].

The MIMO system can increase not only the channel capacity, but also the reliability (Quality of Service) of the wireless system by exploiting different coding schemes. The typical coding schemes, e.g. space time coding, spatial multiplexing and the combination of both schemes, are discussed in the following sections.

1.1.1 Space time coding

Traditionally, multipath propagation is thought of as a source of fading that makes wireless links unreliable. Hence one needs to compensate against the fluctuations in fading channels to have a steady channel gain. One technique to mitigate the multipath fading is to employ the spatial diversity. MIMO systems are well placed to increase the effectiveness of the spatial diversity i.e. diversity gain. Each pair of transmit and receive antennas provides a signal path from the transmitter to the receiver. By sending signals that carry the same information through a number of different paths, multiple independently faded replicas of the data symbol can be obtained at the receiver. By averaging over these replicas, a more reliable reception is achieved. In this case, the idea of MIMO systems is to perform an averaging process over multiple path gains in order to increase its reliability: in a system with $M$ transmitting and $N$ receiving antennas, the maximum order of the diversity gain is $MN$. This can be achieved by exploiting the space time coding technique [9]. Space time coding technique assumes that there is no knowledge of the propagation channel at the transmitter. At the receiver, the structure of the space time codes is exploited to correct the errors [17].
1.1.2 Spatial multiplexing

The spatial multiplexing technique is another alternative method for exploiting MIMO channels. This is based on a different treatment of multipath propagation. In fact, multipath propagation can be beneficial to a MIMO system through increasing the degrees of freedom available for communication [6, 8]. If the path gains between individual transmit-receive antenna pairs fade independently, then the channel matrix is well-conditioned, in which case multiple parallel spatial channels are created. By transmitting independent information in parallel through spatial channels, the data rate can be increased. This effect is known as spatial multiplexing, and is particularly important in the high $SNR$ regime. Foschini reported that the capacity $C$ grows linearly with the number of the antennas for a given fixed transmitter power and bandwidth under ideal conditions in the MIMO systems [6].

1.1.3 Hybrid scheme

Both spatial multiplexing and space time coding offer different advantages for performance enhancement of wireless communication systems. However, depending on the propagation environment in which a system is operated, no single technology is capable of giving the best trade-off between data rate and reliability. Schemes combining the properties of more than one of these different processing techniques, therefore offer potential advantages for deployments in a real environment. The fundamental trade-off between diversity and multiplexing gains has been characterised in the point-to-point scenario in [18, 19].

1.2 Review of the State-of-Art

Research and development on MIMO systems have been advanced globally. In the USA, AT&T Bell laboratories were among the first to report MIMO propagation measurement
results \cite{20, 21}. Later, Lucent Technology presented MIMO propagation measurement results \cite{22} and showed a special interest in the key-hole concept with respect to the BLAST (Bell Labs Layered Space Time) coding \cite{23}. Stanford University, with the support of Lucent Technology in the hardware measurement set-up, has also reported some propagation measurement results \cite{24, 25}. Furthermore, Lucent Technologies has conducted measurements on 16×16 MIMO systems in an urban environment of Manhattan, New York \cite{26}. In the measurement, vertically and horizontally polarised slot antenna elements operating at 2.1GHz were used for both the transmitter and receiver. As shown in Figure 1.1, a laptop was used as the receiver terminal, which contains 16 antenna elements. For a 16×16 MIMO system with a 10dB system SNR, a capacity of 35bits/s/Hz were reported, which is 10 times the corresponding capacity of a 1×1 SISO system (3.5bits/s/Hz), and 4.8 times the corresponding capacity of a 1×16 SIMO (Single Input Multiple Output) system.

In the UK, Ofcom (Office of communication) has funded a MIMO technology research project: Antenna Array Technology and MIMO Systems \cite{27, 28}, which involved QMUL (Queen Mary, University of London), University of Bristol, University of York, BT Exact

Figure 1.1: A laptop used as a MIMO receiver terminal which contains 16 antenna elements \cite{26}.
Technologies, Toshiba Research Europe Limited and Antenova Ltd. The project has shown that MIMO systems can provide significant capacity gain when compared to a SISO system. However, the channel capacity is strongly dependent on the environments as well as the antenna configurations. In the project, QMUL and Antenova Ltd developed two different 4-element antenna arrays on PDAs (Personal Digital Assistants) operating at 5.2GHz. They are a dielectric loaded folded loop antenna array and a DR (Dielectric Resonator) antenna array [28].

In Europe, four projects regarding MIMO systems were supported by the EU (European Commission) in IST (Information Society Technologies) Framework 5 and 6, e.g., METRA (Multi-Element Transmit and Receive Antennas) project [29], I-METRA (Intelligent Multi-Element Transmit and Receive Antennas) project [30], SATURN (Smart Antenna Technology in Universal Roadband wireless Networks) project [31] and the WINNER (Wireless World Initiative New Radio) project [32]. In these projects, several partners have reported their MIMO propagation measurements namely, the University of Bristol [33, 34], France Telecom [35], Royal Institute of Technology (KTH) [36], IMST GmbH [37] and Aalborg University [38]. All these results have shown that the capacity and coverage of MIMO systems (e.g. $2 \times 2$, $4 \times 4$, etc.) are significant improvements over SISO and SIMO systems.

With the successful research on MIMO systems, the industry begins to standardise and commercialise the MIMO technology. Three popular standards involving the MIMO technology are WiMAX (Worldwide Interoperability for Microwave Access, IEEE 802.16 standard) [39], WLAN (Wireless Local Area Network, IEEE 802.11n standard) [40] and LTE (Long Term Evolution of 3GPP: Third Generation Partnership Project) [41]. Among them, the WLAN IEEE 802.11n is the earliest group to draft a MIMO standard. The IEEE 802.11n started in January 2004, and the industry came to a substantive agreement and approved the draft 2.0 of the 802.11n standard by March 2007.

Today, some MIMO products are readily available in the market for WiMAX and WLAN applications, although the standards are still in the drafting stage. With the
introduction of MIMO technology together with the OFDM (Orthogonal Frequency Division Multiplexing) modulation scheme, WiMAX and WLAN can take advantage of high speed broadband internet connections, accommodate bandwidth intensive applications such as video streaming and provide reliable broadband wireless coverage for a business or residential environment. As shown in Figure 1.2, NXP’s latest transceiver employs 2×2 MIMO technology in support of the mass deployment of mobile WiMAX, allowing for twice the data throughput and double the maximum speed of the company’s existing WiMAX transceiver solutions [42]. The WiMAX mobile terminals with UXA23466 chips will be available at the end of 2007. Airgo Networks has also produced the first MIMO chipset for WLAN in 2003. Later, QUALCOMM acquired Airgo Networks, and released the lastest MIMO chipset (AGN400) with IEEE 802.11n draft standard in 2007 [43]. Furthermore, Buffalo, Belkin and Linksys updated their wireless routers with the IEEE 802.11n draft standard this year, as shown in Figure 1.3.

1.3 Motivation

MIMO systems can potentially achieve an extraordinary bandwidth efficiency as long as the environment provides sufficient scattering and each receiving antenna obtains the independent transmission paths from each transmitting antenna. The channel conditions, i.e. the environments and the multiple antenna elements, determine the channel

Figure 1.2: NXP’s lastest transceiver: UXA23466 [42].
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Figure 1.3: Wireless routers: (a) Buffalo (Model: WZR-G300N) [44], (b) Belkin (Model: N1 Wireless Router) [45] and (c) Linksys (Model: WRVS4400N) [46].

capacity of MIMO systems. It is of great interest to characterise and model the MIMO radio channel for different environments in order to predict, simulate and design high performance communication systems. Also, the design of multiple antennas on small mobile terminals is crucial to practically implement MIMO systems. However, the channel modelling and the design of multiple antennas on a small mobile terminal for MIMO systems still remain challenging tasks.

Some research work has been conducted on the MIMO channel modelling. For instance, a MIMO model based on a geometric approach combined with physical arguments about the relevant propagation effects was proposed by Molisch [14]. Both the narrowband and the wideband statistical model for NLOS (Non-line-of-sight) MIMO propagation channels were presented in the following references [47–49]. The narrowband one-ring and the two-ring models were presented by Shiu et al [50] with part of the physical parameters. Gesbert et al [51] proposed a distributed scattering narrowband MIMO model to describe an outdoor MIMO propagation channel. These studies have provided important references for the design of MIMO systems. However, only ideal antenna arrays (most of them are uniform linear arrays), such as ideal dipoles or monopoles by assuming that the antennas radiate omni-directionally in the azimuth plane, have been considered in these MIMO channel models. When two or more dipoles/monopoles
are placed closely to each other on a mobile terminal, the radiation pattern of each
dipole/monopole is no longer omni-directional due to the coupling between them. Fur-
thermore, it is very impractical to implement a number of dipoles/monopoles on a small
mobile terminal. Therefore, there is a need to design an appropriate and practical
antenna array on small mobile terminals, and to account for them in MIMO channel
model to in order to obtain the channel characteristics in a realistic environment.

Diversity antennas are employed at base stations for current mobile communications
to mitigate the fading effects of a multipath environment. They have also been used
on handsets for the PDC (Personal Digital Communications) network system in Japan
[52]. Furthermore, research conducted at Queen Mary has shown that a diversity antenna
array consisting of four antenna elements operating at 5.2GHz on a PDA terminal for
MIMO systems can achieve high channel capacity [53]. Beside the diversity performance
of the multiple antennas for MIMO mobile terminals, the mutual coupling between the
antenna elements is a crucial parameter. Therefore, it is a very challenging task to design
multiple antennas at mobile terminals which can only have good diversity performance
but also undergo low mutual coupling, especially at the lower frequency bands (e.g.
2.5GHz frequency band). In this thesis, a dual-element diversity antenna array on a
handset terminal having good diversity performance and low mutual coupling between
the antenna elements is proposed and investigated so as to improve the reliability and
capacity of MIMO systems. Furthermore, a realistic MIMO channel model is established
based on a ray tracing simulator (i.e. Wireless InSite). Practical aspects such as the
antenna configurations and orientations, radiation patterns, and specific environments
are included into the model when simulating the MIMO channel capacity. Consequently,
the model is used to evaluate the capacity of the dual-element diversity array on a small
mobile terminal.

As additional work, the application of multiple antennas in satellite navigation sys-
tems has been investigated. Generally speaking, the present GPS (Global Positioning
System) does not function well in indoor or urban canyon multipath environments. It is
difficult to deal with the attenuated and multipath faded signals resulting from transmission around and through concrete and steel building structures. Multiple antennas on the user terminal can overcome the multipath fading, and therefore the concept of antenna diversity technology was introduced into the European Galileo navigation system in the GAC (Galileo Advanced Concept) project funded by GJU (Galileo Joint Undertaking) to improve the performance of the Galileo system in an indoor environment.

1.4 Outline of the thesis

Chapter 2 covers an introduction to the capacity of wireless communication channels, e.g. SISO channels and MIMO channels. It shows that these MIMO channels are able to generate higher capacity than SISO channels. This capacity is dependent on the characteristics of the propagation environment and multiple antennas.

Chapter 3 presents a review on small antennas for mobile terminals, and design concept of multiple antennas for small mobile terminals. The two essential requirements, e.g. having a low mutual coupling and good diversity performance, for multiple antennas are discussed in detail. Also, the application of multiple antennas in the Galileo navigation system is proposed to improve the satellite radio link.

Chapter 4 addresses a novel design of the PIFA (Planar Inverted-F Antenna) and its array operating at 5.2GHz and 2.5GHz. The characteristics such as return loss, isolation, radiation pattern and current distribution of the PIFAs are analysed using simulation and measurement. Furthermore, a dual-element helical antenna array for GPS/Galileo application is also investigated.

Chapter 5 evaluates the diversity performance of a dual-element PIFA array operating at 2.5GHz. The calculated diversity gain is compared to that measured. Also, a RT (Ray Tracing) MIMO channel model based on a commercial package (Wireless InSite) is developed and validated. The capacity of the dual-element PIFA array operating at
2.5GHz in an indoor environment is investigated using the RT MIMO channel model.

Chapter 6 describes the details about the application of multiple antennas in the GAC project. The propagation factors of the diversity system for Galileo/GPS system is characterised. Based on the characterised propagation factors, the diversity performance of the dual-helical antenna array for Galileo/GPS systems is evaluated.

Chapter 7 presents the conclusion and some thoughts for future work.

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Chapter 1. Introduction


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Chapter 2

MIMO systems

This chapter provides an introduction to the capacity of wireless communication channels. SISO channels which include AWGN (Additive White Gaussian Noise) and Rayleigh fading channels are considered first. Following the definitions of impulse response and outage capacity, MIMO channels and their capacity are explained. It is shown that these MIMO channels generate higher capacity by supporting parallel data streams. This capacity is dependent on the characteristics of the propagation environment in which the multiple antennas are employed.

2.1 SISO channels

A communication channel is the medium for transmitting a signal from a transmitter to a receiver, and vice versa [1]. The channel describes the various effects between the transmitter and the receiver based on measurements and mathematical models. Channel modelling investigates the influence of various factors on the channel capacity. It is generally divided into two categories, i.e. a deterministic physical model in terms of electromagnetic waves and a stochastic model in telecommunication engineering assuming different channel behaviour using different probability models [2]. In this section, the
capacity of well known stochastic models for a single-user SISO channel either with an
AWGN or with a Rayleigh fading is discussed.

2.1.1 AWGN channel

A single AWGN channel with bandwidth $W$ and received mean power $P$, is taken as an
equation. The capacity of the channel can be expressed as $\left[3\right]$

$$C = W \log_2 \left(1 + \frac{P}{WN_0}\right) \text{ bits/s}$$

where $N_0$ is the noise power spectral density.

The channel capacity can be increased by increasing the mean SNR ($\xi = P/(WN_0)$),
or by increasing the channel’s bandwidth $W$. Figure 2.1 shows that the channel capac-
ity increases monotonically with increasing SNR $\xi$. When the bandwidth $W$ tends to
infinity, channel capacity approaches the asymptotic value $\left[4\right]$

$$C|_{W=\infty} = \frac{P}{N_0} \log_2 e \text{ bits/s}$$

This occurs because the total noise power in the receiver increases with $W$, hence the
signal to noise ratio $\xi$ is correspondingly reduced. For a system with a fixed bandwidth,
channel capacity can only be increased by using higher transmitting power in order to
increase the mean signal to noise ratio at the receiver. Therefore, for a constant $W$
and $\xi \gg 1$, an increase of only 1bit/s in capacity requires a 3dB increase in transmit
power $\left[4\right]$.

2.1.2 Fading channel

For wireless communication channels, multipath propagation results in temporal, spatial
or frequency selective fading which are modelled as a Rayleigh distribution $\left[3, 5\right]$, and
degrades the performance of a SISO system [4]. The expression for channel capacity, equation (2.1), must be modified to include a randomly varying transmission coefficient, also defined as impulse response in literature [3]. The instantaneous capacity of a narrowband fading SISO channel can hence be expressed as [6]

\[ C = W \log_2(1 + |h|^2 \xi) \quad \text{bits/s} \]  \hfill (2.3)
Chapter 2. MIMO systems

where the impulse response $h$ is the complex transmission coefficient between the transmit and receive antennas, normalised to unit mean power,

$$E\{|h|^2\} = 1$$  \hspace{1cm} (2.4)

2.2 Impulse response of a multipath fading channel

As the impulse response is introduced to the instantaneous capacity of a fading SISO channel, it is essential to understand its principle. Impulse response contains all information necessary to simulate or analyse any type of radio transmission through the channel. The impulse response is a useful channel characterisation, since it can be used to predict and compare the performance of many different mobile communication systems and transmission bandwidths for a particular mobile channel condition.

In Figure 2.2, the Rx (Receiver) moves along the ground at some constant velocity $v$. For a fixed position $d$, the channel between the Tx (Transmitter) and the receiver can be modelled as a linear, time-invariant system. However, due to the different multipath waves which have propagation delays that vary over different spatial locations of the receiver, the impulse response of the linear, time-invariant channel should be a function of the position of the receiver. That is, the channel impulse response can be expressed as $h(d, t)$. Let $s(t)$ represent the transmitted signal, the received signal $y(d, t)$ at position

![Figure 2.2: The mobile radio channel as a function of time and space.](image)
$d$ can hence be expressed as a convolution of $s(t)$ with $h(d, t)$ \[3\]

$$
y(d, t) = s(t) \otimes h(d, t) = \int_{-\infty}^{+\infty} s(\tau) h(d, t - \tau) d\tau \tag{2.5}
$$

For a system, $h(d, t) = 0$ for time $t < 0$, thus equation (2.5) is reduced to

$$
y(d, t) = \int_{-\infty}^{t} s(\tau) h(d, t - \tau) d\tau = \int_{0}^{t} s(\tau) h(d, t - \tau) d\tau \tag{2.6}
$$

Since the receiver moves along the ground at a constant velocity $v$, the position of the receiver can be expressed as $d = vt$. And $y(vt, t)$ is a function of $t$. Therefore, equation (2.6) can be expressed as

$$
y(t) = \int_{0}^{t} s(\tau) h(vt, t - \tau) d\tau = s(t) \otimes h(vt, t) = s(t) \otimes h(d, t) \tag{2.7}
$$

From equation (2.7), it is clear that the mobile channel can be modelled as a linear time varying channel, where the channel changes with time and distance.

Since $v$ is assumed to be constant over a short time or distance interval, let $s(t)$ be the transmitted signal, $y(t)$ be the received waveform, and $h(t, \tau)$ be the impulse response of the time varying multipath radio channel. The impulse response $h(t, \tau)$ completely characterises the channel and is a function of both $t$ and $\tau$. The variable $t$ represents the time variations due to motion, whereas $\tau$ represents the channel multipath delay for a fixed value of $t$. The received signal $y(t)$ can be expressed as a convolution of the transmitted signal $s(t)$ with the channel impulse response

$$
y(t) = \int_{-\infty}^{+\infty} s(\tau) h(t, \tau) d\tau = s(t) \otimes h(t, \tau) \tag{2.8}
$$
2.3 Outage capacity

The concept of the Shannon capacity is not applicable to fading channels, since it is defined as the maximum capacity that a channel can support with negligible probability of error [6]. In the limit of Rayleigh fading, this cannot be guaranteed for any capacity greater than zero despite the mean capacity being a finite positive value. It is common in the analysis of fading channels to define a quantity known as the outage capacity [7].

An outage capacity of a channel is defined as a channel capacity that can be guaranteed for a certain percentage of realisations. Therefore, if an ‘outage’ of 10% were to be specified, the outage capacity would be a channel capacity which could be achieved for 90% of channel realisations, as shown in Figure 2.3.

![CDF (Cumulative Distribution Function) of the capacity for a 4x4 MIMO channel with a SNR of 10dB, the MIMO channel is explained in next section.](image)

Figure 2.3: CDF (Cumulative Distribution Function) of the capacity for a 4x4 MIMO channel with a SNR of 10dB, the MIMO channel is explained in next section.

2.4 MIMO channel structure and capacity

As discussed earlier, the multipath fading degrades the performance of wireless channels for a SISO system. However, the MIMO system could exploit the multipath fading to increase the capacity. The MIMO concept is defined as a radio link with \( M \) elements at
the BS (Base Station) and \( N \) elements at the MT (Mobile Terminal) through multipath fading channels. For the uplink, the Tx is at the MT and the Rx at the BS. For the downlink, the roles are reversed, as shown in Figure 2.4. The received signal vector \( Y(t) \) at the BS antenna array is defined by

\[
Y(t) = [y_1(t), y_2(t), \ldots, y_M(t)]^T
\]

(2.9)

where \( y_M(t) \) is the signal at the \( M \)th antenna element and \([\cdot]^T\) denotes the transpose operation. Similarly, the transmitted signals at the MT, \( s_N(t) \) define the vector \( S(t) \)

\[
S(t) = [s_1(t), s_2(t), \ldots, s_N(t)]^T
\]

(2.10)

A usage of such a channel can be described by the following matrix equation, using the relation of the vector \( Y(t) \) to \( S(t) \),

\[
Y(t) = H(t)S(t) + n(t)
\]

(2.11)

where \( n(t) \) is the AWGN and \( H(t) \in \mathbb{C}^{MN} \) is the instantaneous narrowband MIMO radio channel matrix. \( H(t) \) describes the connections between the MT and the BS. The

---

**Figure 2.4:** Two antenna arrays in a scattering environment. Representation of an uplink scenario.
MIMO channel matrix is expressed as

\[
H(t) = \begin{bmatrix}
h_{11} & h_{12} & \cdots & h_{1N} \\
h_{21} & h_{22} & \cdots & h_{2N} \\
\vdots & \vdots & \ddots & \vdots \\
h_{M1} & h_{M2} & \cdots & h_{MN}
\end{bmatrix}
\] (2.12)

where \( h_{MN} \) is the complex narrowband transmission coefficient from element \( n \) at the MT (Tx) to element \( m \) at the BS (Rx).

### 2.4.1 Orthogonal parallel channels

One of the motivations for deploying MIMO technology in wireless systems is the possibility of achieving orthogonal subchannels between the two ends of the path in a rich scattering environment and to increase the offered capacity. Mathematically, the number of independent subchannels between two terminals can be estimated by using the SVD (Singular Value Decomposition) of the matrix \( H \) or the EVD (Eigen Value Decomposition) of the instantaneous correlation matrix \( R \) defined by \([8, 9]\)

\[
R = H^\dagger H \quad \text{or} \quad R = HH^\dagger
\] (2.13)

where \( H^\dagger \) denotes the Hermitian transpose of \( H \).

The derivation of the parallel independent channels is summarised in Figure 2.5, where \( U \) and \( V \) are the diagonal unitary matrices, while \( u \) and \( v \) are the left and right singular vectors, respectively \([10]\). There exists an important relationship between the SVD of \( H \) and the EVD of \( R \) such that \( \sigma_k^2 = \lambda_k \), where \( \sigma_k \) is the \( k \)th singular value, \( \lambda_k \) is the \( k \)th eigenvalue and \( K \) is the maximum number of eigenvalue defined as \( K = \text{Rank}(R) \leq \min(M, N) \). The functions \( \text{Rank}() \) and \( \text{min}() \) return the rank of a matrix and the minimum value of the arguments. Figure 2.5 illustrates the whole process of eigenanalysis for a MIMO channel system.
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Figure 2.5: The process of orthogonal parallel channels for the MIMO system [10].

An example of orthogonal parallel channels is illustrated in Figure 2.6 for a 4×4 MIMO antenna topology. In such a configuration, a MN=16 radio link is created, but only 4 orthogonal subchannels (eigenvalues) are available. The difference in the thickness of the lines emphasises the difference in gain of the parallel subchannels such that $\lambda_1 \geq \lambda_2 \geq \lambda_3 \cdots \geq \lambda_k \geq 0$. These eigenvalues can be used to calculate the MIMO channel capacity as discussed in the following sections.

Figure 2.6: Illustration of parallel subchannels for a 4×4 MIMO topology.
2.4.2 MIMO channel capacity

Assuming that there is no channel state information at a transmitter and that the total transmit power is equally allocated to all $n$ antennas, the channel capacity of a narrow-band $(n, n)$ MIMO system has been derived by Foschini as [7]

$$C(\xi) = \log_2 \det \left[ I_n + \frac{\xi}{n} \cdot HH^\dagger \right] \quad \text{bits/s/Hz} \quad (2.14)$$

where $\xi$ donates the mean SNR at each receiver, $det$ denotes the determinant, and $I_n$ denotes the identity matrix. $H$ is the normalised $n \times n$ channel matrix, which is expressed as $\sum_{j=1}^{n}|h_{ij}|^2 = 1$.

Applying singular value decomposition to $H$, we can obtain a pair of unitary matrices $U$ and $V$ such that

$$H = UDV^\dagger \quad (2.15)$$

where $U \in \mathbb{C}^{n \times n}$ and $V \in \mathbb{C}^{n \times n}$. $D$ is diagonal, and its entries are the non-negative square roots of the eigenvalues of $HH^\dagger$, $\lambda_i$ for $i = 1, 2, \ldots, n$, equation (2.14) becomes

$$C(\xi) = \log_2 \det \left[ I_n + \frac{\xi}{n} \cdot D^2 \right] = \sum_{i=1}^{n} \log_2[1 + \frac{\xi}{n} \cdot \lambda_i] \quad \text{bits/s/Hz} \quad (2.16)$$

According to equation (2.16), when the channel exhibits rich scattering, and its variations can be accurately tracked, the MIMO system can achieve almost $\min(n_T, n_R)$ more bits per hertz for every 3dB increase in SNR. This is an enormous improvement over the single antenna case which achieves one additional bit per hertz for every 3dB increase in SNR [11].
2.5 Physical understanding of MIMO channel

After the introduction to MIMO channels, it is essential to understand how the antenna array and environment affect the MIMO channel capacity. As discussed earlier, the channel matrix $H$ contains the information of the environment and antenna array. The following two examples demonstrate how the channel matrix $H$ characterises the channel in MIMO systems [12].

2.5.1 Antenna spacing effect

Consider $N_{tx} = N_{rx} = 2$ antennas in free space communicating at 2.5GHz as arranged in Figure 2.7. The antenna separation is denoted by $d$ and the distance between transmitter and receiver by $R$. For simplicity, two ideal antennas ($\Gamma_s$) with a single plane wave travelling along a line of sight that connects transmitters and receivers in free space are assumed. The propagation characteristics in this case can be modelled by [12]

$$H_{\text{Norm}} = \frac{1}{\sqrt{2}} H = \frac{1}{\sqrt{2}} \begin{pmatrix} \Gamma_s(R)e^{-j2\pi R/\lambda} & \Gamma_s(\hat{R})e^{-j2\pi \hat{R}/\lambda} \\ \Gamma_s(\hat{R})e^{-j2\pi R/\lambda} & \Gamma_s(R)e^{-j2\pi R/\lambda} \end{pmatrix}$$

where $\lambda = c/f_c = 3e8/2.5e9 = 120mm$ is the wavelength and $R = \sqrt{d^2 + \hat{R}^2}$. Due to the very small difference between $R$ and $\hat{R}$, $\Gamma_s(R) \simeq \Gamma_s(\hat{R}) \simeq 1$ is assumed. Consider $R=250m$ and $d=3m$. The eigenvalues of $H_{\text{Norm}}H_{\text{Norm}}^\dagger$ are $\lambda_1 = 1.59$ and $\lambda_2 = 0.41$. For more details about this analysis, refer to Appendix B. Hence, the capacity formula gives
\[ \log_2 \left( 1 + \frac{\text{SNR}}{2} \lambda_1 \right) + \log_2 \left( 1 + \frac{\text{SNR}}{2} \lambda_2 \right) \text{ bits/s/Hz} \]  

(2.18)

It should be noted that a single transmitter and single receiver system would have the capacity,

\[ \log_2 (1 + \text{SNR}) \text{ bits/s/Hz} \]  

(2.19)

It is instructive to see that when \( d = 0.5 \text{m} \), \( \lambda_1 = 1.9997 \), and \( \lambda_2 = 0.0003 \), the contribution of the second term \( \log_2 (1 + \frac{\text{SNR}}{2} \lambda_2) \) to the channel capacity becomes very small, and the capacities of the two antennas and the SISO system become equivalent, as shown in Figure 2.8. The antenna spacing has a significant effect on MIMO systems.

![Figure 2.8: Capacity of a SISO system compared with those of the MIMO systems in the example when \( d = 3 \text{m} \) and \( d = 0.5 \text{m} \), respectively.](image)

2.5.2 Environment effect

The previous example shows that when the antennas are closely spaced, channel capacity collapses to that of a SISO antenna. In the following example, an explanation of how multipath reflections remove this restriction is demonstrated. Figure 2.9 is the same as Figure 2.7 but with the addition of an ideally perfect reflector.
The reflector creates the presence of two plane waves travelling from the two transmitters to the reflector, and two plane waves travelling from the reflector to the receivers. Assuming that the reflected waves have a reflection attenuation equal to 60% of the attenuations of the nonreflection waves, the propagation matrix $H_R$ can be expressed as

$$H_R = H + 0.6 \begin{pmatrix} e^{-j \frac{2\pi}{\lambda} (P_1 + P_3)} & e^{-j \frac{2\pi}{\lambda} (P_2 + P_4)} \\ e^{-j \frac{2\pi}{\lambda} (P_2 + P_3)} & e^{-j \frac{2\pi}{\lambda} (P_1 + P_4)} \end{pmatrix}$$  \hspace{1cm} (2.20)$$

where $H$ is given in equation (2.17) and $P_1 = \sqrt{R_2^2 + dt^2}$, $P_2 = \sqrt{R_2^2 + (dt + d)^2}$, $P_3 = \sqrt{R_1^2 + dt^2}$, and $P_4 = \sqrt{R_1^2 + (dt + d)^2}$. Consider $R=250$ m and $d=0.5$ m, $R_1 = R_2 = R/2$, and $dt=15$ m. $H_R$ is normalised, so that $\sum_{j=1}^{2} |h'_{ij}|^2 = 1$, then the eigenvalue $\lambda_1 = 1.54$ and $\lambda_2 = 0.46$. For more details about this analysis, refer to Appendix B. It can be seen from Figure 2.10 that the capacity of this two antenna system spaced at only $d=0.5$ m with reflections is significantly higher than that of a SISO system. Therefore, the multipath of the environment increases the capacity of MIMO systems.

Although the two examples above describe a simplified MIMO system, they capture the important concepts of MIMO systems. The first example demonstrates the effect of spacing between antenna elements in MIMO systems. The closer the spacing between the antenna elements is the lower the capacity available. The second example shows that a multipath environment is an important condition to achieve high MIMO channel
capacity. As more multipaths are created in an environment, a higher channel capacity is achieved.

2.6 Summary

This chapter addressed the capacity of wireless channels stemmed from the SISO system to the MIMO system. It has been shown that the MIMO system can increase the channel capacity significantly without increasing the bandwidth and transmission power when compared to the SISO system.

It has been observed that the MIMO channel capacity is affected by both the spacing between antenna elements and the multipath environment. The spacing between antenna elements has a significant effect on MIMO channel capacity. From the MIMO channel point of view, the larger the spacing between antenna elements is, the higher capacity available. From the antenna design point of view, multiple antennas are easy to be employed at the base station because there is enough space. However, it would not be easy to do so at a mobile terminal since the size of the terminal is limited. Mobile
terminal technology in the last few years shows a dramatic reduction in the size and weight of terminals.

On the other hand, it has been shown that multipath environment plays an important role on the MIMO channel capacity. A large number of multipaths created in the environment increases the MIMO channel capacity. It also has been demonstrated that there is a combined effect of the spacing between multiple antennas and the multipath environments. Multiple antennas can benefit from the multipath environment by its different configurations, and therefore the requirements of multiple antennas on mobile terminals are essential for MIMO systems, and these are addressed in Chapter 3.

References


Chapter 3

Small and diversity antennas

As described in Chapter 2, MIMO systems require multiple antennas at both the transmitter and receiver to increase the channel capacity. It is feasible to implement multiple antennas at a base station because there is no strict limitation on the size of the base station. However, there is not much room available for multiple antennas on a small mobile terminal such as a handset or PDA. Therefore, a multiple-element antenna array should be small in order to be embedded into the small mobile terminal. It also should meet some additional requirements, e.g. good isolation and diversity performance for multiple antennas besides the usual requirements of a conventional single antenna such as compact structure, light weight, low profile and robustness. In this chapter, the development and miniaturisation of small antennas are discussed. The requirements for designing multiple antennas on a small mobile terminal are presented along with those of the conventional single antenna. Furthermore, the diversity technology of multiple antennas is explained in detail.
3.1 Development of antennas on small mobile terminals

The rapid growth of the mobile terminal market has been observed globally. The number of mobile phone users worldwide has exceeded 2.5 billions users by the end of 2006 according to the figures from Wireless Intelligence Group of the GSM (Global System for Mobile Communications) Association [1]. One of the trends in mobile terminal technology in the past few years is the reduction in sizes and weights of mobile terminals driven by the development of modern integrated circuit technology and the preference of the users. This remarkable reduction in the terminal’s size has initiated a rapid evolution of the embedded antennas design for mobile terminals. The antennas are required to be small, and yet their optimum performances need to be maintained.

Several reviews of antennas for small mobile terminals have been reported in the past few years [2–6]. The important issues raised by these reviews are summarised in this section. Over the past two decades, the size and weight of the small mobile terminal have been reduced, as shown in Figure 3.1, from the single frequency band voice service only portable cellular phone in 1983 (about 600cc in volume and approximately 850g in weight) to the current multi-band multi-functional small handheld mobile terminal (less than 60cc in volume and a weight of less than 80g). The antennas used for small mobile terminals have evolved from monopole to PIFA, and also other types of antennas such as NMHA (Normal Mode Helix Antenna), microstrip patch, meander line and DR antennas also play important roles in small mobile terminals [2, 3, 7–11]. In this section, the commonly used antennas on mobile terminals such as monopole, NMHA and PIFA are explained in details.

3.1.1 Monopole

Marconi invented the monopole antennas by using the image method in 1896. One pole of a half-wavelength dipole antenna is buried in the ground plane (conducting case) to obtain the quarter wavelength monopole antenna as shown in Figure 3.2. In the early
Figure 3.1: Small mobile terminals: (a) the first portable cellular phone (Dyna-TAC) produced by Motorola in 1983 [12] and (b) a multi-band and multi-functional small handset (T68i) produced by Sony-ericsson in 1999 [13].

1980’s, the quarter wavelength monopole antenna became the most popular antenna for mobile terminals due to its simple structure and relatively short length. However, Fujimoto and Hirasawa discovered that a quarter wavelength monopole antenna spread much larger currents on the terminal case compared to a half-wavelength monopole antenna [7]. For a half-wavelength monopole antenna, the maximum current amplitude occurs around the centre of the monopole. Consequently, the current amplitude around the feed point between the monopole and the terminal case is small, therefore little current flows to the terminal case. In contrast, for a quarter wavelength monopole the maximum current amplitude occurs around feed point and large currents leak into the terminal case. This suggests that the size of the terminal case changes the characteristics of the radiation pattern. In particular, the radiation pattern distortion and the hand-held effect is small with a half-wavelength monopole antenna on a mobile terminal. In practice, the impedance at the feed point of a half-wavelength monopole becomes very
high, and it is difficult to achieve a matching to a feeding cable. Therefore, the 3/8 or 5/8 wavelength monopole antennas has been employed in practice for the mobile terminals [3].

3.1.2 Helix

The monopole antenna can be shortened from quarter wavelength to 4-15% wavelength by introducing a distributed inductive loading, conventionally named NMHA [2]. The NMHA is made of a spiral enclosed in plastic or rubber. It consists of a wire wound $N$ turns around a cylinder in diameter $D$ with spacing $S$, and it is fed against a ground plane at one end of the structure by a coaxial cable, as shown in Figure 3.3. The total length of the helix is $L = NS$ while the total length of the wire is $L_w = NL_0 = N\sqrt{S^2 + C^2}$ where $L_0 = \sqrt{S^2 + C^2}$ is the length of the wire between each turn, and $C = \pi D$ is the circumference of the helix. Another important parameter is the pitch angle $\alpha$ (i.e. the angle formed by a line tangent to the helix wire and a plane perpendicular to the helix axis) is defined by [14]
\[ \alpha = \tan^{-1} \left( \frac{S}{\pi D} \right) = \tan^{-1} \left( \frac{S}{C} \right) \] (3.1)

The helical antenna is a hybrid of two simple radiating elements, the dipole and loop antennas. It can operate in two principal modes: the normal (broadside) and the axial modes (endfire). A helix becomes a linear antenna (like monopole) when its diameter approaches zero or pitch angle goes to 90\(^0\). It is then known as the normal mode helical antenna (NMHA) as defined earlier. A helix of fixed diameter can be seen as a loop antenna when the spacing between the turns vanishes (\(\alpha = 0^0\)). NMHAs with radiation from the broadside of the antenna are used widely in a mobile handset, as mentioned earlier. The axial mode helix antennas are mostly used for satellite communication and GPS due to their high gain and endfire radiation patterns [15].

The NMHA was widely used in the commercial mobile terminals in the middle of 1990s due to its short length, ease of fabrication and low cost. Figure 3.4 shows a prototype of a dual band non-uniform NMHA.

![Figure 3.3: Geometrical configuration for a helix.](image-url)
Figure 3.4: A dual band non-uniform helix, it was invented in 1996 by Z. Ying (Ericsson), Patent number US-6212102, WO-9815028. It has reached the penetration of more than several hundreds million products worldwide [5].

3.1.3 Inverted-L Antenna (ILA) and Inverted-F Antenna (IFA)

The monopole antenna can also be shortened by a top loading. The simplest form of a top loaded antenna is invert-L antenna (ILA). As shown in Figure 3.5, it consists of a vertical element as a short monopole and a horizontal wire element attached at the end of the monopole. It can be seen that the height of the ILA is significantly reduced and essentially low profile when compared to the monopole.

Figure 3.5: Configuration for a wire inverted-L antenna (ILA).
Chapter 3. Small and diversity antennas

The ILA has an inherently low input impedance due to its low profile structure. It consists of a short monopole loaded with a long horizontal wire at the end of the monopole. The input impedance is nearly equal to that of the short monopole plus the reactance of the horizontal wire closely placed to the ground plane. In order to increase the input impedance, another inverted-L shaped element is attached at the end of the vertical element [7]. This modification of ILA is named as an inverted-F antenna (IFA) because its appearance is like a letter F facing the ground plane, as shown in Figure 3.6. Impedance matching of an IFA can be achieved by allocating the position of the feeding point without using any additional circuit.

![Figure 3.6: Configuration for a wire inverted-F antenna (IFA). The ILA with another inverted-L element attached to the vertical element.](image)

A drawback of an ILA or IFA consisting of thin wires, is the narrow impedance bandwidth, typically only one percent or less of the centre frequency. To widen the bandwidth, the wire element of IFA can be replaced by a planar element. This IFA with a planar element is known as planar inverted-F antenna (PIFA), as shown in Figure 3.7, where $LT$ is the length of the PIFA, $WS$ is the width of the PIFA and $W$ is the width of the shorting plate. The fundamental characteristics of the PIFA have been analysed and reported in [16]. The shorting plate is positioned at the corner of the planar element. The narrower the shorting plate width $W$ the lower the resonant frequency of the PIFA. In the case of $W = WS$, the resonant frequency $f$ of the PIFA can be determined from the equation below [16]

$$f = \frac{c}{4(LT + H)}$$  \hspace{1cm} (3.2)
and in the case of $W \approx 0$, it can be determined by

$$f = \frac{c}{4(LT + WS + H)} \quad (3.3)$$

where $c$ is the speed of light in free space. The variation of the resonant frequency can be explained by observing the current distribution on the surface of a planar element, as shown in Figure 3.8. As the width of the shorting plate becomes shorter, the current flow becomes longer. Consequently, the resonant frequency becomes lower, and the current is non-uniformly distributed. When $W$ equals to $WS$, the current is uniformly distributed, and the PIFA can be considered as a kind of short-circuit rectangular microstrip antenna [16].

Figure 3.7: Configuration for a planar inverted-F antenna (PIFA).

Figure 3.8: Variation of surface current flow on planar element for difference width of the shorting plate [16].
Further research on PIFA has been conducted to achieve wideband characteristics. Instead of increasing antenna height, one of the most effective approaches is to add a parasitic element at the open end of the antenna element [17–19]. As the requirements for multiple services for mobile communications systems increase, PIFAs have also been developed to operate at multi-band for the GSM, DCS (Digital Cellular System), DECT (Digital Enhanced Cordless Telecommunications), PCS (Personal Communication Services) and WLAN systems [20–23]. Figure 3.9 shows that PIFA can work at multi-band by cutting a L-shaped slot on the planar element [24, 25]. The planar element is divided into two parts by the slot. The upper part has a longer path for the current flowing which resonates at the low band. While the low part has a shorter path which resonates at the high band. Multi-band PIFAs have been widely used as built-in antennas by most of the mobile handset manufacturers. For example, the popular handset Sony Ericsson T68i as illustrated in Figure 3.1(b) used a multi-band PIFA to operate at GSM 900/1800/1900 and bluetooth band, as shown in Figure 3.10 [5, 26].

It has been reported that the PIFA has a large current flow on the ground plane [16]. Like monopole and helical antennas for mobile terminals, the ground plane of a PIFA has significant effects on the performance of the antenna. By optimising the ground plane, a PIFA can achieve up to 15% increase in bandwidth when compared to a PIFA not on a ground plane [3]. On the other hand, the performance of a PIFA such as return loss, gain and radiation patterns is degraded by the large currents flowing on the ground plane.

![Figure 3.9: Top view of a dual-band PIFA with a L-shape slot on the planar element [24, 25].](image-url)
due to the effect of adjacent objects such as the human hand and head. In addition, the PIFA has to be redesigned for different terminals since it is very dependent on the ground plane.

### 3.2 Requirements for multiple antennas on small mobile terminals

In MIMO systems, more than one antenna are implemented on a small mobile terminal as mentioned earlier. Having multiple antennas on a small mobile terminal while maintaining its performance remains a challenge. As described earlier, the following characteristics are required, a good impedance matching in the interest of bandwidth, omni-directional radiation patterns, small size, compact structure, light weight, low profile and robustness in designing a single conventional antenna for small mobile terminals. However, for multiple antennas on a small mobile terminal the mutual coupling effects and diversity performance are essential considerations.
3.2.1 Mutual coupling effects

Mutual coupling refers to the electromagnetic interactions between the elements of an antenna array. Some of the energy transmitted by a transmit antenna element is transferred to the other elements. Correspondingly, a portion of the energy in the incident field of a receive antenna element is transferred to the nearby elements. As a result, the feed current on each transmitting antenna in an antenna array does not only consist of the current when they are transmitting alone, but also includes the current induced by the other antenna elements in close proximity. The same argument also applies to the receiving elements of the array. Figure 3.11 illustrates the effect of mutual coupling between two dipoles by showing the variation of radiation patterns of Dipole 1 with antenna separation. The other dipole (Dipole 2) is terminated at characteristic impedance $Z_0 = 50\Omega$. It shows that the radiation patterns of Dipole 1 are strongly distorted when Dipole 2 is placed in proximity.

Due to this interchange of energy, the input impedance of a single antenna element changes under the influence of mutual coupling. With no mutual coupling, the input impedance of an antenna is equal to its self-impedance. However, with another antenna in close proximity, the input impedance of the antenna becomes dependent on both

![Figure 3.11: Radiation patterns of a single dipole alone and one dipole (Dipole 1) in a dual-element dipole array for different separations (d).](image)
self-impedance and the mutual impedance. This relation is expressed in the equation below [27]

\[ Z_{in} = Z_{11} + Z_{12}\left(\frac{I_2}{I_1}\right) \]  \hspace{1cm} (3.4)

where \( Z_{11} \) and \( Z_{12} \) refer to the self and mutual impedances respectively, and \( I_1 \) and \( I_2 \) refer to the currents flowing through the respective antennas.

In the case of multi-antenna applications such as diversity antenna array, it is required that the mutual coupling between multiple antennas should be minimised to maintain a high efficiency of the multiple antennas [28–30]. The amount of mutual coupling depends on the separation between antenna elements. It increases when antenna elements are closer to each other. It has been shown that for minimum or no mutual coupling in theory, the element separation between dipoles has to be at least half a wavelength [31]. Therefore, it is a major challenge to bring the two antennas closer than half-wavelength while keeping the mutual coupling levels very low.

### 3.2.2 Diversity technology

Diversity is a technique by which the multipath signals inherently present in a mobile communication channel, are combined to mitigate fading and improve the overall quality of the radio link. The basic principle is that the receiver combines more than one copy of the transmitted signal, where each copy is received through a different branch. Figure 3.12 illustrates two uncorrelated Rayleigh fading signals and the combined signal. If the two multipath fading signals are uncorrelated, it is rare that both will be in a deep null at the same time. It shows that the combined signal has a higher mean SNR at the output when compared to a single branch. The effectiveness of a diversity system is measured by a quantity known as the diversity gain, defined in subsection 3.3.3.

There are several ways of achieving independent fading paths in a radio system, such as frequency diversity, time diversity and antenna diversity. The focus of this thesis is on antenna diversity which can be regarded as a measure of effectiveness for multiple
Figure 3.12: Illustration of two individually received signals and one combined the two signals by diversity technique. The combined signal always has the highest signal level compared to the individual signals.

### 3.3 Antenna diversity techniques

Antenna diversity techniques use multiple antennas, also called diversity antennas, as shown in Figure 3.13. The most important issues in antenna diversity techniques are how multiple antennas are arranged to receive the independent multiple signals uncorrelatedly, and then how the received signals are combined efficiently. These aspects are addressed in this section.

#### 3.3.1 Classification of antenna diversity

Independent fading paths can be obtained by using two or more antennas that are physically separated from each other (spatial diversity), or in different polarisations (polarisation diversity) or have different radiation patterns (pattern or angle diversity). These different diversity approaches are typically used on the mobile terminals [32].
3.3.1.1 Spatial diversity

Spatial diversity utilises more than one antenna separated from each other in space to achieve independent fading paths. Due to the separation between two antennas, the relative phase of the multipath is significantly different at both antennas. In order to achieve sufficient de-correlation, the distance between two antennas usually has a minimum spacing depending on different fading environments. For example, the distance is greater than several wavelengths at the base station in an urban area [33]. Therefore, the spatial diversity was widely applied to the base station at first [34].

Research work on the spatial diversity on a mobile terminal was conducted further. By assuming that there is a uniform angle arrival around the azimuth of the mobile terminal and no elevation angle of arrival, the correlation between two antennas for a separation distance \(d\) can be obtained from the zero order Bessel function [35]

\[
\rho_{12} = J_0(\beta d)
\]  

where \(\beta\) is the phase constant. This equation shows that the minimum distance for negligible correlation (which is one of the measures of the diversity performance and is defined in section 3.3.4.1) between two antennas in space at the mobile terminal is 0.5
wavelength. However, this does not consider the mutual coupling between the antennas. Although reasonable diversity performance such as low correlation can be obtained with horizontal spacing as small as 0.1 wavelength reported in [2, 35], the mutual coupling between the antenna elements degrades the antenna efficiency.

### 3.3.1.2 Pattern (Angle) diversity

The independent paths coming from different directions can be picked up by different radiation patterns of different antennas. This is usually defined as pattern (or angle) diversity. The pattern diversity has been considered at the base station in some cases and compared with spatial diversity [36, 37].

Pattern diversity occurs at mobile terminals when two omni-directional antennas are closely spaced and interacting with each other due to the mutual coupling effect as mentioned earlier. The antennas act as parasitic elements to each other and change their patterns to allow signals to be picked up at different directions. Based on this concept, antennas with beam steering at the mobile terminals (which is realised by changing the feed point impedance of parasitic elements) have been developed [38–40]. This antenna diversity concept allows a much smaller antenna spacing (0.1 wavelength) and provides multiple antenna patterns while keeping low correlations. However, it has been shown that the pattern diversity come at the expense of antennas efficiency due to the mutual coupling [41].

### 3.3.1.3 Polarisation diversity

Polarisation diversity can be achieved when two or more differently polarised antennas are used to obtain the independent fading paths. Typically, antennas with orthogonal polarisation provide low levels of correlation with minimum or no spacing [42]. With the use of polarisation diversity, the size of the multiple-antenna structures can be reduced significantly. Numerous studies on polarisation diversity have been carried out at the base
stations, and they have been widely applied in practice to new base stations providing space and cost savings [43–45]. It has been reported that polarisation diversity integrated with spatial diversity obtains even better diversity performance [46]. The disadvantage of polarisation diversity at base station is 3dB power loss produced by splitting the transmitter power into two polarisations [42].

Polarisation diversity is an attractive option for the mobile terminal because of the reduced size in the multiple-antenna structures [32]. However, the cross-polarisation of a single antenna and the mutual coupling between multiple antennas must be low enough in order to maintain a good diversity performance [47].

Recently, spatial diversity, pattern diversity and polarisation diversity have been applied to MIMO systems. For instance, it has been reported that the MIMO capacity in an indoor environment is still relatively large when the four antennas are closely spaced (i.e, $d=0.2$ wavelength) [48]. With appropriate dissimilarity in the antenna patterns, large MIMO channel capacity can be achieved [49, 50]. Recent studies on the MIMO systems have sought to exploit the MIMO channels by using polarisation diversity [51–54]. These studies mostly ignore the mutual coupling effect between the multiple antennas. However, the performance of multiple antennas on a mobile terminal for MIMO systems not only rely upon a good diversity characteristic but on a low mutual coupling level as well [28–30].

### 3.3.2 Combining methods

After obtaining independent fading paths, the next important consideration in diversity technology is diversity combining. The signals received at each branch can be combined in different ways to mitigate the effects of fading. There are four typical diversity combining methods: switched combining, selection combining, equal gain combining and maximum ratio combining.
3.3.2.1 Selection combining

The selection combining technique monitors the instantaneous SNR at the output of $N$ branches by using $N$ processing modules, as shown in Figure 3.14. The branch with the highest output SNR at any time is selected as the active output signal.

![Figure 3.14: Illustration of selection combining for $N$ antenna elements.](image)

3.3.2.2 Switched combining

The switched combining technique is similar to the selection combining technique. But it only requires one processing module for $N$ branches, as shown in Figure 3.15. Other combining techniques require $N$ receivers for each of $N$ separate antennas. They monitor the received instantaneous signals level of each branch independently. The receiver is

![Figure 3.15: Illustration of switched combining for $N$ antenna elements.](image)
switched to another branch only when the SNR on the current branch is lower than a predefined threshold. The performance of switched combining is lower than that of selection combing since unused branches may have SNRs higher than the current branch.

### 3.3.2.3 Equal gain combining

Since both switched and selection combining techniques only use the signal from one of the branches as the output signal at any given time, the signal energy in the other branches is wasted. In order to improve on this, the signals from all branches can be combined. If this is done directly using the complex signals, their random real and imaginary components would combine incoherently. To obtain effective diversity and the largest available output signal power, the signals must be co-phased so that they add coherently. Before the outputs from each branch are added together, the signal in each branch must be multiplied by a complex phasor having a phase $\theta_i$ where $\theta_i$ is the phase corresponding to branch $i$ as shown in Figure 3.16. When this is achieved, all signals will have a zero phase relationship, and they are combined coherently.

![Figure 3.16: Illustration of equal gain combining for $N$ antenna elements.](image)
3.3.2.4 Maximum ratio combining

In the equal gain combining technique, all the branches may not have the same SNR. Sometimes one of the branches has a much lower SNR than the other branches. Due to the equal weighting in the sum, the overall output SNR may occasionally reduce to a low value. In order to maximise the SNR at the output, a weight $w_i$ is applied to each branch before all the signals are combined coherently, as shown in Figure 3.17. A branch with a lower SNR will be given a lower weighting.

![Figure 3.17: Illustration of maximum ratio combining for N antenna elements.](image_url)

3.3.3 Diversity gain

The effectiveness of diversity technology is usually assessed in terms of diversity gain. Diversity gain is defined as the improvement in the SNR of the signals at the output of the diversity combiner or switch, relative to the SNR from a single branch for a stated reliability. In this thesis, the Rayleigh channel, which is a multipath propagation environment, is assumed. The CDF (Cumulative Distribution Function) of a Rayleigh channel is given as $[35, 55]$

$$P(\gamma < \gamma_s/\xi) = 1 - e^{-\frac{\gamma_s}{\xi}}$$  \hspace{1cm} (3.6)
where $\xi$ is the mean SNR, $\gamma$ is the instantaneous SNR, $P(\gamma < \gamma_s/\xi)$ is the probability that the SNR will fall below the given threshold, $\gamma_s/\xi$. For a selection combiner with $N$ independent branches, assuming that the $N$ branches have independent signals (correlation equal to zero) and equal mean SNRs, the probability of all branches having a SNR below $\gamma_s$ is equivalent to the probability for a single branch raised to the power $N$ as

$$P(\gamma < \gamma_s/\xi) = (1 - e^{-\gamma_s/\xi})^N$$ (3.7)

where $N$ is the number of branches.

Equations (3.6) and (3.7) are plotted in Figure 3.18 to show the reduction of the probability of fading below a given threshold when increasing the number of branches, $N$. In this figure, diversity gain is also illustrated in terms of the increase in normalised SNR of a combined output compared to a single antenna. The normalised SNR is defined as $\gamma_s/\xi$, which is used in the following chapters. Here, the diversity gain is marked off where $P(\gamma < \gamma_s/\xi) = 1\%$ (i.e. 99 \% reliability). The figure shows that there are diversity gains of 10dB and 13dB for the two-branch and three-branch selection combiners, respectively.

Figure 3.18: CDFs of Rayleigh fading signals for a different number of diversity branches. Plotted based on equation (3.7).
For low instantaneous SNR, that is $\gamma \ll \xi$, equation (3.7) can be approximated by [56]

$$P(\gamma < \gamma_s/\xi)^N \approx \left(\frac{\gamma_s}{\xi}\right)^N$$  \hspace{1cm} (3.8)

Therefore, by re-arranging equation (3.8) the diversity gain for a 100% efficiency two branch selection combiner is 10dB with $P(\gamma < \gamma_s/\xi) = 1\%$.

### 3.3.4 Diversity performance analysis

As described earlier, the diversity gain is the measure of the effectiveness of a diversity system. A high diversity gain can be achieved when the received signals from two antennas have low correlations; the power levels of the signals received by the two antennas should not be too different in the diversity system in a multipath environment. The following section explains the relationship between correlation, branch power ratio and mean effective gain. The propagation environment factors are also discussed.

#### 3.3.4.1 Correlation

In order to obtain a high diversity gain, one of the conditions is ensuring a low correlation between the signals received in the branches of the diversity system. The correlation can be described by complex and envelope correlations [57]. The complex correlation $\rho_c$ is described as the complex correlation between two signal envelopes [58]. The instantaneous magnitudes and relative phases of the branch signals are used to calculate the complex correlation. The correlation coefficient of the received signals can be characterised by the complex correlation coefficient $\rho_c$ and the envelope correlation coefficient $\rho_e$ which are related by equation (3.9), assuming that the received signals have a Rayleigh distributed envelope and randomly distributed phase [59]

$$\rho_e \approx |\rho_c|^2$$  \hspace{1cm} (3.9)
In order to evaluate the correlation between two antennas, the complex correlation is computed as follows [59]

$$
\rho_c = \frac{\int_0^{2\pi} \int_0^\pi A_{12}(\theta, \phi) \sin \theta d\theta d\phi}{\sqrt{\int_0^{2\pi} \int_0^\pi A_{11}(\theta, \phi) \sin \theta d\theta d\phi \int_0^{2\pi} \int_0^\pi A_{22}(\theta, \phi) \sin \theta d\theta d\phi}}
$$  \hspace{1cm} (3.10)

where $A_{mn} = XPR \cdot E_{\theta,m}(\theta, \phi) E_{\phi,n}^*(\theta, \phi) P_{\theta}(\theta, \phi) + E_{\phi,m}(\theta, \phi) E_{\phi,n}^*(\theta, \phi) P_{\phi}(\theta, \phi)$, in which $E$ denotes the electric far field of the antenna, and $XPR$ is the ratio of the averaged vertical power to time average horizontal power in the fading environment in linear form [60]. Thus

$$
XPR = \frac{P_V}{P_H}
$$  \hspace{1cm} (3.11)

where $P_V$ is the average vertical power and $P_H$ is the average horizontal power. The $XPR$ is also referred to as the cross-polar power ratio of the incident field.

$P_{\theta}(\theta, \phi)$ and $P_{\phi}(\theta, \phi)$ are the angular density functions of the vertical and horizontal plane respectively. For reference purposes, $\theta$ is the angle relative to the vertical axis $z$ and $\phi$ is the angle in the horizontal plane as shown in Figure 3.19.

The previous discussion of diversity gain in section 3.3.3 assumed that independent signals are received on the diversity branches, i.e. there is no correlation between the signals received ($\rho_c = 0$). However, it is clear that in the majority of cases for a portable receiver this cannot be achieved because of insufficient antenna spacing. If the corre-

![Figure 3.19: The relation of angular coordinates to Cartesian coordinates.](image)
lation coefficient is greater than zero ($\rho_e > 0$), then the diversity gain will be reduced. Therefore, the correlation coefficient should be kept low enough so that the diversity is still effective. The effects of envelope correlation on diversity gain can be found in [56]. The analysis shows that where the correlation is not too close to unity or $\rho_e \leq 0.7$, the degradation of the diversity gain due to envelope correlation is given by the degradation factor ($DF$) of the following equation [56]

$$DF = \sqrt{1 - \rho_e}$$ (3.12)

### 3.3.4.2 Branch power ratio and MEG (Mean Effective Gain)

The other essential condition for achieving a high diversity gain requires that the power levels of the signals delivered by the antennas in the diversity system should not vary significantly from each other. One way of illustrating this is by using the ratio of two branch power levels $k$ as follows in the linear form

$$k = \frac{P_{\text{min}}}{P_{\text{max}}}$$ (3.13)

where $P_{\text{min}}$ is the power from the antenna with the lower power, and $P_{\text{max}}$ is the power from the antenna with the higher power in each pair of antennas.

An alternative method to obtain the branch power ratio of two branches is derived from the MEG (Mean Effective Gain) of the antennas as follows

$$k = \min\left(\frac{MEG_1}{MEG_2}, \frac{MEG_2}{MEG_1}\right)$$ (3.14)

The MEG is the average gain of an antenna in a mobile environment and is defined in [61] as the ratio between the mean received power of the antenna ($P_{\text{rec}}$) and the total mean incident power ($P_V + P_H$). The MEG is a figure of merit for the average performance of an antenna on a mobile terminal taking into account the incident radio waves in
the multipath environment and also the gain patterns of the antenna. This parameter
determines the effectiveness of the diversity antenna in a multipath environment. The
following equation can be used to evaluate the MEG [61]

\[ MEG = \int_0^{2\pi} \int_0^\pi \left[ \frac{XPR}{1 + XPR} P_\theta(\theta, \phi) G_\theta(\theta, \phi) + \frac{1}{1 + XPR} P_\phi(\theta, \phi) G_\phi(\theta, \phi) \right] \sin \theta d\theta d\phi \]

(3.15)

where \(G_\theta\) and \(G_\phi\) are the spherical power gain \((\theta, \phi)\) of the antenna, and \(P_\theta(\theta, \phi)\) and \(P_\phi(\theta, \phi)\) are the angular density functions of the incoming plane waves as used in equation (3.10). The ratio of the MEG between the two antennas must be close to unity to ensure a high diversity gain.

Equation (3.8) is suitable for the ideal case in which \(k\) equals to unity. For a selection combiner the ratio of the powers delivered by the two antennas, \(k\), is multiplied by the diversity gain to obtain a more realistic diversity gain [56]. Hence when \(N = 2\), equation (3.8) becomes

\[ P(\gamma < \gamma_s/\xi)_2 \approx \frac{1}{k} \left( \frac{\gamma_s}{\xi} \right)^2 \]

(3.16)

Assuming the correlation is low enough to achieve high diversity gain, \(k\) should be greater than -3dB to avoid significant loss in diversity gain. An unequal branch power is a disadvantage to the antenna diversity system [32].

### 3.3.5 Propagation factors

In mobile radio communications, the transmitted signals are affected by buildings and other obstacles causing multiple reflections, diffraction and scattering. The incident radio waves arriving at the mobile terminal antennas have various angles of arrival (AOAs) and cross-polar ratio (XPR) in a multipath environment. As evident from the correlation equation (3.10) and MEG equation (3.15) mentioned previously, both equations are dependent on the multipath environment via the angular density functions (AOA distributions) \(P_\theta(\theta, \phi)\) and \(P_\phi(\theta, \phi)\), and cross-polar ratio (XPR). Therefore, the AOA
distributions at both $\theta$ and $\phi$ polarisations, and the cross-polar ratio have an effect on the antenna diversity performance. For simplicity, the angular density function are modelled in elevation and azimuth separately, and they are combined according to

$$
\begin{align*}
P_{\theta}(\theta, \phi) &= P_{\theta}(\theta)P_{\theta}(\phi) \\
P_{\phi}(\theta, \phi) &= P_{\phi}(\theta)P_{\phi}(\phi)
\end{align*}
$$

(3.17)

where $P_{\theta}(\phi)$, $P_{\phi}(\phi)$ are the angular density functions in azimuth and $P_{\theta}(\theta)$, $P_{\phi}(\theta)$ are the angular density functions in elevation for the $\theta$ and $\phi$ polarisations respectively.

In order to evaluate the antenna diversity performance properly, it is necessary to apply a suitable statistical model that is similar to the real environment. Some measurements have been carried out for mobile radio communications on the angular density distribution at the mobile terminal in urban environments [62–64]. Recently, indoor environments have been also considered [65]. It is noted that different types of distribution models for wireless communication systems in an indoor environment have been suggested by different researchers. This is due to different locations and environments setup. In general, a uniform distribution is a reasonable assumption for the azimuthal angular density function as was assumed in [61] when a user of a mobile terminal moves along a random route. This is because the direct path from the base station is usually blocked from the mobile terminals, and there is an equal chance of a local scattering object being at any angle around the terminal. However, the angular density functions in the elevation direction are not uniformly distributed, and therefore the two most common different distributions i.e. Gaussian and Laplacian distributions are proposed.

Gaussian Distribution:

$$
\begin{align*}
P_{\theta}(\theta) &= A_{\theta} \exp \left\{ -\frac{[\theta-(\pi/2-m_\theta)]^2}{2\sigma_\theta^2} \right\} \\
P_{\phi}(\theta) &= A_{\phi} \exp \left\{ -\frac{[\theta-(\pi/2-m_\phi)]^2}{2\sigma_\phi^2} \right\}
\end{align*}
$$

(3.18)
Laplacian Distribution:

\[
P_\theta(\theta) = A_\theta \exp\left(-\frac{\sqrt{2(\theta-(\pi/2-m_V))}}{\sigma_V}\right) \quad 0 \leq \theta \leq \pi
\]

\[
P_\phi(\theta) = A_\phi \exp\left(-\frac{\sqrt{2(\theta-(\pi/2-m_H))}}{\sigma_H}\right) \quad 0 \leq \theta \leq \pi
\]

(3.19)

where \(m_V\) and \(m_H\) are the mean elevation angles of vertical and horizontal polarised wave distribution respectively, \(\sigma_V\) and \(\sigma_H\) are the standard deviations of the vertical and horizontal polarised wave distribution respectively. \(A_\theta\) and \(A_\phi\) are constants determined by the following condition

\[
\int_0^{2\pi} \int_0^\pi P_\theta(\theta, \phi) \sin \theta d\theta d\phi = \int_0^{2\pi} \int_0^\pi P_\phi(\theta, \phi) \sin \theta d\theta d\phi = 1
\]

(3.20)

The cross-polar ratio, \(XPR\), in the scattering environment has an effect on antenna diversity performance especially on the pattern and polarisation diversity. Generally, the value of \(XPR\) is reported between 4dB and 9dB at frequencies around 900MHz in an urban macrocell environment \([49, 66, 67]\). A few different environments have been studied at 2.15GHz, and the \(XPR\) varied between 6.6dB and 11.4dB, being lowest for indoor environments and highest for urban microcell environments \([68]\). All these reported results have shown that the \(XPR\) is not constant due to varying frequencies and environments.

In this thesis, the diversity performance of a dual-element PIFA array is evaluated in both the indoor and outdoor environments in Chapter 5. Uniform distribution is assumed in the azimuth direction whilst Gaussian and Laplacian distributions are assumed in the elevation direction. Also, an isotropic environment is used for the purpose of comparison between the numerical and the measured results. The values of \(AOA\) and \(XPR\) for each environment are summarised in Table 3-A \([69]\).
Table 3-A: Propagation models used for mobile radio systems in this thesis [69]. $m_V$ and $m_H$ are the mean elevation angles of vertical and horizontal polarised wave distribution respectively, $\sigma_V$ and $\sigma_H$ are the standard deviations of the vertical and horizontal polarised wave distribution respectively.

<table>
<thead>
<tr>
<th>Models (Elevation/Azimuth)</th>
<th>Gaussian/Uniform</th>
<th>Laplacian/Uniform</th>
<th>Uniform/Uniform (isotropic)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Statistical distributions</td>
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<td></td>
<td></td>
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<tr>
<td>$P_\theta (\theta) = \text{Gaussian}$</td>
<td>$P_\theta (\theta) = \text{Laplacian}$</td>
<td>$P_\theta (\theta) = 1$</td>
<td>$P_\theta (\theta) = 1$</td>
</tr>
<tr>
<td>$P_\phi (\phi) = 1$</td>
<td>$P_\phi (\phi) = 1$</td>
<td>$P_\phi (\phi) = 1$</td>
<td>$P_\phi (\phi) = 1$</td>
</tr>
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<td>$m_V = 20^\circ$</td>
<td>$m_V = 20^\circ$</td>
<td>$m_V = 20^\circ$</td>
<td>-</td>
</tr>
<tr>
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<td>$m_H = 20^\circ$</td>
<td>$m_H = 20^\circ$</td>
<td></td>
</tr>
<tr>
<td>$\sigma_V = 30^\circ$</td>
<td>$\sigma_V = 30^\circ$</td>
<td>$\sigma_V = 30^\circ$</td>
<td>-</td>
</tr>
<tr>
<td>$\sigma_H = 30^\circ$</td>
<td>$\sigma_H = 30^\circ$</td>
<td>$\sigma_H = 30^\circ$</td>
<td>-</td>
</tr>
<tr>
<td>$XPR = 1\text{dB}$</td>
<td>$XPR = 1\text{dB}$</td>
<td>$XPR = 1\text{dB}$</td>
<td>-</td>
</tr>
<tr>
<td>Scenario parameters</td>
<td>Indoor</td>
<td>Outdoor</td>
<td></td>
</tr>
</tbody>
</table>

3.4 Galileo navigation applications of multiple antennas

Following the study of multiple antennas, it is noticed that with antenna diversity, independent fading paths are realised without an increase in transmitted signal power or bandwidth. Therefore, the concept of antenna diversity was introduced into the Galileo navigation system for the first time in the GAC (Galileo Advanced Concept) project funded by GJU (Galileo Joint Undertaking). Antenna diversity has a minimum impact on the satellite navigation system as it can be independently implemented at the receiver. As discussed in last section, the angular density functions and $XPR$ varies in different operating frequencies and environments, and therefore the propagation models used for mobile radio systems in Table 3-A is unsuitable for the evaluation of the multiple antennas for the satellite navigation systems. Based on the specific environment of the Galileo/GPS navigation systems, the $AOA$ and $XPR$ are characterised in Chapter 6.
3.5 Summary

The requirements (i.e. low mutual coupling and high diversity gain) for multiple antennas have been addressed. By reviewing antennas for small mobile terminals, it has been found that the conventional monopole antenna and PIFA have a large current flow on the ground plane and they are very dependent on the ground plane. It is not feasible and practical to place two conventional antennas in a small terminal while keeping the mutual coupling low. New antennas need to be designed in order to overcome this obstacle, which are discussed in Chapter 4.

The other requirement for multiple antennas design is having a high diversity gain which has also been explained in detail. The design of multiple antennas needs to satisfy two conditions in order to achieve a high diversity gain. It should have low correlations and similar power levels for received signals between multiple antennas. Since mobile terminals are used in different environments and they are in motion most of time, the propagation environment has to be taken into account. It has been summarised according to current literature that different environments and different operation frequencies have different incident power distributions. Based on these reviews, a numerical diversity analysis model is established to evaluate the multiple antennas in Chapter 5.

Following the study and review of multiple antennas, it has been found that the antenna diversity technique can be independently implemented at the receiver to improve the radio link in a multipath environment. Multiple antennas have been successfully proposed to the Galileo navigation system in GAC project. As the operation frequency of the satellite navigation system is different from mobile communications, the power distributions of incident waves could be different. Currently, there is no such study available for the navigation system, and therefore the characterisation of multiple antennas in a navigation system is conducted in Chapter 6.
Chapter 3. Small and diversity antennas

References


K. Kalliola, K. Sulonen, H. Laitinen, O. Kivekas, J. Krogerus, and P. Vainikainen,

Chapter 4

Multiple antenna design

The concept of designing multiple antennas on small mobile terminals has been introduced in Chapter 3. The essential requirements of multiple antennas on a size-limited mobile terminal for MIMO systems are that the multiple antennas should have a good isolation (low mutual coupling) between antenna elements and high diversity gain in order to receive different signals even though they are closely spaced. As discussed in Chapter 3, conventional antennas such as monopoles and PIFAs cannot satisfy the low mutual coupling requirement when they are placed closely on a small mobile terminal. To meet these requirements, a novel design of a PIFA and its array operating at 5.2GHz is proposed and studied. Following investigations of the modified dual-element PIFA array at 5.2GHz, a dual-element modified PIFA array operating at 2.5GHz is studied to further evaluate the design principle. Also, a dual-element helix antenna array is proposed as a test diversity antenna array to demonstrate the improved performance for Galileo/GPS navigation system in this chapter.
4.1 Single modified PIFA at 5.2GHz

A PIFA is commonly used in a mobile terminal due to its compact size and good performance. As mobile terminals, such as handsets and PDA, have become smaller, the design of smaller PIFAs is required, and thus the dimension of a PIFA including a ground plane should be further reduced. Conventionally, the ground plane of a PIFA is the PCB (Printed Circuit Board) of a mobile terminal and the antenna performances are significantly influenced by the size of the PCB [1]. For instance, the performances such as bandwidth and radiation pattern at the frequency of interest are typically varied depending on the size of the PCB [2, 3]. Moreover, conventional PIFAs couple strongly to the PCB and spread a wide RF (Radio Frequency) current on it, as shown in Figure 4.1.

Therefore, it is difficult to employ more than one PIFA on a size-limited PCB of a mobile terminal while obtaining a good isolation and maintaining its performance. In this section, a modified PIFA with a small local ground plane operating at 5.2GHz is designed and optimised by using the CST (Computer Simulation Technology) Microwave Studio® package, which utilises the FIT (Finite Integral Technique) for electromagnetic computation [4]. The modified PIFA not only retains the unique features of a conventional PIFA, but also offers low coupling to the PCB of a mobile terminal.

![Figure 4.1: RF current distribution on the PCB of a conventional PIFA at 5.2GHz.](image-url)
4.1.1 Design of a single modified PIFA

The modified PIFA is designed by shrinking and bending the ground plane of a conventional PIFA to the same size of the top plate. The top plate with the small ground plane constitutes a stand alone structure, as shown in Figure 4.2. The modified PIFA was fed by a coaxial cable and then was mounted on a FR-4 PCB of a mobile terminal with a size of 40mm by 100mm, as shown in Figure 4.3. The relative permittivity of the FR-4 dielectric substrate (PCB) is 4.7. The coaxial feeding cable of the proposed PIFA was drilled through the substrate of the PCB without touching the PEC (Perfect Electric Conductor) layer on the PCB as shown in Figure 4.3(a). As such, the PCB is no longer a ground plane for the PIFA and the influence of the PCB on the antenna is reduced dramatically. The height of the PIFA is reduced to 4mm, which is almost half (6 ~ 10mm) of a conventional PIFA [1], in order to retain the compactness of the whole antenna structure.

Like conventional PIFAs, the performance of the modified PIFA is dependent on a number of design parameters, such as antenna height, length, width and the shorting plate. The modified PIFA was designed carefully through simulation to operate at the 5.2GHz band. The optimised dimensions of the proposed PIFA model are: \( LG=12\text{mm}, \quad LT=10.5\text{mm}, \quad WG=7\text{mm}, \quad WS=5\text{mm}, \quad H=4\text{mm}, \quad HPE=3\text{mm} \) and \( T=0.4\text{mm}. \)

![Figure 4.2: Configurations for (a) a conventional PIFA and (b) a modified PIFA. The dimensions for the modified PIFA operating at 5.2GHz: \( LG=12\text{mm}, LT=10.5\text{mm}, WG=7\text{mm}, WS=5\text{mm}, H=4\text{mm}, HPE=3\text{mm} \) and \( T=0.4\text{mm}. \)]
Chapter 4. Multiple antenna design

Figure 4.3: The configuration for a modified PIFA mounted on a PCB of a mobile terminal: (a) 3D view and (b) side view.

$L_T=10.5\text{mm}$, $W_G=7\text{mm}$, $W_S=5\text{mm}$, $H=4\text{mm}$ and $H_{PE}=3\text{mm}$. A $50\Omega$ RG405 coaxial cable ($d=0.51\text{mm}$, $D=1.7\text{mm}$, $\epsilon_r=2.1$) is used as the feed of the modified PIFA. The feed position with respect to the ground plane is shown in Figure 4.2. The optimal gap between the small ground plane and the PCB is 1mm ($gap=1\text{mm}$). A copper sheet with the thickness ($T$) of 0.4mm is used for the proposed PIFA.

4.1.2 Characteristics of the single modified PIFA

A prototype of the simulated PIFA was fabricated and mounted on the PCB in the Antenna Measurement Laboratory at Queen Mary, University of London (QMUL), as shown in Figure 4.4. A HP8720ES vector network analyser was used to measure the return loss. Figure 4.5 shows a good agreement between the simulated (dotted curve) and measured (solid curve) return losses. The centre resonance of the PIFA is around $5.2\text{GHz}$ in both simulation and measurement. The measured -10dB bandwidth (328MHz)
is slightly narrower than the simulated one (360MHz). These have verified the computer model of the antenna and gave confidence to carry out a further simulation study using the CST Microwave Studio® package.

The radiation patterns of the prototype PIFA were measured inside an anechoic chamber with the transmitting field provided by a quad ridge horn with dual-polarisation capability. Figure 4.6 and Figure 4.7 show the simulated and measured radiation patterns of the modified PIFA at the resonant frequency of 5.2GHz. Generally speaking, the
measured patterns are in good agreement with the simulated patterns, and are like the typical ones of the conventional PIFAs. The slight distortions in the measured radiation patterns (especially the cross-polar patterns) are because the PIFA was placed at a slope when it was measured.

The radiation patterns in Figure 4.6 are plotted along the XZ-plane. The radiation patterns are asymmetrical in the positive and negative X-directions. There are slightly
stronger co-polar radiations in the region above the PCB (e.g. between 0 to 60 degree) than those below the PCB (e.g. between 120 to 180 degree). However, the cross-polar radiation pattern does not show this point.

Figure 4.7 shows the radiation patterns plotted along YZ-plane, which are close to an omni-directional pattern in both simulation and measurement. Like radiation patterns of a conventional PIFA, there are reduced radiation gains on the back of the PCB, e.g. the radiation patterns from $90^\circ$ to $270^\circ$.

The performance of the modified PIFA with different gaps between the small ground plane and the PCB was investigated next. Figure 4.8 illustrates the simulated return loss curves with different gaps. It is noticed that the antenna impedance matching can be substantially improved without significantly sacrificing the bandwidth when the gap is decreased. This result is encouraging since the compactness of the PIFA can be retained with a good performance. The best antenna impedance matching is obtained when an optimal gap is reached at 1mm.

The characteristics of the conventional PIFA, such as return loss, bandwidth and gain at the frequency of interest, strongly depend on the ground plane size [2, 3]. When
the size of the PCB in a mobile terminal is changed the PIFA has to be redesigned. So, further simulations were conducted to study the dependence of the modified PIFA on the PCB. As shown in Figure 4.9, when the width of the PCB (X-direction) is kept constant and the length of the PCB (Y-direction) changed from 30mm to 100mm, the fluctuation of the centre resonant frequency is less than 20MHz. Figure 4.9 also shows that the fluctuation of the centre resonant frequency is less than 55MHz when the PIFA was located in different positions of the PCB. It is evident that the changes of the PCB’s size and the location of the antenna do not affect the modified PIFA performance very much due to its self-contained structure.

The RF current induced by the modified PIFA at 5.2GHz on the PCB is observed in simulation, as shown in Figure 4.10. The RF current of the modified PIFA is mostly confined in the area immediately underneath the antenna on the PCB, when compared to that of the conversional PIFA in Figure 4.1. This is caused by the small ground plane in the modified PIFA acting as part of the radiator and reflecting most of the radiation back towards the PIFA, and therefore the coupling between the PCB and the modified PIFA is dramatically reduced. All the results obtained in this Section gave us an increased confidence to implement more PIFAs on the PCB for MIMO applications.

![Figure 4.9: Centre resonant frequency for different dimension PCBs and different locations of the PIFA on the PCB.](image-url)
4.2 Dual-element modified PIFA array at 5.2GHz

The dual-element modified PIFA array based on the optimised configuration and dimension of the single element was also designed and fabricated. The experimental model and configuration of the PIFA array on a PCB (40mm × 100mm) is shown in Figure 4.11. The two PIFAs are placed 20mm (=0.35λ, λ=57.7mm) apart on the PCB. The relative permittivity of the FR-4 PCB’s dielectric substrate is 4.7. Same as the single PIFA design, the coaxial feed cables were drilled through the substrate of the PCB without
touching the PEC layer on the PCB.

Figure 4.12 shows the simulated and measured S-parameters of Antenna 1 and 2. The simulated S11 and S22 are identical due to the symmetric configuration of the dual-element PIFA on the PCB. There exits a slight shifting (less than 20MHz) from the centre resonance frequency and a small variation of the bandwidth (less than 70MHz) on the measured S11 and S22 observed in Figure 4.12. These are caused by the imperfections during the fabrication process of small antennas. However, as listed in Table 4-A, the measured bandwidths of Antenna 1 and 2 span more than 310MHz, where the lower resonant frequency is less than 5.15GHz and the upper resonant frequency is more than 5.35GHz. These bandwidths have covered the band of IEEE 802.11a WLAN.

The modified dual-element PIFA array has an isolation better than 28dB between two PIFAs in both simulation and measurement, Figure 4.12. The reason for achieving

![Figure 4.12: Measured and simulated S-parameters of Antenna 1 and 2.](image)

Table 4-A: Measured and simulated the lower and upper resonant frequency, and bandwidth at -10dB of Antenna 1 and 2.

<table>
<thead>
<tr>
<th></th>
<th>( f_L ), GHz</th>
<th>( f_U ), GHz</th>
<th>Bandwidth, GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Simulated Antenna 1&amp;2</td>
<td>5.006</td>
<td>5.365</td>
<td>0.359</td>
</tr>
<tr>
<td>Measured Antenna 1</td>
<td>5.081</td>
<td>5.392</td>
<td>0.311</td>
</tr>
<tr>
<td>Measured Antenna 2</td>
<td>5.102</td>
<td>5.480</td>
<td>0.378</td>
</tr>
</tbody>
</table>
such a good isolation is the following. The coupling between the PIFAs that has been caused by the ground plane effect is reduced as each PIFA has its own ground plane respectively.

The radiation patterns of Antenna 1 were measured when Antenna 2 was terminated by a $50\Omega$ load and, vice versa. The simulated and measured radiation patterns of Antenna 1 and 2 at the resonant frequency of 5.2GHz are plotted in Figure 4.13 and Figure 4.14, Figure 4.15 and Figure 4.16, respectively. As predicted, the radiation patterns of the

![Figure 4.13: Measured (-) and simulated (+) radiation patterns of Antenna 1 on the XZ-plane: (a) co-polar and (b) cross-polar.](image)

![Figure 4.14: Measured (-) and simulated (+) radiation patterns of Antenna 1 on the YZ-plane: (a) co-polar and (b) cross-polar.](image)
dual-element modified PIFA array does not have a significant difference with those of the single modified PIFA, which is due to the modified PIFA having the self-contained structure.

The radiation patterns in Figure 4.13 are plotted along XZ-plane of Antenna 1. These patterns are slightly different with those of the single modified PIFA in Figure 4.6. Due to the presence of the second PIFA, the radiation patterns of Antenna 1 in Figure 4.13 are asymmetrical in the positive and negative X-direction, and there are stronger radiations
in the negative X-direction than those in the positive X-direction. However, the YZ-plane radiation patterns of Antenna 1 in Figure 4.14 are almost identical to those of the single modified PIFA (in Figure 4.7). The discrepancies between the measurement and simulation are mainly due to the interference and noise caused by the feeding cable and other supporting materials in the anechoic chamber.

The radiation patterns of Antenna 2 in the dual-element PIFA array are plotted along the XZ-plane, as shown in Figure 4.15. The XZ-plane radiation patterns of Antenna 2 are similar to the patterns of Antenna 1, and asymmetrical in the positive and negative X-directions. But the stronger radiation directions are opposite to those of Antenna 1 due to the presence of Antenna 1. Figure 4.16 shows the radiation patterns plotted along YZ-plane, which are very close to those of Antenna 1 and the single modified PIFA.

A dual-element conventional PIFA array with the same height ($H$), length ($LT$), width ($WT$), the same location and the same size of the PCB as the modified PIFAs was modelled in CST Microwave Studio® for the purpose of comparison. The configuration is shown in Figure 4.17(a). The PCB was acting as the ground plane, i.e. the small ground plane and the parasitic element being removed from the model. The S-parameters of

![Figure 4.17: (a) the configuration and (b) simulated S-parameters of the conventional dual-element PIFA array.](image)

---

**Figure 4.17:** (a) the configuration and (b) simulated S-parameters of the conventional dual-element PIFA array.
the dual-element conventional PIFA array are shown in Figure 4.17(b). The isolation between the two elements is now smaller than 15dB, which is 13dB worse compared to the modified PIFA array. This is primary due to stronger RF current coupled to the PCB in the conventional PIFA. In contrast, the small ground plane in the modified PIFA acts as part of the radiator and so reflects most of the radiation back towards the PIFA. Consequently, there is only a little current coupled to the PCB, and hence better isolation is obtained.

4.3 Modified PIFA and its array at 2.5GHz

Separation between multiple antenna elements is the most critical parameter affecting mutual coupling. As discussed in Chapter 3, analytical studies have shown that for minimal or no mutual coupling the distance between typical antenna elements needs to be at least half wavelength [5]. A dual-element PIFA array operating in 5.2GHz band with a separation of less than half-wavelength (20mm = 0.35λ) was designed to obtain an isolation better than 28dB as presented in the previous section. It is noticed that mutual coupling depends on the frequency of received/transmitted signals since the distance is expressed in terms of the wavelength. Therefore, a dual-element PIFA array operating at lower frequency (2.5GHz band) with an even closer separation (0.17λ) will be studied in this section. The configuration is same as that operating in 5.2GHz band, while the length of each element has been increased in order to obtain the resonant frequency of 2.5GHz.

4.3.1 Single modified PIFA at 2.5GHz

For easy fabrication, the structure of the modified PIFA operating at 2.5GHz is simplified. As shown in Figure 4.18(a), it was made by bending a strip of copper to a rectangle with a slit which acts as capacitive loading. The antenna is fed by a coaxial cable from the bottom of the rectangle. The coaxial feed cable of the PIFAs was drilled through the
substrate of the FR-4 PCB without touching the PEC layer on the PCB, Figure 4.18(b). The ground plane, as small as the antenna, is located between the PIFA and the PCB. As such, the PCB is no longer a ground plane for the PIFA.

The optimized dimensions of the modified PIFA operating at 2.5GHz band are: \( L=23\,\text{mm},\ W=5\,\text{mm},\ H=4\,\text{mm} \) and \( S=0.5\,\text{mm} \). The optimal gap between the small ground plane and the PCB is 1mm (\( \text{gap}=1\,\text{mm} \)), as that of operating at 5.2GHz in Figure 4.8. The thickness of the copper sheet is 0.35mm. A 50\( \Omega \) RG405 coaxial cable (d=0.51mm, D=1.7mm, \( \epsilon=2.1 \)) is used as the feed of the modified PIFA.

A prototype of the modified single PIFA, as shown in Figure 4.19, was fabricated and mounted on the PCB at QMUL. Figure 4.20 shows the simulated (solid curve) and mea-

![Diagram of antenna configuration](image)

**Figure 4.18:** Configurations for (a) the single modified PIFA and (b) the single modified PIFA mounted on the PCB.
Figure 4.19: Photograph of the prototype PIFA operating at 2.5GHz.

Figure 4.20: Measured and simulated return loss curves.

The measured and simulated radiation patterns at the resonant frequency of 2.5GHz are plotted in Figure 4.21 and Figure 4.22, respectively. They have similar properties to those of the modified PIFA at 5.2GHz. As shown in Figure 4.21, the radiation patterns are asymmetrical in the positive and negative X-directions. In Figure 4.22, the nearly
omni-directional patterns are achieved in both simulation and measurement.

The effect of the slit ($S$) on the modified PIFA was studied. The simulations show that the slit strongly affects the resonant frequency. As shown in Figure 4.23, the narrower slit the lower resonant frequency. The slit is more likely a capacitive load for the modified PIFA. By reducing the slit to 0.5mm$^2$ the dimension is reduced to 5mm×17mm×4mm for operating at 2.5GHz from the simulation results. However, the capacitive load
Chapter 4. Multiple antenna design

Figure 4.23: Return losses for different size slits of the modified single PIFA.

Further simulations were conducted to check the dependence of the modified PIFA on the main PCB (which for this antenna is no longer the ground plane). Figure 4.24 shows the resonant frequencies when the length of the PCB is reduced. The fluctuation of the resonant frequency is less than 3MHz. It corresponds to that operating at 5.2GHz in Figure 4.9. The changes of the PCB’s size do not affect the modified PIFA performance due to its self-contained structure.

A conventional PIFA with the same height ($H$), length ($L$), width ($W$), the same location and the same size of the PCB as those of the modified PIFAs was modelled for the purpose of comparison. The RF current distributions induced by the modified PIFA and the conventional PIFA on the PCB at 2.5GHz were observed in simulation, as shown in Figure 4.25 (a) and (b), respectively. The results indicate that there is less current coupled to the PCB in the modified PIFA than in the case of the conventional PIFA. The RF current is localised underneath the antenna on the PCB, and the coupling between the PCB and the modified PIFA is dramatically reduced. The location flexibility and current reduction on the PCB of the modified PIFA operating at 2.5GHz cohere well to
Figure 4.24: Resonant frequencies for different size PCB of the modified single PIFA. Vertical lines represent the -10dB bandwidth and triangles represent the central resonant frequency.

Figure 4.25: RF current distributions on the PCB at 2.5GHz: (a) the conventional PIFA and (b) the modified PIFA.

those operating at 5.2GHz as discussed in section 4.1. Therefore, more PIFAs on the PCB operating at 2.5GHz band are employed in the following section.
4.3.2 Dual-element modified PIFA array at 2.5GHz

The dual-element modified PIFA array at 2.5GHz was fabricated based on the configuration of the single modified PIFA. A prototype is shown in Figure 4.26(a). Both PIFAs are placed 2mm from top and 5mm from side on the PCB with a separation of 20mm (0.17λ), as shown in Figure 4.26(b). The S-parameters were measured and compared to the simulation results in Figure 4.27. The simulated return losses of both PIFAs are identical due to the symmetric configuration. The measured return loss of Antenna 1 is slightly different with that of Antenna 2. This is caused by the imperfection in the fabrication of two PIFAs.

It is also shown in Figure 4.27 that an isolation better than 20dB can be achieved in both simulation and measurement. Compared to the isolation of 28dB obtained at 5.2GHz, there is a 8dB isolation decrease. This is mainly caused by the distance between two PIFAs which was shorten from 0.35λ to 0.17λ. However, compared to the conventional dual-element PIFA array with the same separation, the dual-element modified PIFA has improved the isolation for more than 10dB, as shown in Table 4-B.

The 3D radiation patterns of Antenna 1 and 2 at the resonant frequency of 2.5GHz

![Figure 4.26: (a) photograph and (b) configuration for the dual-element PIFA array on the PCB operating at 2.5GHz.](image)
Figure 4.27: Measured and simulated S-parameters of the dual-element PIFA array.

Table 4-B: Summary of the isolations of the dual-element conventional and modified PIFA array operating at 5.2GHz with separation of $0.35\lambda$ and 2.5GHz with separation of $0.17\lambda$.

<table>
<thead>
<tr>
<th>Separation between two PIFAs</th>
<th>0.35\lambda</th>
<th>0.17\lambda</th>
</tr>
</thead>
<tbody>
<tr>
<td>Isolation of the conventional dual-element PIFA array *</td>
<td>15dB</td>
<td>8dB</td>
</tr>
<tr>
<td>Isolation of the modified dual-element PIFA array +</td>
<td>28dB</td>
<td>20dB</td>
</tr>
</tbody>
</table>

* Simulated
+ Simulated and measured

were measured at Sony Ericsson Mobile Communications AB while the dual-element PIFA array on the PCB was positioned vertically in a Satimo chamber. The results are plotted in Figure 4.28 and Figure 4.29, respectively. It has been found that both co-polar and cross-polar radiation patterns of Antenna 1 and 2 are complementary to each other. Hence, it is indicated that the dual-element PIFA array has a good pattern diversity. It is also noticed that Antenna 2 has slightly higher gain patterns than those of Antenna 1. This may degrade the diversity gain due to the unequal branch power.

### 4.4 Dual-helical antenna array for Galileo/GPS terminals

In this thesis, a dual-element helix antenna array has been proposed and studied as a diversity antenna array to demonstrate the potential performance enhancement for
Figure 4.28: Measured 3D radiation patterns of Antenna 1: (a) co-polar and (b) cross-polar.

Figure 4.29: Measured 3D radiation patterns of Antenna 2: (a) co-polar and (b) cross-polar.

Galileo/GPS navigation system. To our knowledge, this is the first time to introduce multiple antenna technology to the Galileo/GPS navigation system. The proposed design of the diversity antenna array consists of two normal mode helical antennas operating
at 1575.42MHz, L1 band. Though a normal-mode helix exhibits linear polarisation, it is noticed that RHCP (Right Hand Circular Polarisation) is not strictly required for a GPS link in a highly scattered environment [7].

4.4.1 Design of helical antennas

The principle of a helical antenna has been explained in Chapter 3. In this section, a normal mode helical antenna with radiation from the side of the antenna is used for each element of the proposed diversity antenna array. The design of the helical antennas was carried out by using CST Microwave Studio®. Figure 4.30(a) shows the configuration of each of the proposed helical antennas. It is optimally designed to operate in the L1 band and is made by winding 5 turns of 0.5mm diameter copper wire around a core of 3.5mm diameter with a pitch of 2.5mm. It is fed by a 50Ω coaxial cable.

Two helical antennas were soldered to the top edge of the ground plane of a typical mobile terminal. The antennas are symmetrically placed 50mm apart on the ground

![Figure 4.30](image-url):

Figure 4.30: Configurations for (a) the single helical antenna and (b) two helical antennas mounted on a ground plane. Dimensions are in mm.
plane. The size of ground plane is 60mm by 100mm, as shown in Figure 4.30(b).

4.4.2 Performance of helical antennas

The dual-helical antenna was fabricated and measured in the Antenna Measurement Laboratory at QMUL. The photographs of a prototype of the antenna array is shown in Figure 4.31. Figure 4.32 shows the S-parameters of Antenna 1 and Antenna 2.

The return losses (S11 and S22) of both antennas are almost identical in both simulation and measurement and the dual-helical antenna array can fully cover the bandwidth (1559MHz - 1591MHz) required by Galileo system. The simulated S-parameters are slightly different from those measured because of imperfections in the fabrication of small helical antennas. It is noted in Figure 4.32 that the isolation between these two helical antennas is less than 6dB in both simulation and measurement. As discussed in Chapter 3, multiple antennas with low isolation will result in a low efficiency of the whole diversity
system, but could achieve low correlation and high diversity. Consequently, the dual-helical antenna array is selected only as a test diversity antenna array for Galileo/GPS navigation system and its diversity performance are evaluated in Chapter 6.

The radiation patterns of the dual-helical antennas were measured in the Antenna Measurement Laboratory at QMUL. The radiation patterns of Antenna 1 were measured when Antenna 2 was terminated by a 50Ω load, and vice versa. A comparison of measured and simulated radiation patterns of Antennas 1 and 2 at the resonant frequency of 1.57GHz is illustrated in Figure 4.33 and Figure 4.34, Figure 4.35 and Figure 4.36, respectively. The measured and simulated co-polar radiation patterns agree well, and the ripples in the measured radiation patterns are due to the effect of the long feeding cable. However, the difference between the measurements and simulations for the cross-polar radiation patterns is increased, and the signal detected is very weak in XZ-plane plane, as shown in Figure 4.33(b) and Figure 4.35(b). Due to the coupling effect, the co-polar and cross-polar radiation patterns of Antenna 1 have symmetrical displacements relative to those of Antenna 2. Therefore, the dual-helical diversity antenna array has both spatial and pattern diversity characteristics.
Figure 4.33: Measured (-) and simulated (+) radiation patterns of Antenna 1 on the XZ-plane: (a) co-polar (b) cross-polar.

Figure 4.34: Measured (-) and simulated (+) radiation patterns of Antenna 1 on the YZ-plane: (a) co-polar (b) cross-polar.
Figure 4.35: Measured (−) and simulated (+) radiation patterns of Antenna 2 on the XZ-plane: (a) co-polar (b) cross-polar.

Figure 4.36: Measured (−) and simulated (+) radiation patterns of Antenna 2 on the YZ-plane: (a) co-polar (b) cross-polar.

4.5 Summary

Extensive studies have been conducted both numerically and experimentally on a modified PIFA with a small ground plane and its dual-element array mounted on a PCB of a mobile terminal operating at 5.2GHz. It has been shown that the compactness and performance of the PIFA can be well retained with a small gap between the small ground
plane and the PCB. It has also demonstrated that the dual-element modified PIFA array can achieve an isolation better than 28dB with a separation of 20mm (0.35λ) in the case of operating at 5.2GHz because the RF current on the PCB has been reduced dramatically with the introduction of the small ground plane between the PIFA and the PCB.

A dual-element PIFA array operating at 2.5GHz has been investigated with the same configurations as those operating at 5.2GHz but closer separation of 0.17λ. An isolation better than 20dB between each element in the case of operating at 2.5GHz can be obtained. This further confirms that the design of the small ground PIFA array can reduce the coupling effect to obtain the good isolation performance.

As discussed in Chapter 3, the good isolation performance is only the first design criteria of multiple antennas on a mobile terminal for MIMO systems. The diversity performance and channel capacity of the multiple antennas are also crucial parameters for the MIMO system. Although the dual-element modified PIFA array has shown the preliminary pattern diversity characteristics by observing the 3D radiation patterns, the dedicated analysis of its diversity performance and channel capacity will be conducted in next chapter.

Furthermore, a dual-element helical diversity antenna system was designed through extensive simulations and a prototype was fabricated and measured. Good agreement was shown between the measurements and simulations. From the radiation pattern characteristics, it was found that the two helical antennas working together show pattern diversity characteristic. However, the dual-helical antenna array has shown a poor isolation performance (less than 6dB), and therefore only selected as a test diversity antenna array for Galileo/GPS navigation system and its diversity performance will be also evaluated in Chapter 6.
References


In Chapter 4, dual-element PIFA arrays were investigated and it was shown that they can achieve low mutual coupling characteristic in both operating at 5.2GHz and 2.5GHz frequency bands. Essentially, its diversity performance and channel capacity are the crucial properties for the characterisation of multiple antennas as addressed in Chapter 3. In this chapter, the dual-element PIFA array on the PCB operating at 2.5GHz is selected to study its diversity performance for the purpose of validation. Also, its channel capacity is evaluated in an indoor environment.

5.1 Diversity performance of the dual-element PIFA array

The diversity performance of the proposed dual-element PIFA array on the PCB operating at 2.5GHz is studied in this section. The diversity performance has been evaluated by calculating the correlation, MEG and diversity gain of the antennas. The measurement on diversity gain in a scattering field chamber has been conducted and compared to those calculated.
5.1.1 Correlation and MEG

The correlations for the dual-element PIFA array on the PCB are evaluated using equation (3.9) and (3.10) with the measured 3D radiation patterns which have been measured when the dual-element PIFA array on the PCB was vertically positioned, and statistical propagation models depicted in Table 3-A. The results are summarised in Table 5-A. The impact of indoor and outdoor environments on the envelope correlation has been evaluated using two different statistical models as discussed in Chapter 3 (i.e. Gaussian/Uniform and Laplacian/Uniform distributions). Table 5-A shows that an envelope correlation of less than 0.3 (as evaluated by two different statistical models) has been achieved in both indoor and outdoor environments. These low correlation values result in very small degradation of diversity gains, as evident from equation (3.12).

The MEG of each antenna within the different environments is evaluated using equation (3.15) and the results are also tabulated in Table 5-A. It is noticed that the MEG values of each antenna within the different environments can vary up to 1.5dB. The difference of the MEG values between Antenna 1 and 2 is due to the Antenna 2 having a slightly better radiation gain than the Antenna 1 as discussed in Section 4.3.2 Chapter 4. However, the difference is less than 1dB, and therefore it leads to a small degradation on the diversity.

Table 5-A: Envelop correlation coefficient and MEG of the dual-element PIFA array in different propagation models.

<table>
<thead>
<tr>
<th>Propagation environments</th>
<th>$\rho_{e12} \equiv \rho_{e21}$</th>
<th>MEG, dB</th>
<th>MEG, dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>Antenna 1</td>
<td>-3.90</td>
<td>-3.02</td>
<td></td>
</tr>
<tr>
<td>Antenna 2</td>
<td>-4.08</td>
<td>-3.41</td>
<td></td>
</tr>
<tr>
<td>Gaussian (indoor)</td>
<td>0.02</td>
<td>-4.65</td>
<td>-3.75</td>
</tr>
<tr>
<td>Gaussian (outdoor)</td>
<td>0.24</td>
<td>-5.26</td>
<td>-4.39</td>
</tr>
<tr>
<td>Laplacian (indoor)</td>
<td>0.08</td>
<td>-3.86</td>
<td>-3.03</td>
</tr>
<tr>
<td>Laplacian (outdoor)</td>
<td>0.02</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Isotropic</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
Chapter 5. Diversity performance and channel capacity of the dual-PIFA

5.1.2 Diversity gain

After evaluating the correlation and MEG results, the diversity gain in an isotropic environment and more realistic environments such as indoor and outdoor environments as addressed in Chapter 3 are used to statistically assess the diversity performance of the dual-element PIFA array using a selection combiner (SC). The diversity gain results have included the degradation factor ($DF$) in equation (3.12), taking into account the correlation and the branch power power difference in terms of ratio ($k$) in equation (3.14). Figure 5.1 shows that SC diversity gain of the dual-element PIFA array in Gaussian/Uniform and Laplacian/Uniform statistical models for both indoor and outdoor environments compared with those of both of single and dual ideal antennas, and a dual-element PIFA array in an isotropic environment. As indicated in Figure 5.1, the SC diversity of the dual-element PIFA array is slightly less than that of 2 ideal antennas (10dB) in any case at 99% reliability. The degradation in the diversity gain is due to the correlation and unequal mean antenna branch power, as shown in Table 5-A.

As predicted, the SC diversity gain of the dual-element PIFA array in an isotropic environment is higher than that in indoor or outdoor environments using the Gaussian/Uniform or Laplacian/Uniform statistical models as shown in Figure 5.1. However, the differences are very small. The SC diversity gains in different environments using different statistical models have less than 1dB variation. Therefore, the dual-element PIFA array can work well in different situations.

5.1.3 Measurement on diversity gain

The dual-element PIFA array was also tested in an in-house scattering field chamber at Sony Ericsson Mobile Communications AB to examine its diversity performance [1, 2]. The scattering field chamber is a method to reproduce a three-dimensional isotropic random field environment by using three rotating stirrers and a phantom torso, as shown in Figure 5.2. The 4000 samples were measured in the chamber for the evaluation of the
Figure 5.1: Calculated SC diversity gain of the dual-element PIFA array: (a) in Gaussian/Uniform statistical model and (b) in Laplacian/Uniform statistical model for indoor and outdoor environments compared with those of 2 ideal antennas and the dual-element PIFA array in an isotropic environment, and a single antenna.

The measured SC diversity gain in the isotropic environment are plotted in Figure 5.3. This measured result is very close to the calculated result. The measured SC diversity gain at 99% and 50% reliability compared to those calculated in an isotropic environment are tabulated in Table 5-B. There is only 0.1dB difference of SC diversity gain between
Figure 5.2: A diagram showing a top view of the Sony Ericsson in-house scattering field chamber [1, 2].

Figure 5.3: Measured SC diversity gain of the dual-element PIFA array in the isotropic environment compared with those of 2 ideal antennas and a single antenna.

Table 5-B: Measured and calculated SC diversity gain at 99% and 50% reliability in an isotropic environment.

<table>
<thead>
<tr>
<th>SC Diversity Gain</th>
<th>99% reliability</th>
<th>50% reliability</th>
</tr>
</thead>
<tbody>
<tr>
<td>Measured</td>
<td>9.3dB</td>
<td>2.3dB</td>
</tr>
<tr>
<td>Calculated</td>
<td>9.4dB</td>
<td>2.4dB</td>
</tr>
</tbody>
</table>

the measurements and calculations. This has verified the previous numerical modelling and analysis.
Chapter 5. Diversity performance and channel capacity of the dual-PIFA

5.2 Channel capacity of the dual-element PIFA array

As discussed earlier, the dual-element PIFA array has exhibited low mutual coupling and good diversity performance. Now, the major concern is whether it can perform well in a realistic MIMO channel model. In this section, a RT (Ray Tracing) MIMO channel model based on a commercial package (Wireless InSite) is developed. Validations of the RT MIMO channel model are obtained by comparing the simulation channel results with the equivalent IEEE 802.11 MIMO channel model. Further, the RT MIMO channel model is used to study the channel capacity of the dual-element PIFA array and other antenna arrays.

5.2.1 Design of a RT MIMO channel model

A ray tracing approach [3] is used to estimate the capacity of a wireless MIMO system with an objective of clarifying the general principles involved in MIMO indoor channels, rather than accurately determining channel capacities for a specific system. Moreover, the theoretical capacity evaluation of a MIMO system based on site-specific ray tracing, using a uniform linear array of vertical dipoles with half-wavelength, has been reported [4]. However, both ray tracing approaches for the MIMO channel modelling did not consider the practical antenna array which plays an important role in MIMO systems. The development of more accurate channel models, which take into account practical antenna arrays as well as propagation environments, is the key to evaluate the realistic gains of MIMO systems. Therefore, a ray tracing MIMO channel model which can consider practical antenna arrays (e.g. the dual-element PIFA array) and realistic propagation environments is proposed. It is based on a full 3D ray tracing simulation tool, Wireless InSite, which is used to obtain the time delay, phase angle and received power for characterising the MIMO channels.

This work can be divided into five stages, as shown in Figure 5.4. These five stages are introduced in the following sections, respectively.
Chapter 5. Diversity performance and channel capacity of the dual-PIFA

5.2.1.1 Testing antenna arrays and an indoor environment

An ULA (Uniform Linear Array) of vertical dipoles operating at 2.5GHz was first set up for the purpose of validating the RT MIMO channel model. Figure 5.5(a) shows four dipoles arranged in a linear array with a equal gap \( d \) of half-wavelength between them. There is no mutual coupling between the dipoles. This set-up of the ULA is same as the one used in the 802.11 MIMO channel model [5]. The same four dipoles were also arranged in a circular array (CA) with a space of half-wavelength between each dipoles without considering the mutual coupling, as shown in Figure 5.5(b).

![Flow chart showing the design of the RT MIMO channel model.](image)

![Figure 5.4: Flow chart showing the design of the RT MIMO channel model.](image)

Figure 5.4: Flow chart showing the design of the RT MIMO channel model.

![Figure 5.5: Configurations for the receiving antennas: (a) an ULA and (b) a CA, with a space of \( d = \lambda/2 \) between dipoles.](image)

Figure 5.5: Configurations for the receiving antennas: (a) an ULA and (b) a CA, with a space of \( d = \lambda/2 \) between dipoles.
Chapter 5. Diversity performance and channel capacity of the dual-PIFA  

An indoor environment considered in this thesis is the second floor of the Department of Electronic Engineering at QMUL, as shown in Figure 5.6. The indoor environment was modelled using Wireless InSite. In Figure 5.7, three different materials such as concrete, wood and glass have been used to model the indoor environment and all materials are assumed to be homogenous. The properties of the materials at 2.5GHz [6, 7] are summarised in Table 5-C.

Figure 5.6: The floor plan of the second floor of Department of Electronic Engineering at QMUL.

Figure 5.7: 3D schematic diagram of the indoor environment (i.e. second floor of the Department of Electronic Engineering building at QMUL) modelled using Wireless InSite as used in this thesis. The transmitter is placed at the corridor (green dots) and the receivers (red dots) are scattered randomly in Rooms A and B. The ceiling has been removed for visual purpose.

Table 5-C: Material properties used for the indoor environment model at 2.5GHz [6, 7].

<table>
<thead>
<tr>
<th>Material</th>
<th>Permittivity</th>
<th>Conductivity</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wall, floor and ceiling (Concrete)</td>
<td>7.00</td>
<td>0.0814</td>
</tr>
<tr>
<td>Door (Wood)</td>
<td>5.00</td>
<td>0.0084</td>
</tr>
<tr>
<td>Window (Glass)</td>
<td>2.40</td>
<td>0.0000</td>
</tr>
</tbody>
</table>
The transmitter consisting of the ULA is placed on the ceiling of the corridor outside Room A, as shown in Figure 5.7. The transmit power level from the transmitter is 20dBm and the bandwidth is 20MHz. The receivers (e.g. ULA and CA of 4 dipoles, and the dual-element PIFA array) are clustered randomly in 1000 positions near the desktop height in Rooms A and B.

5.2.1.2 Full 3D ray tracing tool

Wireless InSite provides a full 3D ray-based propagation model [8]. This model combines ray tracing algorithms with the UTD (Uniform Theory of Diffraction) [9–11]. The ray tracing procedure is used to find the propagation paths from each transmitter to each receiver, and the UTD is used to evaluate the complex electric field associated with each ray path. In the full 3D propagation model, the SBR (Shooting and Bouncing Ray) method is used to find ray paths which have been transmitted from the transmitters to the receivers via reflection from various vertical surfaces. The ground reflection points on these ray paths are found analytically using the method of images. It is noticed that there are two ways to identify diffracting edges when the SBR method is used to find the geometrical propagation paths. These methods such as SBR, method of images and identifying diffractions, which are employed in the Wireless InSite, are discussed in Appendix C.

5.2.1.3 Data collection

Example of the simulated rays arriving at the receivers in Room A and B are shown in Figure 5.8 and Figure 5.9. For clarity, only a few ray paths are drawn in Figure 5.9. It is shown that the signals launched from the transmitter are reflected off and/or transmitted through obstacles before reaching the receivers. A maximum of 20 ray paths can be simulated from the transmitter to each receiver with a power threshold of -200dBm. The information of each path, i.e. the time delay, the length of the path and the received
5.2.1.4 Channel matrix $H$

After collecting the data, the channel response is modelled as the vector sum of all the rays arriving at the receiving antenna locations. A narrowband channel is assumed in power at the receiver, are recorded in simulation. These results are collected and analysed using a Matlab program. The whole post-process of the data is outlined in Figure 5.10.
Each link has a maximum of 20 ray paths. The time delay, the length and the received power of each path are collected from Wireless InSite M-Elements N-Elements

\[ h_{ij} = \sum_{k=1}^{M} \sqrt{P_k} \cdot e^{i(2\pi/\lambda)l_k} \cdot e^{i2\pi f_0 \tau_k} \]  

(5.1)

where \( M \) is the number of rays, \( f_0 \) is the carrier frequency, \( P_k \) is the received power, \( l_k \) is the length of the \( k^{th} \) ray and \( \tau_k \) is the time delay of the \( k^{th} \) ray. The elements \( h_{ij} \) of the channel matrix \( H \) is computed by using equation (5.1) and \( P_k, l_k \) and \( \tau_k \) are obtained from Wireless InSite simulations.

### 5.2.1.5 Eigenvalue and capacity

After obtaining the channel matrix \( H \) of the indoor environment, the eigenvalues from the channel matrix \( H \) are evaluated mathematically by the SVD method as shown in Figure 2.5 and the channel capacity is then computed using equation (2.16). The details have been described in Chapter 2.

### 5.2.2 Comparison with the IEEE MIMO channel model

A MIMO channel model in indoor environments based on the experimental data was proposed by IEEE 802.11 group in 2003. For the comparison, the RT MIMO channel
model uses the same parameters as those of the IEEE 802.11 MIMO channel model to simulate the channel capacity in the indoor environment. The simulation parameters between them are summarised in Table 5-D. Therefore, the simulated result in Room A for the ULA antenna configuration from the RT MIMO channel model is compared with those from the IEEE 802.11 MIMO channel model E.

The CDFs of a narrowband capacity for the RT MIMO channel model, IEEE MIMO channel model E and the iid (independent and identically distributed) case are plotted in Figure 5.11 with SNR $\xi = 10$dB, 1000 channel realisations, and NLOS conditions. In the iid case, elements of channel matrix $H$ are independent and identically distributed zero-mean unit-variance complex Gaussian random variables. The iid case is the theoretically

<table>
<thead>
<tr>
<th>Environments</th>
<th>802.11 MIMO model [5]</th>
<th>RT MIMO channel model</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Model E with NLOS for Large office.</td>
<td>Offices of Department of Electronic Engineering at Queen Mary.</td>
</tr>
<tr>
<td>Simulation Parameters</td>
<td>4×4 set-up</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Uniform linear array (ULA)</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$\lambda/2$ adjacent antenna spacing</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Isotropic antennas</td>
<td></td>
</tr>
<tr>
<td></td>
<td>No antenna coupling effect</td>
<td></td>
</tr>
<tr>
<td></td>
<td>All antennas with same polarisation (Vertical)</td>
<td></td>
</tr>
</tbody>
</table>

Figure 5.11: The channel capacity comparison of the RT MIMO channel model in Room A and the IEEE MIMO channel model E.
ideal case. It can been seen that the capacity result from the RT MIMO channel model is in good agreement with model E of the IEEE indoor MIMO channel. The results of IEEE MIMO channel model E and the RT MIMO channel model are 2bits/s/Hz lower than that of the iid case.

Furthermore, channel capacities of the ULA and CA antenna configurations are investigated in Room A and B, respectively. The CDFs of a narrowband capacity for ULA and CA antenna configurations are plotted in Figure 5.12. The solid curves represent the channel capacities for the ULA antenna configuration in room A and B. The dotted curves represent the channel capacities for the CA antenna configuration in Room A and B. The channel capacities for the ULA and CA in Room B are slightly higher than those in Room A, which is due to the presence of more scattering paths in the more distant and larger Room B. It has been observed that a slightly higher capacity can be achieved in the CA antenna configuration when compared to the ULA antenna configuration. This is because the CA antenna configuration creates more omni-directional radiation patterns in the azimuth plane than the ULA configuration. Overall, the results between the RT MIMO channel model and the IEEE MIMO channel model E are in a good agreement.

![Figure 5.12: Channel capacities of the ULA and CA antenna configurations in the RT MIMO channel model in Room A and B, respectively.](image-url)
5.2.3 Investigation of practical antenna arrays in the RT MIMO channel model

One of the objectives of this project is to investigate the capacity enhancement of practical antenna arrays in a realistic environment MIMO channel model. In the previous sections, the RT MIMO channel model has been introduced and validated, and therefore the dual-element PIFA array as the receiver is applied to the RT MIMO channel model instead of the receiver which employed 4 ideal dipoles. The measured 3D radiation patterns of the dual-element PIFA array as shown in Figure 4.28 and 4.29 are imported to the model. These radiation patterns have included the effects of mutual coupling. The other parameters of the RT MIMO channel model, such as the environment and the location of the receivers, are kept same as the previous set-up.

Figure 5.13 shows the MIMO channel capacity obtained from the realistic propagation environments (i.e. Room A and Room B) with the proposed dual-element PIFA array on the handset PCB as the receivers. The channel capacity results from Rooms A and B are compared to those of the conventional dual-element PIFA array on the PCB and two ideal dipoles, and that within the SISO system which consists of only a single dipole at both the transmitter and receiver. It can be seen in Figure 5.13 that the capacity of the proposed dual-element PIFA array (which have low mutual coupling shown in Figure 4.27) is close to that of the two ideal dipoles (which have no mutual couple included), and higher than that of the conventional dual-PIFA (which have strong mutual coupling shown in Table 4-B). In the two ideal dipoles model, the mutual coupling had not been taken into account. It is shown that the channel capacity is slightly smaller in the proposed dual-PIFA diversity antenna array when the mutual coupling effect is taken into consideration. However, the difference is not significantly big. When SNR = 20dB, the channel capacity increased from 6.1bits/s/Hz in the SISO system to 9.8bits/s/Hz in the MIMO system in both Room A and Room B using the proposed dual-PIFA diversity antenna array, as shown in Figure 5.13. It is also noted that the channel capacity performances at Room A and B are similar despite the different sizes of the rooms.
Figure 5.13: Channel capacities of the proposed dual-PIFA compared to two ideal dipoles and the conventional dual-PIFA and within a SISO system: (a) in Room A and (b) in Room B.

5.3 Summary

The diversity gain of the proposed dual-element PIFA array has been evaluated by assessing the correlation and MEG of the antennas in both indoor and outdoor environments.
Generally speaking, the envelope correlations of the dual-element PIFA array are less than 0.3. The MEG values of each antenna within the different environments can vary up to 1.5dB. The difference of the MEG values between two PIFAs is less than 1dB. As a result, it has been shown that the calculated diversity gain of the proposed dual-element PIFA array is close to the 2 ideal antennas (10dB) in both indoor and outdoor environments. It also has been shown that the dual-element PIFA array can operate well in different fading environments. Furthermore, the diversity gain of the dual-element PIFA array has been measured and verified at Sony Ericsson.

A RT MIMO channel model has been developed based on the Wireless InSite ray tracing simulator. This propagation model has been validated with the IEEE 802.11 MIMO channel model using an ULA of four ideal dipoles at both the transmitter and receiver for a $4 \times 4$ set-up. The RT MIMO channel model has been used to investigate the capacity of different antenna configurations such as the CA, the proposed dual-element PIFA array, the conventional dual-element PIFA array and the 4-element diversity antenna array.

A CA of four ideal dipoles has been applied to the RT MIMO channel model. It has been shown that the CA antenna configuration has achieved slightly higher capacity than the ULA. Furthermore, the performance of the proposed dual-element PIFA array, the conventional dual-element PIFA and two ideal dipoles were evaluated in the RT MIMO channel model in the indoor environment. It has been shown that the channel capacity of the proposed dual-element PIFA array is slightly decreased when the mutual coupling effect is taken into consideration, compared to two ideal dipoles without mutual coupling included. However, the channel capacity of the proposed dual-element PIFA array is higher than that of the conventional dual-PIFA which has the strong mutual coupling.
Chapter 5. Diversity performance and channel capacity of the dual-PIFA

References


Chapter 6

Diversity performance of the dual-helical antenna array

As discussed in Chapter 3, the concept of multiple antennas was introduced into the Galileo navigation system in the GAC (Galileo Advanced Concept) project. The work in the GAC project is described in details in this chapter. The propagation factors such as AOA and XPR of the diversity system for Galileo/GPS system are characterised. Based on these characterised propagation factors, the diversity performance of the dual-helical antenna array for Galileo/GPS systems is evaluated.

6.1 Introduction to the WP2160 of the GAC project

The GAC project funded by GJU aims to evolve and/or consolidate the critical aspects of the Galileo system, in order to anticipate the future context of satellite navigation technology and to sustain the competitiveness of Galileo. Within this context, a sub-work package, WP2160: Modelling of Difficult Environments, proposes and investigates receiving diversity technology to improve the performance of the Galileo system in indoor environments. The subwork package lasted for 8 months from February to October 2006.
Generally speaking, the present GPS does not function well in indoor or urban canyon multipath environments, which are characterised as difficult environments. It is difficult to deal with the attenuated and multipath faded signals resulting from transmission around and through concrete and steel building structures. Currently, the solution for GPS operating in an indoor environment is A-GPS (Assisted-GPS) \[1\]. The terrestrial wireless networks can potentially provide timing and ephemeris assistance to a GPS receiver. However, it is more desirable for GPS service providers to minimize their dependence on other networks, and also to improve location reporting performance and robustness.

Receive diversity is a technique to combine multipath signals in a mobile communication receiver, as discussed in Chapter 3. Multipath fading of the GPS downlink signal is caused by the transmitted signal from the satellite following multiple paths to the receiving antenna. At a single point these signals may add constructively, enhancing the received signal, or destructively, reducing the received signal below a useable threshold. Therefore, multiple antennas on a Galileo/GPS terminal are proposed to enhance received signals from the satellite.

6.2 Characterisation of propagation factors for Galileo/GPS system

Propagation models shown in Table 3-A of Chapter 3 specified the $AOA$ and $XPR$ values for wireless communications. However, the $AOA$ and $XPR$ are frequency dependent. Antenna diversity technology is introduced to Galileo/GPS navigation system in this work for the first time. There are almost no equivalent studies and measurements for L-Band satellite systems. It is essential to determine the parameters for the Galileo system. It can be concluded that antenna diversity performance in terms of diversity gain for the Galileo system is determined by following two factors:
Chapter 6. Diversity performance of the dual-helical antenna array

- Antenna configurations

- Propagation environments, as described by $AOA$ and $XPR$

The dual-element helical antenna array described in Chapter 4 is used as a test antenna for the evaluation of diversity performance. A deterministic multipath channel model based on ray tracing is used to characterise the $AOA$s for the evaluation of the diversity performance of a dual-element helical antenna array for Galileo/GPS navigation system. The models are designed as close to the real Galileo/GPS propagation environment as possible by using the same parameters as measured by the German Aerospace Centre (DLR) [2].

6.2.1 Set-up of metrics

In 2002, DLR launched a measurement campaign to establish the expected characteristics of the received signal from Galileo in a cluttered indoor environment [2–6]. The satellite was simulated by a helicopter flown around the building where the receiver was located inside, operating at distances from 1500m to 4000m from the receiver. Elevation angles of around 10°, 30° and 60° were considered. A signal with a power of 10W was transmitted from the helicopter on the L1 frequency band (the centre frequency is 1575.42MHz with 40MHz bandwidth). The metrics of our deterministic multipath channel model are based on the parameters determined from these measurements. The structure of the Electronic Engineering building at QMUL has been established in the multipath simulator (Wireless InSite), as shown in Figure 6.1. For the purpose of clarity, the ceiling of the building has been set invisible in the figure. The bright grey parts are the external walls, the dark grey parts are the internal walls, the brown parts are the door-ways, the light blue parts are the portions between floors and the deep blue parts are the glass windows. Table 6-A shows the list of material properties used for the QMUL building structure in the simulation, which are same as those used by the DLR group [2].

The satellite transmitting antennas shape the directional beam to cover the surface
Chapter 6. Diversity performance of the dual-helical antenna array

Figure 6.1: The electronic engineering building structure of QMUL. For the purpose of clarity, the roof of the building has been set invisible.

Table 6-A: Material properties used for the modelled QMUL building structure [2].

<table>
<thead>
<tr>
<th>Material</th>
<th>Permittivity</th>
<th>Conductivity</th>
<th>Thickness (cm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Doors</td>
<td>5.8</td>
<td>0.05</td>
<td>4</td>
</tr>
<tr>
<td>Internal walls</td>
<td>2.9</td>
<td>0.15</td>
<td>10</td>
</tr>
<tr>
<td>Partitions between floors</td>
<td>2.9</td>
<td>0.001</td>
<td>17</td>
</tr>
<tr>
<td>External wall</td>
<td>6</td>
<td>0.34</td>
<td>25</td>
</tr>
<tr>
<td>Windows</td>
<td>6</td>
<td>0.19</td>
<td>1</td>
</tr>
</tbody>
</table>

of the earth [7, 8]. Our simulation used directional antennas with RHCP to simulate the satellite transmitters operating in the Galileo L1 band. The radiated power of the antenna is 40dBm (10W). The simulated transmitters are distributed on a hemisphere above the modelled building structure at a radius of 1500m, as shown in Figure 6.2. It also can be seen in Figure 6.2(a) that the transmitters (green dots) are located at elevation angles of 10°, 30°, 60° and 80°, where 0° is the zenith and 90° the horizon as used in the measurements [4]. In Figure 6.2(b), it shown that the transmitters are spaced in 6 planes. The overall distribution of the transmitters simulates a generic case of the satellites in space.
Finally, omni-directional RHCP antennas operating at the Galileo L1 band were used as the receivers. A maximum of 10 ray paths can be simulated to each receiver with a power threshold of -200dBm. The receivers, represented by red points in Figure 6.3, are placed inside the building. For the purpose of clear views of the receivers, the building structure of the first floor including the partition between first floor and ground floor were...
rendered invisible but they are present in the propagation model. The simulated receivers are placed in different locations on different floors so various penetration conditions are included. For example, some receivers are located on the ground floor inside the offices, some receivers are placed in the corridor of the second floor, some are on the ground floor inside a big lab area, close to the windows, and others are in middle of big lab area with several partitions on the second floor. After setting up all the parameters, the ray tracing multipath model is ready for simulations.

### 6.2.2 Full 3D ray tracing

Wireless InSite [9] provides a full 3D ray-based propagation model, as introduced in Chapter 5. Based on the full 3D ray tracing methods, the multipath channel model established previously can be simulated. Figure 6.4 shows an example of the propagation paths in the 3D building model after a full 3D ray tracing simulation.
6.2.3 **AOA data collection and analysis**

The first simulation case concerns receivers with RHCP antennas, while the remaining parameters were kept the same as those described above. The data obtained from the multipath model are the AOAs and the associated angular power distribution. Figure 6.5 shows the received power versus the AOA in the azimuth plane. More than 20,000 sample path rays were collected. It can be seen in Figure 6.5 that the received power is mostly between -70dBm and -110dBm. The minimum, maximum, mean and standard deviation of the received power are summarized in Table 6-B. The received power density is almost evenly spread in the azimuth plane (0° to 360°). It is noticed that the probabilities for the power received between 210° to 360° are slightly less than other angles.

The level of received power in the elevation plane is similar to that in the azimuth plane (between -70dBm and -110dBm), as shown in Figure 6.6. The received power is mostly confined within a particular range of elevation angles (60° to 100°, where 0° is

<table>
<thead>
<tr>
<th>$min$</th>
<th>$max$</th>
<th>$mean$</th>
<th>$std$</th>
</tr>
</thead>
<tbody>
<tr>
<td>-177.03dBm</td>
<td>-56.88dBm</td>
<td>-94.82dBm</td>
<td>15.97dBm</td>
</tr>
</tbody>
</table>
the zenith and $90^\circ$ the horizon). These results indicate that the angles of the received power in the elevation plane are nearly unchanged, no matter where the receivers are located, so it is of interest to identify the distribution of the AOAs in the elevation plane as well as in the azimuth plane.

Figure 6.7 shows the histogram of received signals in azimuth plane. The received power distribution is not quite uniformly distributed as one would expect. It has been
shown that there are more signals coming from $30^\circ$ to $210^\circ$ than any other angles. This is mainly due to the fact that only a single indoor environment has been investigated here. Further investigations on this issue, i.e. modelling a large number of different indoor environments are necessary. However, the uniform distribution of AOA at azimuth plane is adopted in this study, as used for wireless communication.

The probability density of received signals in the elevation plane, together with the Gaussian distribution curve, is shown in Figure 6.8. The histogram is reasonably fitted to the Gaussian distribution. However it has a higher and sharper peak compared with that of Gaussian distribution and the Laplacian distribution is better fitted to the results. The mean and standard deviation are $11^\circ$ and $16^\circ$, respectively.

After identifying the AOA distributions, a set of parameters as used for wireless mobile communications in Chapter 3 has been established to evaluate the diversity performance of the Galileo system in an indoor environment. Table 6-C shows the values of AOA and XPR in the case of an indoor environment for the Galileo system. Due to the limitations of the simulation tool, the powers received in the horizontal polarisation plane ($P_H$) and vertical polarisation plane ($P_V$) cannot be separately determined. For
Figure 6.8: Histogram of received signals along with Gaussian distribution (red curve) and Laplacian distribution (green curve) in elevation plane where $0^\circ$ is the zenith and $90^\circ$ the horizon.

Table 6-C: Values of AOA and XPR in an indoor environment for the Galileo system.

<table>
<thead>
<tr>
<th>Scenario</th>
<th>Statistic model (Elevation/Azimuth)</th>
<th>Laplacian/Uniform</th>
</tr>
</thead>
</table>
| Galileo Indoor  | $m_V = 11^\circ$  
                 | $m_H = 11^\circ$  
                 | $\sigma_V = 16^\circ$  
                 | $\sigma_H = 16^\circ$  
                 | $XPR = 1\text{dB}$  
|                 | $m_V = 11^\circ$  
                 | $m_H = 11^\circ$  
                 | $\sigma_V = 16^\circ$  
                 | $\sigma_H = 16^\circ$  
                 | $XPR = 1\text{dB}$  |

this reason, the XPR presented in Table 6-C is same as that for wireless mobile communications. However, the effect of different XPR values on the diversity performance are discussed in the next section.

### 6.3 Diversity performance of the dual-helical antenna array for Galileo/GPS systems

The diversity performance of the dual-helical antenna array has been evaluated by calculating the correlation, MEG and diversity gain of the antenna system.
6.3.1 Correlation and MEG of the dual-helical antennas

3D radiation patterns are required to evaluate the correlation using equation (3.9) and (3.10) in Chapter 3. This work is mainly focused on receiving diversity technology for the purpose of improving the performance of the Galileo system in an indoor environment. Because of the time limitations of this work, only simulated results have been used for this analysis. From earlier discussions in Chapter 4, the results of simulations using CST Microwave Studio® have good agreement with the measured results. It is therefore considered to be reliable to use the simulated 3D radiation patterns to evaluate the diversity performance of the antennas. These patterns have been computed with 5° step elevation cuts while the dual-helical antennas is positioned vertically in free space. The correlations of signals received by the dual-element helical antennas are summarised in Table 6-D. The impact of the Galileo indoor environment on the envelope correlation has been evaluated using two different statistical models as discussed in Chapter 3, i.e. Gaussian/Uniform and Laplacian/Uniform distributions. It is noted that there is not much difference between the correlation values when evaluated by the two statistical models. Table 6-D shows that an envelope correlation of less than 0.4 has been achieved for the dual-helical antenna array.

The MEG of each antenna is evaluated from the 3D gain patterns using equation (3.15) and the results are tabulated in Table 6-D. The MEG values for the Gaussian/Uniform and Laplacian/Uniform statistical models have less than 1dB difference in the Galileo indoor environment, indicating that the dual-element helical diversity antenna array works well in environments with either Gaussian/Uniform or Laplacian/Uniform

<table>
<thead>
<tr>
<th>Galileo Indoor Propagation Models</th>
<th>Envelope correlation, $\rho_e$</th>
<th>MEG, dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gaussian/Uniform</td>
<td>0.3091</td>
<td>-2.2320</td>
</tr>
<tr>
<td>Laplacian/Uniform</td>
<td>0.3151</td>
<td>-3.0271</td>
</tr>
</tbody>
</table>

Table 6-D: The envelope correlation and MEG of the dual-element helical diversity antenna array.
distribution. Also, the MEG values of two helical antennas are almost identical due to its same polarisation and symmetric simulated radiation patterns.

### 6.3.2 Diversity gain of the dual-helical antennas

After assessing the correlation and MEG results, the diversity gain of the dual-element helical antenna array can be evaluated as described in Chapter 3. The diversity gains using selection combining are shown in Table 6-E. The diversity gain results include the degradation factor (DF) in equation (3.12) due to the signal correlation, and the branch power ratio (k) in equation (3.14).

In an ideal two-branch selection combiner, the diversity gain at 99% reliability is 10dB, as shown in Figure 3.18 in Chapter 3. Table 6-E shows that the diversity gains of the dual-element helical diversity antenna array are less than 10dB. As the variation

<table>
<thead>
<tr>
<th>Galileo Indoor Propagation Models</th>
<th>DF, dB</th>
<th>K, dB</th>
<th>Diversity Gain, dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gaussian/Uniform</td>
<td>-0.8029</td>
<td>-0.0082</td>
<td>9.1889</td>
</tr>
<tr>
<td>Laplacian/Uniform</td>
<td>-0.8219</td>
<td>-0.0098</td>
<td>9.1683</td>
</tr>
</tbody>
</table>

Figure 6.9: Diversity gain of two ideal antennas and the dual-element helical diversity antenna system compared to a single antenna.
of the MEG between two helical antennas is very small in all cases as shown in Table 6-D, the diversity gain is slightly degraded by the correlation. The degradation is less than 1dB. In Figure 6.9, the diversity gain of two ideal antennas and the dual-element helical diversity antenna system are illustrated in terms of the increase in the SNR of a combined output compared to a single antenna. There is only 0.9dB degradation of the diversity gain of the dual-element helical diversity antenna array compared to that of the two ideal antennas at most reliabilities.

6.3.3 The effect of XPR on the diversity performance

In the previous section, an XPR of 1dB has been assumed for the Galileo indoor environment. The XPR could vary in different fading environments and frequency bands \cite{10}, so the effect of the XPR on the diversity performance is studied here. Figure 6.10 shows a comparison of MEG values using Gaussian/Uniform and Laplacian/Uniform distribution against XPR varying from -10dB to 10dB. When the XPR is greater than 0dB, the average power of vertically polarised RF signals is stronger than the average power.
of horizontally polarised signals, and vice versa. There is no more than 1dB difference in MEG using Gaussian/Uniform and Laplacian/Uniform distributions, as shown Figure 6.10, which correlates well to that in the case of the dual-element PIFA array in the Table 5-A of Chapter 5. The comparison results in Figure 6.10 show that both Antenna 1 and 2 have higher MEG in the vertical polarisation dominated region. This is due to both helical antennas being vertically placed on the terminal. The MEG values of Antenna 1 and 2 are almost identical when the XPR is greater than 1dB. Therefore, an XPR of 1dB is a reasonable value selected for the Galileo indoor environment.

6.4 Summary

As a preliminary study in the GAC project for 8 months, a statistical antenna diversity analysis model for the Galileo indoor propagation environment has been established. The key parameters, such as AOA and XPR of this model were specified for the Galileo system and the antenna configuration studied is representative of one that could be used in practice for a small terminal.

The AOA distributions were characterised using a multipath channel model based on a ray tracing method using a set of parameters based on measurements for the Galileo system. From the statistical analysis of the AOA obtained, the mean and standard deviation of AOA distributions are 11° and 16°, which are different with those in wireless communications (20° and 30°). It was noticed that a Laplacian distribution has a slightly better fit to the AOAs in the elevation plane, but the practical difference was shown to be small.

A statistical antenna diversity analysis model was used to demonstrate the potential performance improvement that could be provided by the use of a simple diversity receiving system in the Galileo indoor environment. It has been shown that the dual-element helical array obtained a diversity gain of up to 9dB in the Galileo system compared to a single-element antenna at 99% reliability assuming XPR=1dB. Following a study on
the effect of XPR on the MEG values, it was shown that the MEG values of Antenna 1 and 2 are almost identical when the XPR is greater than 1dB.

References


Chapter 7

Conclusions and future work

7.1 Summary

The MIMO system has shown its ability to significantly increase channel capacity and enhance reliability of wireless channels without increasing the transmitter signal power and spectrum usage. The data rate and coverage of a MIMO system can be improved by using spatial multiplexing and space time coding, respectively, to exploit the MIMO channels. Therefore, WLAN and WiMAX have employed MIMO technology in their systems and mobile communication systems such as 3G and 4G will implement it in the future. Although significant amounts of research have been carried out on signal processing algorithms and channel modelling for MIMO systems by assuming ideal half-wavelength dipoles, it has been indicated that practical multiple antennas on a small mobile terminal play an important role on implementing the MIMO system in mobile communication systems. However, only limited research has been conducted on multi-antenna designs and channel modelling involving the practical multi-antenna for MIMO systems. Therefore, the design of multiple antennas on a small terminal and the model of MIMO channels including these multiple antennas were carried out in this thesis.
systems, the requirements for designing multiple antennas such as having a high diversity gain and low mutual coupling was addressed in this thesis. The diversity technique which uses more than one antenna to receive or transmit signals, being a well known method for solving the signal fading problem in a multipath environment, is utilised in space time coding to exploit the MIMO channels. The diversity gain is a measure of the effectiveness of the diversity technique. High diversity gain can be achieved when the received signals from two or more antennas have low correlations; the power levels of the signals received by two or more antennas should not be too different in the diversity system in a multipath environment. However, there is no clear design criterion to achieve low mutual coupling in a sized-limited mobile terminal at the moment. For minimum coupling, the separation between antennas needs to be at least half-wavelength in theory.

By studying the three diversity techniques such as spatial, pattern and polarisation diversity, it has been shown that there is no direct relation between the mutual coupling and diversity gain. For instance, pattern diversity antennas obtain different radiation patterns through the mutual coupling effect. Low correlation and high diversity gain can be achieved, although the pattern diversity antennas have a strong mutual coupling. However, mutual coupling between antennas degrades not only the efficiency of the independent antennas, but also the efficiency of the diversity system, especially for a power-constrained mobile terminal. Therefore, multiple antennas on a small mobile terminal should satisfy the requirements of having both high diversity gain and low mutual coupling.

A conventional PIFA is commonly used nowadays in a mobile terminal due to its compact size and good performance. However, it generates strong current flow on the ground plane which is the PCB of a mobile terminal, so the ground plane is acting as a radiator rather than a reflector. A novel PIFA design and its array, which generate only a small amount of current on the ground plane, were proposed in this thesis. By introducing a small ground plane between the PIFA and the PCB, both the coupling between the PIFA and the PCB and the coupling between the PIFAs have been significantly reduced.
The dual-element PIFA array operating at 2.5GHz with a separation of 0.17 wavelength can achieve an isolation of 20dB. Furthermore, the diversity performance was evaluated with selection combiner technique and it was found that a diversity gain of 9.3dB at 99% reliability was achieved in both simulation and measurement.

A RT MIMO channel model in an indoor environment was established to investigate the channel capacity of practical antenna arrays on a mobile terminal such as the dual-element PIFA array for MIMO systems. The model was verified with the IEEE 802.11 MIMO channel model. The channel capacity performances of the proposed dual-element PIFA array, the conventional dual-element PIFA array and two ideal dipoles were studied in the RT MIMO channel model in an indoor environment. The channel capacity achieved by dual-element PIFA array is more than 1.5 times than that of a SISO system. Compared to two ideal dipoles without including their mutual coupling, the channel capacity of the proposed dual-element PIFA array was slightly lower when the mutual coupling was taken into consideration. However, the channel capacity of proposed dual-element PIFA was higher than that of the conventional dual-element PIFA which has a strong mutual coupling. Therefore, the mutual coupling effect slightly degraded the MIMO channel capacity in this case.

For the application of multiple antennas on Galileo/GPS terminals, a dual-helical antenna array was proposed. To properly evaluate the diversity performance of the dual-helical antenna array, the propagation factors such AOA and XPR were characterised for the Galileo/GPS system by a multipath channel model. The mean and standard deviation of AOA distributions were 11° and 16° for the Galileo/GPS indoor environment, which are different from the wireless indoor environments (20° and 30°). Based on the characterised propagation factors, the diversity performance of the dual-helical antenna array was evaluated. Although the mutual coupling of the dual-helical antenna array was as high as -6dB, a diversity gain of 9.1dB was achieved at 99%. Therefore, the dual-helical antenna array is not an ideal candidate from the system point of view.
7.2 Key contributions

The major contributions in this thesis are detailed in the following three sections:

**Dual-element PIFA arrays**

- A novel and compact PIFA design was proposed and realised. The ground plane, as small as the antenna, was located between the PIFA and the PCB so that the PCB was no longer acting as a ground plane for the PIFA. The modified PIFA has little coupling to the PCB, and therefore is suitable for multiple antenna implementation on small terminals.

- A dual-element modified PIFA array on a handset PCB operating at 5.2GHz was proposed and studied in both simulation and experiment. Though the separation between two PIFAs was $0.35\lambda$, an isolation of more than 28dB was achieved. The modified dual-element PIFA array provides a good solution for mobile terminals that employ the MIMO technology.

- A dual-element modified PIFA array on a handset PCB operating at 2.5GHz was also investigated, and the results have further confirmed the design at 5.2GHz. An isolation of 20dB was achieved even the two PIFAs were only separated by $0.17\lambda$. The diversity performance of the proposed dual-element PIFA array was evaluated through simulation and measurement. A maximum SC diversity gain of 9.3dB at 99% reliability was achieved in an isotropic environment.

**RT MIMO channel model**

- A RT MIMO channel model in an indoor environment was established and verified with the IEEE MIMO channel model. The model had the ability to evaluate the channel capacity performance of different practical antenna arrays such as the proposed dual-element PIFA array, the conventional dual-
element PIFA array, two ideal dipoles, etc.

- The RT MIMO channel model has shown that the proposed dual-element PIFA array can achieve a channel capacity 1.5 times more that of a SISO system. It is also shown that the mutual coupling of the multiple antennas degraded the MIMO channel capacity slightly.

Application of multiple antennas on Galileo/GPS terminals

- The concept of antenna diversity technology was introduced into the Galileo navigation system for the first time in the GAC project. The propagation factors such as AOA and XPR were obtained for the Galileo/GPS system.

- A simple dual-helical antenna array was proposed and its diversity performance was evaluated based on the characterised propagation factors. A SC diversity gain of up to 9.1dB was obtained in the Galileo system compared to a single-element antenna system at 99% reliability.

7.3 Future work

MIMO technology is a vast research topic at the moment. The work done in this thesis only covers a small part of it due to the time constraint. Many aspects of this work can be further improved. Also, there are many important areas in MIMO technology that need to be further investigated. Future work could be carried out in the following areas.

- In the proposed PIFA design, there was an air gap between the top plate and the small ground plane. A dielectric filling could be inserted to further reduce the size of the antenna.

- The dual-element PIFA array was considered for a mobile terminal. Therefore, the user’s effect on the diversity antenna’s performance should be studied. Different situations have to be considered, e.g. talking position and inside the pocket. In
the talking position, the SAR (Specific Absorption Rate) value inside the user’s head caused by the antenna arrays should also be considered.

- Recently, the DAIC (Dual Antenna Interference Cancellation) has been considered for GSM networks with an aim to reduce the impact of radio interference. This would increase average bit rates and radio coverage. Therefore, research on design of a dual-element receive diversity array on a GSM handset could be carried out.

- The MIMO channel capacity presented in Chapter 5 was only evaluated in an indoor environment. The channel capacity of the dual-element PIFA could be investigated in different environments such as the outdoor environment.

- The dual-helical antenna array was not a ideal candidate for the Galileo/GPS application. Therefore, a dual-element diversity antenna array which has a low mutual coupling level and high diversity performance should be considered.

- A larger number of elements of the proposed PIFA design could be considered for PDA terminals and laptops, where more space are available.
Appendix A

Author’s publications

Journal papers


Conference papers


**Project report**


Appendix B

Solutions for the Examples in Chapter 2

B.1 Example 1: Antenna Spacing Effect

**Step 1:** Get the $H_{\text{norm}}H_{\text{norm}}^\dagger$, set $2\pi R/\lambda = \omega_1$ and $2\pi(R)/\lambda = \omega_2$ ($\lambda$ is the wavelength), so

$$H_{\text{norm}} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{-j\omega_1} & e^{-j\omega_2} \\ e^{-j\omega_2} & e^{-j\omega_1} \end{pmatrix}$$  \hspace{1cm} (B.1)

Note: $H_{\text{norm}}^\dagger = (H_{\text{norm}})^T$ and $e^{j\theta} = \cos\theta + js\sin\theta$

$$H_{\text{norm}}H_{\text{norm}}^\dagger = \frac{1}{2} \begin{pmatrix} e^{-j\omega_1} & e^{-j\omega_2} \\ e^{-j\omega_2} & e^{-j\omega_1} \end{pmatrix} \begin{pmatrix} e^{j\omega_1} & e^{j\omega_2} \\ e^{j\omega_2} & e^{j\omega_1} \end{pmatrix} = \frac{1}{2} \begin{pmatrix} 1 + 1 & e^{-j(\omega_1-\omega_2)} + e^{-j(\omega_1-\omega_2)} \\ e^{-j(\omega_1-\omega_2)} + e^{-j(\omega_1-\omega_2)} & 1 + 1 \end{pmatrix}$$  \hspace{1cm} (B.2)

$$= \begin{pmatrix} 1 & \cos(\omega_1 - \omega_2) \\ \cos(\omega_1 - \omega_2) & 1 \end{pmatrix}$$
Step 2: Get the eigenvalue of $H_{\text{norm}}H_{\text{norm}}^\dagger$, here set $\lambda$ as eigenvalue, then,

$$
\begin{pmatrix}
1 - \lambda & \cos(\omega_1 - \omega_2) \\
\cos(\omega_1 - \omega_2) & 1 - \lambda
\end{pmatrix} = 0
$$

(B.3)

and $\omega_1 - \omega_2 = 2\pi R/\lambda - 2\pi(\bar{R})/\lambda = -54^\circ$, so $\lambda = 1\pm\cos(\omega_1 - \omega_2)\approx1\pm0.59$. Hence, $\lambda_1 = 1.59$ and $\lambda_2 = 0.41$.

B.2 Example 2: Environment Effect

Step 1: set $2\pi/\lambda = \omega$, then

$$
H_{\text{LOS}} = \begin{pmatrix} e^{-j\omega R} & e^{-j\omega \bar{R}} \\
e^{-j\omega R} & e^{-j\omega \bar{R}} \end{pmatrix} \quad \text{and} \quad H_{\text{NLOS}} = 0.6 \begin{pmatrix} e^{-j\omega (P_1+P_3)} & e^{-j\omega (P_1+P_4)} \\
e^{-j\omega (P_2+P_3)} & e^{-j\omega (P_1+P_4)} \end{pmatrix}
$$

Step 2: get the $H_R$,

$$
H_R = H_{\text{LOS}} + H_{\text{NLOS}} = \begin{pmatrix} -0.6114 - i1.4556 & -0.3942 - i0.2665 \\
-0.3942 - i0.2665 & -0.6452 - i1.4482 \end{pmatrix}
$$

Step 3: normalise $H_R$, $\sum_{j=1}^2 |h'_{ij}|^2 = 1$,

Set $H_R = \begin{pmatrix} a & b \\ c & d \end{pmatrix}$ and $H_{\text{modulus}} = \begin{pmatrix} A & B \\ C & D \end{pmatrix} = \begin{pmatrix} 1.5788 & 0.4758 \\ 0.4758 & 1.5854 \end{pmatrix}$

So $H_{\text{Norm}} = \begin{pmatrix} \frac{a}{\sqrt{A^2 + B^2}} & \frac{b}{\sqrt{A^2 + B^2}} \\ \frac{c}{\sqrt{C^2 + D^2}} & \frac{d}{\sqrt{C^2 + D^2}} \end{pmatrix}$

Proof for the Normalisation:

$$
\left| \frac{a}{\sqrt{A^2 + B^2}} \right|^2 + \left| \frac{b}{\sqrt{A^2 + B^2}} \right|^2 = \frac{|a|^2 + |b|^2}{A^2 + B^2} = 1 \quad , \quad \left| \frac{c}{\sqrt{C^2 + D^2}} \right|^2 + \left| \frac{d}{\sqrt{C^2 + D^2}} \right|^2 = \frac{|c|^2 + |d|^2}{C^2 + D^2} = 1
$$
Step 4: the normalized $H_R$ is,

$$H_{Norm}H_{Norm}^\dagger \approx \begin{pmatrix} 1 & 0.4650 + i0.0044 \\ 0.4650 - i0.0044 & 1 \end{pmatrix}$$

Step 5: get the eigenvalue of $H_{Norm}H_{Norm}^\dagger$: $\lambda_1 = 1.54$ and $\lambda_2 = 0.46$. 
Appendix C

Introduction to the methods used in Wireless InSite

Wireless InSite provides a full 3D ray-based propagation model \[1\]. This model combines ray tracing algorithms with the UTD. The methods which have been used in Wireless InSite (i.e. SBR, method of images and identifying diffractions), are described below.

C.1 Shooting and bouncing ray method

The SBR method is a high frequency electromagnetic simulation technique for predicting the radar scattering from realistic targets and propagation in complex environments. The SBR method is based on geometrical optics. Given the geometrical description of a target, a large set of rays is shot towards the target. The rays are traced according to the laws of geometrical optics as they are scattered around the target. At the exit point of each ray, a ray-tube integration is performed to sum up its contribution to the total scattered field.

The SBR method is employed to trace the ray paths through the two-dimensional
Appendix C. Introduction to the methods used in Wireless InSite

building geometry [2, 3]. Ray paths are first traced from the source points with the rays reflecting specularly from the building walls and transmitting through the building walls with no change in the direction as shown in Figure C.1.

![Figure C.1: A single ray emitted from the transmitter generates many rays as it splits into reflected and transmitted rays at each exterior surface [1].](image)

C.2 Two different diffraction situations

In this part, the procedure of identifying diffracting edges in the full 3-D ray tracing of Wireless InSite is described. Diffraction occurs at the points where the field becomes discontinuous in the GTD (Geometrical Theory of Diffraction). The first order diffraction edges are found by searching for adjacent rays which follow different paths through the building geometry, since such occurrences identify discontinuities in the GO (Geometrical Optics) field. A diffraction edge can then be located between these rays.

There are two different kind of ways to identify diffracting edges in Wireless InSite. Figure C.2 shows two examples of how the rays transmitted from the transmitter are used to identify diffracting edges. For instance, the two adjacent SBR rays in Figure C.2(a) both reflect from Building 1 face 11, but only one reflects from Building 2 face 22. This means that a diffracting edge lies between these two rays, and it is then quite simple to locate the diffraction point on edge \{21-22\} and to construct the path followed by
Appendix C. Introduction to the methods used in Wireless InSite

the incident field. A somewhat different diffraction situation is shown in Figure C.2(b), where the edge \{61-62\} would be identified as a diffracting edge for the incident field which first reflects from Building 4 face 41. The two situations differ in that the latter example has two reflection shadow boundaries, whereas the diffraction in Figure C.2(a) has an incident shadow boundary and a reflection shadow boundary.

![Diagram](a)

![Diagram](b)

Figure C.2: Rays which identify edge (a) faces \{21-22\} and (b) faces \{61-62\} as a diffraction edge. Two different diffraction situations [1].

C.3 Method of images

There are a number of ray tracing methodologies available. One of these is SBR, in which a finite number of rays is launched from the transmitter in predetermined directions, and traced until they reach the vicinity of the receiver, as described above. The method of
images is described in this section.

Figure C.3(a) shows the three simplest multipath routes in a corridor model, involving zero, one and two reflections. Clearly an infinite number of such routes exit, and indeed this scenario is of interest because it is the simplest in which such infinite reflections can occur. Figure C.3(b) shows how these multipath routes can be determined using the method of images. The points marked in pale green are images of the transmitter, of increasing order as we move outward from the actual transmitter (marked in green). The dotted lines from the images to the receiver (marked in red) show the virtual paths of the reflected rays. The length of the actual multipath routes are equal to the lengths of these virtual routes [4].

![Figure C.3](image_url)

Figure C.3: (a) Simple ray tracing example with transmitter and receiver symmetrically placed between two parallel reflecting surfaces, showing three multipaths. (b) Prediction of multipath routes using method of images [4].

References

