Abstract

In this digital era, the usage of smart phones and mobile devices is becoming a norm in society with mobile communication quickly transitioned from voice oriented transmission to picture transmission to a more complex live video streaming. This latest development has demanded more capacity and higher bandwidth in communication links. Static links, which are the focus of this thesis, are an integral part of this mobile system in delivering high capacity data transmission using backhauls or nomadic links. Multi polarised antennas with multiple-input multiple-output (MIMO) multiplexing can be employed to greatly enhance the capacity of a mobile system, especially at frequencies lower than 6 GHz, using their compact size.

A practical antenna inherently exhibits elliptical polarisation though it may be designed to form linear or circular polarisation. Little attention has been given to this aspect of polarised waves as they have always been deemed as unwanted polarisation, although in practice, any antenna is elliptically polarised as it can never be perfectly circularly or linearly polarised. This work therefore aims to deliberately exploit this opportunity by forming antennas with elliptical polarisation to identify the advantages of doing so in order to improve orthogonality in comparison with linear polarisation. It was found that in order to achieve perfect orthogonality, it was more practical to set the magnitudes and phases of the co-polar and cross-polar linear components, which resulted in an improved co to cross polar ratio more than 20 dB better than linear polarisation in free space.

A dual elliptically polarised antenna prototype was designed and evaluated in this work, which was evaluated both in free space and within an indoor measurement campaign. Results concluded that at short distances with low scattering in the channel and directional antennas, elliptically polarised antennas provide improved multiplexing gain over dual linear polarisations.

Key words: Elliptical Polarisation, MIMO, Multiplexing, static wireless links, degrees of freedom, dual polarised antenna.

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<tbody>
<tr>
<td>$E$</td>
<td>Electric-field</td>
</tr>
<tr>
<td>$M$</td>
<td>Number of transmitter</td>
</tr>
<tr>
<td>$N$</td>
<td>Number of receiver</td>
</tr>
<tr>
<td>$P_r$</td>
<td>received power</td>
</tr>
<tr>
<td>$P_t$</td>
<td>transmit power</td>
</tr>
<tr>
<td>$G_t$</td>
<td>transmit gain</td>
</tr>
<tr>
<td>$G_r$</td>
<td>receive gain</td>
</tr>
<tr>
<td>$\lambda$</td>
<td>wavelength</td>
</tr>
<tr>
<td>$R$</td>
<td>distance between TX and RX</td>
</tr>
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<td>$H$</td>
<td>channel</td>
</tr>
<tr>
<td>$\rho$</td>
<td>SNR</td>
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<td>$\rho_{12}$</td>
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<td>$\mathbf{x}$</td>
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<td>$AR$</td>
<td>axial ratio</td>
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<td>$\phi$</td>
<td>phase shift</td>
</tr>
<tr>
<td>$k_1$</td>
<td>co-polar branch imbalance</td>
</tr>
<tr>
<td>$k_2$</td>
<td>cross-polar branch imbalance</td>
</tr>
<tr>
<td>$l$</td>
<td>length of dipole arm</td>
</tr>
<tr>
<td>$t$</td>
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<tr>
<td>$S_{11}$</td>
<td>return loss at Port 1</td>
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<td>$S_{22}$</td>
<td>return loss at Port 2</td>
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<tr>
<td>$S_{21}$</td>
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<td>$\lambda$</td>
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<tr>
<td>$XPR_C$</td>
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List of abbreviations

- 2D: Two-dimensional
- 3D: Three-dimensional
- 5G: fifth generation
- AUT: antenna under test
- BS: base station
- CDF: cumulative distribution function
- CP: circular polarisation
- CSI: channel state information
- dB: decibel
- DCPM: Dual Circular Polarisation Multiplexing
- DoF: Degrees of Freedom
- EP: elliptical polarisation
- ESPAR: Electronically Steerable Parasitic Array Radiator
- HP: horizontal polarisation
- IEEE: Institute of Electrical and Electronic Engineers
- LHCP: left hand circular polarisation
<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>LOS</td>
<td>line-of-sight</td>
</tr>
<tr>
<td>LP</td>
<td>linear polarisation</td>
</tr>
<tr>
<td>MIMO</td>
<td>multiple-input multiple-output</td>
</tr>
<tr>
<td>mmWave</td>
<td>millimetre wave</td>
</tr>
<tr>
<td>MPtoMP</td>
<td>multipoint-to-multipoint</td>
</tr>
<tr>
<td>MU-MIMO</td>
<td>multi-user MIMO</td>
</tr>
<tr>
<td>NLOS</td>
<td>non-line-of-sight</td>
</tr>
<tr>
<td>OLOS</td>
<td>obstructed-line-of-sight</td>
</tr>
<tr>
<td>PSTN</td>
<td>Public Switched Telephone Network</td>
</tr>
<tr>
<td>PtoMP</td>
<td>point-to-multipoint</td>
</tr>
<tr>
<td>PtoP</td>
<td>Point-to-Point</td>
</tr>
<tr>
<td>QHA</td>
<td>quadrifilar helical antenna</td>
</tr>
<tr>
<td>RHCP</td>
<td>right hand circular polarisation</td>
</tr>
<tr>
<td>RX</td>
<td>Receiver</td>
</tr>
<tr>
<td>SIMO</td>
<td>single-input-multiple-output</td>
</tr>
<tr>
<td>SISO</td>
<td>single-input single-output</td>
</tr>
<tr>
<td>SNR</td>
<td>signal-to-noise-ratio</td>
</tr>
<tr>
<td>SPA</td>
<td>switched parasitic antenna</td>
</tr>
<tr>
<td>SVD</td>
<td>singular value decomposition</td>
</tr>
<tr>
<td>TDM</td>
<td>time-division multiplexing</td>
</tr>
<tr>
<td>TV</td>
<td>Television</td>
</tr>
<tr>
<td>TX</td>
<td>Transmitter</td>
</tr>
<tr>
<td>ULA</td>
<td>Uniform Linear Array</td>
</tr>
<tr>
<td>VP</td>
<td>vertical polarisation</td>
</tr>
<tr>
<td>WLAN</td>
<td>Wireless local area network</td>
</tr>
<tr>
<td>XPR</td>
<td>Cross polarisation power ratio</td>
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</tbody>
</table>
Chapter 1

1 Introduction

Demand for high bandwidth and capacity in wireless links has exploded in recent years due to the increased usage of smart phones and mobile devices. As a result, these demands have led to the massive growth of wireless data traffic, which are comprised of both mobile and static wireless links. Static links, which are used to deliver high capacity data transmission, are the focus of this thesis as this research not only examines applications that are permanently fixed, but are also stationary for a certain period of time and may move to another static position (i.e. a nomadic wireless link). Therefore, in this thesis, different types of static links are elaborated: large scale fixed radio links (sub 6 GHz and millimetre wave (mmWave) backhauling); medium scale nomadic links (Small cell and Wireless local area network (WLAN) access links); and small scale nomadic links (short range device to device).

To backhaul is to transmit data from the wireless mesh network at the edge of the system to the core wired network and to aggregate all the traffic over one or more high speed lines. Traditionally, backhaul has been carried out using the sub 6 GHz frequencies. However, current and future cellular deployments exploit mmWave frequency bands to deliver packet microwave links in the backhaul, where multi-Gigabit capacity can be found. Macro cells can be integrated with small cells to enhance the coverage for a dense area as small cells add extra spectrum reuse by serving a smaller number of users in each cell. This requires a more effective, scalable and flexible backhauling solution to deploy small cells in dense areas where fibre is not available.

WLAN is one of the popular solutions for delivering wireless network links especially for commercial users because it is easier to be installed and used. It also employs multiple-input multiple-output (MIMO) techniques to improve on the throughput, whereby multiple data streams can be transmitted simultaneously and at the same time and interference between users can be nullified. In short range static links with a line of sight, MIMO systems are able to enhance the throughput capacity by designing the antennas in such a way to optimise the multiplexing gain of the system. Methods such as spatial, angular and polarisation can be utilised to improve the
multiplexing gain. Spatial multiplexing can exploit the spatial dimension of the antenna arrays to produce maximum capacity but the antenna arrays have wide antenna spacing which is undesirable at sub 6 GHz. Angular multiplexing with flexible beams removes the minimum spacing requirement of MIMO systems but compared with antenna arrays with fixed beams providing an array gain in a line of sight, they provide lower data throughput. At frequencies lower than 6 GHz, polarisation in a line of sight can help to substantially improve spectral efficiency whilst being a compact antenna solution. By making use of dual orthogonal polarised antennas, interference from other nearby antennas with opposite polarisations can also be greatly reduced, if not eliminated, thus improving the overall quality of the system.

1.1 Motivation

After comparing all methods of MIMO multiplexing, a dual polarisation multiplexing technique is found to be the best solution to stream multiple data simultaneously at sub 6 GHz in a line of sight due to its compactness. There are three main types of polarisation: elliptical, linear and circular polarisation. For any electromagnetic wave, the locus of its electric-field, $E$, is an ellipse and therefore, it is an elliptically polarised wave. The ellipse can theoretically form either a straight line or circular locus. In these conditions, the polarisation state is linear or circular, respectively. A wave then inherently exhibits elliptical polarisation if it is not modified to become linear or circular polarisation and hence has an elliptical locus. Little or no attention has been given to this aspect of polarisation which has always been deemed as an unwanted polarisation, yet in practice any antenna is elliptically polarised as it can never be perfectly circularly or linearly polarised. This work aims to deliberately exploit this opportunity and identify the advantage of using elliptical polarisation by investigating the improvement in orthogonality required for multiplexing.

Therefore, there is a further degree of freedom in the phase components of elliptical polarisation that can be exploited as a potential method to carry out MIMO multiplexing, with the improvement of better orthogonality in comparison to linear polarisation. Moreover, any practical antenna may be designed to be linear or circular, but now the focus must be to design antennas to be elliptical in order to enhance multiplexing gain. This research forms metrics to evaluate elliptically polarised antennas at the transmitter and receiver in order to identify where they have the advantage to multiplex better than linear or circular polarisation.
In a real environment, scatterings will arise due to ground reflections, diffractions and refractions from surrounding obstacles. Consequently, there will be line-of-sight (LOS) and non-line-of-sight (NLOS) propagations. Static wireless links will undoubtedly also suffer from these scatterings and multipath. In a NLOS scenario, there will be significantly more scattering that deconstructs the orthogonality of the signals of any dual polarised antennas, however, this provides an opportunity to exploit the spatial, angular and polarisation diversity richness in the radio environment.

Taking these findings into consideration, a dual elliptically polarised antenna candidate that is co-located was designed and evaluated in this work. At short distances with lower multipath, a directional elliptically polarised antenna is found to contribute to good multiplexing. Elliptical polarisation multiplexing fails if these conditions are not fulfilled, leading to weaker orthogonality between the polarisations. The increased multiplexing gain then results in higher capacity in the 2x2 MIMO system formed. Alternatively, an elliptically polarised antenna can act as an angular diversity antenna in a real environment with severe multipath such as in a NLOS scenario. As the criteria required for elliptical polarisation multiplexing can be found in mmWave communication links, the latter part of this work investigates the feasibility of using elliptical polarisation in such bands.

### 1.2 Original contributions to knowledge

The following are the original contributions to knowledge included in this thesis:

- Derivation of cross polarisation power ratio for elliptical polarisation ($XPR_e$) as a measure of orthogonality between two ellipses.
- Design and fabrication of novel elliptically polarised STAR dipole antenna as a prototype to be used to evaluate elliptical polarisation multiplexing.
- Evaluation of elliptically polarised antennas in free space to prove the ability of elliptical polarisation to increase the orthogonality more than 100 times that of linear polarisation and reach the capacity bound of a 2x2 MIMO link.
- Channel characterisation of elliptically polarised antennas in a real environment to show the ability of elliptical polarisation multiplexing for static links in environments with directional antennas, short range and low scattering.
• Analysis of the feasibility of elliptical polarisation multiplexing potential for static links at mmWave frequency bands.

1.3 Publications

The following papers are being planned for publication:

• “Exploitation of Elliptical Polarisation for Substantially Improved Orthogonality”. In preparation for submission to the IEEE Transactions on Antennas and Propagation.
• “Feasibility of Elliptical Polarisation Multiplexing at Millimetre wave Frequency Band”. To be submitted to a conference publication.

1.4 Structure of thesis

The rest of this thesis is structured as follows. Chapter 2 presents a literature review of the application of this research, the static wireless link communication. MIMO multiplexing and diversity techniques comprising of spatial, angular and polarisation techniques are also discussed as background theory. Dual polarised antenna design candidates at sub 6 GHz for static link are also reviewed, including patch, dipole, helix, monopole and slot antennas. This is followed by the discussion on multiplexing techniques for mmWave backhaul.

Chapter 3 introduces the theory of elliptical polarisation and investigates from a theoretical perspective how exploiting the degrees of freedom in the elliptically polarised wave provides a suitable multiplexing technique with strong orthogonality. This chapter also derives the cross polarisation power ratio for elliptical polarisation ($XPR_{E}$).

In chapter 4, a dual elliptically polarised antenna candidate known as STAR dipole, which is a dipole design re-arranged into a star shape, and generates elliptical polarisation, is presented with simulated and measured results of a prototype. By comparison with spatially separated vertical and horizontal linearly polarised dipoles, an evaluation of multiplexing capability of linear and elliptical multiplexing capability was performed in a free space environment.
Chapter 5 demonstrates the multiplexing potential of elliptical and linear polarisations in a real environment, carried out with indoor propagation measurements. This chapter investigates the MIMO channel behaviour at line of sight and non-line of sight propagation as well as the effects of transmitter-receiver distance on the elliptical and linear polarisations.

Further to the research work carried out in previous chapters at sub 6 GHz, Chapter 6 explores the feasibility of employing elliptical polarisation at mmWave frequency band. This chapter also discusses the phased array beamforming technique where there is a potential for conducting further research as a hybrid beamformer with elliptical polarisation. This chapter aims to give an insight into what other work can be carried out in the future regarding elliptical polarisation. Finally, chapter 7 concludes the thesis.
Chapter 2

2 Background to static wireless links and multiplexing techniques

Wireless communication links are comprised of mobile and static communication links. The mobile links are the wireless link from a base station or access point to a moving mobile device. Static links are communication links that are stationary for at least a certain period of time. When wireless data traffic experienced explosive growth, the demand for static links also grew to cater for high bandwidth capacity lines where fibre or wire connections could not be easily deployed. In this chapter, applications based on static links are discussed, ranging from large scale such as fixed wireless access, backhauls and nomadic links to medium scale that includes small cell and WLAN architecture as well as smaller range, for instance device to device communication. In recent developments of mobile telecommunication, the millimetre wave band is introduced as a new frequency spectrum and is frequently researched upon as the latest solution for high capacity applications. The introduction to millimetre wave backhauling and how it uses MIMO multiplexing is also covered in this chapter.

The usage of multiple antennas at transmitter and receiver are proven to be highly beneficial to wireless communications, particularly in enhancing the network capacity, data rates and quality of service as compared to single-input single-output systems. There are two common techniques in implementing MIMO; either using diversity or multiplexing techniques. In the case of channel propagation, multiplexing methods are employed in line of sight scenarios while diversity techniques are used in non-line of sight environments. Three methods are known to achieve MIMO multiplexing benefits and are presented in this literature review; angular, spatial and polarisation. This chapter also discusses the fundamentals of a MIMO system, the capacity of MIMO system to and how to measure the multiplexing potentials by using singular value decomposition. At the end of the MIMO section, it is established that the polarisation method is most effective when multiplexing multiple simultaneous streams. Therefore, dual polarised antenna designs compiled from the literature are deliberated in this chapter.
2.1 Fixed links in communications

Mobile communications have progressed significantly in the past three decades especially since the first generation analogue cellular technology was launched. Subsequently, since 1990s, many mobile operators examined fixed wireless access to provide speedy implementation, flexibility and economical provisions of their networks. Fixed radio links are intended to support high bandwidth capacity communication links to meet the ever increasing demand for mobile network infrastructures. The demand is mostly to provide for customer connections and supporting infrastructure for public mobile networks. This demand is due to fixed radio network infrastructures that have been used for a long time, in some instances, they are an integral part of the Public Switched Telephone Network (PSTN) network or national broadcast distribution networks. It is also noted that the public mobile service is currently one of the most significant users of spectrum in Europe and will continue to be for the next 10 years as some forecasts predict [1]–[5]. In [2], it is reported that the number of Point-to-Point (PtoP) links, on average, increased by more than 24.5% per year between 1997 and 2010 which is mainly due to the expected growth in data traffic. The typical users for fixed links are telecom operators (e.g. for mobile network infrastructure), corporate users (e.g. private data network) and private users.

Apart from using wireless methods, there are other options in deploying fixed services such as using cable and fibre. Nevertheless, fixed wireless radio links are frequently the preferred solution due to fundamental considerations of constraints and practicality such as cost, local topography (e.g. highland terrains and rivers) and remote rural regions. Installing fibre or cable links necessitate substantially more work and cost. For instance, digging up roads in order to lay down fibres or cables can cause severe disruption in dense urban areas. Consequently, fibre is only employed in mobile networks when there is a high number of users, such that the cost of payload reaches that of fibre and radio. In addition, by using radio as transmission media, mobile operators can roll out a network rapidly and with more flexibility to install and scale transmission paths when and where they are required, which promotes better, more focussed distribution of investment.

In general, there is line-of-sight (LOS), obstructed-line-of-sight (OLOS) or non-line-of-sight (NLOS) type of channel propagation. A LOS propagation is defined as a direct and unobstructed path from transmitter (TX) to receiver (RX) with clearance within 60% of the first Fresnel zone [6]. If this criterion is not met, an OLOS link is obtained due to minor diffraction or partial blockage. A worst case scenario is the NLOS case when the propagation is totally blocked, allowing the signal
to reach its destination only via penetration through an obstacle, reflections off other objects or diffractions off the obstacle. Diffraction, a phenomenon that is often described as the bending of the signal. In reality, the energy of the wave is scattered in the plane perpendicular to the edge of the building. Reflection, in particular random multipath reflection, is another phenomenon that is essential for mobile broadband using wide-beam antennas. Penetration loss depends on the thickness of the material and its electromagnetic properties. In urban areas where there are a lot of buildings or an indoor area, deploying LOS solutions for fixed links can be very challenging. When the signals arrive at the RX due to the multipath resulting from these phenomena, they have resulting delay spreads, attenuation, and de-polarisation, relative to the transmitted signal. Hence, to mitigate impairments in the propagation channel, techniques such as directional antennas, transmit and receive diversity, adaptive modulation, error correction and power control are deployed.

Fixed radio links are able to transmit between two or more fixed points to accommodate telecommunication services, such as voice, data or video transmission. Nevertheless, static links are put in focus because the research in this thesis not only examine applications that are permanently fixed, but are also stationary for a certain period of time and may move about later. In this thesis, different types of static links are discussed as follows:

a) Large scale fixed radio links – Backhauling, specifically at sub 6GHz and millimetre wave frequency
b) Medium scale nomadic links – Small cell and WLAN access links
c) Small scale nomadic links – Short range device to device

2.1.1 Sub 6GHz backhauls and fixed links

Backhaul links refer to intermediate links between the core network or backbone network and the small sub networks at the edge of the entire hierarchical network. Indeed, every connection from the base station (BS) back to the core network can be considered as backhaul. It implies a high-capacity line; for example, to backhaul from a wireless mesh network to the wired network means aggregating all the traffic on the wireless mesh over one or more high-speed lines to a private network or the Internet. Backhaul can be either be wired, wireless (mobile) or fibred, just like in fixed wireless services.
Evolution of the wireless backhaul links begins with time-division multiplexing (TDM) and then progressing to packet microwave technology as the former method could no longer cope with the increasing demand of speed and inefficient usage of bandwidth. In the future generation of mobile communications, it is believed that in highly dense urban areas, base stations will have smaller footprints in order to combat the speed and bandwidth issue. Studies have suggested that equipment may be installed on light poles at street level which will affect the way base stations are deployed at cellular level. An example of this is a lamp street deployment is drawn on Figure 2-1 whereby smaller cells are utilised to connect two main busy streets in Guildford, England. These BSs must not have large aesthetic impact and thus, driving the need of smaller, integral and/or adaptive antennas. Consequently, the fixed services backhauling link hop will need to be considerably decreased.

Figure 2-1: Example of urban backhauling connecting two busy streets in Guildford, England

Three main configurations are identified to be used in deploying fixed radio links; point-to-point (PtoP), point-to-multipoint (PtoMP) and multipoint-to-multipoint (MptoMP). In the PtoP system, a radio and antenna is required at the end of every wireless link. Among the major applications for PtoP links are mobile networks backhauling and evolution of radio traffic nature from TDM to packet data. Therefore, due to the high amount of traffic taking place in these type of applications, it is most certain that very high capacity systems will be required. PtoP microwave is a cost-efficient technology for flexible and rapid backhaul deployment in most locations. It is the dominant backhaul medium for mobile networks and is expected to maintain this position as mobile broadband evolves; with microwave technology that is capable of providing backhaul capacity of the order of several gigabits-per-second [7]. The main focus of this research will be on PtoP configuration.
Chapter 2. Background to static wireless links and multiplexing techniques

Figure 2-2: PtoMP configurations typically used in deploying fixed radio links (a) star topology (b) mesh topology (c) tree topology

The second configuration, which is the PtoMP usually employs either the star, mesh or tree topology, as shown in Figure 2-2 and has a central base station that collects and aggregates traffic from other cells. Thus, in comparison to PtoP topology, one radio and antenna at an aggregation point are sufficient to serve a number of cell sites. PtoMP networks need to transmit high data rates between base and terminal stations in a dense manner while simultaneously minimising the intra-system interference between different cells or sectors of network. MPtoMP configurations are mainly used to serve large number of highly packed fixed terminal stations and thus, would provide an alternative for PtoMP networks. As opposed to PtoMP, this type of network does not require a central base station for communications between terminal stations. As a substitute, each and every terminal station may act as a repeater and pass on the traffic to or from the next terminal station. Both PtoP and PtoMP networks are already deployed in the market and are commercially available but MPtoMP networks are still under development. This is primarily because of investment cuts that obstruct the testing of MPtoMP networks in real time deployment.

There are many challenges to implement backhaul. One of the main issues for backhaul networks is to reduce the cost per bit when the traffic volume is increasing. Cellular industry suggests that
among the costs needed to spend, the biggest portion is the transport cost, a similar case in leasing backhaul communication capacity. More specifically, information from the cellular industry seems to suggest that transport equipment, excluding administrative costs, for example, can amount to around 40% of the construction cost of a backhaul network, whereas in the case of leasing backhaul communication capacity, transport costs could account for up to half of the total network operational costs, with backhaul contributing to three-quarters of this cost [8]. Unlike fixed traditional TDM technology, packet microwave needs to improve the backhaul efficiency by using adaptable methods such as statistical multiplexing and adaptive modulation to ensure it is more cost effective [9].

One important parameter in this technology is that it utilises frequency bands, mainly categorised into two groups; licensed and unlicensed bands. Frequency spectrum is a very precious resource in all wireless systems especially at sub 6GHz, and as demand for higher data rates grows, it only becomes more precious. Licensed bands can be expensive but is much preferred by mobile operators as they have good propagation characteristics, higher reliability and performance, for example, better latency, jitter, and spectral efficiency as opposed to unlicensed bands. However, licensed bands, e.g. at 2.4 GHz and 5.725 GHz have less bandwidth available due to incumbent usage and higher market competition [10]. There are different licensing schemes in different countries which factor in any of a number of variables that include in addition to the band of operation, the channel bandwidth, path length, availability and data rate [11]. For backhaul, simplicity and licensing cost are important issues. Light licensing or technology-neutral block licensing are attractive alternatives to other approaches such as link licensing as they provide flexibility [2]. In contrast, using unlicensed bands can reduce capital expense but raises radio interference issues and have lower penetration. Unlicensed, free to use spectrum is another appealing option for network service provider as it also poses less political obstacles. Many countries such as US, Europe, Japan and Australia have allocated unlicensed spectrum at around 5.8 GHz and at from 57 GHz to 66 GHz (varies between countries) and this has created a new found interest among researchers and mobile network operators. Table 2-1 summarises the licensing, application and configuration for most used frequency spectrum at sub 6 GHz in Europe.
Chapter 2. Background to static wireless links and multiplexing techniques

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Licensing schemes</th>
<th>Application</th>
<th>Configuration</th>
<th>Additional details</th>
</tr>
</thead>
<tbody>
<tr>
<td>Below 2 GHz</td>
<td>Mostly licensed</td>
<td>Mobile (IMT, GSM 900/1800, UMTS, HSPA, LTE, etc.)</td>
<td>Mainly PtoP</td>
<td>Potential for NLOS backhauling applications</td>
</tr>
<tr>
<td>2.025 – 2.4 GHz</td>
<td>Licensed</td>
<td>Mobile service (e.g. IMT, MSS)</td>
<td>PtoP</td>
<td>The use of this band for the fixed service seems to be in reduction or stable in almost all countries. Potential for NLOS backhauling applications</td>
</tr>
<tr>
<td>5.725 – 5.95 GHz</td>
<td>Unlicensed/light licensed from 5.725 to 5.850 GHz to 5.95 GHz</td>
<td>Fixed services, infrastructure and broadcasting.</td>
<td>PtoP and PtoMP from (5.85 to 5.95 GHz)</td>
<td></td>
</tr>
</tbody>
</table>

Table 2-1: Summary of licensing, application and configuration for popular frequency bands below 6 GHz in Europe [2]

In 2007, the Office of Communications (OfCom) first introduced the television (TV) white space as an unlicensed spectrum to be used for wireless communication [12]. From 2013 to 2014, various pilot studies, which were undertaken by OfCom and industry, were launched to study the feasibility of deploying TV white space and the approach to authorising deployment of devices in UK. The TV white space constitutes the frequency bands in the TV broadcasting spectrum that has been left unused after digital transmission migration from analogue broadcast. With today’s demand for wireless connectivity, the time is thus ripe to consider expanding broadband capacity and improving access for many users in this underutilised bands. More focus is put into ultra-high frequency bands for TV broadcasting because they have favourable propagation characteristics and the existing usage is relatively static and readily characterized, for example, in UK it is between 470 to 790 MHz [12]. Mechanisms based on a geolocation database are employed whereby devices report their location to a database which returns channels available at that location.

In the trials organized by OfCom, outdoor measurements were performed in the London urban area [13]. System throughput via the TV white space between user and server was among the parameters measured and the maximum downlink throughput recorded was about 19 Mbps. Propagation models and measurements were also carried out to conform the TV white space performances and...
benefits in LOS and NLOS environment [14] [15]. Other studies also show the advantage of the TV white space for wireless backhaul links which can be incorporated in a mobile terminal device within a WLAN or small cell as well as the access point [16]–[22]. TV white space is thus another interesting alternative solution for sub 6GHz backhaul and fixed links.

2.1.2 mmWave backhauls

In fifth generation (5G) of mobile communication, a new topic of interest that is constantly recurring is millimetre wave technology. This emerging technology attracts a lot of attention mainly due to its capabilities to have very wide bandwidth, i.e. ten times larger channels and the multi gigabit per second wireless personal area connectivity. The frequency band ranges from 30 GHz to 300GHz. 60 GHz radio, a popular choice of unlicensed spectrum, is mainly limited for indoor applications because of the increased path loss and higher oxygen absorption of 15dB/km around this band [23]. Hence, several applications are envisaged to suit this technology, to exemplify, high definition multimedia interface cable replacement/ uncompressed high definition video streaming, wireless gaming, and wireless gigabit Ethernet. More recent applications have been used for outdoor purposes, e.g. in stadiums, theme parks and airports. Since Federal Communications Commission released the unlicensed use of 60GHz frequency band, more and more intensive researches and industry interests, including the IEEE standardization of millimetre-wave personal area network [24], have been done for the applications of high speed and short distance wireless communication.

From Friis equation in (2-1) (where \(P_r\), \(P_t\), \(G_t\) and \(P_t\) are the received power, transmit power, transmit gain and receive gain respectively, \(R\) is the distance between TX and RX and \(\lambda_{\text{wavelength}}\) is the wavelength), the path loss for 60GHz carrier increases by nearly 30 dB in free space as compared to 2GHz carrier.

\[
P_r = P_t + G_t + G_r + 20\log_{10}\left(\frac{\lambda_{\text{wavelength}}}{4\pi R}\right)
\] (2-1)

Path loss together with transmission loss poses significant challenges at mmWave. Therefore, there is high path loss at mmWave and together with high transmission loss, they pose a significant challenge at this frequency band. These issues unfortunately limit the mmWave communication to be mostly for indoor and short-distance outdoor backhauls. It can however, play a prime role in supporting high-data-rate wireless applications over short, last-mile range backhaul, such as in dense urban settings. For instance, the latest application for millimetre-wave technology involves a pico-cell structure in the 60 GHz band which is easily applied to indoor applications within a large
building. In order to compensate for the larger path loss at this particular band, a high gain antenna (more than 10dBi) is required to considerably reduce the delay spread of the radio channel, particularly when TX and RX antennas are aligned. Other options include having a beam steering feature in such a system to provide enough system link budgets [25] [26] or equipping massive number of antennas at both link ends to produce high beamforming gain [27].

Nonetheless, higher path loss combines with high gaseous attenuation, e.g. oxygen absorption issues at millimetre wave frequency band open up the opportunity for spatial reuse of channels. These inherent properties in the frequency band are not well suited for long-distance links and short range translates to less interference among adjacent cells, causing it to be highly beneficial for spectrum reuse [28]. Additionally, channels in the indoor or open areas show a strong multipath behaviour because of easy reflection. Therefore, 60 GHz is usually envisaged for communication confined to a room or an open area where LOS signals from the antennas can be expected.

At 60 GHz, the wavelength is just 5mm long. With such a small wavelength, the radio system is capable of dense spatial communications. Typical cell site density of today’s networks is not sufficiently high to fit with the low ranges supported by these products. The situation is likely to change in the future where the provision of ever-increasing capacity to end users will be met with a considerable increase in the number of sites deployed. Although the antenna at this band is physically small, it can be electrically large and can be designed to steer towards the strongest propagation path. These beams have the benefits of being flexible and can adapt to the changing interference and propagation conditions in real time. Therefore, the highly directional antenna is mostly suitable for LOS propagation at V-band [29].

When dimensions are smaller, the radio design also becomes more compact which is greatly favourable. Furthermore, the small size of antennas and RF circuits, allows the integration of 60 GHz technology in small devices. The relatively compact size of the equipment and the antenna is an added advantage for positioning in urban environments. Many antenna structures at 60 GHz come in the form of beamforming arrays and can become quite complex. Hence, more research is needed to meet the desirable criteria of compact in size but with high gain steerable antenna array.

It has also been researched that the performance of backhaul at millimetre wave can exceed the performance of microwave backhaul [30]. The backhaul links were tested against diffraction,
reflection and penetration scenarios. The results demonstrated that the throughput at millimetre wave is higher or comparable to the throughput at sub 6GHz. This is mainly because of the higher antenna gain and narrow beam pattern at millimetre wave that experiences fewer multipath caused by the obstacles such as buildings, trees and lamp post. Nonetheless, antennas at millimetre wave can be sensitive to alignment errors due to the narrow beam pattern and higher rain attenuations are experienced [31].

2.1.3 Small cell and WLAN access links

2.1.3.1 Small cell

With the evolution of the wireless networks, it also poses challenges to the current cellular system. It is known that macro cellular systems have reached maturity and operators in many countries are reaching the maximum price/performance benefit for this type of infrastructure within their cellular networks. With the increasing demand for data traffic in mobile communications, smaller cells, for instance micro, pico and femto cells, are introduced to help macro cellular networks cope with this demand. Therefore, more effective, scalable and flexible backhauling is required to carry the traffic. An example of an indoor scenario of small cell viewed by an access point in a busy airport is portrayed in Figure 2-3.

Figure 2-3: Example implementation of small cell scenario in the busy Amsterdam Schiphol airport

In the case of small cell backhaul, it is expected that when PtoP microwave systems are used, they will operate in the higher frequency bands which would have universally lower licensing fees than the lower bands and also smaller antenna size. Nevertheless, the cost of employing PtoP microwave
systems are higher than PtoMP due to the fact that the spectrum cost is fixed on a per link basis and therefore will have limited scalability. Current fourth generation communication technology employs orthogonal frequency division multiplexing for better tolerance to multipath, and even though it does not tolerate high Doppler spread, this is not a major issue in backhaul.

Small cells pose a great benefit for next generation of cellular systems. Small cells add extra spectrum capacity by utilizing different spectrum bands than the macro cell and some techniques allow for frequency reuse to be deployed easily. Another advantage of small cell deployment is to extend the coverage in weaker areas such as at the cell edge [31]. Small cells are also advantageous to minimize the operating costs in terms of transferring data to the core network. Figure 2-4 is an example of how small cell is integrated together with current macro cells in the Long Term Evolution-Advanced where it helps to interconnect the backhaul links to each other and to core network [32].

![Image](image.png)

**Figure 2-4**: A macro cell integrated together with smaller cells. Blue, green and red regions represent macro cell, micro cell and pico cell respectively.

Small cell systems can incur additional hardware and operational cost as well as the complexity of the installation is certainly higher. An overall trend for smaller size cells is also expected in any geographical area; therefore, the upgrading or new deployment of mobile backhauling networks will, in general, require significantly shorter hops, either on the lower layer (connections between base stations using higher frequency bands e.g. 23 GHz to 42 GHz) and on the higher layer (between larger and more distant exchange stations using lower frequency bands e.g. 15 GHz down to 6 GHz). Nevertheless, continuous development of small cell solutions will bring down the cost in future and are aimed to be fully implemented for 5G communications and beyond [33].
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Small cells are the best answer to dense area coverage and capacity requirements, but also come with particular challenges that must be addressed. Deploying high-performance microwave backhaul in places where there is no direct LOS brings new challenges for network architects. It is observed that in the case of urban areas where more small cells are intended to be deployed, the installation height for base stations is normally lower than rooftop and thus is not easily backhauled using LOS microwave or even wired connections. The challenges posed by locations without a clear LOS are not new to microwave-backhaul engineers, who use several established methods to overcome them. In mountainous terrain, for example, passive reflectors and repeaters are sometimes deployed. However, this approach is less desirable for cost-sensitive small-cell backhaul, as the cost increases together with the increasing number of sites.

Therefore, in order to cope with the growing cost of small cell implementation, more flexible and cost effective methods need to be developed. In addition, for small cell deployments, the scalability aspect is very important in order to grow with increasing demand and to be cost effective. NLOS wireless backhaul is one of the techniques that is inherently scalable due to its main characteristics: block spectrum pricing, PtoMP configuration, ease of deployment and installation. Moreover, NLOS wireless backhaul is more suited than other wireless backhaul solutions for integration with the small cell base station. The result is a diminishing marginal cost per link which allows small cell base station deployments to scale.

2.1.3.2 Wireless Local Area Network (WLAN)

WLAN is governed by IEEE 802.11 standards, which is by far one of the most successful standards in wireless communication systems. This type of wireless computer network links two or more devices using a wireless distribution method within a limited area such as a home, school, computer laboratory or office building. It gives users the ability to move around within a limited area while still connecting to the network and can provide a connection to the wider Internet. Most modern WLANs are marketed under the Wi-Fi brand name. Wireless LANs have become so popular in the home due to ease of installation and use and in commercial complexes offering free wireless access to their customers that it has become such a norm to society.

The first IEEE specification was published under the 802.11 working group in 1997. Initially, the specification defined only the data rate of 1 or 2 Mb/s operated at 2.4 GHz band. Later, they decided
to expand the standard to support up to 11 Mb/s in 802.11b, and 802.11a to support up to 54 Mb/s. Later in 2009, the IEEE 802.11n [34] was introduced to improve the throughput by adding MIMO systems, particularly to focus on beamforming and compressed feedback schemes. The goal of 802.11n was to achieve the throughput of 100 Mb/s at the MAC layer. 802.11ac was proposed around four years later to add more evolutionary changes on MIMO methods. More advanced MIMO techniques such as downlink multi-user (MU)-MIMO that allowed up to 160 MHz bandwidth and up to 8 streams were suggested. These techniques then enable the network to obtain more than 1 Gb/s of capacity.

802.11n employed MIMO techniques including spatial-division multiplexing, space-time block coding and transmit beamforming in order to realize such a throughput enhancement [35]. Utilising multiple antennas at the TX and the RX, spatial-division multiplexing allows multiplexing of multiple data streams across spatial dimensions, and the throughput increases as much as the number of data streams. In addition, transmit beamforming improves the received signal strength by emphasizing the dominant modes of transmission for the channel. In beamforming, antenna coordination for directional beams is enabled with an aid of channel state information (CSI) feedback, and thus it is essential to efficiently deliver such information from a transmitter to a receiver. Beamforming can also be applied to transmit multiple streams simultaneously to multiple users, by nullifying interference between users. These techniques are essential to maximize the downlink system throughput for multiple user transmission as well as for single user transmission [36].

In future, promising techniques to further enhance the spectrum efficiency of WLAN systems include, but are not limited to, orthogonal multiplexing division multiple access, uplink MU-MIMO and full duplex transmission. Currently, 802.11 technologies only adopt half duplex transmission, whereby separate transmission and reception occur at different time. Alternatively, full duplex transmission allows simultaneous transmission and reception over the same time and frequency. In this case, spectrum efficiency could be significantly improved up to two times higher but self-interference cancellation is critical to realisation of this technique.
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2.1.4 Short range device to device links

With current development of technology, all mobile devices are now able to connect to each other at all time. An example of this connectivity is via tethering. It is the ability to share a smartphone's Internet connection with computers or other devices which can be accomplished by connecting the devices wirelessly using a Bluetooth wireless link or a Wi-Fi connection. Most mobile devices, e.g. laptop, tablet and mobile phone have Wi-Fi built in nowadays. If these devices are connected together at a static position for a certain period of time, for instance, in an office during working hours as can be seen in Figure 2-5, temporary fixed links can be established.

![Example of short range device to device connection via Wi-Fi tethering at a static position for a certain period of time, establishing temporary fixed links](image)

2.2 MIMO Multiplexing and Diversity

2.2.1 Introduction to MIMO

The use of MIMO systems has been proven to be highly beneficial to wireless communications, particularly in enhancing the network capacity, data rates and quality of service. By using multiple antennas at the TX and RX, not only does it mitigate the interference between cells but also increases the data rate efficiently. This type of system is able to operate using angular, spatial and/or polarisation properties of the multipath channel and by exploiting these properties the performance of mobile communication can be greatly enhanced [37].

As can be seen from Figure 2-6, multiple antennas at both end of the communication system are capable to resolve multipath components, thus enabling higher data throughput. In a MIMO
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channel, NLOS multipath scattering is a common phenomenon, creating rich multipath environments which can further improve the spectral efficiency of a wireless link.

A rich multipath environment is created as a result of the impairments of the wireless channel. It is highly desirable to improve the channel impairments using a multi antenna system. The transmission of data over a wireless channel can be mainly categorised into either diversity techniques or spatial multiplexing techniques.

Using spatial multiplexing techniques, multiple signals are being sent over the channel and the RX will learn the channel matrix as well as inverting it to separate the data. As the term spatial is being used here, this technique requires multiple antennas at both ends of the wireless link. The data rate is increased by transmitting independent information streams simultaneously on different antennas but using same frequency as well as exploiting the rich scattering and fading to gain more advantage [38]. There is no requirement for channel knowledge at TX when employing spatial multiplexing method. If scattering is rich enough (i.e. high rank channel $H$), several spatial data pipes are created within the same bandwidth. One of the many advantages of spatial multiplexing is also that the gain comes at no additional bandwidth or power due to the use of multiple antennas at the TX and RX. In spatial multiplexing, an $M \times N$ MIMO channel opens up $m$ independent single-input single-output (SISO) channels between TX and RX, where $m = \min (M, N)$. Therefore, theoretically, a maximum of $m$ different information symbols can be transmitted simultaneously over the channel at any given time. It is necessary for the antennas to have some spacing between them so that they can multiplex data. Multiplexing needs low correlation between branches, but high coupling will prevent low antenna correlation and weaken the efficiency of the MIMO antennas.
Exploitation of the multiplex rich channel requires signal processing such as the spatial division multiple access (SDMA) and the Bell Labs Layered Space-Time (BLAST) techniques. The SDMA protocol exploits the spatial diversity among different terminals to selectively transmit and receive signals in the same time slot and same frequency band [39]. The SDMA protocol is a packet reservation protocol to significantly increase both the uplink and downlink throughput. Narrow beams from array antennas are steered toward desired users to filter out interference caused by co-channel users located in the same cell and from neighbouring cells. Its advantages include simple implementation, adaptability to multimedia traffic with diverse bandwidth requests, network security and guaranteed fairness in bandwidth sharing.

The BLAST technique, developed by Gerard Foschini at Lucent Technologies' Bell Laboratories, also exploits the spatial diversity aspect to achieve higher capacity [40]. Using multiple antennas at both TX and RX, the signals in this transceiver architecture are layered in space and time as suggested by a tight capacity bound. The BLAST architecture is designed largely to uncouple distinct adjacent spatial modes. The data capacity of the system then increases linearly with the number of antennas. The space-time architecture is layered diagonally in which code blocks are dispersed across diagonals in space-time, therefore, it is also referred as diagonal-BLAST or D-BLAST technique. The vertical-BLAST or V-BLAST was later developed to simplify the D-BLAST architecture, whereby the vector encoding process is simply a demultiplex operation followed by independent bit-to-symbol mapping of each sub stream[41].

On the other hand, a MIMO diversity gain technique is the improvement in link reliability obtained by transmitting the same data on independently fading branches. In terms of power consumption, diversity techniques use a conservative approach to minimise power for reliability and Quality-of-Service to combat fading, as opposed to multiplexing techniques which uses a more optimistic approach to maximize transmission rate. Each radio link between a TX and RX antenna delivers a signal path from one end to the other end of the link. Multiple independently-faded replicas of the data symbol can be achieved at the RX by transporting the same information through different paths, thus resulting in more consistent and robust propagation [38]. The gain for diversity technique suggests that power $P_c$ decays at a rate of $1/\rho^d$ (where $\rho$ is the signal-to-noise-ratio (SNR) and $d$ is the diversity gain in the high SNR region which is the opposite of $1/\rho$ for a SISO system. Consequently, the maximal diversity gain $d_{\text{max}}$ is the total number of independent signal paths that exist between TX and RX. The total number of signal paths for a diversity method is $1 \leq d \leq d_{\text{max}} = NM$ for an $(M,N)$ system.
Nevertheless, a system design has to compromise between both techniques, enhancing either one or a little of both [42]–[45]. In comparison to a diversity system, the spatial multiplexing gain is defined as the additional gain of a spatial multiplexing system [46]. Subsequently, the spatial multiplexing gain depends on the SNR and the channel conditioning. Therefore, as compared to a diversity system, a spatial multiplexing system can be considered less reliable, which is singly dependent on the energy of the channel coefficients.

2.2.1.1 Capacity of MIMO system

Shannon’s theorem describes that the capacity is a measure of transmission rate for dependable communication on a given channel. According to this theorem, a communication system needs to be reliable so that data is transmitted at such a speed with randomly small error probability. There are many factors affecting the capacity of wireless communication system, to exemplify, the background noise, number of transmit and receive antennas as well as number of multipath components produced in the propagation channel.

Let us compare with the spectral efficiency or more well known as capacity of a single-input-single-output (SISO) system first. Using only a single antenna at TX and RX, the capacity of the channel, $C$, is given by:

$$C = \log_2 (1 + \rho_0) \quad \text{bits/s/Hz}$$

(2-2)

Where assuming an additive white Gaussian noise interference (AWGN), $\rho_0 = \frac{P_D, av}{N_0}$ is the average SNR at the RX, $N_0$ is the noise power of the additive noise, and $P_D, av$ is the average received power.

An input signal enters an encoder at the TX. The encoded signal then gets forwarded to the $M$ transmit antennas. The transmit antennas will send the signal through a wireless channel that contains scatterers, which will reflect, diffract and refract the incoming signal. Upon reaching the RX side, the $N$ receive antennas will capture the signal and forward it to the decoder. This MIMO system is also known as an $M \times N$ system. The system can also be described as

$$y = Hx + n$$

(2-3)

Where $y$ is the receive signal vector, $x$ is the transmit signal vector, $n$ is the additive noise vector and finally, $H$ is defined as $M \times N$ channel matrix as follows:
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\[
H = \begin{pmatrix}
    h_{11} & \cdots & h_{1M} \\
    \vdots & \ddots & \vdots \\
    h_{N1} & \cdots & h_{NM}
\end{pmatrix}
\]  

(2-4)

$H$ is also a channel matrix that contains magnitude and phase information of the propagation paths between the $M$ transmit and $N$ receive ports. Another way of expressing $H$ is after propagating through the channel, as the ratio of received voltage waves at the receive antennas, to the incident voltage waves at transmit antenna.

If there is no channel information at the TX, the transmitted power is distributed equally among the transmit antennas. Hence, the spectral efficiency of the MIMO system is expressed by:

\[
C = \log_2 \det \left( I + \frac{\rho}{M} HH^H \right) \text{ bits/s/Hz}
\]

(2-5)

Where $\rho = \frac{p_{\text{SNR}}}{N_0}$ is the average SNR at the TX. Equation (2-5) is a general expression of the capacity when no channel coding techniques is applied.

Transmitting power equally to all antennas is a less efficient method and wasting valuable resources. It is however more useful to distribute power according to water-filling method which is based on channel current state whenever there is channel information at TX. Therefore, spectral efficiency can be further optimized.

2.2.1.2 Singular value decomposition of MIMO channel, $H$

In order to understand MIMO systems, the singular value decomposition of the channel matrix is fundamentally important as it extracts the equivalent independent sub channels and gives maximum number of streams that can be multiplexed simultaneously [47]. The method of singular value decomposition (SVD) is a method of diagonalising the channel matrix $H$ and finding the eigenvalues [48]. It is particularly useful for interpretation in the antenna context.

In addition, SVD provides a very simple way to compute the capacity based on a sum of the eigenvalues. When multiple antennas are present simultaneously at both the TX and RX, the capacity achieving scheme consists of sending multiple symbols per transmission period. Using SVD allows the extraction of spatial or polarised sub channels. In MIMO systems where multiple
spatial or polarised routes are present and independent of each other, the MIMO system becomes a set of independent SISO channels, which are the sub channels. Therefore, the capacity of a MIMO system becomes the sum of the capacity of each SISO channel.

In the literature, an analytical approach has also been carried out where the eigenvalues of the transmission system lead to a definition of the maximum diversity order or array gain (depending on whether it is a diversity rich or beamforming rich channel) as the largest eigenvalue [48], which is achieved through applying the SVD.

Real measurements have been carried out to characterize the MIMO channel and its corresponding capacity [47], proven to be in agreement with the theoretical analysis in [48].

The SVD of a 2 x 2 channel matrix $\mathbf{H}$ is [47]:

$$
\mathbf{H} = \mathbf{USV}^H = [\mathbf{u}_1 \quad \mathbf{u}_2] \begin{bmatrix}
\mathbf{s}_1 & 0 \\
0 & \mathbf{s}_2
\end{bmatrix} \begin{bmatrix}
\mathbf{v}_1^H \\
\mathbf{v}_2^H
\end{bmatrix}
$$

(2-6)

where matrix $\mathbf{U}$ is a unitary and orthogonal matrix $M \times N$

matrix $\mathbf{V}$ is a unitary and orthogonal matrix $M \times M$

matrix $\mathbf{S}$ is an $N \times M$ diagonal matrix with nonnegative singular values

### 2.2.2 Spatial

MIMO multiplexing gain can be improved by employing several techniques. This includes using a spatial dimension or a geometrical model and changing the antenna spacing of the TX or RX array. Studies in the literature were found to employ uniquely designed antenna arrays in order to achieve orthogonality between spatially multiplexed signals of MIMO systems in LOS channels [49], [50].

A MIMO system with full rank capacity is greatly desired and by optimising the antenna array design, maximum capacity criterion can be satisfied. In [49], criteria for the maximum theoretical capacity is derived in terms of the positioning of the antenna elements for optimal MIMO architectures. Typical applications for LOS MIMO systems are fixed wireless access and radio relay systems. A representation of a 2x2 MIMO system using spatial dimensions is described in Figure 2-7.
To obtain the maximum capacity of a MIMO channel in free-space, $\mathbf{HH}^H$ matrix is used. A 2x2 example is given as follows:

$$\mathbf{HH}^H = \begin{bmatrix} e^{jk(r_{11}-r_{21})} + e^{jk(r_{12}-r_{22})} & e^{jk(r_{21}-r_{11})} + e^{jk(r_{22}-r_{12})} \\
\end{bmatrix}$$  \hspace{1cm} (2-7)$$

Where $k$ is a wavenumber to the corresponding carrier frequency and $r_{11}$, $r_{12}$, $r_{22}$ and $r_{21}$ are the distances between TX 1 and RX 1, TX 2 and RX 2, TX 2 and TX 2 and RX 1. Using $\mathbf{HH}^H = 2\mathbf{I}_2$, where $\mathbf{I}_2$ is a 2 x 2 identity matrix, Equation (2-7) is simplified to:

$$|r_{11} - r_{12} + r_{22} - r_{21}| = (2p + 1) \frac{\lambda_{\text{wavelength}}}{2}$$  \hspace{1cm} (2-8)$$

where $\lambda_{\text{wavelength}}$ is the wavelength and $p \in 0, 1, 2, \ldots$. Full derivation of Equation 2-8 is added in Appendix A. Equation (2-8) is the criteria to achieve full orthogonality and thus the maximum capacity criterion for a 2 x 2 MIMO system. To simplify the criterion, a few assumptions are made; assume the transmit and receive array are in parallel with inter-element spacings $d_1$ and $d_2$, then $r_{11} = r_{22}$ and $r_{12} = r_{21}$. Hence, the criterion is given as:

$$|r_{11} - r_{21}| = (2p + 1) \frac{\lambda_{\text{wavelength}}}{4}$$  \hspace{1cm} (2-9)$$

From the geometry of the arrays, it is observed that:

$$d_1 d_2 = r_{21}^2 - r_{11}^2 = (r_{21} - r_{11})(r_{21} + r_{11})$$  \hspace{1cm} (2-10)$$
Since the distance $R$ between the TX and the RX is much bigger than $d_1$ and $d_2$, it is safe to assume that $(r_{11} + r_{12}) \approx 2R$. Thus, replacing into Equation (2-10), the final condition is:

$$d_1d_2 \approx 2R(r_{21} - r_{11}) \approx (2p + 1) \frac{R\lambda_{\text{wavelength}}}{2} \quad (2-11)$$

Inter-element spacing can then be determined from Equation (2-11) to achieve the maximum $2 \times 2$ MIMO capacity. Using the above equation, a spacing of 0.85m at both TX and RX is required for a 10m TX-RX distance at 2 GHz while only 0.15m spacing is required at 60 GHz for the same setting. For most indoor applications, this is reasonable even though the spacing is larger than the conventionally used $\lambda_{\text{wavelength}}/2$. Taking WLAN as an example, an access point is installed at 3m above a RX terminal. The spacing required for both arrays is 0.3m at 5 GHz which could fit easily on top of a computer.

Envelope correlation coefficient (ECC) characterises the mutual correlation between channels in a MIMO system. It is beneficial for a MIMO system to have low ECC value so that there is low correlation between adjacent antennas. Different diversity techniques, such as spatial diversity, pattern diversity, and polarization diversity are employed to ensure low value of ECC in the working bandwidth of the MIMO antenna system. The general formula is as follows [51]:

$$\rho_e = \left| \frac{\int_{4\pi} \left( \int F_1^*(\theta, \phi) \cdot F_2(\theta, \phi) d\Omega \right)^2}{\int_{4\pi} \left( \int \left| F_1(\theta, \phi) \right|^2 d\Omega \right) \int_{4\pi} \left( \int \left| F_2(\theta, \phi) \right|^2 d\Omega \right)^2} \right|^2 \quad (2-12)$$

where $F_i(\theta, \phi)$ is the field radiation pattern of the antenna system when port $i$ is excited and $\cdot$ denotes the Hermitian product. The general formula of ECC includes the radiated far fields of the individual antennas, when other antennas were terminated with matching loads. Therefore, the formula indicates that ECC is about how independent two antennas' radiation patterns are.

An evaluation was conducted to investigate the relationship between the inter element spacing at the RX and the complex correlation of MIMO channels. The transmitted inter element spacing, $d_1$ is fixed at 4m while the received array inter-element spacing, $d_2$ is varied. The TX-RX distance, $R$ is also fixed at 1 wavelength. Figure 2-8 shows the relationship between complex correlation of the MIMO system and $d_2$. It can be seen in this figure that the signal envelope does not change in a spatial case whereby the correlation was maintained at a value of $|\rho_{12}| = 1$ even though the $d_2$ spacing was different and the channels are found to be correlated. The result is similar for both TX.
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However, Figure 2-9 shows that it is not the magnitude that changed but rather the change of phase by 180° at both TX that makes the MIMO channels orthogonal.

![Figure 2-8: Complex correlation vs spacing in a line of sight spatial multiplexing system where $d_1$ is 4m at a wavelength of 1m](image)

![Figure 2-9: Phase of complex correlation vs spacing in a line of sight spatial multiplexing system where $d_1$ is 4m at a wavelength of 1m](image)

This finding was also proven in Figure 2-10 when the $d_2$ spacing was plotted against eigenvalues of the MIMO system. Figure 2-10 that examines the orthogonality of MIMO channel is plotted based on Equation (2-11). The parameters in this evaluation is similar as in Figure 2-8 and 2-9 with additional parameter added which is $p = 5$. When $d_2 = 1.4$ m, both eigenvalues are equal and overlap with each other, thus enabling maximum capacity at this point. Eigenvalue decomposition are further discussed in Chapter 3.

![Figure 2-10: Eigenvalues vs spacing in a line of sight spatial multiplexing system where $d_1$ is 4m at a wavelength of 1m](image)
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For a $M$ number of transmit antenna and $N$ number of receive antenna, a Uniform Linear Array (ULA) is employed at both ends of the communications link. The simplified maximum capacity criterion then becomes:

$$d_1 d_2 \approx (2p + 1) \frac{R \lambda_{\text{wavelength}}}{M}$$  \hspace{1cm} (2-13)

It can be seen that when there are more antennas at TX, the antenna spacing is further reduced, resulting in more efficient use of space. It is also interesting to note that there is no limit on the linear increase of MIMO capacity since higher order maximum capacity array design does not depend on scatterers, unlike in NLOS systems. Additionally, for this type of architecture, the orthogonality between different columns is equal to zero. Under the condition in Equation (2-13), the channel matrix hence becomes orthogonal.

Another technique based on optimisation of antenna placement in a uniform linear array was later introduced in [52], [53]. By using a ray tracing model, a new geometry is presented that allows arbitrary orientation of the TX and RX arrays as shown in Figure 2-11. Hence, unlike in the previous work in [53], this design is not restricted to parallel orientation. Designed for ULAs, the product of the inter-element spacings, $d_1 d_2$, which is also the maximum capacity criterion, includes a function of sender and receiver separation ($R$), wavelength ($\lambda_{\text{wavelength}}$), number of antenna at RX ($M$), and the spherical angles at the local coordinate systems at the TX ($\theta_t$) and RX ($\theta_r$), is given as [52]:

$$d_1 d_2 = \frac{\lambda_{\text{wavelength}} R}{M \cos \theta_t \cos \theta_r}$$  \hspace{1cm} (2-14)

Experiments were also conducted to further research the maximum capacity of LOS MIMO, using other parameters and in different environment. An extension to Sarris work is the study of capacity behaviour in outdoor MIMO channel [54]. In [55], the results suggest that for frequency reuse method, three sectors in a conventional cellular system can be combined to form one “edge-excited”
(inward-facing) cell. This is important to improve the capacity for receivers that are not close to a base station. In order to calculate the spectral efficiency of a LOS MIMO system, a distributed-source model was presented in [56]. There is also a study to compare the performance of ULA vs Uniform Circular Array (UCA) whereby the evaluation showed that to get the same capacity, the spacing of UCAs is greater than ULAs [57]. Another method to maximize the capacity is by increasing the slope of the far-field phase response of the antennas as a function of incident angle [58].

These MIMO systems are able to exploit the spatial dimension of the antenna arrays in order to produce maximum capacity using LOS propagation that successfully surpasses the capacity of i.i.d Rayleigh fading model. Nevertheless, the system does not come without its disadvantages. The transmit and receive arrays consist of widely spaced discrete antenna elements, more than half of a wavelength. At lower frequencies, particularly at sub 6 GHz, this is not desirable due to higher transceiver complexity, thus making it only suitable for mm-wave systems which exhibit very small wavelength. The wide spacing and subsequently the bigger dimension that is required to orthogonalise the channel matrix also results in reduced array gain, increased interference and compromised security due to grating lobe.

2.2.3 Angular

An alternative technique to improve spectral efficiency of MIMO systems uses pattern or angular diversity. Pattern diversity allows radiation pattern to be reoriented in the space as well as the ability to switch between multiple directional beam patterns. The basic principle of this type of diversity, as shown in Figure 2-12 is to have patterns that are ideally orthogonal, or uncorrelated in practice, over the incident multipath field distribution. Two type of scenarios are depicted in Figure 2-12; (a) is a diversity rich scenario whereby one strong and one weak path exist illustrated by a thicker and thinner arrow and (b) is an angular rich scenario where two paths are of the same strength (similar arrow thickness is portrayed). The switching concept enables a single antenna port to receive signals from different patterns except that only one pattern can be used at a time. Therefore, in order to benefit from the multipath generated by the beams, optimisation of the degrees of freedom of the beam angles is deemed necessary.
A method to carry out the angular diversity was presented using a switched parasitic antenna (SPA), which is a novel technique for electronically directing the radiation pattern in a MIMO system [59]–[61]. The idea is to have central active antenna as a radiating element and a set of parasitic elements that are strongly coupled to the active one using a baseband control signal. The active antenna is driven by a high frequency RF signal and connected to a radio transceiver, operating near resonance. The parasitic elements together with the radiating element form an array, similar to the well-known Yagi-Uda antenna. By changing the termination impedance of the passive elements and hence changing the current flow in those elements, the radiation pattern of the parasitic array is altered. When the parasitic elements are shorted to the ground plane using PIN diodes, they become reflectors and affects the antenna’s radiation characteristics. The parasitic antennas can be designed using monopoles on a ground plane as seen in Figure 2-13 or as parasitic patch antennas [62]. In order to get four different directions with 90° separation, it is assumed that three parasitics were always shorted.
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The five-element monopole SPA shown in Figure 2-13 consist of an active element in the centre and four parasitics surrounding it. SPA design has potential to have higher if not comparable throughput to that of conventional MIMO systems with the same number of driven elements. With this architecture, SPA has been successfully utilised for spatially multiplexing several data streams at 2.6 GHz [59]. This was performed by switching the SPA far-field at the modulation rate. Additionally, beam patterns of the antenna array can be manipulated by controlling the reactance via a dc control.

A more advanced technique is later proposed to improve the performance of MIMO systems using angular diversity. Electronically Steerable Parasitic Array Radiator or popularly known as ESPAR antenna was originally engineered to combat the multipath fading issue which is a low cost solution compared to adaptive array antenna at that time [63]–[65]. Authors in [64] and [65] later used ESPARs at the transmission and/or receiving side to build a novel beamspace MIMO. Beam space MIMO has proven that it can deliver comparable capacity and diversity gain to that of traditional MIMO system using multiple antenna elements. The ESPAR antenna is a variable array antenna consisting of one radiator and parasite(s) that are loaded with varactor diodes, unlike conventional methods that uses PIN diode switches. These varactor diodes act as control devices. The ESPAR antenna offers extremely low power consumption because of these varactor diodes are always in a condition of reverse bias, so they do not consume forward direction current.

Figure 2-14 shows the configuration for an ESPAR antenna. A coaxial feed is used to excite a single monopole radiator which is oriented vertically and a quarter-wavelength with omnidirectional radiation pattern. Passive radiators are then added to provide directivity. These are arranged in equal intervals and loaded with variable reactors in the bottom. Varactor diodes together with two fixed inductors are used to vary the reactance and subsequently control the directivity of the patterns electronically. When the reverse bias voltages are switched, the directivity of the ESPAR antenna is also adjusted. In MIMO systems, an algorithm based on the received signal strength can be implemented for switching control. The ESPAR antenna is utilised at the receiver to provide orthogonal directional beam patterns by changing the reactive loads of passive elements. Moreover, strong mutual coupling among the ESPAR antenna elements are also exploited at the transmitter side to create a linear combination of orthogonal patterns. Thus creating great potential for delivering better angular diversity for users.
Pattern diversity removes the minimum spacing requirement of MIMO systems and substantially reduces the size and saves valuable space, especially in the mobiles as well as provides array gain over the SISO systems. SPA successfully reduces the complexity of transceiver as compared to array antenna, however at the cost of lower throughput than an array of antennas with fixed beams [60]. This was also confirmed with another research when more number of elements, i.e. passive reflector is needed to enhance the spatial multiplexing MIMO gain when compared with conventional ULA which in turn increase the complexity at frequencies lower than 6 GHz [67].

2.2.4 Polarisation

A MIMO system that employs dual polarisation technique is very beneficial to mobile communication. A dual polarisation system that is orthogonal enables the antennas to be co-located to each other and hence producing more compact design. In [68], the performance of dual polarised MIMO system was evaluated and compared to single polarisation system. It was discovered that the MIMO system capacity was improved when dual polarisation is employed compared to single polarisation antenna at close distance and mostly LOS condition. This is due to the enhanced matrix rank, i.e. capacity increases linearly with number of transmit/receive antenna, although there is cross polarisation discrimination loss of 3dB for every antenna added. Note that [60] only compared the spatially separated vertical and horizontal polarisations instead of making them co-located.
Ideally dual polarisation proposes an improved separation between channels, through a large decorrelation at both transmitter and receiver. By employing the same channel frequency but with two different polarisations on the same antenna results in double capacity. Analysis has been conducted that shows using dual-polarised arrays, large multiplexing gains are feasible in Ricean or highly correlated Rayleigh fading channels [69], [70]. Figure 2-15 shows how a dual linear polarised MIMO system is configured.

\[ H_{DLP} = \begin{bmatrix} h_{VV} & h_{VH} \\ h_{HV} & h_{HH} \end{bmatrix} \]  \hspace{1cm} (2-15)

Dual polarisation can be achieved using several methods, such as ±45° Linear Polarisation and Dual Circular Polarisation (Right-hand/Left-hand). In the literature, it has been shown that by combining orthogonal polarisation-induced autonomous channels, Dual Circular Polarisation Multiplexing (DCPM) offers lower complexity where scattering is low [71]. An example of the DCPM system is presented in Figure 2-16.
The channel is optimally weighed such that the weights are passive and do not consume additional energy. At the receiver, the two single-input-multiple-output (SIMO) outputs are combined. The deployment of dual polarisation schemes also intends to reduce the effect of coupling between orthogonal polarisations. One of the crucial advantages of the DCPM technique is that channel state information is not required to be sent back to transmitter, as opposed to traditional MIMO system [71]. For wireless channels, the channel varies in time and frequency and therefore experiences wideband fading. In traditional MIMO systems, maximum capacity is achieved when the transmitter varies the data rate sent to each user as their channels vary [72]. This implied that channel estimates are required at the transmitter. In [71], the DCPM technique was developed for the land mobile satellite (LMS) application and it is difficult and costly to feedback the channel state information back to the satellite in space. When the scattering is low and there is a LOS, the orthogonal circular polarisations will minimise the co-channel interference and maximum capacity can be achieved without having to send back the channel estimates to the satellite, rather a zero forcing is applied at the receiver on the ground only. Therefore, the DCPM technique has the advantage of simpler receiver based processing.

Several measurement campaigns have been carried out to model dual polarised MIMO channel [71], [73]. The channel matrix for dual circular polarised MIMO channel is presented here:

\[ \mathbf{H}_{DCPM} = \begin{bmatrix} h_{RR} & h_{RL} \\ h_{LR} & h_{LL} \end{bmatrix} \]  \hspace{1cm} (2-16)

$h_{RR}$ and $h_{LL}$ refers to two co-polar components: right hand to right hand and left hand to left hand respectively. While $h_{LR}$ and $h_{RL}$ refers to two cross-polar components: left hand to right hand and right hand to left hand respectively. A DCPM system can achieve perfect orthogonality when $\mathbf{H}_{DCPM} = \sqrt{2} \mathbf{I}$ which is observed in [71] when there is at least 10 dB cross polar rejection. A further examination of channel orthogonality is portrayed in Figure 2-17, whereby the relationship between eigenvalues from SVD and XPR is studied. It is assumed that the RHCP and LHCP antennas are co-located. In this study, the analysis is not frequency-related, therefore the relationship between eigenvalue and XPR does not depend on frequency. The co-polar channel components, i.e. $h_{RR}$ and $h_{LL}$ were varied with values from 1 to 50 while the cross-polar channel components, i.e. $h_{RL}$ and $h_{LR}$ were minimised with value equal to 1. This is to simulate various channel conditions, for example, from the cross-polar components are as high as co-polar components (when $h_{RR} = h_{LL} = h_{RL} = h_{LR} = 1$) to the cross-polar components are very low and can be deem as negligible to the co-polar components (when $h_{RR} = h_{LL} = 50$ and $h_{RL} = h_{LR} = 1$).
From Figure 2-17, both eigenvalues showed the tendency to be closer to each other as the cross polar rejection gets higher, thus illustrating better orthogonality. At high XPR, there is a presence of strong LOS channel whereby co-polar components have similar magnitudes as compared to negligible cross-polar components. Therefore, the eigenvalues got closer to each other. This in turn indicates that higher capacity can also be attained.

Figure 2-17: XPR vs eigenvalue for a dual circular polarised MIMO (DCPM) system, where the co-polar components of channel, $H$ were varied while the cross-polar components were minimised and not varied

In [73], the research studied the characteristics of DCPM using measured and modelled data. From the measurement, the channel matrix can then be expanded to be:

$$
\begin{bmatrix}
    h_{RR} & h_{RL} \\
    h_{LR} & h_{LL}
\end{bmatrix}
= \begin{bmatrix}
    \alpha + \sigma \exp(j\theta_{RR}) & \beta + \sigma \exp(j\theta_{RL}) \\
    \beta + \sigma \exp(j\theta_{LR}) & \alpha + \sigma \exp(j\theta_{LL})
\end{bmatrix}
$$

(2-17)

Where $\alpha$ and $\beta$ are the mean signal levels of the measured co- and cross-polar channels respectively, $\sigma$ is the multipath components and $\theta_{ij}$ are the zero mean randomly distributed elements with unit standard deviation. $\sigma$ represents standard deviation extracted separately for the co- and cross-polar channels while $\theta_{ij}$ represents the phase angles of the individual channels. The research also compared DCPM and equal power allocation MIMO. The results showed that at lower SNR, DCPM fares better in terms of capacity performance than equal power MIMO [73].

In another research, a new modelling on the dual circular polarised MIMO was presented [74]. From the model validation results, it was discovered that the cumulative distribution function (CDF)
plots of measured and modelled data were similar. Also, by analysing the distributions of eigenvalues, it is found that there is strong polarisation multiplex rich environment which was not the case with traditional MIMO system. Therefore, the use of lower complexity DCPM is very beneficial for a multiplex rich channel.

Polarisation based MIMO is considered to be a better solution for MIMO multiplexing technique for frequencies at sub 6 GHz due to its compactness. From previous section, it was discovered that large antenna spacing is needed to achieve high capacity gains in MIMO wireless systems. Alternatively, a low cost solution using dual-polarized antennas is promising where two spatially separated antennas can be easily replaced by a single antenna element exhibiting orthogonal polarisations [70], [75], [76]. In another research work, when comparing polarisation and spatial diversity, for the same antenna size, the capacity of the LOS MIMO channel is significantly higher for polarisation diversity in the case of 4x4 MIMO [77]. It was also found that the capacity of 2 x 2 polarisation diversity MIMO is also comparable to the capacity of spatially separated antennas. This is due to the decreasing correlation between the antenna elements of the co-polar submatrices.

### 2.3 Sub 6GHz Dual polar antennas for fixed links

Ideally, antenna designs for static wireless links must have the capability to provide sufficient bandwidth, directionality, low cross polarization level, low backward radiation, compact size and high port isolation. Unfortunately, in the real case, it is very difficult for all of these requirements to be incorporated simultaneously in a compact antenna. The following subsection discusses a number of techniques that have been adopted in the literature to form a dual polarised antenna using linear and circular polarisations. Since there is no elliptically polarised antenna that has been built so far, these linearly and circularly polarised antennas then become the state of the art reference in forming a suitable candidate for a dual elliptically polarised antenna.

There are also other techniques exploiting the antenna polarisation diversity in enhancing the base station antenna performance. A number of researches use two orthogonal linear polarisations to achieve polarisation diversity. The proposed designs either combined vertical / horizontal polarisations or +45° / -45° linear polarisation. The advantage of this polarisation diversity design is it permits antennas to be co-located and improves the isolation between adjacent antennas in the
array. Moreover, a more advance yet complicated technique is to combine linear and circular polarisation together to achieve triple polarisation diversity. The miniaturisation and compactness of an antenna design is also an essential consideration for dual polarisation method as existing base stations use inefficient methods to overcome the high cost of installing wireless communication equipment.

A microstrip patch antenna being printed has generally simple fabrication and it is fitted into small devices, hence making it a popular choice among antenna engineers. On the other hand, it has several serious drawbacks, such as relatively narrow bandwidth, poor polarisation purity and limited efficiency due to dielectric loss [78]. This therefore requires a number of modifications to be implemented on a patch antenna to overcome these critical issues. Narrow bandwidth can be mitigated by inserting thick substrate layers in a multilayer or stacked structure [79] or by designing an aperture-coupled patch [80]. When designing a patch antenna, it is known that there is a trade-off between port isolation, bandwidth and return loss, which require careful consideration. To reduce port isolation or coupling, different shape of slots is usually added to the radiating patch [79]–[82]. Alternatively, in [83], unbalanced power dividers are utilised to improve the isolation characteristic and hence reducing the complexity of the structure.

In the literature, patch antennas can also be designed to switch between orthogonal polarisations. To produce dual polarisation, different feeding patch are individually fed by unbalanced power dividers with a phase difference of 90° [83], [84]. Another popular choice is to use PIN diodes to carry out the switching electronically [82], [85]. Most dual polarised patch antennas employ stacked or multilayer structure in order to accommodate the requirements for fixed links system, including low cross-polarisation [79], high gain [80] and high front-to-back ratio [81]. This architecture can easily add the level of complexness to a MIMO system.

Another favourite choice for dual polarisation implementation is the dipole antenna. Constructed either using wire, rod or printed, dipole antennas offer wide bandwidth, good radiation characteristics, are easily integrated into circuits and are able to be extended to form an array. In order to perform dual polarisation operation, two pairs of dipoles are combined and placed orthogonally to each other to create a crossed dipole design. When printed, the shapes of dipole arms can differ according to the requirement of the design, instead of the normal rectangular patch [86], for instance, a trapezoidal dipole is introduced in [87], flag-shaped radiators to fit on top of circular patch for impedance matching [88], square loop dipole arms for coupling improvement [89] and arc-shaped dipole arms [90].
Despite having the advantage of a wider bandwidth than patch antenna, one cannot depend solely on the design of dipole antenna. Enhancements to the design need to be added to fulfil the wide bandwidth criteria. The broadband performance is achieved, for example, by adjusting the dimension of conical probe in [90], by introducing parasitic elements to suppress the reactance of the antenna [88], [89], as well as by employing U-shaped coupling feeding [86]. In addition, isolation or coupling reduction issue also need to be addressed by, a few methods such as by adjusting the gap of the feeds [87], adding a balun [84], [88], and integrating a metallic cube [86].

Dual sense (right-hand/left-hand) circular polarisation (RHCP/LHCP) is also accumulating more interest with time due to the benefits it can offer to MIMO communications. With orthogonal polarisations, it is able to combat the channel interference issue. Further to that, not only can it reduce the multipath effects, circular polarisation can also mitigate the transmitter and receiver antenna misalignment problem. Dual sense circular polarisation also allows for the antennas to be co-located which leads to a more compact design. Hence, it is not surprising that it is also popular in other applications such as satellite communications, radar and radio frequency identification.

Among the antenna designs used to create dual circular polarisation are monopole, slot, helix and patch antenna. Authors in [91] proposed a triple band dual circular polarisation antenna using monopole design, whereby the monopole lengths are adjusted to give the required 90° phase difference. Patch antennas, a default choice by researchers, can be used to reconfigure the circular polarisation by inserting switches such as MEMS or PIN diode in the feeding network like in [92], [93] and [94]. Feeding networks consisting of a 90° branch-line coupler and 3 dB Wilkinson power divider can also be used in patch antenna to achieve equal magnitude and 90° phase difference [95]. In slot antenna, the number of the slots and positioning of the slots [96] as well as the size of the slots [97] play a crucial role in producing the switchable circular polarisation. Helix antenna is another favourite selection to produce circular polarisation due to their symmetry of their geometry, balanced feeding and wide angular beamwidth, with quadrifillar helical antenna (QHA) being a famous form of helix antenna. In order to switch between right and left hand circular polarisation, two QHAs which have opposite tuning direction are deployed in [98] while in [99], a QHA is combined with a crossed dipole. A dual sense circular polarisation can also be achieved by feeding the antenna at different ports like in [100].

There is also some research that combines linear and circular polarisations for polarisation reconfigurability purposes. Structures such as an array of microstrip patches on top of a parasitic
layer are designed that can be used for beam steering purposes too [101]. In another piece of work, an interdigital capacitor together with a meandered line inductor on a coplanar waveguide-fed antenna perform both dual band and linear/circular polarisation [102]. Other work in this area used the microstrip patch antenna as a basis and went on to modify them for polarisation switching, for instance, by truncating corners of a triangular or square patch in order to make the structure asymmetrical [103], [104] as well as adding complementary split ring resonator [105], U-shaped slot [106] and V-shaped slits [107].

The antenna designs presented in this subsection are the dual or multi polarised antennas based only on linear and circular polarisations. No elliptically polarised antenna was found in the literature as of the writing of this thesis. This indicates that there is a gap in the literature whereby elliptically polarised antenna has never been built before. Therefore, a new and novel elliptically polarised antenna candidate has to be designed and fabricated in order to perform elliptical polarisation multiplexing. The design for elliptically polarised antenna, which will be presented in Chapter 4, will be based on the dual polarised antennas discussed in this subsection.

2.4 Multiplexing for mmWave backhauls

There have been some studies that research the possibility of utilising MIMO at millimetre wave (mmWave) frequencies. For wireless links at mmWave, it is an interesting question which method will perform best in terms of throughput; is it multiplexing, diversity or beamforming? The channel at mmWave frequency is extremely sensitive to small changes in the environment, therefore would spatial multiplexing still work in this condition?

There is already plenty of throughput at this frequency band due to the high bandwidth that it has offered and is therefore sufficient for the near future demand. However, according to computer simulations conducted in [108], the theoretical throughput can go as much as 10 Gbit/sec in indoor environment, if the channel bandwidth is of the order of GHz, when a 2x2 MIMO spatial multiplexing was deployed. This MIMO system was then able to double the peak data rate of normal wireless link at mmWave. However, since this is a simulation study, the question of how to realise such links in practical remains.
Chapter 2. Background to static wireless links and multiplexing techniques

An outdoor LOS MIMO channel measurement was performed in [109] to evaluate the potential of spatial multiplexing in pure LOS scenario at 60 GHz. It was found that orthogonal channel vectors can be obtained if the spacing criterion is met, therefore enhancing spatial multiplexing over fairly long distances (up to TX-RX distance of 30 m). It was estimated that maximum capacities of 16.9bit/s/Hz and 29.2bit/s/Hz at SNRs of 24.3dB and 24.6dB for the 2 × 2 and 3 × 3 setup, were achieved respectively in this study. This finding then shows that spatial multiplexing is viable in a LOS mmWave communication.

A paper evaluated the MIMO potential by comparing all three different techniques of MIMO; diversity gain, multiplexing gain and beamforming gain at 60 GHz according to IEEE 802.11ad standard [110]. Space-time block coding is used to maximize the spatial diversity, spatial multiplexing is used to increase the spectral efficiency by transmitting several data streams simultaneously while beamforming is used to enhance the directivity of the signal transmitted and received in this research. Upon comparing all the three different methods together with a SISO system, the throughput simulation results in LOS showed that the beamforming gain is able to achieve the same throughput but with lower SNR, 5-6 dB less than a SISO system.

Multiplexing and beamforming methods were also compared in [111], which evaluated the channel capacity that depends on the antenna aperture size in an indoor environment. At high antenna aperture size, the capacity of spatial multiplexing is higher than capacity of beamforming because of the enhanced gain able to focus on the strongest propagation path. At low antenna aperture size, the capacity of spatial multiplexing is comparable to beamforming. Spatial multiplexing is advantageous when the antenna can afford a large aperture size but when there is a design constraint, beamforming can be a substitute as its capacity approximates to the capacity of spatial multiplexing. This then suggests that there is a potential to combine the benefits of spatial multiplexing and beamforming to form a hybrid approach, which was also suggested in [112].

Another paper studied the trade-off between achieving multiplexing gain and diversity gain [113]. An asymmetric subarray structure was built to assess the channel in a LOS scenario. The proposed scheme can afford both multiplexing and diversity gains through carefully designed antenna spacing at both TX and RX. However, the disadvantages of the scheme are it is difficult to know in advance the optimal geometrical settings and the antenna elements separation cannot be dynamically adjusted once it is deployed.
Chapter 2. Background to static wireless links and multiplexing techniques

It is discovered that at higher SNR and larger antenna aperture size, MIMO spatial multiplexing can outperform beamforming and diversity to achieve high capacity. At lower SNR, beamforming has higher potential at millimetre wave frequencies in terms of gaining higher throughput. This is probably due to the fact that spatial multiplexing relies on the richness of multipath propagation and there is not as much multipath to be exploited at 60 GHz as compared to lower frequencies.

2.5 Summary

This review of literature presented in this chapter is focused on fixed or more precisely, static wireless communications. Fixed radio links are able to transmit between two or more fixed points to accommodate telecommunication services, such as voice, data or video transmission. In this chapter, different types of static links were discussed, from larger scale such as backhaul, small cell, WLAN to smaller scale which includes device to device communication. The discussion also analysed and compared both sub 6 GHz and millimetre wave frequency bands.

Multiple antenna techniques are a promising topic area to be researched for static wireless links. MIMO systems can be used to enhance the quality of communication links as well as increasing the data throughput of the system. All three methods i.e. spatial, angular and polarisation for conducting MIMO were presented. In particular, combined together with MIMO methods, polarisation can help in improving the spectral efficiency especially at frequencies lower than 6 GHz. For instance, by using orthogonal polarization, interference from other nearby antennas can be greatly reduced if not eliminated. Hence, the quality of the system is further improved but the ability to form good orthogonality between the polarisations is essential.

Furthermore, examples of dual polarisation antenna design at sub 6 GHz were discussed. Dual linear polarisation antenna is mostly designed using microstrip patch and dipole antennas. Other than linear polarisation, dual circular polarisation as well as linear and circular polarisations combined together were found in the literature. Finally, recent research interest in millimetre wave has prompted a lot of research to study MIMO in this particular frequency band. Analysis of the MIMO methods comparing diversity, multiplexing and beamforming was also presented.
Chapter 3

3 Elliptical Polarisation Multiplexing

As previously elaborated in chapter 2, it is highly beneficial to employ MIMO with polarisation system to create a multiplex rich channel. A major objective of this thesis is therefore to exploit elliptically polarised antennas as a potential method to conduct MIMO multiplexing, with the improvement of better orthogonality in comparison to linear polarisation. This also exploits the fact that any practical antenna is elliptically polarised, even though it may be designed to be linear or circular, practical imperfections will prevent it from ever reaching such an ideal state. The chapter begins with a background to elliptically polarised waves, the fundamental parameters and criteria of elliptical polarisation (EP). The second part of Chapter 3 will explore the technique for implementing EP and investigating its potential for MIMO multiplexing, using metrics including the definition of cross polar ratio (XPR) for EP cases, evaluation and further theoretical analysis to determine the criteria for a strong elliptical XPR, which will result in strong orthogonality.

3.1 Background to elliptical polarisation

Wave polarisation is defined as the shape and locus of the tip of the Electric field, \( E \) vector at a given point in space as a function of time [114]. The time dependent electric field plane is orthogonal to the direction of propagation. In any practical situation, the locus of \( E \) is an ellipse, and the wave is elliptically polarised. However, in a theoretically ideal case, the ellipse, under certain criteria, may form a circular locus or simply a single straight line. In these conditions, the polarisation state is circular or linear, respectively, where the two cases are illustrated in Figure 3-1. A circular locus is made normal to the direction of propagation (i.e. the \( z \) direction) in the circular case, while in the linear case, the polarisation is vertical in this case and so the locus is parallel with the \( x \) axis. In circular polarisation, the electric field components, \( E_x \) and \( E_y \) are equal in magnitude with 90° phase difference, while in linear polarisation, only one of the electric field components has a certain value and the other is equal to zero.
The history of elliptically polarized antennas studies began in the 1950s [116]. In [117], a theorem of reciprocity to calculate the power received when an elliptically polarized antenna intercepts an elliptically polarized wave of different polarization was presented. The calculations were made in terms of the distant field that would be produced by the antenna when radiating a known power. Another publication discussed the concept of equivalent length of an antenna, which is the length of a vector whose scalar product with the electric vector of an incident plane wave gives the open-circuit voltage at the terminals of the antenna when used for reception. This concept became complex but is able to be adapted to antennas radiating and receiving elliptically polarised waves when used appropriately [118].

Figure 3-2 : An example of polarisation with ellipse shape aligned at $\psi$ tilt angle with $E_x$ and $E_y$

Electric-field components
## 3.1.1 Parameters in Elliptical Polarisation

For an elliptically polarised locus looking from the direction of propagation in the z direction, Figure 3-2 illustrates the electric field vector \( \vec{E}(z) \), which includes an x-component, \( \vec{E}_x(z) \) and y-component, \( \vec{E}_y(z) \):

\[
\vec{E}(z) = \hat{x}\vec{E}_x(z) + \hat{y}\vec{E}_y(z),
\]

(3-1)

With

\[
\vec{E}_x(z) = E_{x0}e^{jkz}
\]

(3-2a)

\[
\vec{E}_y(z) = E_{y0}e^{jkz}
\]

(3-2b)

Where \( E_{x0} \) and \( E_{y0} \) are the complex magnitudes of \( \vec{E}_x(z) \) and \( \vec{E}_y(z) \), respectively. In addition, the diagram portrays the rotation angle \( \psi \) which is the angle between the major axis of the ellipse and x-axis (which is chosen to be the reference direction), with a range of \(-\pi/2 \leq \tau \leq \pi/2\). For simplicity, the phase of \( E_{x0} \) is chosen as reference, i.e. assigning \( E_{x0} \) a phase angle of zero and the phase of \( E_{y0} \) is denoted relative to \( E_{x0} \), as \( \delta \). Hence, \( \delta \) is the phase difference between the y-component of \( \vec{E} \) and its x-component. Therefore,

\[
E_{x0} = a_xe^{j\theta} = a_x
\]

(3-3a)

\[
E_{y0} = a_ye^{j\delta}
\]

(3-3b)

Where \( a_x = |E_{x0}| \), \( a_y = |E_{y0}| \) and \( \delta \) is the phase difference between the y-component of \( \vec{E} \) and its x-component as shown in Figure 3-3. \( E_{x0} \) is assigned a phase angle of zero as a reference. Substituting into Equation (3-1), it is computed that the total electric field phasor \( \vec{E}(z) \) is given by

\[
\vec{E}(z) = (\hat{x}a_x + \hat{y}a_ye^{j\delta})e^{jkz}
\]

(3-4)

Major axis \( a_\xi \) along the \( \xi \)-direction and a minor axis \( a_\eta \) along the \( \eta \)-direction are also shown in Figure 3-3. In order to determine the shape and direction of the ellipse, the ellipticity angle \( \chi \) is used here:

\[
\tan \chi = \pm \frac{a_\eta}{a_\xi} = \pm \frac{1}{AR}
\]

(3-5)
Where $AR$ is the axial ratio and $\chi$ are bounded within the range $-\pi/4 \leq \chi \leq \pi/4$. The axial ratio, a crucial parameter when a wave polarisation is involved, is the ratio of major axis to minor axis, and thus given by $AR = a_x/a_\eta$. Rumsey et al in [119] suggested that the concept of axial ratio is similar to impedance ratio. This is because the concept of field impedance involves the ratio of a component of electric field to another component of magnetic field whereas in axial ratio, it involves the ratio of a component of electric field to another component of electric field. When compared together, both involved the ratio of two different components of the electromagnetic field, thus they can be treated the same way. Furthermore, Rumsey et al also showed how the orientation and shape of the polarisation ellipse are represented using the impedance Smith Chart.

![Figure 3-3: Polarisation ellipse with rotation at z = 0 as a function of time with major axis and minor axis shown (that is used to determine axial ratio)](image)

Normally, the locus of the Electric field is an ellipse, where $a_x \neq 0$, $a_y \neq 0$ and $\delta \neq 0$, hence the wave is elliptically polarised. In another interpretation, elliptical polarisation can be achieved only when the time-phase difference between the $E$-components is odd multiples of $\pi/2$ and their magnitudes are not equal or when the time-phase difference is not multiples of $\pi/2$ (irrespective of their magnitudes) [120].
That is,

\[ |\vec{E}_x| \neq |\vec{E}_y| \]  \hspace{1cm} (3-6a)

when \( \Delta \phi = \phi_y - \phi_x = \left\{ \begin{array}{l} \left( + \frac{1}{2} + 2n \right) \pi \text{ for } \text{CW} \\ \left( - \frac{1}{2} + 2n \right) \pi \text{ for } \text{CCW} \end{array} \right. \)  \hspace{1cm} (3-6b)

\[ n = 0, 1, 2, ... \]

or:

\[ \Delta \phi = \phi_y - \phi_x \neq \pm \frac{n}{2} \pi = \left\{ \begin{array}{l} > 0 \text{ for } \text{CW} \\ < 0 \text{ for } \text{CCW} \end{array} \right. \]  \hspace{1cm} (3-6c)

\[ n = 0, 1, 2, 3, ... \]

In addition to the parameters explained previously, the size of the ellipse can be determined by the amplitude \( \vec{E} \) and is defined by:

\[ \vec{E}^2 = |\vec{E}_x|^2 + |\vec{E}_y|^2 \]  \hspace{1cm} (3-7)

where \( \vec{E}_x \) and \( \vec{E}_y \) are complex numbers [121]. Amplitude \( \vec{E} \) then represents the power density (W/m²) in the wave when multiplied by the intrinsic admittance of space. It is also discovered that the relationship between \( \tau, AR \) and \( \delta \) is given by [122]:

\[ \tan \delta = \frac{2AR}{(1 - AR^2)\sin2\tau} \]  \hspace{1cm} (3-8)

Left-handed and right-handed rotation is demonstrated in Equation (3-5) by the plus sign and minus sign respectively. It is right-hand elliptically polarised if the field vector rotates clockwise, and it is left-hand elliptically polarised if the field vector of the ellipse rotates counter clockwise. Elliptical polarisation can then be treated similarly to circular polarisation, where there is a dual sense, i.e. left or right polarisation.

### 3.1.2 Special cases and forms of elliptical polarisation

Under special circumstances, other types of polarisation can be generated. Circular polarisation is generated when the axial ratio is 1 and \( \chi = \pm 45^\circ \) and the ellipse shape changed to a circle. Linear polarisation is produced if the value of \( AR \) is \( \infty \) and \( \chi = 0 \) and the ellipse is reduced to a straight
line. To create circular polarisation, another set of criteria can then be added, whereby the electric field components must be orthogonal to each other. Additionally, they must possess the same magnitude and the time-phase difference between the two components must be odd multiples of 90°. Linear polarisation requires two orthogonal electric field components with the different magnitude and time-phase differences equal to 0° or multiples of 180°. Therefore, it can be said that a wave intrinsically exhibits elliptical polarisation if it is not modified to become linear or circular polarisation. In practice, elliptical polarisation normally refers to polarisation other than linear or circular and the latter polarisations are considered special cases of elliptical. Various ellipses with different tilt and ellipticity angles are depicted in Table 3-1.

Study of the relationship between differently polarised waves is also deemed useful in some instances, such as for a radar’s transmitted and the target’s reflected waves, the uplink and downlink of a mobile communications system, or the interception and jamming of an electronic warfare signal. In 1892, a graphical aid using a sphere shape was developed to analyse the multiple dimensionality of polarised waves by Poincare and is hence named Poincare sphere [123]. Initially introduced to help picturing elliptical polarisation in optics, the Poincare sphere can be utilised as a visual and analytic tool, similar to the Smith Chart for circuit analysis [121]. The special sphere provides a three dimensional (3D) plotting surface, and by using correct spherical-coordinate equations, intra-system calculations can be easily performed.
Ellipses Criteria

- $\psi = 0^\circ$
- $\chi = 0^\circ$
- $AR = \infty$
- Horizontal polarisation

- $\psi = 80^\circ$
- $\chi = 5^\circ$
- $AR = 29$
- Elliptical polarisation

- $\psi = 60^\circ$
- $\chi = 20^\circ$
- $AR = 6.5$
- Elliptical polarisation

- $\psi = 90^\circ$
- $\chi = 25^\circ$
- $AR = 3$
- Elliptical polarisation

- $\psi = 90^\circ$
- $\chi = 45^\circ$
- $AR = 1$
- Circular polarisation

Table 3-1: Different ellipses forming different type of polarisations (EP, LP and CP) when the tilt angle and ellipticity angle were changed
Figure 3-4 portrays a visual representation of orthogonal polarisations in a Poincare sphere. It also describes the major polarisations that are located diametrically opposed on the sphere’s surface which are orthogonal to each other. The sphere’s radius is unity, which relates to the normalised power of the represented polarisations. Latitude and longitude on the sphere illustrate the shape and orientation of the ellipse. Points along the same latitude possess the same axial ratio but a different tilt angle ranging from -90° to 90° depending on the longitude angle. Points with same longitude represent the same tilt angle but a different axial ratio ranging from 1 to \( \infty \). The relationship between the tilt angle, \( \psi \) and the longitude angle, \( \phi \) as well as the relationship between axial ratio, \( AR \) and latitude angle, \( \theta \) are given as:

\[
\psi = \frac{\phi}{2} \tag{3-9}
\]

\[
AR = \frac{1}{\cos \theta} \tag{3-10}
\]

Along the equatorial line, the points exhibit pure linear polarisation. Additionally, the upper hemisphere represents left hand elliptical polarisation with left hand circular polarisation at the top zenith, while lower hemisphere represents right hand elliptical polarisation and subsequently the lower zenith represents right hand circular polarisation. If there is a 90° arc between the polarisations at transmitter and receiver, this represents an infinite signal loss, also known as polarisation mismatch which in turn helps to minimize or even eliminate interference between channels.
3.1.3 Cross Polarisation of antennas

When the terms of co-polarisation and cross-polarisation of an antenna were first explored, there was explicable ambiguity as to how they would be defined. In 1973, Ludwig reported three definitions that were taken from the literature so as to provide meaningful representations of the polarisation [124]. The first definition simply stated that the radiated antenna E-field components were projected onto one-unit linear vector as the direction of reference polarisation while the other as the direction of cross polarisation in a rectangular coordinate system. The difference in the second definition is that the spherical coordinate is used, where the unit vectors tangent to a spherical surface are taken as the reference. Unfortunately, both of these definitions were later rejected by the engineering community [125].

Ludwig went on to deliver the third definition which is widely accepted that better resolves the ambiguity and later adopted as the Institute of Electrical and Electronic Engineers (IEEE) definition [125], [126]. The cross polarisation is given by the polarisation orthogonal to a specified reference polarisation, in this case the co-polarisation, in a specified plane containing the reference polarisation ellipse [127]. In this method, the antenna pattern is measured using an elevation over azimuth mount, whereby a single linearly polarized illuminator must first be aligned with the antenna under test when the azimuth and elevation angles are both 0°. This is crucial so that in this direction, co-polarisation and cross-polarisation are appropriately defined using the fields from a dual orthogonal-port antenna.

![Antenna pattern measurement](image)

*Figure 3-5: Antenna pattern measurement in defining the third Ludwig’s definition of cross polarisation [124]*
Figure 3-5 depicts how the measurement is carried out, by firstly mounting the antenna under test (AUT) on an elevation-over-azimuth positioner with an auxiliary polarisation axis at the origin of the system. Angle $\theta$ is the angle where the AUT is rotated from $z$-axis, while angle $\phi$ is the rotation angle relative to the $x$-axis. Each pattern cut begins at $\theta = 0^\circ$ along the azimuth plane. The $\phi = 0^\circ$ position is defined such that the probe is aligned parallel to the polarisation of the AUT in order to obtain the principal co-polarisation pattern. Subsequently, the $\phi = 90^\circ$ when $\theta = 0^\circ$ is set such that the probe is aligned orthogonal to the AUT polarisation to obtain the cross-polarisation. Therefore, the measured co- and cross-polarisation patterns are given by

$$E_{\text{copol}}(\theta, \phi) = E(\theta, \phi) \{\sin \phi \hat{\imath}_\theta + \cos \phi \hat{\jmath}_\phi\} \quad (3-11)$$

$$E_{\text{xpol}}(\theta, \phi) = E(\theta, \phi) \{\cos \phi \hat{\imath}_\theta - \sin \phi \hat{\jmath}_\phi\} \quad (3-12)$$

where $\hat{\imath}_\theta$ and $\hat{\jmath}_\phi$ are unit vectors, $E_{\text{copol}}$ is the reference or co-polarisation pattern and $E_{\text{xpol}}$ is the cross-polarisation pattern.

### 3.2 Elliptical polarisation multiplexing

#### 3.2.1 Orthogonality between elliptically polarised waves

Two elliptical waves are said to be in the same state of polarization if the two corresponding ellipses are of similar size, similarly oriented and described in the same sense. When orientated orthogonal to each other and with opposite rotations, the waves are said to be orthogonal to each other, as previously described in Chapter 2. The concept of orthogonality of the waves is very useful when applied in a MIMO system, particularly to help improve MIMO multiplexing capabilities. Channels with orthogonal polarisations are able to separate and can minimise if not eliminate the interference between them, thus allowing multiple streams to be multiplexed simultaneously. This section will examine the orthogonality of an elliptically polarised MIMO system and also discusses the correlation between the elliptical modes.
Chapter 3. Elliptical Polarisation Multiplexing

The evaluation initially considers a 2x2 MIMO system. The components of two elliptical polarisations are illustrated in Figure 3-6, where the vertical and horizontal E-fields are defined as $E_y$ and $E_x$ respectively. The ellipses are orientated relative to each other with angle $\alpha$. It is assumed that in this case, the axial ratio, $AR$ is similar for both ellipses and is defined as:

$$AR = \frac{E_y}{E_x}$$  \hspace{1cm} (3-13)

The orthogonality of the ellipses can be evaluated by investigating the correlation between both ellipses. The vectors for the two ellipses are defined as follows, where the first ellipse, $v_1$ is rotated while the second, $v_2$ remains vertical:

$$v_1 = \begin{pmatrix} E_{H1} \\ E_{V1} \end{pmatrix} = \begin{pmatrix} \cos \alpha + \sin \alpha \frac{E_x}{AR} \\ \cos \alpha \end{pmatrix}$$  \hspace{1cm} (3-14)

$$v_2 = \begin{pmatrix} E_{H2} \\ E_{V2} \end{pmatrix} = \begin{pmatrix} \frac{1}{AR} E_x \\ E_y \end{pmatrix}$$  \hspace{1cm} (3-15)

where $E_x$ and $E_y$ are the carrier signals and are defined as phasors in Equation (3-14) and (3-15).

As the ellipse, $v_2$ remains vertical, so $\alpha = 0^\circ$. Equation (3-15) then came from substituting $\alpha = 0^\circ$ into Equation (3-14). In order to define the rotation of the ellipse being either right hand or left hand, $E_y$ can be phase inverted.
Thus the correlation between the two ellipse modes, $\rho_{12}$ is defined as follows [128]:

$$\rho_{12} = \frac{v_1^H v_2 - |v_1||v_2|}{\sqrt{v_1^H v_1 v_2^H v_2}}$$  

(3-18)

whereby, the vector products are defined as:

$$v_1^H v_2 = \frac{1}{R} \left( \cos \alpha + \sin \alpha \right) + \left( \cos \alpha + \frac{\sin \alpha}{AR} \right)$$  

(3-19)

$$v_1^H v_1 = \left( \cos \alpha + \sin \alpha \right)^2 + \left( \cos \alpha + \frac{\sin \alpha}{AR} \right)^2$$

$$= 1 + \frac{1}{AR^2} + \frac{4 \cos \alpha \sin \alpha}{AR}$$  

(3-20)

$$v_2^H v_2 = 1 + \frac{1}{AR^2}$$  

(3-21)

$$|v_1| = |v_2| = 0$$  

(3-22)

Hence, incorporating Equation (3-19) – (3-22) into the previous correlation Equation (3-18):

$$\rho_{12} = \frac{1}{AR} \left( \cos \alpha + \sin \alpha \right) \pm \left( \cos \alpha + \frac{\sin \alpha}{AR} \right) \sqrt{\left(1 + \frac{1}{AR^2} + \frac{4 \cos \alpha \sin \alpha}{AR}\right) \left(1 + \frac{1}{AR^2}\right)}$$  

(3-23)

Equation (3-23) can then be related to the envelope correlation coefficient (ECC) by squaring the complex correlation, $\rho \approx |\rho_{12}|^2$ as was presented by Clarke in [129].

Figure 3-7(a)-(d) are plotted to show the evaluation of orthogonality based on the correlation between two elliptical modes when AR is 4, 10, 50 and 100. AR less than two is considered circular
polarisation and AR more than two is considered elliptical polarisation, therefore only AR more
than two is considered in the correlation evaluation. Let us consider first when it is the case of $AR = 4$. When two ellipses are co rotated, for example when both are right hand elliptical polarised, the
MIMO system will have 50% correlation when separated by 90°, i.e. $\rho_{12} = 0.5$. For meaningful
comparison, the absolute value for a contra rotated correlation is considered. It can be seen that
when the ellipses are contra rotated, i.e. when they have opposite polarisations, the system will have
similar orthogonality when separated by 45°, but when separated by 90° there is complete
orthogonality. As the $AR$ was increased, the co-rotated and the absolute contra-rotated curves get
closer to each other until the point where they became similar in values as can be seen in the case
of $AR = 100$. At this point, the ellipses have similar correlation regardless of whether they have
the same orientation or not because the elliptical polarisation is close to that of linear polarisation
representing two linear fields becoming orthogonal at 90°.

![Figure 3-7: Evaluation of orthogonality based on the correlation between two ellipses when they are
co rotated and contra rotated and the AR is varied](image-url)
Figure 3-8 is drawn to better illustrate this result. The results from the plot in Figure 3-7 also indicate that there are possibilities for other MIMO configurations to be orthogonal. When separated at correct angles, the channels can be perpendicular and thus separated from each other. Figure 3-8 display examples of these configurations up to 20 x 20 MIMO system. The $AR$ for each configuration is different in this figure for purpose of clarity.

In such a case, to exemplify the channel matrix for these configurations, a 4x4 MIMO is defined as follows, whereby the correlation in this case also represents the relative ratio of magnitude in the branches between the polarisations and $\phi_{m,n}$ is the random phase alterations in each antenna port:

$$
H = \begin{pmatrix}
\rho_{11}e^{j\phi_{11}} & \rho_{21}e^{j\phi_{21}} & \rho_{31}e^{j\phi_{31}} & \rho_{41}e^{j\phi_{41}} \\
\rho_{21}e^{j\phi_{21}} & \rho_{22}e^{j\phi_{22}} & \rho_{32}e^{j\phi_{32}} & \rho_{42}e^{j\phi_{42}} \\
\rho_{31}e^{j\phi_{31}} & \rho_{32}e^{j\phi_{32}} & \rho_{33}e^{j\phi_{33}} & \rho_{34}e^{j\phi_{34}} \\
\rho_{41}e^{j\phi_{41}} & \rho_{42}e^{j\phi_{42}} & \rho_{34}e^{j\phi_{34}} & \rho_{44}e^{j\phi_{44}}
\end{pmatrix} \quad (3-24)
$$
Figure 3-8: Example elliptical constellations from 2 x 2 to 20 x 20 MIMO configurations where neighbouring ellipses have opposite sense of rotation.
3.2.2 MIMO Multiplexing using Elliptical Polarisation

Capacity or spectral efficiency of an environment is one of the measures to determine the multiplexing capability of a MIMO system. It depends on the antenna structures at each end, knowledge of channel information at the ends, correlations between antennas and between paths, distribution of scatterers and most importantly on the SNR level. Expressed in the well-known capacity formula [130], capacity for a single-input single-output (SISO) channel is:

\[
C = \log_2(\det(I + \left(\frac{P}{Mn}\right)HH')) 
\]  

(3-25)

where \(P\) is the average received power at the receiver with additive noise, \(n\) while \(M\) is the number of transmit antennas.

A comparison of the resulting capacity for different topologies of MIMO used here is shown in Figure 3-9. The AR is set at 4 and the simulations were performed using 10 000 random samples. The channel matrix was set as follows; the co-polar coefficients were set to be random samples but the cross-polar coefficients were set to be zero, hence no cross-polarisation in the MIMO system. The gradient of the cumulative distribution is shown to increase towards vertical as the number of ellipses increase towards 20. At 20, the curve is vertical and there is no change in the capacity due to the random phase changes in the source of each ellipse. At such a point, the correlation is high which does not allow any multiplexing to be possible. Hence the gain in capacity beyond this point is representative of a beamforming gain through which greater capacity is reached with a larger number of elements.
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Figure 3-9: Comparison of capacity for SISO and $M \times N$ MIMO systems where the AR is set at 4 and there is no cross-polarisation in the simulated MIMO channel model

To identify the loss in multiplexing, another method of comparison is also evaluated. Investigation the degrees of freedom (DoF) found in a MIMO system was performed in several publications. The DoF measure, as stated in [131] includes time, frequency, the spatial dimension at the transmitter and the spatial dimension at the receiver. In [132] it was discovered that the spatial DoF depends on multipath richness of the propagation channel as well as antenna aperture size, where antenna aperture is the radius or area of the two-dimensional (2D) or three-dimensional (3D) multipath observation regions. Spatial DoF was observed to get higher when using larger antenna aperture at the TX and RX for a given multipath richness. It was also reported that the DoF indicates the number of effective antenna elements for the spatial multiplexing on the antenna aperture, in this case, at most 4 eigenmodes is feasible using an electrically small mobile antenna. Further to this, other research concluded that if the antenna aperture size is increased, extra DoFs can be effectively obtained in OLOS and NLOS channels as compared to LOS channel [133]. Moreover, other papers have also investigated the relationship between the DoF and the angular correlation of multipath [134], [135]. These works were showing how the DoF are affected by multipath from a restricted range of angles and they confirmed that the multipath richness is proportional to the angular spread of multipath power. It was also suggested that random multipath scattering, regardless of the number of receiver antenna elements, has an intrinsic limited richness [136] and proposed that the receiver antenna placements in MIMO systems are optimised in order to maximise capacity.

It is of interest to measure the number of multipaths that exist in the environment, also known as the multipath richness, in order to compare and evaluate different elliptical polarisation
constellations. Multipath richness can be quantified using a metric similar to the effective DoF [137], [138]. For a single channel realization as effective DoF,

\[
\text{Multipath Richness} = \sum_{k} |S_k| / \max_k |S_k| \tag{3-26}
\]

Where \( S_k \) is the \( k \)th singular value of the channel matrix. This value lies on the range \([1, \min(M, N)]\) and represents the effective number of parallel channels that can be formed.

Another way to compute intrinsic multipath richness is by using the normalised eigenvalues to determine the volume of eigenmodes and by express them as a cumulative sum of the log of the eigenvalues richness, independent of the power level [139]. The number of significant eigenvalues or singular values determines this.

In the case of the unknown channels at the transmitter, the richness curve, \( \zeta(k) \) is now defined as the cumulative sum of the log of the eigenvalues, \( \lambda_i \),

\[
\zeta(k) = \sum_{i=1}^{k} \log_2(\lambda_i) \tag{3-27}
\]

The richness curve contains a significant amount of information concerning the degree of useful multipath.

![Figure 3-10 : Comparison of richness in DoF for \( M \times N \) MIMO systems defined at the integer values for each eigenvalue, computed based on Equation (3-26)](image)
Hence, by knowing the number of elliptical modes that have significant capacity gain, the most promising elliptical mode solutions can be designed. In Figure 3-10, the richness in DoF is plotted as a curve defined at the integer values for each eigenvalue, computed based on Equation (3-26). Note it is defined here as richness in DoF because there is no scattering present as such, but separate eigenmodes are created due to the orthogonality of the elliptical modes. This plot is very useful in measuring the scale of DoF when perfect LOS propagation is assumed. In general, it can be observed that the higher the polarisation mode is, the higher the maximum richness is but all the curves have decreased gradients. The scattering richness does not grow as the number of elliptical modes are increased. At high polarisation modes, i.e. 10 x 10 onwards, there is minimal increase in richness due to a larger number of modes indicating that beyond this point there is only a a beamforming gain possible rather than a multiplexing gain.

An array of twenty elements, for instance, can be used in the 20 x 20 case, however, since the curve plot is saturated at a certain maximum value, it can be seen that only six or seven elements are contributing to the substantial increase in multiplexing gain. In the other cases, the gradient of the richness curves starts to decrease around the third eigenvalues for any constellation, which means there is very marginal increase in multiplexing from beyond the 4 polarisation states. Ideally, if all the modes were able to provide multiplexing, then richness would increase beyond the 4th eigenvalue. However, in this case, higher order constellations caused the lower eigenmodes to be weak (as is the case in a beamforming rich channel). The number of antenna branches using elliptical polarisation then have little case to extend beyond four branches in order to be multiplex rich. This is also in agreement with the practical constraint for real antenna design in relation to the size of the antenna. Therefore, for the reasons given above, the optimised MIMO setup to gain the highest richness in DoF would be a 4 x 4 MIMO system.

### 3.2.3 Cross Polarisation Power Ratio (XPR) Evaluation

Several studies to investigate a cross-polarised antenna system using computed XPR were reported in [69], [140]–[144]. In this chapter, the multiplexing capabilities of the elliptically polarised MIMO system are evaluated by comparing the XPR derivation for different types of polarisations. It is of interest to determine how great orthogonality can be obtained in a practical case with elliptical polarisation when in theory it is perfectly orthogonal. In this thesis, 2 x 2 MIMO configuration is
chosen as a start to examine this measure of orthogonality. Simulations to determine the XPR were conducted, assuming a free space environment.

Firstly, the XPR for linear polarisation is presented. The setup for dual linear polarisation consisting of vertically and horizontally polarised antennas was shown in Figure 2-12 in chapter 2. It is assumed that all the antennas are perfectly aligned to each other, i.e. VP antenna at transmitter is aligned to the VP antenna at receiver and the same goes to HP antenna. The channel matrix, $\mathbf{H}$ was computed previously in Equation 2-15.

The cross-polarisation power ratio received at vertically polarised antenna, $XPR_V$ is commonly defined as follows:

$$XPR_V = \frac{\text{power of co-polarised channel}}{\text{power of cross-polarised channel}}$$

$$XPR_V = \frac{|h_{VV}|^2}{|h_{HV}|^2}$$

(3-28)

While the cross-polarisation power ratio received at horizontally polarised antenna, $XPR_H$ is:

$$XPR_H = \frac{|h_{HH}|^2}{|h_{HV}|^2}$$

(3-29)

![Figure 3-11: A MIMO channel model for generating circular polarisation at post processing that is used to derive XPR$_C$](image-url)
Section 3.1.1 previously mentioned the criteria required to create circular polarisation. Equation (3-28) can then be expanded to produce circular polarisation at receiver by introducing a 90° phase difference between the vertical and horizontal components. The channel matrix for circular polarisation (CP) was given in Equation (2-16). The derivation of XPR for CP is based on Figure 3-11. A phase change of $e^{j\theta}$ and $e^{j\theta'}$ is added at each transmitter and receiver components to generate circular polarisation. Ideally, when there is only direct link of VP to VP antenna or HP to HP antenna and there is no transmission from HP antenna to VP antenna and vice versa, the cross-polarisation power ratio for circular polarisation, $XPR_{C,\text{ideal}}$ is determined to be:

$$XPR_{C,\text{ideal}} = \frac{\text{power of sum of co-polarised signals}}{\text{power of sum of cross-polarised signals}}$$

$$XPR_{C,\text{ideal}} = \frac{|h_{VV}\text{|C} + (-j)(j h_{HH})|^2}{|h_{VV}\text{|C} + (j)(j h_{HH})|^2}$$

$$= \frac{|h_{VV}\text{|C} + h_{HH}|^2}{|h_{VV}\text{|C} - h_{HH}|^2}$$

(3-30)

where $h_{VV}\text{|C}$ is $h_{VV}$ adjusted to be co-phased with $h_{HH}$.

However, in real world, there will be some cross-polarisation where some of the signals transmitted by a HP antenna are received by a VP antenna and vice versa. Hence, the cross-polarisation power ratio for circular polarisation, $XPR_{C}$ is:

$$XPR_{C} = \frac{|h_{VV}\text{|C} + j h_{HV} + (-j)(j h_{HH}) + (-j)(h_{VH})|^2}{|h_{VV}\text{|C} + j h_{HV} + j(j h_{HH}) + j h_{VH}|^2}$$

(3-31)

Re-arranging Equation (3-31),

$$XPR_{C} = \frac{|h_{VV}\text{|C} + h_{HH} + j(h_{HV} - h_{VH})|^2}{|h_{VV}\text{|C} - h_{HH} + j(h_{HV} + h_{VH})|^2}$$

(3-32)

The case of EP is more complicated due to the freedom of phase differences that it can employ. An EP can utilise any value of phase except for 0° and 90° as mentioned before. Figure 3-12 portrays a 2x2 MIMO system using EP where XPR value can be calculated from.
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Figure 3-12: MIMO channel model for generating elliptical polarisation at post processing that is used to derive $XPR_E$

Based from the Figure 3-12, there are two elliptical modes at transmitter, $P_{in1}$ and $P_{in2}$ as well as at the receiver, $P_{out1}$ and $P_{out2}$. Four different phase changes, $e^{j\phi_{T1}}$, $e^{j\phi_{T2}}$, $e^{j\phi_{R1}}$ and $e^{j\phi_{R2}}$ are introduced at the transmitter and receiver in order to create the required ellipses. The channel matrix for EP is given by:

$$H = \begin{bmatrix} h_{11} & h_{12} \\ h_{12} & h_{22} \end{bmatrix} \quad (3-33)$$

Since there are two elliptical modes, there will then be two cases of XPR:

$$XPR_{E1} = \frac{P_{out1}}{P_{out2}} \bigg| \frac{P_{in2}}{P_{in1}} = 0 \quad (3-34a)$$

$$XPR_{E2} = \frac{P_{out2}}{P_{out1}} \bigg| \frac{P_{in1}}{P_{in2}} = 0 \quad (3-34b)$$

In the first case based on the Equation (3-32a), when $P_{in2} = 0$ is visualised in Figure 3-13.
$P_{in1}$ is set so that it is unity power at the branch inputs. The power at the output branches can therefore be derived as:

$$P_{out1} = |(h_{11} e^{j0} e^{j0} + h_{12} e^{j0} e^{j\phi_{R1}}) + (h_{22} e^{j\phi_{T1}} e^{j\phi_{R1}} + h_{21} e^{j0} e^{j\phi_{T1}})|^2$$  \hspace{1cm} (3-35a)

$$P_{out2} = |(h_{11} e^{j0} e^{j0} + h_{12} e^{j0} e^{j\phi_{R2}}) + (h_{22} e^{j\phi_{T1}} e^{j\phi_{R2}} + h_{21} e^{j0} e^{j\phi_{T1}})|^2$$  \hspace{1cm} (3-35b)

Therefore, the first definition of XPR, $XPR_{E1}$ for EP is:

$$XPR_{E1} = \frac{|h_{11} + h_{12} e^{j\phi_{R1}} + h_{22} e^{j(\phi_{T1} + \phi_{R1})} + h_{21} e^{j\phi_{T1}}|^2}{|h_{11} + h_{12} e^{j\phi_{R2}} + h_{22} e^{j(\phi_{T1} + \phi_{R2})} + h_{21} e^{j\phi_{T1}}|^2}$$  \hspace{1cm} (3-36)

Figure 3-14 is a depiction of the second case, $XPR_{E2}$, when $P_{in} = 0$. 
$P_{in2}$ is set so that it is unity voltage at the branch inputs when $P_{in1} = 0$. Hence, at the output branch:

$$P_{out1} = \left| (h_{11}e^{j\phi_{T1}}e^{j\phi_0} + h_{12}e^{j\phi_{T2}}e^{j\phi_R1}) + (h_{22}e^{j\phi_{T2}}e^{j\phi_R1} + h_{21}e^{j\phi_0}e^{j\phi_{T2}}) \right|^2 \quad (3-37a)$$

$$P_{out2} = \left| (h_{11}e^{j\phi_{T1}}e^{j\phi_0} + h_{12}e^{j\phi_{T2}}e^{j\phi_R2}) + (h_{22}e^{j\phi_{T2}}e^{j\phi_R2} + h_{21}e^{j\phi_0}e^{j\phi_{T2}}) \right|^2 \quad (3-37b)$$

The second XPR is then computed as:

$$XPR_{E2} = \frac{\left| h_{11} + h_{12}e^{j\phi_R2} + h_{22}e^{j(\phi_{T2}+\phi_R2)} + h_{21}e^{j\phi_{T2}} \right|^2}{\left| h_{11} + h_{12}e^{j\phi_R1} + h_{22}e^{j(\phi_{T2}+\phi_R1)} + h_{21}e^{j\phi_{T2}} \right|^2} \quad (3-38)$$

Upon comparing all three types of polarisation, it is not easy to design and practically build a perfect linearly polarised antenna with $0^\circ$ phase shift or $90^\circ$ phase shift for circular polarisation. Any phase deviation from these criteria will simply generate elliptical polarisation which is far from the ideal case. Therefore, there is an opportunity to exploit elliptical polarisation with DoF in phase and also imperfections of the cross polarisation in an antenna to create better orthogonality, and hence multiplexing, between two polarisations.
3.3 Theoretical analysis of elliptical polarisation multiplexing

Further theoretical analysis was performed to find out the conditions to produce a good XPR. Assuming this is a practical EP antenna, a high value of XPR, if not infinity, can only be obtained when the denominator of Equations (3-36) and (3-38) is equal to zero or at least minimised. Thus, based on this statement, the criteria for a good XPR is derived as follows.

For maximum $XPR_{E1}$, assuming that $h_{11}$ and $h_{22}$ have the same magnitude as well as $h_{21}$ and $h_{12}$ also have the same magnitude, these terms in the denominator of Equation (3-36) is made equal to zero:

\[
\begin{align*}
    h_{11} + h_{22}e^{j(\phi_{T1}+\phi_{R1})} &= 0 \\
    h_{12}e^{j\phi_{R2}} + h_{21}e^{j\phi_{T1}} &= 0
\end{align*}
\] (3-39a) (3-39b)

Solving these equations will get the required phase shift, $\phi_{R2}$:

\[
e^{j2\phi_{R2}} = \frac{h_{11}h_{21}}{h_{22}h_{12}}
\] (3-40)

Then, having known $\phi_{R2}$, $\phi_{T1}$ can be resolved:

\[
e^{j\phi_{T1}} = \frac{-h_{12}e^{j\phi_{R2}}}{h_{21}}
\] (3-41)

Similarly, maximum value of $XPR_{E2}$ are obtained when the denominator terms of Equation (3-38) are also equal to zero:

\[
\begin{align*}
    h_{11} + h_{22}e^{j(\phi_{T2}+\phi_{R1})} &= 0 \\
    h_{12}e^{j\phi_{R1}} + h_{21}e^{j\phi_{T2}} &= 0
\end{align*}
\] (3-42a) (3-42b)

Upon solving the equations above, the following phase shifts are derived:
\begin{align}
    e^{j2\phi_{r2}} &= \frac{h_{11}h_{12}}{h_{22}h_{21}} \\
    e^{j\phi_{r1}} &= \frac{-h_{21}e^{j\phi_{r2}}}{h_{12}}
\end{align}

Thus all the phase differences can be set to meet these criteria.

In another set of evaluations to examine the criteria for good XPR, these phase shifts were put to a test. A simulation was performed with random phase offsets applied to the four polarisation components (i.e. VV, HH, VH and HV) that works for any vertical to horizontal ratio. They are randomly offset from each other by virtue of the antenna phase shifts at boresight. Therefore, this can exploit the fact that antennas have a low cross polar ratio, which is characteristic of an elliptically polarised antenna. The distance between the TX and RX in the simulation is set to be 2m. An assumption of a 0 dB co-polarisation and 6 dB cross-polarisation existing in the antenna was made in order to resemble a practical scenario. A set of synthetic data generated using the random offsets are shown in Figure 3-15.

![Figure 3-15](image)

**Figure 3-15 : Synthetic phases generated using random offsets and low cross polarisation to be used in evaluation to examine the criteria for good XPR**

Using these synthetic data, the XPR for EP is calculated and presented in the plot in Figure 3-16. The XPR value for LP at 6 dB is also added for comparison purposes. Indeed, the result shows that the XPR of the elliptical mode goes to infinity.
Following the synthetic data results, it is important to analyse the conditions required to achieve perfect orthogonality. In order to take a closer look into the conditions required to achieve good orthogonality and hence good multiplexing gain, co-polar and cross-polar branch imbalances, $k_1$ and $k_2$ are introduced. One of the branch was impaired by offsetting one of the channel components against the other, i.e. constant $k_1$ is introduced to weaken one of the co-polar components while $k_2$ is introduced to weaken one of the cross-polar components. Therefore, $h_{11}$ will not have the same magnitude as $h_{22}$ and similarly, $h_{12}$ will not have the same magnitude as $h_{21}$. The constants are defined as follows:

$$h_{22} = k_1 h_{11} e^{j \phi_1} \quad (3-45)$$
$$h_{21} = k_2 h_{12} e^{j \phi_2} \quad (3-46)$$

The elliptical XPR is plotted against co-polar and cross-polar branch imbalance in Figure 3-17. It can be seen that when both constants $k_1$ and $k_2$ are 0 dB, which correspond to the magnitudes of $h_{11}$ to be equal to $h_{22}$ and $h_{12}$ to be equal to $h_{21}$, indeed both elliptical XPR become very high at around 100 dB. When there is a slight branch imbalance or when the components do not have similar magnitude, $XPR_{E1}$ and $XPR_{E2}$ quickly drop to a much lower value at approximately 20 dB. As a result, this finding shows that when there is no branch imbalance, EP will exhibit extremely high polarisation purity and so, the antennas will be able to clearly distinguish between two orthogonal elliptical modes at the RX.
The XPR evaluation reveals only the condition required for the magnitude of the channel coefficients in order to achieve strong orthogonality. Another investigation is then required to study the condition necessary for the phase of the channel coefficients. This investigation was conducted using another eigenvalues derivation by Vaughan and Bach Andersen [145]. The equations used are:

\[ G = H^H H \]  \hspace{1cm} (3-47)

\[ G = \begin{bmatrix} |h_{11}|^2 + |h_{12}|^2 & h_{11}^* h_{21} + h_{12}^* h_{22} \\ h_{11}^* h_{21} + H_{12}^* H_{22} & |h_{22}|^2 + |h_{21}|^2 \end{bmatrix} = \begin{bmatrix} a & c \\ c^* & b \end{bmatrix} \]  \hspace{1cm} (3-48)

\[ \lambda_{\text{max}} = \frac{1}{2} \left( a + b + \sqrt{(a - b)^2 + 4|c|^2} \right) \] \hspace{1cm} (3-49)

\[ \lambda_{\text{min}} = \frac{1}{2} \left( a + b - \sqrt{(a - b)^2 + 4|c|^2} \right) \] \hspace{1cm} (3-50)

whereby \( G \) is the inner Gram matrix. From Equation (3-49) and (3-50), it can be seen that the phase and magnitude conditions need to meet the criteria of \((a - b)^2 + 4|c|^2\) so that the first and second eigenvalues can be as equal as possible. Therefore, the SVD was then performed using these set of conditions; \(|h_{11}| = |h_{22}|, \ |h_{22}| = |h_{12}|\) and \(\phi_{11} + \phi_{22} - \phi_{12} - \phi_{21} = \pm 180^\circ\) in order to fulfil this criterion. The ratio of eigenvalues produced when the phase shifts were set in this way is then depicted in Figure 3-18. It can be seen that the ratio of eigenvalues dropped to 0 dB when there is no branch imbalance, i.e. when \(k_1\) and \(k_2\) are 0 dB. This also verifies that more than 30 dB \(XPR_E\) of and less than 0.5dB of ratio of eigenvalues can be reached when \(k_1\) and \(k_2\) are above -1dB. The MIMO channel obtains perfect orthogonality when the eigenvalues are equal. Therefore, this
finding then proved that, the magnitudes of the co-polar components have to be equal, so does the magnitudes of the cross-polar components and phase shift between the co-polar components has to be \(180^\circ\). These three conditions then necessitate the conditions to achieve perfect orthogonality and thus high multiplexing gain.

![Figure 3-18: Evaluation of ratio of eigenvalues vs branch imbalance \(k_1\) and \(k_2\) where conditions for perfect channel orthogonality can be investigated](image)

### 3.4 Summary

Chapter 3 introduced the basics of elliptically polarised electromagnetic waves. Various parameters to generate elliptical polarisation were described, including the definition, axial ratio, ellipticity and tilt angle. It has also been established that other special polarisations, specifically linear and circular polarisations may arise from elliptical polarisation. Hence, it can be said that the radiation from all antennas is essentially elliptically polarised. Nonetheless, comparatively little attention has been given to this aspect of radiation. Elliptically polarised antennas also promote different problems of representation and measurement that do not happen with straightforward linearly polarized antennas, to exemplify, phase measurements.

The second part of this chapter presented how elliptical polarisation can be used to perform MIMO multiplexing. A comparative analysis was done to evaluate other \(M \times N\) MIMO configurations in
order to achieve orthogonality. The higher the order of the MIMO constellation, the weaker multiplexing capability for MIMO is due to the higher correlation between elliptical modes. Elliptical polarisation has little case to extend beyond $M = N = 4$ due to decreased gradient of the richness plot, and thus, very marginal increase in richness. Therefore, in this research work, a constellation of $2 \times 2$ is being focused so as to reduce the complexity and investigate the degree to which full orthogonality can be reached. Furthermore, cross-polar ratio was also defined for all three polarisations that were being compared; linear, circular and elliptical polarisations. This is crucial to analyse the MIMO multiplexing capabilities. Finally, it was found that in order to achieve perfect orthogonality and high multiplexing gain, these conditions needed to be fulfilled; magnitudes of the co-polar components have to be equal, so does the magnitudes of the cross-polar components and phase shift between the co-polar components has to be $180^\circ$. 


Chapter 4

4 Elliptically Polarised Antenna

In the previous chapter, the theory of performing MIMO multiplexing using two orthogonal elliptical polarisations was discussed. This chapter subsequently progresses to purposely design an elliptically polarised antenna. A candidate for an elliptically polarised antenna is introduced in this chapter, suitable for analysis in measurement. The elliptically polarised antenna is based on a crossed dipole design and is known as STAR dipole antenna (due to its star shape). Its design and method of construction are illustrated in this chapter. The simulated and measured antenna characteristics are also shown. As a reference linearly polarised antennas were also built for comparison purposes. The performance of both antennas were compared and summarised later in the chapter. A preliminary measurement data taken in isolated free space environment is also presented to verify the means by which elliptical polarisation has improved orthogonality based upon the antenna design.

4.1 STAR Dipole Antenna

4.1.1 Design and fabricated STAR dipole antenna

Building two orthogonal elliptically polarised antennas can be made possible by deliberately creating cross polarisation between their linear components, whereas in a circular case the cross polarisation is expected to be negligible. Therefore, the design here aims to make a consistent cross polarisation between two antennas that will enable the possibility to form orthogonal ellipses based in the phasing applied to the linear elements. An antenna prototype was constructed using wire and re-shaping dipole antennas because of its flexibility to be shaped into different forms to deliberately generate cross polarisation. Figure 4-1 further explains how the first antenna candidate is re-formed.
Chapter 4. Elliptically Polarised Antenna

The first antenna candidate is the STAR dipole antenna. The design is based on two dipole antennas; one horizontal and one vertical, spaced apart at a distance. These spaced dipoles will later be used as a reference antenna throughout the thesis. Using two antennas spaced at a distance is undesirable as they consume unnecessary space particularly at lower frequencies. Therefore, a more compact design was formed which is termed as crossed dipole with the two linear elements co-located as can be seen in Figure 4-1(a) [146]. However, the co-located elements are subject to cross coupling which reduces the polarisation purity of the vertical and horizontal elements. Hence, some degree of cross polarisation is generated, which is exacerbated by the presence of feeders in a practical scenario. The crossed dipole design for these reasons is unable to transmit perfect dual linear polarisation or circular polarisation due to the presence of cross polarisation. In the practical case, antenna designs are intended to attempt to minimise the cross coupling in order to generate a polarisation as close to dual linear or circular polarisation as possible. Furthermore, circular polarisation requires a phase difference of $90^\circ$ between the vertical and horizontal elements [147] but this is difficult to achieve in practise with the mutual coupling present.

![Figure 4-1: Step by step of how STAR dipole is created; (a) Building a crossed dipole (b) Forming STAR dipole from crossed dipole](image)
As opposed to attempting to suppress the cross polarisation in a crossed dipole design, the aim is now to exploit this issue by reforming the linear antennas to deliberately generate cross polarisation, which will subsequently make the polarisation highly elliptical. This is achieved by reducing the four arms in a crossed dipole down to three arms to form a STAR antenna as illustrated in Figure 4-1(b). The STAR antenna has two radiating arms that would each be connected to a corresponding inner core of two separate coaxial feeds, while the third arm would be connected to the ground branch of both the coaxial feeds. Hence there is now just one ground arm to the two dipoles. The antenna is now re-formed to look like a star shape, hence the name STAR dipole as a novel design to deliberately generate elliptical polarisation far removed from linear or circular polarisation. When the ground arm is taken away, one of the electric field components for both polarisations will be weakened and no longer be zero, as in the case of linear polarisation. The structure then loses its symmetry.

Figure 4-2 visualises the elliptical polarisations that would be created by the results of this method. In Figure 4-2(a), one of the radiating arm is excited and some of the currents will couple to the other radiating arm, creating two electric-fields, $E_{V1}$ and $E_{H1}$ in the vertical and horizontal fields respectively. These electric-fields will in turn generate the first elliptical mode. The second elliptical mode, as seen in Figure 4-2(b), is produced using the same method, whereby the other radiating arm is excited and some of the currents are coupled onto the adjacent radiating arm, thus creating $E_{V2}$ and $E_{H2}$. It is important to note that all electric-fields, $E_{V1}$, $E_{H1}$, $E_{V2}$ and $E_{H2}$ are of different magnitude so that elliptical polarisations can be made. Therefore, the remaining ground arm together with two radiating arms will produce two elliptical modes at the same time as in Figure 4-3. It is important to note that the radiating arms are perpendicular to each other so that two orthogonal polarisations can be achieved, even if they are not aligned vertically and horizontally. The STAR dipole possesses physical advantages such as a more compact in size compared to the spaced dipoles as well as being less complicated in construction compared to crossed dipole. The performance of the STAR dipole in executing MIMO multiplexing using dual elliptical polarisation will be compared with dual linear polarisation in subsequent sections and chapters.

The STAR dipole was later simulated in Computer Simulation Technology Studio Suite (CST) software to find an optimised design. The construction of simulated STAR dipole antenna is shown in Figure 4-4. There are a few types of dipole, the one used here is a half-wave dipole, so the length of a single radiating arm is a quarter-wavelength and constructed using copper wire to obtain maximum current flow. The STAR dipole was fed using coaxial cable. After optimising the design, the length of all dipole arms, $l_1 = l_2 = l_3$ is found to be 23.7 mm. The thickness of the wire, $t_1$ is
1 mm, which is the same as thickness of the inner coaxial cable, $t_2$. The insulator thickness, $t_3$ is 1.2 mm while the outer coaxial cable thickness, $t_4$ is 1.7 mm. The inner coaxial cables are connected to the radiation arms of Port 1 and 2 while the outer coaxial cable is connected to the ground arm.

A dipole radiates an omnidirectional pattern. To improve on directivity, a square ground plane is placed at a quarter-wavelength behind the STAR dipole, also made from copper. The optimised gap between dipole and ground, $g$, which is supposed to be around a quarter-wavelength, is 28.7 mm, whereby $g_1$ is 25 mm and $g_2$ is 3.7 mm. The size of the ground plane is $l_4 = l_5 = 160$ mm while the thickness of the ground plane, $t_5$ is 1 mm.

A prototype of the STAR dipole was built and presented in Figure 4-5. Two antenna prototypes were constructed, one to be used at TX while the other at RX, but due to the near similarity in the characteristics of both prototypes, only one is discussed here. After taking into account the coupling effects between the dipole arms and the effect of the coaxial cable, the dipole arm length is adjusted to get the wanted scattering parameters (S-parameter).

Figure 4-2: A visualisation of the elliptical polarisations that would be created by the results of reforming the crossed dipole to form STAR dipole (a) Illustrated current and phase constellation 1 (b) Illustrated current and phase constellation 2
Figure 4-3: A visualisation of the dual orthogonal elliptical polarisations that would be created by the result of reforming the crossed dipole to form STAR dipole

Figure 4-4: Simulated design of STAR Dipole antenna with dimensions (a) Top view (b) Close-up view (c) Side view (d) Perspective view
4.1.2 Simulation and measurement results

Figure 4-6 compares the simulated and experimental S-parameters of both prototypes of STAR dipole. It is apparent from this figure that the antenna is well matched at the operating frequency, which is at 2.3 GHz. The simulated return loss at resonant of port 1, $S_{11}$ and port 2, $S_{22}$ are both -33 dB while the 10 dB bandwidth is 380 MHz. However, it can be seen that the isolation loss between port 1 and 2 or $S_{12}$ is significantly high, at around -6 to -8 dB. Therefore, there is a huge coupling between ports to overcome in practical context. On the other hand, the resonant of measured $S_{11}$ and $S_{22}$ are slightly shifted to lower and higher frequencies as compared to the simulated return loss. This is due to the necessity to compromise practically between port 1 and 2 so that matching at the nearest operating frequency can be achieved. Nevertheless, there is a
somewhat good agreement between the simulated and measured 10 dB impedance bandwidth. It can also be seen that both prototypes showed a consistency in s-parameter characteristics.

Figure 4-6: Simulated vs measured S-parameter of STAR dipole prototypes
Chapter 4. Elliptically Polarised Antenna

(a) when only Port 1 was excited

(c) When only Port 2 was excited

Figure 4-7: Simulated current distribution of STAR dipole antenna

Figure 4-8: Simulated current distribution of STAR dipole when excited simultaneously

(a) STAR dipole 1

(b) STAR dipole 2

Figure 4-9: Measured S-parameter of STAR dipole prototypes with and without hybrid coupler. A hybrid coupler was attached to the STAR dipole to minimise port coupling
The simulated current distribution of STAR dipole is displayed in Figure 4-7. The highest amount of current is shown by the red colour whereas the least current is marked by the blue colour. In Figure 4-7(a), the current is strong and distributed well in the radiating arm, with some of the current flowing into the ground arm when only port 1 is excited. Significantly less current has flown into the second radiating arm. Similar observations can be found in Port 2 in Figure 4-7(b). Thus, based on the good current distribution, the STAR dipole antenna will potentially exhibit high gain. Additionally, this observation can also be used to explain the high isolation loss seen in Figure 4-6, which is caused by some of the current coupled to the adjacent radiating arm. Figure 4-8 illustrated the current distribution when both arms were excited simultaneously in simulation. It can be observed that there is stronger current flowing in the second port in comparison to the first port.

Consequently, an external Midisco hybrid coupler (frequency range from 2 to 8 GHz), was attached to the STAR dipole antennas in the measurements to minimise the coupling between port 1 and 2. Hybrid couplers are normally used in applications such as antenna feed networks, power dividers, combiners, mixers and digital phase shifters. Hybrid couplers use the distributed properties of microwave circuits, in which case, the coupling process generally occurs within a quarter-wavelength or multiple quarter-wavelength portions of the device. The hybrid coupler used in this measurement is a rat race coupler. A diagram of a rat race coupler is shown in Appendix B whereby, if Port 2 and 4 are the input ports, then Port 1 and 3 are the 3 dB output ports, and vice versa. The Midisco hybrid coupler is a 180° hybrid and a 3 dB coupler, whereby when a signal is applied to one of the ports, that signal is equally divided at the two opposite ports (through port and coupled port) with a relative phase shift of 180°. Port 1 of STAR dipole is connected to Port 2 of the hybrid coupler and Port 2 of STAR dipole is connected to Port 4 of hybrid coupler. The measured S-parameter of the hybrid couplers, Coupler 1 and Coupler 2, are also shown in Appendix B.

Figure 4-9 shows the measured S-parameter of STAR dipole prototypes with and without hybrid coupler. From Figure 4-9(a) and (b), we can see that the usage of the hybrid coupler dramatically reduces $S_{12}$ to around -26 dB from -10 dB in previous measurement. There is high isolation between port 1 and 2 of STAR dipole as compared from before using hybrid coupler. When Port 2 is the input port, the signals in Port 4 were 180° out-of-phase with Port 2, therefore, no signal comes out of Port 4, making it highly isolated from Port 2. The same goes for when Port 4 is the input port.

However, it can be seen that the reflection coefficient for both ports are now increased to around -10 dB whereas it was much lower before using the external hybrid coupler. The $S_{14}$ increased and did not show any resonance at the centre frequency. Figure B-1 in Appendix B shows the diagram
of the hybrid coupler that is attached to the STAR dipole antenna. The increased $S_{11}$ is caused by the high coupling then reflecting back to the source constructively, which worsens the return loss. This is portrayed by the blue and red lines in Figure B-1 in Appendix B. Other antenna studies also recorded similar observations when the antennas were attached to an external hybrid coupler [148]–[154].

Nonetheless, the antenna has a close to 10 dB return loss, which is still an acceptable return loss with minimal mismatch loss for the purpose of this study. The hybrid coupler, that was used to decouple the antennas, degraded the return loss but the patterns still radiated and there is still isolation between ports as seen in Figure 4-9(a) and 4-9(b). A stub matching technique could have been used to improve the return loss of the STAR dipole as in [148]. However, such matching would have been carried out at the centre frequency and it would have compromised the antenna bandwidth to be lower than the 200MHz required for channel sounder measurements. Therefore, to enable a maximum bandwidth in this instance while still having acceptable return loss, the stub was not used in this instance.

In Figure 4-10, the 3D radiation patterns of both ports of STAR dipole were simulated and plotted. The aim of simulating radiation patterns is to do further analysis such as to identify the phase angle in simulation that will help predict the form of ellipse of the simulated STAR dipole. It is also important to simulate the 3D pattern as this could not be captured in measurement. The results showed that the patterns are directional, provided by the ground plane that acts as a reflector. The
highest gain is 7.82 dB which is similar for both ports and as expected from the simulated current distribution. The efficiency of the STAR dipole antenna is calculated at 99.95%.

The ellipses radiated by the STAR dipoles were measured in receive mode in the anechoic chamber and shown in Figure 4-11 and 4-12. The linear horn antenna at the transmit end was tilted every 9° to capture the elliptical polarisation from the STAR dipole at the receive end for the whole range of 360°. The AR of STAR Dipole 1 is calculated to be 3.52 dB at 2.3 GHz while the AR of STAR Dipole 2 is calculated to be 2.55 dB at 2.3 GHz. For both prototypes of STAR dipole antenna, the ellipses radiated are fully orthogonal to each other, as observed from the orthogonal tilt angles. The ellipses in Figure 4-11 has odd ellipse shapes. It is likely due to the two ellipses, that were generated from each port of hybrid coupler, are superimposing on each other, thus creating this type of ellipses.

As both ports are being fed by the hybrid, that must mean that two ellipses are being generated from each port then super imposing on each other to give this kind of pattern. The ellipses will change at every frequency but for the frequencies where there is good capacity, it is interesting to see the ellipses and what type of radiation that was happening. In this case, the frequency of interest is the centre frequency where the antennas resonate which is at 2.3 GHz.

Comparisons of simulated and measured radiation patterns of STAR dipole 1 and STAR dipole 2 are shown in Figure 4-13 and 4-14 respectively. The radiation patterns of the STAR dipoles were measured in an anechoic chamber against a reference horn antenna, operating from 2-8 GHz. The gain recorded in the radiation pattern measurement is the absolute gain in dBi. In the measurement, Port 1 was positioned horizontally while Port 2 was positioned vertically. Therefore, the radiation pattern was measured along the azimuth plane of Port 1 of STAR dipole while Port 2’s radiation pattern was measured along the elevation plane. The measured co-polar radiation patterns follow closely the simulated ones in co-polar plots at boresight. The gain at boresight is recorded as 6.8 dBi. On the contrary, the measured and simulated cross-polar radiation patterns differs more, at around 7-8 dBi, with the simulated ones overestimate their values at some azimuth angles. In the simulation, discrete ports were used on the coaxial cables and these discrete ports likely caused the difference between simulated and measured radiation patterns, especially around ±90° onwards. Overall, it can be observed that both prototypes showed a consistency in radiation pattern characteristics.

The radiation patterns of the circular and linear components of the antenna was also measured and presented in Figure 4-15 and 4-16 respectively. There was a reasonable consistency in most of the
radiation patterns in both plots with the exception of some of the circular polarisation patterns. It can be seen that in Figure 4-15, the STAR dipole 1 Port 2 RHCP and STAR dipole 1 Port 1 LHCP have slightly different patterns which is caused by the fact that they are not completely identical antennas as they only mirror each other. Additionally, the orthogonality of the ellipses can be noted in the radiation pattern plot in Figure 4-15 whereby Port 1 is RHCP while Port 2 is LHCP.

Figure 4-11: The ellipse measured from STAR dipole 1. The AR of STAR Dipole 1 is calculated to be 3.52 dB at 2.3 GHz.

Figure 4-12: The ellipse measured from STAR dipole 2. The AR of STAR Dipole 2 is calculated to be 2.55 dB at 2.3 GHz.
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Figure 4-13: Simulated and measured radiation patterns of STAR dipole antenna 1

Figure 4-14: Simulated and measured radiation patterns of STAR dipole antenna 2

Figure 4-15: Measured radiation pattern comparison of circular polarisation components (RHCP and LHCP) of STAR dipole

Figure 4-16: Measured radiation pattern comparison of linear polarisation components (VP and HP) of STAR dipole
4.2 Spaced Dipoles

4.2.1 Design and fabricated spaced dipole antennas

As explained in section 4.1, spaced dipoles are used as a reference antenna that radiates dual linear polarisations, particularly horizontal and vertical polarisations. Figure 4-17 depicts the polarisations produced by the spaced dipole antenna. From this figure, it is apparent that a horizontal dipole creates horizontal polarisation while a vertical dipole generates vertical polarisation. The spaced dipole design is then simulated and optimised in CST as shown in Figure 4-18. The optimised length of all dipole arms, $l_1$, are the same and follow the length of STAR dipole arms, which is 23.7mm and is equal to a quarter-wavelength. The gap $g_1$ which is the gap between the horizontal and vertical dipole is set at one wavelength apart, i.e. 125 mm. A reflector is positioned at a quarter-wavelength from the dipoles and is optimised at $g_2 = 31.25$ mm with $l_2 = 160$ mm. The thickness of dipole arm and ground plane is $t_1 = t_2 = 1$ mm. The coaxial cable connection for the simulated spaced dipole antenna follows similar dimension as in STAR dipole coaxial feed. The spaced dipole design was fabricated and tested to get the antenna’s characteristics. Two prototypes were constructed to be used at TX and RX. The fabricated antenna is displayed in Figure 4-19.

Figure 4-17: Spaced dipole antenna with horizontal and vertical polarisations. Horizontal dipole creates horizontal polarisation while a vertical dipole generates vertical polarisation
Figure 4-18: Simulated spaced dipole antenna with dimensions
4.2.2 Simulation and measurement results

The S-parameters from simulation and measurement for both prototypes of spaced dipole are plotted in Figure 4-20. Both $S_{11}$ and $S_{22}$ for simulated and measured data showed similar trend and are well matched at 2.3 GHz. In addition, the 10 dB bandwidth for simulated and measured results is almost equal at around 200 MHz. At resonant frequency, the simulated S-parameters are better than the measured S-parameters with -19 dB and -11 dB respectively. On the other hand, the simulated $S_{21}$ is significantly below the measured $S_{21}$ by 40 to 60 dB throughout the frequency band. This is due to the presence of the feeder cable in the measurement, which was not simulated in CST. Nonetheless, a 20 dB of isolation was achieved in measurements and this is sufficient to avoid coupling between ports. It is also important to observe that the isolation loss is clearly less than the return loss and hence, a hybrid coupler is not required. Both prototypes showed a consistency in s-parameter characteristics.
Figure 4-20: Simulated vs measured S-parameter of spaced dipole

(a) Spaced dipole 1

(b) Spaced dipole 2

Figure 4-21: Simulated current distribution of spaced dipole antenna

(a) when the Horizontal dipole was excited  
(b) when the Vertical dipole was excited
The current distributions of the horizontal and vertical dipoles are shown in Figure 4-21. When the current distribution is computed, the effects from the adjacent antenna is taken into account. However, since the antennas are one wavelength apart, the coupling between them are very low. Therefore, the effects from the adjacent antenna can be omitted as evidenced from the low isolation loss. The highest current distribution, which can be observed from the red colour region, is recorded at 29.1 dB for horizontal dipole and 29.2 dB for vertical dipole. It can thus be suggested that the spaced dipole antenna will likely produce a good amount of gain. The efficiency of the antenna is calculated to be 97.59%.

(a) Vertical Polarisation. Highest gain recorded at 7.87 dB
(b) Horizontal Polarisation. Highest gain recorded at 7.86 dB

Figure 4-22: Simulated 3-D radiation pattern of spaced dipole antenna. The efficiency of the antenna is calculated to be 97.59%.

In Figure 4-22, the 3D radiation patterns of both dipoles are illustrated. Overall, there is a good gain produced by the antenna, 7.87 dB for vertical polarisation and 7.86 dB for horizontal polarisation. Both patterns showed high directivity as a result from the reflector. A comparison between simulated and measured radiation pattern is performed in Figure 4-23. In the co-polar plot, there is only a slight difference of around 2 dBi in simulation and measurement pattern at antenna boresight but more pronounced difference of 20 to 30 dBi at the side lobes. It is noted that the simulated radiation pattern for both dipoles are 2 dBi higher than the measured ones, particularly at antenna boresight. In contrast to the co-polar data, the vertical dipole cross-polar data for both simulated and measured are quite comparable with a few exceptions. There are a few inconsistencies observed with the horizontal dipole cross-polar simulation and measurement. At some azimuth angles, simulation is better than measurements and vice versa. This may be attributed to the slightly different way of constructing both vertical and horizontal dipole in practice, i.e. the position both
antennas might be somewhat shifted. On average, the cross-polar results are 20 dBi lower than the co-polar results. Overall, both prototypes showed a consistency in radiation pattern characteristics.

![Simulated vs measured radiation pattern of spaced dipole antenna](image)

(a) Spaced Dipole 1

(b) Spaced Dipole 2

Figure 4-23: Simulated vs measured radiation pattern of spaced dipole antenna

### 4.3 Comparison of elliptically and linearly antennas

In this section, the characteristics of both antenna designs presented in this chapter are summarised and compared. STAR dipole antenna showed from 15 to 20 dB measured return loss at the resonant frequency. On the other hand, the spaced dipole recorded lower measured return loss, at only 11 dB. The measured value is much lower than the simulated one because the effects of coaxial cable was not simulated and taken into account, therefore there will be coupling effects with the ground.
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Nevertheless, the measured return loss is greater than 10 dB return loss which is still acceptable. Then, the current distribution is simulated and displayed for all antennas. All current is distributed well in the radiating and ground dipole arms. It was found that only a small amount of current is leaked into the adjacent dipole arm. 2D and 3D radiation patterns were computed in simulation as well as in anechoic chamber measurements. All 3D patterns observed a pair of orthogonal polarisations. The gain for STAR dipole and spaced dipole are comparable at around 7.8 dBi.

All measured 2D patterns follow similar overall trend except for the case of cross-polar data of spaced dipole. In most co-polar and cross-polar plots, the simulation results tend to overestimate their value. The efficiency of all dipoles is comparable and consistent between the elliptically and linearly polarised antennas as plotted in Figure 4-24. In Figure 4-24, there is no efficiency loss observed as they have the same peak gain. Therefore, it can be concluded that compacting and decoupling them does not compromise the dipole efficiency.

Figure 4-24: Comparison of measured radiation pattern of STAR dipole and spaced dipole antenna prototypes. The efficiency of all dipoles is comparable and consistent between the elliptically and linearly polarised antennas.
4.4 Anechoic Chamber Evaluation – Free Space Multiplexing

After examining the elliptically and linearly polarised antennas’ characteristics, the dual polarised MIMO channels were measured and compared in a free space environment, i.e. an isolated environment where it is assumed that there is no other interference in the channel. The free space measurements were performed in an anechoic chamber, which is considered as the best case scenario pure LOS channel conditions. This is crucial in order to study the effects of polarised antenna onto the MIMO system without any reflected or diffracted signals. The measurement setup, whereby one dual polarised antenna is placed at the TX and one at RX, was shown in previous chapter in Figure 3-12, and the measurement photos are shown in Figure 4-25. Vector Network Analyzer (VNA) was used to measure the signals received from the antennas. The VNA was calibrated using a full 2-port calibration. The calibration was conducted by connecting the VNA’s ports to be used to open, short and 50\( \Omega \) terminations. By performing the calibration, the cable losses were eliminated and not taken into account during measurement, therefore ensuring accuracy of the measurement performed. This process was repeated for both VNA Port 1 and 2. The channel matrix, \( H \) was then measured by connecting VNA’s Port 1 to TX antenna and VNA’s Port 2 to the RX antenna. The first channel coefficient measured, \( S_{11} \) is from transmitter’s Port 1 to receiver’s Port 1. This was followed by \( S_{12} \) whereby transmitter’s Port 1 was connected to receiver’s Port 2. All the channel coefficients were then measured using this method. During the whole process of calibration and measurement, the cables were carefully moved from one port to another to avoid phase instability from the cables.

Figure 4-25: Measurement setup in the anechoic chamber (a) at TX end where STAR dipole is attached to hybrid coupler for port coupling minimisation (b) STAR dipole at both TX and RX ends.
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The normalised measured channel $H$ for elliptically polarised STAR dipole and linearly polarised spaced dipoles are portrayed in Figure 4-26 and 4-27 respectively. It is observed that the co-polarised components, $h_{11}$ and $h_{22}$ for STAR dipole are very close for frequency from 2.2 GHz to 2.4 GHz, similar to spaced dipole where $h_{11}$ and $h_{22}$ are also close to each other throughout the frequency band. Meanwhile, the cross-polarised components, $h_{12}$ and $h_{21}$ for STAR dipole differs from each other by as many as 3.5 dB except at 2 GHz and 2.32 GHz where they completely overlap. 

In comparison to STAR dipole antenna, the cross-polarised components for spaced dipole are not equal throughout the frequency band of interest. It is expected that when the co-polarised and cross-polarised components are very similar in value, there is potential that very high orthogonality can occur and this leads to higher polarisation purity.

Hence, the STAR dipole has higher potential at orthogonality due to both co-polar and cross-polar components are very similar at very close to the resonant frequency, i.e. 2.32 GHz, as compared to spaced dipole where the cross-polar components are very far apart at 2.3 GHz. On average, there is a 10 dB difference between co- and cross-polarised elements for spaced dipole but the cross-polarisation rise dramatically for STAR dipole, at only 3 dB below co-polarisation. This shows that STAR dipole design has managed to create significant cross-polarisation between the antennas. Figure 4-28 shows the phase response of channel $H$ for STAR dipole antenna. The phase is shown in all cases to be changing linearly without having any influence on the singular vectors.

![Figure 4-26: Normalised measured channel matrix $H$ for elliptically polarised STAR dipole antenna](image)
Previously, section 3.2.3 described the method to create elliptical polarisation using four phase shifts. These phase shifts were then calculated and presented in Figure 4-29. It is noted that from 2.2 GHz to 2.4 GHz, the phase shifts are almost constant with very small fluctuation. To exemplify, $\phi_{Tz}$ is $150^\circ$ at resonant frequency but as the frequency gets higher, the angle gets lower but only by $20^\circ$. Furthermore, from this plot, we are able to determine the phase angles needed to create EP with the best multiplexing potential. For instance, at the operating frequency, $\phi_{T1} = -49^\circ$, $\phi_{T2} =$
150°, $\phi_{R1} = -115°$ and $\phi_{R2} = 60°$. These phase shifts will then be applied at post-processing for XPR calculation.

Figure 4-29: Four phase shifts computed at TX and RX to create dual orthogonal elliptical polarisation at post processing

In Figure 4-30, the XPR values were calculated for dual EP and dual LP using Equation (3-36) and (3-38). It can be seen that throughout the frequency band, the XPR for EP is higher than LP by 10 to 30 dB. On average, XPR for EP is around 50 to 60 dB while XPR for LP is merely 20 to 40 dB. This shows that it is possible to set all four phase shifts to a value where EP has an XPR nearly 10 to 40dB better than what could be achieved with linear polarisation at that frequency. This therefore proves there is an opportunity to exploit elliptical polarisation and also imperfections of the cross polarisation in an antenna to create better channel orthogonality, and hence multiplexing, between two polarisations.

Figure 4-31 displays the relationship between $\phi_{R2}$, $\phi_{T1}$ and XPR for EP, $XPR_{E}$ at a single frequency. Referring to Equation (3-38) where $XPR_{E2}$ was calculated, $\phi_{R2}$ and $\phi_{T1}$ is in the denominator of the equation. In order to reach maximum $XPR_{E2}$, the denominator has to be made equal to zero. It can be noted that the denominator is equal to zero when $\phi_{R2} = -140°$, $\phi_{T1} = \ldots$
−45° and when \( \phi_{RZ} = 70°, \phi_{T1} = 150° \). As a result, maximum \( XPR_{E2} \) can be achieved using these phase shifts.

![Figure 4-30: Comparison of XPR between EP (STAR dipole) and LP (spaced dipole)](image)

Figure 4-30: Comparison of XPR between EP (STAR dipole) and LP (spaced dipole)

![Figure 4-31: Elliptical phase shifts vs \( XPR_E \) plot computed to investigate the phase shifts at which maximum \( XPR_E \) can be achieved](image)

Figure 4-31: Elliptical phase shifts vs \( XPR_E \) plot computed to investigate the phase shifts at which maximum \( XPR_E \) can be achieved

Another crucial evaluation is the singular value decomposition of channel matrix \( \mathbf{H} \). The singular values were obtained using Equation (2-6) and are plotted as eigenvalues, \( \lambda \) in Figure 4-32 by taking the square of the singular values. It was explained previously in chapter 3 that SVD is used to resolve the MIMO channel and to determine the multiplexing richness of streams whereby, there are two streams in the channel, shown by two comparative eigenvalues in the range of 2.2GHz to 2.4GHz. This is shown more clearly by plotting the ratio of the eigenvalues in Figure 4-33. The first
and second EP eigenvalues are closer to each other at precisely 2.32 GHz and 2.4 GHz compared to LP. At these particular frequencies where the STAR dipole is well matched, this data coincides with Figure 4-26, where the co- and cross-polarised components were of similar values as well as Figure 4-32, where the singular values of EP are very close to each other. As evidenced from these results, it can be shown that EP is able to achieve good channel orthogonality at these frequencies. Hence, with such orthogonality, a STAR dipole antenna is capable in delivering comparable multiplexing gain.

Figure 4-32: Comparison of normalised eigenvalues for EP (STAR dipole) and LP (spaced dipole)

Figure 4-33: Comparison of normalised eigenvalue ratio for EP (STAR dipole) and LP (spaced dipole)
Interestingly, based on the observation on the SVD results in Figure 4-32 and 4-33, there are EP multiplexing potential at two different frequencies, i.e. at 2.32 GHz and 2.4 GHz. To further examine the multiplexing in elliptical polarisation at both frequencies, the $U$ matrices and $V$ matrices of the STAR dipole antenna are plotted in Figure 4-34. The results are separated in different sub-figures for better understanding as some of their values are equal and overlap with each other, e.g. $u_{11} = u_{22}$ and $u_{12} = u_{21}$. In the first case i.e. at 2.32 GHz, it can be observed that the magnitude of all $U$ and $V$ components are equally divided at 2.32 GHz. When the branch power is divided equally between the ports, then $u_{11} = u_{22} = u_{12} = u_{21} = v_{11} = v_{22} = v_{12} = v_{12} = \frac{1}{\sqrt{2}} = 0.7071$ or equivalent to 3dB. This is then crucial to achieve best multiplexing and orthogonality.
The second case i.e. at 2.4 GHz, is best explained using Figure 4-35. Two eigenmodes are shown in Figure 4-35, whereby the STAR dipole antenna together with the hybrid coupler is used to produce the ellipses. In Eigenmode 1, more power is supplied to one branch of the hybrid coupler at the TX, which is shown by higher value of $v_{11}$ and lower value of $v_{12}$ in Figure 4-34(a). At the RX, both coupler branches and U-components were activated in order to optimally receive and match the EP produced by TX. The same goes to Eigenmode 2, shown by higher value of $v_{22}$ and lower value of $v_{21}$ in Figure 4-34(b). Hence, in this case, only one of the STAR dipole branches is powered by the U-components in each mode, which creates an elliptical mode, but then the V-components are used to form a co-polarisation at the receiver with the transmitted ellipse while rejecting the orthogonally transmitted ellipse at the same time.

Additionally, a capacity plot comparing EP, LP and the capacity limit is presented in Figure 4-36. The capacity limit is the maximum capacity limit for two perfectly orthogonal antennas in a 2 x 2 MIMO which is calculated as follows:

$$ C_{\text{lim}} = \log_2 \left[ \det \left( I + \left( \frac{\rho}{2} \mathbf{I}^H \right) \right) \right] $$

(4-1)

From 2.2 GHz to 2.4 GHz, capacity for LP displays very close resemblance to the capacity limit. In contrast to LP, the result for EP shows slightly greater fluctuation. Nonetheless, from 2.3 GHz to 2.4 GHz, the capacity for EP is comparable, if not slightly higher than LP, and is very close to the capacity limit. Therefore, it can be concluded that EP’s performance is comparable, if not slightly
better than LP. This also shows that there is an opportunity to utilise the elliptical polarisation concept to produce a high multiplexing MIMO system.

![Figure 4-36: Comparison of capacity for EP (STAR dipole), LP (spaced dipole) and theoretically perfect orthogonal 2x2 MIMO](image)

**4.5 Summary**

In summary, STAR dipole antennas was presented as the candidates for elliptical polarisation. Their designs are based on a crossed dipole structure. Instead of having four dipole arms as in the usual case for crossed dipole, STAR dipole only have three dipole arms. The structure consists of two radiating arms and one ground arm. This is crucial so as to generate significant cross-polarisation within the antenna and thus creating elliptical polarisation. A reference linearly polarised antenna, which is the spaced dipole antenna, was also constructed as a comparison to the elliptical polarisation’s performance. From the measured radiation patterns, it was shown that the efficiencies of the elliptically polarised STAR dipoles were not affected from compacting them and are consistent with the linearly polarised spaced dipoles. A fair comparison of performance for both antennas can then be performed in subsequent analysis.

A set of channel measurements was performed in anechoic chamber using the elliptically and linearly polarised antennas. In order to study the potential of orthogonality for MIMO multiplexing, XPR was calculated and channel was decomposed using SVD. The results showed that, at 2.3 GHz which is very close to the resonant frequency and at 2.4 GHz, EP has better XPR and closer singular values than LP which leads to better orthogonality. A capacity plot was also presented and it was observed that at resonant frequency, where the antennas are well matched, the capacity of EP is slightly better than LP and is very close to the capacity limit for a 2 x 2 MIMO system.
Chapter 5

5 Dual Polarisation Multiplexing Measurement Campaign

In previous chapter, the STAR shaped dual dipole antenna was designed and constructed as a prototype of an elliptically polarised antenna. The antenna was then tested in an ideal and isolated free space environment to study the effects of elliptically polarised (EP) multiplexing without any interference. However, this experiment was inadequate. Hence, an evaluation to examine the performance of EP in a multiplex rich channel as well as in a diversity rich channel was conducted in a real environment. A channel sounder was used to carry out the channel measurement. By using the channel sounder, channel properties can be characterised and thus they can be utilised for channel modelling and system simulation. The measurement campaign took place in an indoor office environment where LOS, obstructed LOS and NLOS scenarios were studied. In the analysis of the measurement data, the eigenvalues, the ratio of eigenvalues (\( \lambda \)), cross polar power ratio (XPR) and capacity are discussed to evaluate and to compare the multiplexing performance of EP to linear polarisation (LP).

5.1 Measurement Setup

5.1.1 Antenna configuration

Both dual elliptical and linear polarisation antennas were placed vertically at the top of the channel sounder’s transmitter and receiver units side by side, as seen in Figure 5-1. This vertical orientation is necessary to enable the antennas at TX and RX to face each other in a direct LOS transmission for maximum power transfer. The separation between the STAR dipole and spaced dipole antennas is approximately one wavelength to avoid mutual coupling, and the height of the antenna from ground is about 2.4 m. At the TX, the antennas were labelled as \( EP_{1 \text{port1}} \) and \( EP_{1 \text{port2}} \) for STAR
dipole, and $HP_1$ and $VP_1$ for the spaced dipole, while at the RX, the antennas were labelled as $EP_{2\text{Port1}}$, $EP_{2\text{Port2}}$, $HP_2$ and $VP_2$, as displayed in Figure 5-2. Note that in Figure 5-2 the antennas are arranged in a “mirror” format so that for example the STAR antennas are at opposite sides at the TX and RX respectively. This ensures that the STAR antennas are aligned with and facing each other when in a line of sight. The same applies to the VP and HP dipoles.

Figure 5-1: Indoor measurement campaign setup whereby the STAR dipole and spaced dipole antennas were placed at the top of the channel sounder TX unit

Figure 5-2: Antennas at TX and RX aligned in a “mirror” format so that they face each other appropriately in a line of sight
5.1.2 Measurement equipment

A channel sounder was used to characterise the channel properties and record the measurement data. The equipment used in this measurement campaign was the Elektrobit Propsound wideband MIMO channel sounder which can measure frequencies from 2.2 – 2.6 GHz and 5.2 – 5.4 GHz with up to 200 MHz of null-to-null bandwidths. Additionally, it can measure wideband MIMO channels with up to 56 x 32 TX and RX elements. The channel sounder includes a TX unit and a RX unit, whereby the RX unit is linked to a data storage box using one RF chain. It can also be powered using a main supply or a battery for mobile operation. The signal transmitted from the channel sounder is a direct-sequence spread spectrum signal which was produced from binary phase-shift-keyed modulated pseudo-noise codes; by evaluating their correlation vs time delay, the wideband impulse response can be established. By using fast switching TDM techniques, each MIMO channel was obtained sequentially. In order to exceed the Nyquist sampling, around three MIMO channel matrices should be taken in every wavelength at the maximum Doppler frequency shift. In other words, the sampling frequency should be three times the maximum Doppler shift or more. Moreover, each MIMO matrix must remain well within the channel coherence time so that the data can be captured and the channel could be considered constant [155]. A picture of the Elektrobit Propsound Channel Sounder is shown in Figure 5-3 with corresponding schematic drawings of the TX and RX in Figure 5-4 and 5-5 respectively.

![Elektrobit Propsound Channel Sounder consist of TX unit, RX unit, Antenna Switching Unit (ASU) and Data Storage Box](image)

Figure 5-3: Elektrobit Propsound Channel Sounder consist of TX unit, RX unit, Antenna Switching Unit (ASU) and Data Storage Box
Chapter 5. Dual Polarisation Multiplexing Measurement Campaign

Figure 5-4: Schematic drawing of the TX architecture in the Elektrobit Propsound Channel Sounder that consists of TX System Integration Unit, TX PN Sequence Generator Unit, TX RF Unit and ASU

In the TX unit architecture, shown in Figure 5-4, the TX System Integration Unit provides the main control over the transmission and the timing for switching sequence. The PN sequence is first generated in the TX PN Sequence Generator Unit, which is then modulated into Intermediate Frequency (IF). Using a local oscillator, the TX Radio Frequency (RF) unit then up-converted the IF to the desired RF frequency before transmission. The TX RF unit also has a power amplifier to amplify the transmitted power to the Antenna Switching Unit (ASU). The connection to the array of antennas is controlled by the ASU.

On the RX side, which is shown in Figure 5-5, similar architecture to the TX can be seen but with a reversed process. The RX is based on a super heterodyne architecture. At the ASU, a low noise amplifier (LNA) is positioned to minimise receiver noise figure and maximise the SNR. The received complex samples are down-converted to IF and post processed in the RF unit. It is then sampled using an Analogue-to-Digital Converter (ADC) to get the raw baseband I/Q-data and sent to the Post Processing Unit. Using the sliding window cross-correlator technique, the impulse response (IR) of the channel is then extracted from the raw I/Q-data after which the data is stored.
One crucial factor to note is the issue of jitter or variation on the PN sequence. This issue is resolved through synchronisation using high stability rubidium clocks. As seen in the Timing blocks in Figure 5-4 and 5-5, the rubidium is tuned to receive the clock sequence with the right timing. Before commencing a measurement campaign, the clocks are synchronised by connecting the TX and RX boxes. Once synchronised, their MIMO branch switching and phase tuning are subsequently set. The lack of switching synchronisation would otherwise cause the power delay profile to exit the sounder’s power delay profile display window during measurement. Additionally, there is a separate synchronisation for switching between the MIMO branches using the timing circuit which is pre-programmed so over time it is known what is transmitting and what is receiving.

This measurement campaign consists of 4 transmitters and 4 receivers, making a total of 16 combinations. A delay domain spatial resolution of 100 MHz is used to capture all relevant multipath components. The maximum Doppler shift is calculated as:

$$\text{Max Doppler Shift} = \frac{v_{\text{walk-max}} f_c}{c}$$  

where $v_{\text{walk-max}}$ is the assumed maximum walking speed, $c$ is the speed of light and $f_c$ is the centre frequency. In this evaluation, the $v_{\text{walk-max}} = 1.111$ m/s (corresponding to 4km/h) and $f_c = 2.3$ GHz, hence the maximum resolvable Doppler shift is 8.5177 Hz. In order for the signal to be accurately sampled, the Nyquist criterion states that the sampling frequency needs to be at least three times the maximum Doppler shift, in this case, it is set at 30.934 Hz. The transmit power is 10 dBm. The null-to-null bandwidth, $B_{rf}$ with two samples per chip is computed as: 

$$B_{rf} = \frac{v_{\text{walk-max}} f_c}{c}$$
$B_{rf} = 2R_c$  \hspace{1cm} 5-2

where $R_c$ is the chip rate. The channel sounder is set to have a bandwidth of 200 MHz, so the chip rate is 100 MHz. This corresponds to a 10 ns for the duration of a single chip.

### 5.1.3 Measurement scenario

The measurement campaign was performed in an indoor environment: specifically, in an office setting. The setting included a seminar/meeting room, a short corridor and an atrium with a spiral staircase on the lower ground level of the Institute of Communication Centre (ICS), University of Surrey. An indoor setting is very suitable to represent static wireless links scenario which are mostly direct LOS paths with fixed TX and RX as well as moving scatterers. The transmitted signal will encounter more obstacles and reflective surfaces that results in higher amount of multipath from a lot of direction at the RX. Nine scenarios in this measurement campaign that are described in Figure 5-6 together with room dimensions and explained as follows. In all scenarios, the measurements were completed twice to ensure repeatability of the measurement. Since the repeatability is proven, only one set of data is discussed for each scenario in the following sections except for the atrium moving (AM) scenario. For the AM scenario, AM1 and AM2 are presented to compare repeatability of the measurement campaigns.

i. **LOS Edge Walking (EK)**

This is a LOS measurement conducted in the seminar room, between a concrete wall and a few tables at the edge of the room as in Figure 5-6(a). The measured TX-RX distance was 14.5 m. In this scenario, the TX and RX were stationary as two people walked in between the TX and RX. Hence, it was predicted that there would be a stronger LOS and moderate multipath caused by the moving scatterers.

ii. **LOS Edge Waving (EV)**

This is the same measurement as in EK except that the two people were walking and waving their hands in the air so as to provide more obstruction and create more multipath. The multipath arrived at different time intervals caused by moving scatterers and their hand movements. There were also some reflected propagation paths caused by the ground, walls and tables.
iii. LOS Edge Moving Run (EM)
The positions of TX and RX remained the same as in EV and EK. However, in this setting, there were no moving scatterers, but the TX moved towards the RX steadily. This is to capture the effects of the time/space variance where scattering would change significantly in a line of sight dependency on the separation between the TX and RX and their position. It is important to study the effects of different spatial positions where the distance between TX and RX is varied as there may be cases where the signal received varies according to the channel condition at different spatial positions. Therefore, when the terminal is moved, different spatial positions with different channel conditions will be created. Although the terminal is moved, the communication link is still considered as a static link.

iv. LOS Mid Room (MR)
A LOS measurement was also performed in the middle of the seminar room rather than at the edge. This was a more open space compared to the edge of room measurements. The aim here was to provide a different scattering scenario from the edge of the room as well as to determine the effect of scatterers being further away from the line of sight.

v. OLOS Door Diffraction (DD)
Measurements were done in another setting as shown in Figure 5-6(b). The TX was placed at the door of the seminar room while the RX was moved to the front of the room, creating a separation distance of only 6 m. The aim was to create an obstructed LOS where the TX and RX did not directly face each other and some of the signals were diffracted from the door. In this case, the received signal power was expected to be lower with significantly more scattering. The DD measurement is pictured in Figure 5-7.

vi. NLOS Round Door (RD)
In the RD measurement, the TX was positioned outside the seminar room, i.e. in the Atrium while the RX was moved nearer to the door, as shown in Figure 5-6(b). Hence, the propagation path was completely blocked due to the short corridor in between TX and RX. The transmitted signal would have travelled via reflected and diffracted paths in order to reach the RX. This scenario was necessary to examine the capabilities of EP in a NLOS situation, where the channel is mainly diversity rich. This scenario was necessary to examine the capabilities of EP multiplexing in a NLOS situation, where the channel is mainly diversity rich. Although the LOS was blocked, the communication link still
maintained a static link as the terminals were stationary. The RD measurement is portrayed in Figure 5-8.

vii. **LOS Atrium Walking (AK)**
Both the TX and RX were moved to the Atrium just outside of the seminar room, as displayed in Figure 5.6(c). The measured distance between TX and RX was 19.8 m. The Atrium is a vast open space, with a ceiling above the spiral staircase located four storeys up from the lower ground level. Therefore, it was expected to find scattering with more time delay to arrive at the RX, caused by the longer distance. In AK, again, two people walking were in between fixed TX and RX to act as the moving scatterers.

viii. **LOS Atrium Waving (AV)**
The settings in AV was similar to AK, except that, the moving people were waving their hands to create more multipath than in AK.

ix. **LOS Atrium Moving (AM)**
In the AM measurement, the TX was no longer stationary but was moving towards the RX in order to evaluate the effects of varying time and distance in a more open space, in contrast to the seminar room. The Atrium measurements are pictured in Figure 5-9.
Figure 5-6: Indoor measurement campaign scenario plan in all three locations with room dimensions

(a) Seminar room scenario (MR, EK, EM, EV)
(b) Corridor scenario (DD, RD)
(c) Atrium scenario (AK, AV, AM)
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Figure 5-7: OLOS Door Diffraction (DD) measurement photos where the TX was placed at the corridor and the RX was positioned in the seminar room

Figure 5-8: NLOS Round Door (RD) measurement photos where the TX was placed at the Atrium and the RX was positioned in the seminar room

Figure 5-9: Atrium (AK, AV and AM) measurements photos where both TX and RX were placed at the Atrium
5.1.4 Post Processing

Before the data from the channel sounder can be used in the evaluation of EP for MIMO multiplexing potential, it needs to undergo several post-processing stages as illustrated in Figure 5-10. Firstly, the raw I/Q data is converted from the sliding correlator and saved as measured impulse responses (IRs). In the recorded data, some will be recovered as measured delay taps, while any other delay taps that exceed the minimum detectable delay path will be recorded as noise. This noise is determined by setting a minimum measured threshold at 54 dB down from the peak value that was at the shortest distance in LOS so that all delay taps of interest are captured, in this case around 350 ns from the LOS delay tap. After taking into account 63.2 dB of free space path loss, 10 dB of reflection loss and 6 dB of antenna gain, there is around 32.8 dB of fade margin that can be allowed for fading from NLOS measurement and the threshold value is set at -100 dB. Therefore, any delay tap below this threshold value of -100 dB is filtered out to minimise any noise effects. In order to analyse the narrowband channels across the frequency bandwidth, the IRs need to be converted into the frequency domain, which can be achieved using the Fast Fourier Transform (FFT) operation. The narrowband data for each frequency carrier is subsequently extracted from the wideband data so that the results where the antenna is best matched can be obtained. Finally, the narrowband data that is of interest is normalised before it is ready to be utilised in subsequent analysis. In this measurement, the Frobenius normalisation technique is used to remove the path loss included in the data so that the scattering can be specifically analysed. The resultant narrowband data can be analysed to specifically observe the effects of scattering and the orthogonality of the antennas.

![Diagram](image)

**Figure 5-10:** Block diagram for analysing the data obtained from the channel sounder at post-processing
5.2 Physical channel evaluation

The data from the channel sounder measurement was analysed to compare the orthogonality of elliptical and linear polarisations in a static MIMO channel, and are presented over a range of frequency, from 2.2 to 2.4 GHz within which the frequency range of interest is 2.22 to 2.38 GHz as the region in which the antennas were impedance matched as established in chapter 4. The analysis shows results of average performance in the time variant channels at an individual frequency carrier while it also examines more closely at selected frequency carrier instances, which can show the time variant behaviour at the best and worst cases for EP. Thus for this part two frequency carriers were identified for the best and worst cases respectively. The presented results consist of first ($\lambda_1$) and second eigenvalues ($\lambda_2$), the ratio of the two eigenvalues, which will assist in analysing the multiplexing richness of the channels. XPR is used to analyse the orthogonality of the antennas and finally capacity is compared to show the resultant effects on the communication.

5.2.1 Line of sight

5.2.1.1 LOS Edge Walking (EK)

The measurement results in the frequency domain for EK are displayed in Figure 5-11. The first and second eigenvalues of EP and LP are shown in Figure 5-11(a). Overall, it can be seen that the eigenvalues become quite close to each other in this frequency range, which is also shown by a low eigenvalue ratio in Figure 5-11(b) at this frequency. These findings are in agreement with the free space measurements observed in Chapter 4. Previously, it was found that EP had good orthogonality at 2.32 GHz in the free space measurement. However, in this outdoor scenario, the eigenvalues drifted further apart at 2.32 GHz, which shows that the EP starts to be affected by the presence of multipath in a real environment. Interestingly, at 2.34 GHz, EP successfully overcomes the multipath effects that the spaced dipoles are suffering from. Due to the spatial separation in the spaced dipole antennas, LP was adding de-constructively at the RX and thus were susceptible to the channel fluctuations. This problem has been resolved by using EP as co-located antennas and hence, created multiplex rich orthogonal ellipses due to the STAR dipole antennas and the multipath environment.
In Figure 5-11(c), it is apparent that the XPR for EP is substantially higher than XPR for LP. With strong XPR, this indicates that the EP does not have its orthogonal channels substantially affected by the multipath at a single frequency. A capacity comparison between EP, LP and a perfectly orthogonal 2x2 MIMO channel is plotted in Figure 5-11(d). The capacity of a perfect 2x2 MIMO channel is calculated to be approximately 6.9 bit/s/Hz as was shown in Equation (4-1). It is also shown that at 2.34 GHz, there is an improvement in capacity of EP as compared to LP, which shows that EP is able to increase its multiplexing gain when the eigenvalues were closer and the polarisation purity was higher. The existence of multipath in the channel therefore helps EP to achieve multiplex rich orthogonal ellipses. Although the STAR dipole antenna loses its diversity behaviour, it illustrated overall multiplexing improvement by exploiting EP.

Table 5-1 shows the evaluation of instantaneous eigenvalues, eigenvalue ratio, XPR and capacity at the frequency for best and worst case scenarios of EP which is at 2.34 GHz and 2.315 respectively. The cumulative distribution function (CDF) plot of instantaneous eigenvalues of EP and LP is first compared. It can be observed that in the best case scenario, the first and second eigenvalues of EP are substantially closer to each other as compared to eigenvalues of LP, while in the worst case scenario, the LP eigenvalues are still comparable to EP; however, there is a minor improvement from LP, which is seen more clearly when comparing the eigenvalue ratios below. Hence when comparing the EP and LP in this measurement, there is always an improved multiplexing from EP and never a case when it is worse than LP. The third metric is the average XPR of EP in both scenarios is calculated to be around 50 to 60 dB while the average XPR of LP is merely 20 dB. At a few time instances, XPR for EP is seen faltering due to the multipath, but it is noted that there are more fading in the worst case scenario as compared to the best case. Lastly, the fourth metric is the average capacity plot. It is found that at these specific frequencies, the capacity for EP is comparable to the capacity for LP.
Figure 5-11: Frequency Domain EK mean evaluations of (a) eigenvalues, (b) eigenvalue ratio, (c) XPR and (d) capacity averaged over time.
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<table>
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<tr>
<th>Instantaneous</th>
<th>Best Case (2.34GHz)</th>
<th>Worst Case (2.315GHz)</th>
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Table 5-1: Instantaneous evaluations for EK measurement scenario of eigenvalues, eigenvalue ratio, mean XPR and mean capacity
5.2.1.2 LOS Edge Waving (EV)

Similar observations can be noted in EV as in EK measurements as portrayed in Figure 5-12 and Table 5-2. The experiments showed repeatable results with a slightly worse performance in EP and LP. The best case scenario is taken at frequency 2.34 GHz while the worst case scenario is taken at 2.31 GHz. Due to the waving hands occasionally interrupting the LOS between TX and RX, the worst case scenario for EP shows a lot more depolarisation effect from the XPR and thus portrays more diversity rich behaviour, as seen in LP. These increased obstructions of LOS also cause the channel to experience more multipath and higher fluctuations in the instantaneous XPR. As a consequence, it makes two equal power elliptical modes impossible to be distinguished at the RX.
Figure 5-12: Frequency Domain EV mean evaluations of (a) eigenvalues, (b) eigenvalue ratio, (c) XPR and (d) capacity averaged over time
Table 5-2: Instantaneous evaluations for EV measurement scenario of eigenvalues, eigenvalue ratio, mean XPR and mean capacity
5.2.1.3 LOS Edge Moving Run (EM)

In the EM measurement, the TX was moved towards to the RX during the measurement in order to study the effect of time/space variant on EP and LP. The EM measurement data is illustrated in Figure 5-13 and Table 5-3. In general, it is found that LP suffers higher degradation across frequency in comparison to EP when the TX is mobile. In Figure 5-13(a), it is observed that the elliptical eigenvalues are closer, which is more apparent in Figure 5-13(b) where the ratio of EP eigenvalues experienced less fluctuation as compared to the ratio of LP eigenvalues. The reason in less fluctuation is because of the merging of two EP antennas together in the same location generates two elliptical orthogonal modes which stabilises the channel. Therefore, creating higher multiplex rich channel. Another observation can be seen in Figure 5-13(b) whereby more fluctuations occur in the frequencies above 2.3 GHz for EP while less fluctuations occur below 2.3 GHz. On the other hand, LP suffer from fluctuation with more multipath throughout the frequency range of interest due to higher spacing between LP antennas. The elliptical XPR is found to be higher than linear XPR by around 20 to 40 dB as found in the other edge of the room measurements with occasional fading at some frequencies. When compared with all edge of the room results, there are even more fading for EP in this XPR analysis. This finding shows that the increased amount of multipath in the moving channel causes the orthogonality of EP to fail at given time instants. Furthermore, the capacity of EP and LP dropped even further than the capacity for perfect 2x2 MIMO channel. At the resonant frequency, the capacity for both polarisations are only around 6.7 bit/s/Hz.

Table 5-3 presents the best and worst cases for EP in EM measurement. The frequencies are 2.31 GHz and 2.295 GHz for the best and worst cases respectively. It is evident from the instantaneous eigenvalue CDF plot that in the LP case, more multipath in the channel causes increased diversity gain in comparison to the EP case where the behaviour is more inclined towards multiplexing gain. This therefore explains why EP is less susceptible to multipath.
Figure 5-13: Frequency Domain EM mean evaluations of (a) eigenvalues, (b) eigenvalue ratio, (c) XPR and (d) capacity averaged over time.
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<tr>
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| Instantaneous XPR vs Time | | Instantaneous Capacity CDF |
|---------------------------|-------------------------------------------------|
| ![Graph](image3.png)     | ![Graph](image4.png)  |

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<th>Table 5-3: Instantaneous evaluations for EM measurement scenario of eigenvalues, eigenvalue ratio, mean XPR and mean capacity</th>
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5.2.1.4 LOS Mid Room (MR)

The measurement campaign then took place in the middle of the room. Figure 5-14 describes the analysis for MR measurement. From Figure 5-14(a), it can be observed that both polarisations experienced fluctuations at different frequency and significant gaps between their eigenvalues due to significantly more multipath at this part of the room. Upon taking the ratio of the eigenvalues, it is noted in Figure 5-14(b) that the ratio of LP is significantly lower than EP by around 20 dB at 2.3 GHz where the antennas are well matched. The drop in EP ratio is caused by the lower channel coefficient in one of the STAR dipole antenna ports due to multipath, thus making it more inclined towards diversity behaviour. The elliptical XPR shows consistent value at approximately 50 dB throughout the frequency band except for at 2.3 GHz where the XPR is reduced to zero, which results from lack of polarisation orthogonality due to the multipath at this frequency. As a consequence, it can be noted that in order to achieve EP multiplexing, there needs to be a lower amount of multipath or more directional antenna so that multipath vulnerability can be avoided.

In Table 5-4, the instantaneous parameters for MR measurement are analysed at 2.33 GHz and 2.3 GHz for the best and worst cases respectively. It is found that even in the best case for EP, the eigenvalues for EP are still further away from each other when compared to the eigenvalues for LP. Meanwhile, the worst case for EP turns out to be the best case for LP as it can be seen that the mean eigenvalues for LP are comparatively near to each other. This finding is also evident in the mean ratio of eigenvalues whereby the ratio for LP is quite close to zero in contrast to the higher ratio for EP in both best and worst case scenarios. In the XPR evaluation of MR measurements, the XPR for EP and LP are fairly consistent with values of 50 dB and 20 dB respectively, particularly in the best case scenario. However, in the worst case scenario, elliptical XPR suffers major reduction down towards 0dB due to the depolarisation at the centre frequency where no orthogonality is achieved. The mean capacity is rather comparable for both EP and LP in the best case scenario while in the worst case scenario, elliptical capacity is lower than the linear capacity. In summary, there is a deterioration in EP performance due to the semi directional antenna as well as the open space of the environment that leads to a lot of scattering and multipath from many directions. Nonetheless, when there is a shorter range and good antenna directionality, EP has managed to show good multiplexing performance and effectively combated the multipath issue.
Figure 5-14: Frequency Domain MR mean evaluations of (a) eigenvalues, (b) eigenvalue ratio, (c) XPR and (d) capacity averaged over time
Table 5-4: Instantaneous evaluations for MR measurement scenario of eigenvalues, eigenvalue ratio, mean XPR and mean capacity
5.2.1.5 LOS Atrium Walking (AK)

The measurement campaign then moved to another LOS setting which is in the atrium. Several measurements were completed in this bigger open space environment with larger distance to study the orthogonality of EP in a static link scenario. Figure 5-15 illustrates the analysis data for AK measurement campaign. The first and second eigenvalues for EP and LP are comparable to each other as observed in Figure 5-15(a) but EP fluctuates more at the operating frequency. In Figure 5-15(b), generally, there is higher fluctuation in ratio of EP eigenvalues in comparison to previous edge of the room measurement campaign. However, below 2.3 GHz, both eigenvalue ratios remain stable as compared to above 2.3 GHz. The XPR for EP mainly stays at about 60 dB as can be seen in Figure 5-15(c) except for some fading in the region where the STAR dipole antennas are not well matched. The capacity plot in Figure 5-15(d) shows that the capacity for both polarisations is slightly degraded as compared to edge of room measurements. In all the plots across frequency domain, it is also noted that there is a significant degradation in performance for both EP and LP at 2.325 GHz which is possibly due to a reduction in the received power that was caused by a higher significant multipath in the Atrium area. Overall, there is a higher number of fluctuations in this scenario which is quite similar to MR measurements. Due to the larger distance between TX and RX, EP is exposed to higher multipath vulnerability and has affected its multiplexing capability.

In Table 5-5, all the metrics used in the analysis of orthogonality of EP is portrayed. Frequencies 2.29 GHz and 2.32 GHz are chosen to represent the best and worst cases respectively. The instantaneous eigenvalue and eigenvalue ratio plots demonstrate the diversity rich channel whereby one eigenvalue is shown to be stronger than the other. Both EP and LP therefore show more diversity behaviour that is produced due to higher multipath and larger distance. Table 5-5 also presented the mean XPR for EP and LP in both cases. The elliptical XPR maintains at approximately 55 dB while the linear XPR is quite consistent at about 20 dB. Nevertheless, there are some fades at the selected frequency tap for the best case scenario. In the worst case scenario for EP, the XPR plot shows that there is even more fading for EP where the XPR values drop down to zero which result from increased vulnerability of multipath. The last metric of the evaluation shows that the capacity of EP and LP are fairly comparable in both cases.
Figure 5-15: Frequency Domain AK mean evaluations of (a) eigenvalues, (b) eigenvalue ratio, (c) XPR and (d) capacity averaged over time.
Table 5-5: Instantaneous evaluations for AK measurement scenario of eigenvalues, eigenvalue ratio, mean XPR and mean capacity
5.2.1.6 LOS Atrium Waving (AV)

The analysis results for AV measurements are displayed in Figure 5-16 and Table 5-6. These AV measurement results are almost similar when compared to AK measurements, thus suggesting a repeatable and consistent data. Nevertheless, there are more fluctuations observed in the AV measurements, which is caused by more obstructions to LOS when hands were waving in the air. As a result, the higher occurrence of obstructions causes more XPR de-polarisation as seen in Figure 5-16(c). It also affects the ability of the RX antennas to distinguish between two orthogonal ellipses.

In Table 5-6, it is shown that 2.29 GHz and 2.32 GHz are selected to represent the best and worst cases respectively, which are the same frequencies shown in AK measurements. The instantaneous plots reported that there is a diversity behaviour displayed from both EP and LP due to the larger distance and higher multipath in the Atrium area. Thus strengthening the argument that EP possesses higher multiplexing gain when there is shorter TX-RX distance and higher antenna directivity.
Chapter 5. Dual Polarisation Multiplexing Measurement Campaign

Figure 5-16: Frequency Domain AV mean evaluations of (a) eigenvalues, (b) eigenvalue ratio, (c) XPR and (d) capacity averaged over time.
Table 5-6: Instantaneous evaluations for AV measurement scenario of eigenvalues, eigenvalue ratio, mean XPR and mean capacity
5.2.1.7 LOS Atrium Moving 1 (AM1)

The AM1 measurement results are presented in Figure 5-17. It is noted from the mean eigenvalues and mean eigenvalue ratio plots, that both polarisations are vulnerable to severe multipath and thus suffer from heavy fluctuations due to the movement of TX. Nonetheless, it is observed from this measurement that EP is still comparable with LP. Based on Figure 5-17(c), it can be observed that EP possesses higher polarisation purity as the value of elliptical XPR is significantly higher than linear XPR.

Frequencies 2.32 GHz and 2.29 GHz are displayed as the best and worst cases respectively in Table 5-7. In Table 5-7, the mean XPR over time domain is also presented. When there is no fading, the XPR of EP maintains at around 50 dB. The elliptical XPR also suffers greater fluctuations in the worst case scenario in comparison to the best case scenario for EP. It is apparent that the multipath, that is created due to more distance between TX and RX as well as the movement from TX, is too severe in this measurement that EP does not work as well as diversity gain from LP. The instantaneous capacity of EP is seen to be equivalent to the instantaneous capacity of LP in the best case scenario.
Figure 5-17: Frequency Domain AM1 mean evaluations of (a) eigenvalues, (b) eigenvalue ratio, (c) XPR and (d) capacity averaged over time
Table 5-7: Instantaneous evaluations for AM1 measurement scenario of eigenvalues, eigenvalue ratio, mean XPR and mean capacity.
5.2.1.8 LOS Atrium Moving 2 (AM2)

In addition to AM1 measurement, the AM2 measurement is discussed in order to ensure that the measurements are repeatable. The AM2 data is presented in Figure 5-18 and Table 5-8. With higher presence of multipath in the channel due to the higher distance and moving TX, it is observed that there are greater fluctuations across the frequency range of interest in the mean plots. Therefore, the co-located STAR dipole antennas do not perform as well as the spaced dipole antennas with spatial separation.

Further analysis of AM2 measurements in the time domain are presented in Table 5-8. The best case scenario is taken at 2.295 GHz while the worst case scenario is taken at 2.3 GHz. Again, it is evident that the EP antennas created a more diversity rich channel instead of multiplex rich due to severe multipath and the movement from TX.

Upon comparing the performance of the measurements in the seminar room and the atrium, there is a slight degradation in EP performance in the atrium measurements which is caused by the much larger distance between TX and RX. The time delay of the multipath also became longer due to the ceiling height which is four storeys above the ground floor. However, it is found that EP performed similarly to LP and in fact, has a comparable diversity to LP. Thus, it can also be said that the STAR dipole antenna is also a good diversity antenna.
Figure 5-18: Frequency Domain AM2 mean evaluations of (a) eigenvalues, (b) eigenvalue ratio, (c) XPR and (d) capacity averaged over time
Table 5-8: Instantaneous evaluations for AM2 measurement scenario of eigenvalues, eigenvalue ratio, mean XPR and mean capacity
5.2.2 Obstructed Line of sight

5.2.2.1 Door diffraction (DD)

The measurement campaign then moved to a different scenario, i.e. from LOS to obstructed LOS. Using the door of the seminar room, an obstructed LOS scenario is created whereby the TX and RX are not directly facing each other. The presence of the door will also act as a big obstruction situated near TX that will hugely reflect and diffract some of the multipath to the RX. In practice, there will be some instances that obstructed LOS scenario like this may happen in a static link communication. Figure 5-19 shows the analysis results for DD measurement. In this scenario, the gap between the first and second eigenvalues are comparable for EP and LP as seen in Figure 5-19(a). On the other hand, it is found that there is a bigger difference between the eigenvalues in DD measurement in contrast to previous LOS measurements. For instance, in the EV measurement, the gap between the EP eigenvalues was as close as 1 dB in comparison to DD measurement whereby the gap is as huge as 6 dB. In Figure 5-19(b), both the eigenvalues ratios are comparable but nevertheless both polarisations suffer from major fluctuations at every frequency due to a higher degree of multipath. In comparison to the LOS measurements, the ratio of eigenvalues for EP and LP in DD measurement is noted to be as high as around 10 dB. Despite maintaining the XPR value at almost 50 dB, it can be observed in Figure 5-19(c) that the elliptical XPR experienced greater fading as compared to linear XPR due to the major fluctuations in Figure 5-19(b). In Figure 5-19(d), the capacity of LP is observed to be slightly higher than the capacity of EP which resulted from the absence of antenna boresight alignment and therefore less ideal setup. Despite that, both EP and LP deliver similar diversity gain and thus comparable capacity.

Table 5-9 presents the best case at 2.31 GHz and the worst case at 2.29 GHz of EP for DD measurement. It is noted that the average eigenvalues, the average ratio of eigenvalues and the average capacity for both polarisations are comparable in both cases. Due to fading that occur in Figure 5-19(c), some fading can also be observed in the mean XPR plot for elliptical XPR, although the XPR value is consistent at around 50 dB. From Table 5-9, we can also conclude that there is a diversity rich channel based on the result of the mean eigenvalues, and both EP and LP have comparable, if not equivalent, diversity performance.
Figure 5-19: Frequency Domain DD mean evaluations of (a) eigenvalues, (b) eigenvalue ratio, (c) XPR and (d) capacity averaged over time.
### Table 5-9: Instantaneous evaluations for DD measurement scenario of eigenvalues, eigenvalue ratio, mean XPR and mean capacity
5.2.3 Non-line of sight

5.2.3.1 NLOS Round Door (RD)

RD measurement is a totally blocked or non-LOS (NLOS) scenario. It is also important to measure the performance of EP multiplexing in this type of environment to see the difference between LOS and NLOS propagation. In comparison to a multiplexing rich channel in LOS environment, the NLOS environment possesses a diversity rich channel. In Figure 5-20, it is seen that the channel behaves like a frequency selective channel whereby at a different frequency, EP performs better than LP. This is apparent in Figure 5-20(a), (b) and (d) whereby at frequency 2.31 until 2.325 GHz, the eigenvalues of EP, the ratio of eigenvalues of EP and the capacity of EP perform better than LP. In Figure 5-20(c), the XPR value of EP remains at about 50 dB except for when there is some fading, although the fading is not as heavy as in DD measurement. Hence, the co-located EP performs similarly to the spatial LP, if not better, in a NLOS case, and either a spatial or angular diversity system were obtained from the de-correlated channels.

Table 5-10 presents the important metrics in time domain for RD measurement. The best case is taken at 2.32 GHz while the worst case is taken at 2.29 GHz. A diversity behaviour is shown by both polarisations as apparent from the instantaneous eigenvalue and eigenvalue ratio plots. In the mean XPR plot, the elliptical XPR maintains at 50 dB but there is less fluctuation in the best case as compared to the worst case which are caused by the non-existence of LOS signal and purely NLOS propagations. The mean capacity plot shows that the capacity of EP remains greater than the capacity of LP in both cases. Therefore, it can be confirmed that even in a NLOS environment, the performance of EP is comparatively close to the performance of LP, if not better in some cases.
Figure 5-20: Frequency Domain RD mean evaluations of (a) eigenvalues, (b) eigenvalue ratio, (c) XPR and (d) capacity averaged over time
Table 5-10: Instantaneous evaluations for RD measurement scenario of eigenvalues, eigenvalue ratio, mean XPR and mean capacity
Chapter 5. Dual Polarisation Multiplexing Measurement Campaign

5.3 Summary

Chapter 5 presented and discussed the measurement campaign performed to evaluate the multiplexing capability of EP in a static wireless link. The elliptically and linearly polarised antennas that were built in Chapter 4, i.e. the STAR dipole and the spaced dipole, were used in this experiment. The measurement equipment used to capture the channel in this campaign was the Elektrobit Propsound Wideband MIMO Channel Sounder. The campaign was executed in an indoor environment, specifically in an office environment consisting of a seminar room, a short corridor and an atrium with a spiral staircase. LOS, obstructed LOS and NLOS environments were simulated in this setting with the inclusion of people walking and waving as moving scatters while also in other measurements the TX is moved.

An evaluation was carried out to study the effects of exploiting elliptical polarisation in a multiplex rich channel. A comparison analysis with a reference linear polarisation was also performed. Four crucial metrics were evaluated: mean eigenvalues, mean ratio of eigenvalues, mean XPR and mean capacity across a range of frequencies. An in-depth investigation was conducted by choosing instantaneous metrics in the best and worst cases for every measurement scenario. In the LOS scenario, measurements were performed at the edge of seminar room, in the middle of the room and in the atrium. Comparing all three locations, in general, edge of room results showed the best EP multiplexing performance as evident by the least fluctuations and multipath occurrence. The reason is that EP uses its co-located antennas to overcome the multipath effects as compared to the spatially separated LP antennas and thus created multiplex rich orthogonal ellipses. Additionally, it is apparent that short distances with lower multipath as well as directional antennas contributed to good EP multiplexing. When these conditions were not met, EP multiplexing started to fail and the received antennas were not able to distinguish between two strong orthogonal elliptical modes that are multiplex rich as indicated by the faltering XPR.

In the obstructed LOS and NLOS scenarios as well as when the TX was moved, it was observed that both polarisations suffer severe multipath and showed diversity behaviour. It was observed that multiplexing technique does not work in the diversity-rich environment where it was difficult to get perfect orthogonality. However, it was important to show that the STAR dipole antennas can be used as angular diversity antennas alternatively in this type of environment and could achieve high diversity gain. It is also a more compact antenna in comparison to conventional spatial and polarised diversity antennas. The antenna efficiencies are also comparable as shown in Chapter 4; hence, the two antennas experienced same channel path losses.
Chapter 6

6 mmWave Backhaul Links – Rationale for Improving Multiplex Richness

Previously, it was found that elliptical polarisation has the ability to achieve good multiplexing when there is low scattering within the radio environment, highly directional antennas and relatively short distances between transmitter and receiver. These criteria are exactly found in static mmWave links, for example in backhauls and some nomadic links in very small cellular radius such as high throughput WLAN (WiGig) and small cell access applications in next generation mobile. This therefore gives reason to investigate the feasibility of using elliptical polarisation multiplexing at mmWave frequencies.

This chapter will present a discussion of how a phased array antenna design can mitigate the inter-backhaul interference at mmWave frequency. The design can therefore be further optimised such that there will be different phase constellations on the phased array that will allow the main lobe as well as the side lobes to be steered in order to avoid interference between static links. Evaluations of the potential of utilising a hybrid of polarisation multiplexing and phased array beamforming, based on measured wideband channels for a 60GHz backhaul link are presented here.

6.1 Hybrid Beamforming and Polarisation Multiplexing for mmWave Backhaul

A discussion on the high gain capability at mmWave frequencies such as 60 GHz, which leads to extremely high capacity was presented in Chapter 2. It was found that at this frequency, a communication system is able to utilise a wide of bandwidth and thus allow multi Gb/s throughput. The antenna beam at 60 GHz is highly directional and thus translates to higher gain but comes with
a narrow beamwidth, which substantially reduces any multipath. In order to optimise these characteristics, a phased array antenna design is used to further steer the beam to obtain best gain in the direction-of-interest, while suppressing any interference from other directions [156]–[158].

Phased array antennas with a large number of elements will have the increased capability to null any interference coming from other directions while steering the beam heavily in the desired direction and forming a high signal to interference and noise ratio (SINR), which will be achieved by forming the desired phase constellations in the array in order to change and produce several discrete array modes with different polarisations. An example of a phase shifter design that can be integrated in a phased array antenna was presented in [159] and this phase shifter has a switching capability of $22^\circ$. Deploying two of these phase shifters, a total of $11^\circ$ of phase shift can be obtained. With three phase shifters, a phase shift of $7^\circ$ can be reached. This is therefore moving near to full analogue beamforming. One solution to achieve full analogue beamforming is to use mechanical waveguide phase shifters but they are not common at 60 GHz due to the small physical size.

Many antennas employed at the mmWave frequencies for backhaul applications produce side lobes. Even though the main beam of the antennas is steerable in order to avoid substantial interference to other links, the side lobes are harder to control and can still cause interference on a lower scale to other channels. An example scenario of short distance backhauls for small cells on an urban street is portrayed in Figure 6-1. Due to high path loss at 60 GHz, the signals are unable to travel very far and thus a single street requires several backhauls. In Figure 6-1, the first backhaul is on one side of the street and is represented by the blue 16-element antenna arrays at the TX and RX while the second backhaul is across the street with the red antenna arrays. Assuming that at 60 GHz, the channel follows a two-path model as will be verified later in this chapter, there will be one direct LOS path and one reflection multipath that bounces off from the wall of a nearby building. The reflection that comes off the ground is insignificant due to the angle at which it is transmitted and arrives at the receiver, meaning that with low antenna gain at these angles and reflection loss it is significantly weakened.

Since the phased array will generate side lobes in addition to the main beam, these side lobes will also radiate weaker signals and these might be picked up by other backhauls such as those illustrated in Figure 6-1. As seen in the diagram, based on the position of the backhauls relative to each other, these side lobes can align with each other and start to interfere with each other. Therefore, the side
lobes of phased arrays have also the potential to cause performance degradation in the backhaul. The best solution to then combat the interference caused by the side lobes is to steer not only the main beam but the side lobes as well, as depicted in Figure 6-2. The phasing of the arrays can be setup such that the blue side lobe and the red one are rotated in order to cleverly null the interference. In the future, there needs to be a clever dynamic linking between the mmWave backhauls such that they can be self-configuring to ensure all the surrounding backhauls in the street and not interfere with each other.

Figure 6-1: Example scenario of short distance backhauls using phased array beamforming for small cells on an urban street where interference caused by side lobes can occur

Figure 6-2: Phased array beamforming with side lobe steering as a solution to combat the interference caused by the side lobes
From another perspective, the phased array can be also designed as a hybrid system to combine beamforming gain together with multiplexing gain. As proposed throughout the thesis, the dual polarisation concept can be exploited to increase the multiplexing capability of an antenna design in a line of sight link. In the future, as phased array antennas develop their capability for 60 GHz backhaul applications, dual polarised antenna arrays can be utilised to form a hybrid elliptically polarised multiplexed and beamforming solution. Figure 6-3 proposes one of the approaches to produce such a solution. By selecting every other antenna element, half of the antenna elements can be used to generate one elliptical mode and the other half is used to create another elliptical mode which are orthogonal to each other. Therefore, by constantly changing the polarisations each way, with a directional antenna, this is one of the possible constellations in order to create hybrid multiplex and beamforming antenna.

There could be many other constellations within this beamforming link while simultaneously forming elliptical orthogonal links. For instance, eight adjacent elements in a horizontal or vertical position can be selected to radiate one EP while the rest of the 8-elements radiate another orthogonal EP. Increasing the number of elements on a phased array can form an increased number of different constellations. As a result, the antenna beams can still be formed and steered while allowing some cross-polarisation rejection for a hybrid solution. As has also been established from results presented in chapter 5, multipath does cause degradation to the polarisation multiplexing and this needs to be ascertained through measurement in the next section to analyse its impact from which synthetic analysis can be formed.

![Figure 6-3: Proposed hybrid dual elliptically polarised multiplexing and beamforming antenna array solution to further increase the throughput in a LOS link](image)
6.2 Scaled Propagation Measurement at 60 GHz

A real environment measurement to measure the channel coefficients in a backhaul scenario at 60 GHz would have required measuring in a street over a distance of 40m, which is impractical as well as limited by the dynamic range of the measurement equipment. For this reason, the propagation measurement was scaled down to 6.6 m in an indoor environment to form a replica scenario. The scaled down measurement in particular will be able to ascertain the scattering characteristics off rough surfaces, which have significant impact on the magnitude and phase of the reflected paths.

6.2.1 Measurement Setup

The measurement setup, shown in Figure 6-4 and 6-5 took place in an indoor environment and is a scaled down version of a common backhaul outdoor environment with typically 40m TX-RX distance between two lamp posts. The Fresnel zone and all other parameters are scaled down in line with the 6.6m TX to RX distance, which were set up in an open indoor environment. The scaled down ratio is therefore roughly 6 times smaller than a real scenario. The Fresnel zone radius can be calculated as follows [6]:

\[
    r_{\text{Fres}} = \sqrt{\frac{n\lambda \text{wavelength} d_{\text{obst-TX}} d_{\text{obst-RX}}}{(d_{\text{obst-TX}} + d_{\text{obst-RX}})}}, \tag{6-1}
\]

where \( n \) represents the \( n^{\text{th}} \) Fresnel zone, \( d_{\text{obst-TX}} \) is the distance of the obstructed signal from the TX and \( d_{\text{obst-RX}} \) is the distance of the obstructed signal from the RX. At TX-RX distance, \( d_{\text{TX-RX}} = 6.6 \text{ m} \), where \( d_{\text{obst-TX}} \) is equal to \( d_{\text{obst-RX}} \), the maximum first Fresnel zone radius, \( r_{\text{Fres}} \) is then calculated to be 0.0908 m at 60GHz from direct LOS path between TX and RX. A Vector Network Analyser (VNA) was used to measure the static wideband channel from 45 GHz to 67 GHz. However, the useful measurements were limited by the waveguide cut off of the horn antenna and also the bandwidth of the amplifiers used to compensate against cable loss. Therefore, the range in which useful information could be obtained was from 56 GHz to 66 GHz. 60 GHz horn antennas with a 10° beamwidth and 25 dBi of gain were utilised as the transmitting and receiving antennas. The azimuth and elevation pattern of the horn antenna is shown in Figure 6-4.
In a real application, the TX and RX antennas are normally placed on a street lamp post, which are typically positioned 2m, 3m or 4m away from buildings on the same side of the street. The measurement setup is shown in Figure 6-5 and 6-6. In this measurement, the TX and RX were placed at proportionate positions, \( d_{\text{wall1}} = 0.318 \) m and \( d_{\text{wall2}} = 0.61 \) m from the wall (thus representing distances of approximately 2m and 3m when scaled up to the real street scenario) to determine the effects of the wall distance from the TX-RX link. The wall of a street building will typically contain a certain degree of roughness at mmWave frequencies and cause diffuse instead of specular reflections. Therefore, a wall in the test setup here as shown in Figure 6-6 (a) was used, where the door is at the mid-point with a suitably rough surface comparable with a street building. When the antennas were positioned at \( d_{\text{wall1}} \) and \( d_{\text{wall2}} \) away from the wall, the wall is therefore outside of the first Fresnel zone of the antennas, creating a clear LOS propagation path between the TX and RX.

A measurement was firstly taken to obtain a direct LOS from the TX to the RX without any obstruction. Subsequently, a vertical pole was placed in the middle of the LOS propagation path, at \( d_{\text{obst}} = \frac{d_{\text{RX-RX}}}{2} \) to get an obstructed LOS measurement. A metal sheet 210mm wide and 300m high was also attached to the pole, to resemble a road sign or floral decoration that is put on a lamp post. This setup reflects the scenario where a backhaul link is typically positioned between two lamp posts 40m apart with another lamp post in the middle directly within the line of sight, which may impact the link. Therefore, the link has to rely on some partial diffraction around the lamp post or any objects attached to it.

![Figure 6-4: Radiation pattern of 60 GHz horn antenna used in the mmWave backhaul measurement](image-url)
Chapter 6. mmWave Backhaul Links – Rationale for Improving Multiplex Richness

Figure 6-5: Indoor measurement setup (a scaled down version of a common backhaul outdoor environment, TX-RX distance from 40m to 6.6m) for 60 GHz backhaul performed in the Atrium

(a) View from TX end

(b) View from RX end

Figure 6-6: mmWave backhaul measurement setup in the Atrium photos consisting of TX, RX and vertical pole as an obstacle in the middle
6.2.2 Results

The measured channel coefficients of the backhaul measurement are presented in Figure 6-7. Note that the impulse responses for the LOS at 0.61m are offset by -0.5ns for purposes of clarity. It is observed that the LOS channel coefficients remain the same even when the distance is varied. As expected, the channel coefficients dropped when there is an obstruction to the LOS path. The delay domain channel coefficients were then calculated from a frequency range at 55 GHz to 66 GHz and the normalised magnitudes are plotted in Figure 6-8.

Figure 6-7: Measured channel coefficients, $h$ in the frequency domain from scaled 60GHz backhaul measurement in LOS and obstructed LOS links at different distances from the wall (0.318m and 0.61m)

Figure 6-8: Derived delay domain wideband channel responses from scaled backhaul measurement in LOS and obstructed LOS links at different distances from the wall (0.318m and 0.61m)
Two maximum delay taps are then extracted from Figure 6-8 and tabulated in Table 6-1. The magnitudes of the delay taps were normalised to the maximum value. The power delay profile for distance 0.61 m was purposely offset at 0.5 ns from distance 0.318 m since the LOS path will be overlapping with each other. At a distance of 0.318 m from the wall, the maximum delay taps are separated by 0.1 ns. At a distance of 0.61 m from the wall, the maximum delay taps are theoretically separated by 0.38 ns but due to the surface roughness of the wall, a slightly shorter delay was recorded, at about 0.2 ns.

Rough surfaces are found to have significant impact on the magnitude and phase of reflected paths and is not a simple specular reflection. It is also noteworthy that the reflected path off the wall is also attenuated when the obstruction is in place. This is owing to reflections off the post with rough surfaces, which also cause further multipath components that cannot be identified by the resolution of the delay bins. Furthermore, this finding verifies that the channel follows a 2-path delay tap model at 60 GHz as illustrated in Figure 6-9.

<table>
<thead>
<tr>
<th>Distance (m)</th>
<th>Delay Taps</th>
<th>Measured Delay (ns)</th>
<th>Normalised Delay (ns)</th>
<th>Normalised Magnitude (dB)</th>
<th>Normalised Phase (°)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.318</td>
<td>LOS path</td>
<td>20.7</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td></td>
<td>Reflected path</td>
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<td>0.1</td>
<td>-8.33</td>
<td>12.87</td>
</tr>
<tr>
<td>0.610</td>
<td>LOS path</td>
<td>21.2</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td></td>
<td>Reflected path</td>
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<td>0.2</td>
<td>-6.2</td>
<td>-169.02</td>
</tr>
<tr>
<td>0.318</td>
<td>LOS obstructed path</td>
<td>20.7</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td></td>
<td>Reflected obstructed path</td>
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<td>0.1</td>
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</tr>
<tr>
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<td>0</td>
<td>0.07</td>
<td>0</td>
</tr>
<tr>
<td></td>
<td>Reflected obstructed path</td>
<td>21.4</td>
<td>0.2</td>
<td>-7.03</td>
<td>192.27</td>
</tr>
</tbody>
</table>

Table 6-1: Magnitude and phase of the extracted maximum delay taps for scaled propagation measurement
6.3 Measurement of Dual Linear Polarisation Multiplexing Capability

Another measurement was performed in order to investigate the dual linear polarisation multiplexing capability at 60 GHz. The same measurement setup used in the previous section was used here though the antennas were only positioned at 0.318m from the wall.

6.3.1 Measurement Setup

The measurement setup shown in Figure 6-10 is the same setup as in section 6.2.1. The antennas were placed at $d_{\text{wall}} = 0.318$ m from the wall. Rectangular horn antennas were used as the TX and RX antenna. In order to get vertical polarisation, the horn antennas were aligned vertically. The antennas were then aligned horizontally in the same position to obtain horizontal polarised channel. Finally, the cross polarisations vertical to horizontal (V-H) and horizontal to vertical (H-V) were obtained. A Vector Network Analyser (VNA) was used to measure the static wideband channel from 45 GHz to 67 GHz. However, due to the waveguide cut off and amplifier’s dynamic range to measure both co- and cross polarisations, the useful range of frequency was limited to 50 GHz to 55 GHz.
6.3.2 Results

The channel coefficients for co-polarised components, VV and HH, as well as cross-polarised components, VH and HV, are plotted in Figure 6-11. The noise floor level in the measurement was also recorded and it is shown that most of the polarised components are above the noise floor. It is observed that the cross-polar components are 20 dB below the co-polar components across the frequency range which shows that the signals received at the RX has an acceptable level of polarisation purity.

Additionally, it can be seen that both co-polar components, VV and HH have somewhat equivalent magnitudes, particularly from 52 GHz to 53 GHz. The same is also true for the cross-polar components. It is also observed in Figure 6-12 and 6-13 that the eigenvalues show the closest magnitudes and thus the lowest eigenvalue ratio at these frequencies. Figure 6-14 shows the linear XPR which has an average value of 25 dB. These findings show that the polarisations transmitted have the expected level of orthogonality for linear polarisation and thus have the potential to be exploited in order to achieve higher multiplexing gain. However, EP can also be exploited in this scenario and this is analysed in the next section.

Figure 6-11: Measured co-polar and cross-polar channel coefficients from Dual LP Multiplexing mmWave backhaul measurement
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Figure 6-12: Computed first and second eigenvalues of dual LP channel, analysed from the Dual LP Multiplexing mmWave backhaul measurement

Figure 6-13: Computed eigenvalue ratio of dual LP channel, analysed from the Dual LP Multiplexing mmWave backhaul measurement

Figure 6-14: Computed XPR of dual LP channel, analysed from the Dual LP Multiplexing mmWave backhaul measurement
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Figure 6-15: Derived delay domain wideband channel responses from the Dual LP Multiplexing mmWave backhaul measurement

Figure 6-15 depicts the derived delay domain with wideband channel coefficients from the dual linear polarisation measurement. From this figure, the maximum delay taps were then extracted which are recorded in Table 6-2. In order to account for phase error between VV and HH due to cables, the phases for the first taps of VV and HH were maintained the same while the phases for the second taps were corrected accordingly. The delay taps are separated by 0.1 ns.

<table>
<thead>
<tr>
<th>Polarisation</th>
<th>Delay Taps</th>
<th>Measured Delay (ns)</th>
<th>Normalised Magnitude (dB)</th>
<th>Normalised Phase (°)</th>
</tr>
</thead>
<tbody>
<tr>
<td>VV</td>
<td>LOS path</td>
<td>21.2</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td></td>
<td>Reflected path</td>
<td>21.3</td>
<td>-6.351351dB</td>
<td>134.4</td>
</tr>
<tr>
<td>HH</td>
<td>LOS path</td>
<td>21.2</td>
<td>0</td>
<td>-108.83</td>
</tr>
<tr>
<td></td>
<td>Reflected path</td>
<td>21.3</td>
<td>-0.591792dB</td>
<td>176.74</td>
</tr>
</tbody>
</table>

Table 6-2: Normalised magnitude and phase of the maximum delay taps for dual polarisation measurement

6.4 Synthesis of the elliptical polarisation at mmWave

The 2-path delay model from the measurement results derived in section 6.3 and Table 6-2 were used here to analyse the capabilities of elliptical polarisation at mmWave frequency. The 2-path delay model was simulated and displayed in Figure 6-16. The magnitudes and phases of the LOS and reflected paths of VV were used to create the first and second delay taps of VP while the
magnitudes and phases of the LOS and reflected paths of HH were used to create the first and second delay taps of HP. Using the model in Figure 6-16, a set of synthetic elliptically polarised channels with scattering were simulated, as presented in Figure 6-17. The delay taps in Figure 6-16 were used as the impulse response of the signals and were then Fourier Transformed to convert them into signals in frequency domain, creating newly synthesised VV and HH channel coefficients. The magnitude of the cross polarisation was set to -6dB to generate newly synthesised HV and VH channel coefficients. In order to create elliptically polarised waves, appropriate phase shifts were introduced at TX and RX. The resultant elliptical modes are illustrated in Figure 6-18. The vertical and horizontal components are assumed to be very near to each other, if not co-located, giving out the same reflection from the wall.

An eigenvalue and XPR analysis was then computed using this set of synthetic narrowband data following the calculations in Equation (2-6), (3-36) and (3-38), as plotted in Figure 6-19, 6-20 and 6-21 respectively. It is observed that the eigenvalues are more than 2dB apart and also the XPR is found to fail at a number of frequencies. This is owing to the rough surface reflections, which are prohibitive to the measurement. Therefore, to overcome the issues caused by surface roughness, a dedicated absorber can be placed at the central specular reflection point, which will cover over any diffuse scattering and form a single path line of sight. This is simulated in Figure 6-22 to 6-26, whereby the second delay taps for VV and HH signals were attenuated by 49 dB that closely resembles a direct LOS case. It can be seen that the channel coefficients have similar values, the eigenvalues got closer to each other without any frequency selectivity and the XPR has less fluctuations.

![Figure 6-16: 2-path delay model derived from section 6.3](image1)

![Figure 6-17: Synthesised VV and HH channel coefficients from the 2-path delay model derived from section 6.3](image2)
Chapter 6. mmWave Backhaul Links – Rationale for Improving Multiplex Richness

Figure 6-18: Illustration of how dual orthogonal elliptical modes synthesised from dual LP Multiplexing mmWave backhaul measurement

Figure 6-19: Synthesised eigenvalues of EP using the 2-path delay model derived from section 6.3

Figure 6-20: Synthesised eigenvalue ratio of EP using the 2-path delay model derived from section 6.3

Figure 6-21: Synthesised XPR of EP using the 2-path delay model derived from section 6.3
Figure 6-22: 2-path delay model after 49 dB attenuation of second delay tap

Figure 6-23: Synthesised VV and HH channel coefficients after 49 dB attenuation of second delay tap

Figure 6-24: Eigenvalues of EP using synthetic data at mmWave after 49 dB attenuation of second delay tap

Figure 6-25: Eigenvalue ratio of EP using synthetic data at mmWave after 49 dB attenuation of second delay tap

Figure 6-26: XPR of EP using synthetic data at mmWave after 49 dB attenuation of second delay tap
6.5 Summary

This chapter discussed the potential of doing a mmWave backhaul using elliptical polarisation multiplexing. Previous chapters found that elliptical polarisation multiplexing is able to achieve good performance when there is low scattering within the radio environment, highly directional antennas and relatively short distances between transmitter and receiver. These criteria can be found in static mmWave links. Therefore, there is a high potential that EP can also work at mmWave frequency.

It was proposed that a hybrid beamforming phased array antenna with multiplexing technique is useful in mitigating inter-backhaul interference. This hybrid antenna can easily null any interference coming from another direction while steering the beam in wanted direction. A further work was also suggested that by changing the phase constellations of phased array, several discrete array modes with different polarisations can be produced. This new design of phased array may also have the capability of steering not only the main lobe but also the side lobes in order to combat the interference from other backhauls.

Evaluations of mmWave backhaul were conducted using real environment measurements and then was followed by a synthesis evaluation. It was found that the measured linear XPR at mmWave is around 25 dB and the measured eigenvalue ratio has an average of 1.2 dB. A 2-path delay model was then derived from the real measurements. The maximum delay taps were extracted from this 2-path delay model and a synthesis of elliptical polarisation at mmWave was performed using the extracted delay taps. It was shown that the eigenvalues were more than 4dB apart and the XPR was failing within the band, which was caused by rough surface reflections of the wall. Therefore, surface roughness has significant impact on the reflected delay tap at this frequency band and this scenario is not a simple specular reflection case. Such issues could be overcome by forming an absorber to cover the diffuse scattering points and reduce a two path propagation scenario into one, which will therefore not inhibit the ability to form elliptical polarisation multiplexing.
Chapter 7

7 Conclusion and Further Work

7.1 Summary of Insights and Conclusion

A suitable antenna candidate to carry out the analysis of multiplexing using dual elliptical polarisation for a 2 x 2 MIMO system was designed, simulated and fabricated. This antenna candidate is based on a crossed dipole design and is known as STAR dipole antenna. It is a known issue that crossed dipole design has high cross-polarisation, therefore, instead of attempting to suppress the cross polarisation, the aim was to exploit this issue. The polarisation produced was highly elliptical and orthogonal when cross polarisation was deliberately created by re-forming the linear wire antennas which formed like a star and hence the name STAR dipole. The STAR dipole was well matched at 2.3 GHz and had a 10 dB bandwidth of 380 MHz. The STAR dipole had a gain of 7.81 dBi and an efficiency of 99.95%. The STAR dipole was compared against spaced dipoles with dual linear polarisation to act as a reference antenna. In order to conduct a fair like-to-like comparison, both antennas radiation patterns were measured and it was justified that the efficiency of the antennas were maintained even though the antennas were compacted.

A propagation channels with full orthogonality could be achieved by using two elliptical polarisations which had an orthogonal tilt angle and opposite rotations (i.e. left-hand and right-hand elliptical polarisations). This minimised the interference between the two polarisations and thus enhanced the multiplexing gain. In order to evaluate the orthogonality, cross polarisation power ratio (XPR) was derived as a metric to show the multiplexing capabilities of the elliptically polarised MIMO system. Using XPR, a measure of polarisation purity of a certain polarised wave can be determined, such that at the receiver, the antenna will be able to distinguish between two orthogonal polarisations. Using theoretical analysis and simulations with synthetic data, it was found that the criteria for producing high value of XPR were co-polarisation components which have the same magnitude (< 0.5 dB different), while cross polarisation components must also have the same magnitude (< 0.5 dB different) and the phase difference between the co-polarisation components,
when comparing the right hand and left hand ellipses, is $180^\circ$. Four different phase shifts at transmitter and receiver was then derived by minimising the denominator of the XPR equations. This was proven from the anechoic chamber measurements where the XPR of elliptically polarised antennas were about 30 dB higher than the linearly polarised antenna. Additionally, results from a measurement campaign in an indoor radio environment showed that in LOS, the co-located STAR dipole has managed to combat channel fluctuations due to multipath in the channel. The spaced dipoles, on the other hand, were susceptible to channel fluctuations due to their spatial separation. Higher elliptical XPR observed in the measurements demonstrates that the orthogonal EP channels were not affected by the multipath and they possess larger polarisation purity. Subsequently, there was also improvement in the capacity of EP whenever there is high XPR, whereby the measured capacity can be as close as the capacity of a $2 \times 2$ channel with perfect orthogonality.

In order to investigate the number of multiple streams that can be multiplexed from a MIMO system, a singular value decomposition (SVD) of the channel matrix, $H$ was carried out. SVD is important as it is able to extract the equivalent independent sub channels of the MIMO system. SVD works by diagonalising the channel $H$ and subsequently finds its singular values. In this research, a 2x2 MIMO channel was studied and therefore, two multiplex streams were extracted. As a result, two eigenvalues were produced and for higher clarity of the SVD result, a ratio of two eigenvalues was calculated to show the closeness of the eigenvalues. As previously explained in Chapter 3, the eigenvalue ratio dropped to 0 dB when the phase conditions for perfect orthogonality are achieved. In the anechoic chamber measurement results, the eigenvalue ratio was shown to be zero at the points where these conditions were precisely met. At these frequencies, this finding coincides with the results of the measured channel coefficients that have similar magnitude and therefore, multiplex-rich orthogonal ellipses were produced. Similar observations can be seen in the measurement campaign. In line of sight (LOS), the elliptical eigenvalue ratio was nearer to 0 dB and there was strong orthogonality when there were low scatterings, short TX-RX distance and highly directional antennas. Significant degradation in the eigenvalue results were observed when there is severe multipath and no LOS propagation. As a consequence, there was a slight drop in capacity and thus reduction in the multiplexing gain. In non-line of sight (NLOS) scenarios, the STAR dipole antennas can be used as angular diversity antennas with comparable diversity gain to conventional spatial and polarisation diversity antennas which are not as compact.

As a conclusion, the research carried out in this thesis will set a new path for antenna design, which has found a new starting point on degrees of freedom for MIMO multi antenna design by exploiting elliptical polarisation for static wireless links. In addition, the research has the potential to be
extended into performing backhaul at mmWave frequencies that have extremely high bandwidth. Further development of new novel antennas for this purpose, which have a hybrid of beamforming and polarisation multiplexing, will need to be formed.

The key conclusions from the analysis of this research work are:

- Two elliptical polarisations, with orthogonal tilt angles and opposite rotations (i.e. left-hand and right-hand elliptical polarisations), can be exploited to achieve full channel orthogonality and minimise the interference between adjacent MIMO sub channels.

- A STAR dipole was designed, simulated and fabricated as an example antenna candidate to carry out the analysis of multiplexing using dual elliptical polarisation for a 2 x 2 MIMO system. It was based on a crossed dipole design. The aim of utilising a crossed dipole design is to exploit the high cross polarisation issue. By re-forming the linear wire antennas with only three dipole arms, a high cross polarisation was deliberately created, which could be more readily exploited. A dual linear polarised spaced dipole antenna was used as a reference antenna. The STAR dipole had similar antenna characteristics to the spaced dipole, therefore, it was justified that the efficiency of the antennas was maintained even though the antennas were compacted.

- A cross polarisation power ratio (XPR) metric was derived to show the multiplexing capabilities of the elliptically polarised MIMO system and help to evaluate the orthogonality of the MIMO channels. A measure of polarisation purity of a certain polarised wave can then be determined, such that at the receiver, the antenna will be able to distinguish between two orthogonal polarisations.

- Using theoretical analysis and simulations with synthetic data, it was found that the criteria for producing high value of XPR are:
  i. co-polarisation components which have the same magnitude (< 0.5 dB different)
  ii. cross polarisation components must also have the same magnitude (< 0.5 dB different)
  iii. phase difference between the co-polarisation components, when comparing the right hand and left hand ellipses, is 180°.

- Real environment measurement campaign showed that in LOS, the co-located STAR dipole has managed to combat channel fluctuations due to multipath in the channel whereas the spaced dipole, with spatial separation, was susceptible to channel fluctuations. Additionally, it was difficult for the spaced dipole to achieve two perfectly orthogonal polarisations, unlike the elliptically polarised STAR dipole. The capacity of EP improved
whenever there is high XPR and can be as close as the capacity of a 2 x 2 channel with perfect orthogonality.

- In LOS, there was strong orthogonality when there were low scatterings, short TX-RX distance and highly directional antennas. The multiplexing gain and consequently, the capacity of the MIMO system was increased when these conditions were met in the real environment. Elliptical polarisation thus has the potential to be exploited at mmWave frequencies where these conditions can be easily found. In NLOS, the STAR dipole antennas can be used as angular diversity antennas with comparable diversity gain to conventional spatial and polarisation diversity antennas which are not as compact.

### 7.2 Further Work

- An elliptically polarised antenna design was presented in this thesis, which was suitable for demonstration purposes in showing the ability of elliptical polarisation to form polarisation multiplexing. Alternatively, other elliptically polarised antenna designs could be formed to improve multiplexing gain, for example patch antenna design [160]. Another recommendation is to print the STAR dipole antenna design onto a substrate and make it a planar structure. The printed STAR dipole would then be more accurate and easier to be manufactured.

- The STAR dipole antenna design discussed in the thesis achieved high isolation through use of a hybrid coupler. Other decoupling techniques could be considered that form a more suited and compacted method of achieving such isolations.

- Elliptically polarised antennas can also be made reconfigurable whereby they are reconfigured to generate an ellipse in a desired direction and avoid adjacent channels. The antenna can also be designed to work at several desired frequencies.

- It was shown in Chapter 6 that EP has a reasonably high multiplexing capability at mmWave frequency. However, the investigations were conducted using synthetic data simulations and linearly polarised horn antennas in order to derive suitable elliptically polarised waves. The challenge is then to build an elliptically polarised antenna with a hybrid beamforming solution to evaluate the capability of elliptical polarisations in a real backhaul environment.

- Upon building elliptically polarised antenna with a hybrid beamforming solution that can work at mmWave, different topology possible can then be investigated. These different topologies will be able to radiate different orthogonal polarisations in order to combat inter-backhaul interference.
Appendix A

In free space (i.e. time invariant), the complex response between a transmit element \( m \) and a receive element \( n \) is 
\[
e^{-jkr_{nm}}/r_{nm},
\]
where \( k \) is the wavenumber corresponding to the carrier frequency and \( r_{nm} \) is the distance between the two elements. It is assumed that there is no mutual coupling between the elements and that the relative differences in the path losses are negligible. Therefore, the normalized free-space response matrix of an \( M \times N \) MIMO system can be written as
\[
H_{\text{LOS}} = \begin{bmatrix}
e^{-jk_{11}} & \ldots & e^{-jk_{1N}} \\
\vdots & \ddots & \vdots \\
e^{-jk_{N1}} & \ldots & e^{-jk_{NN}}
\end{bmatrix}
\] (A-1)

Assuming a transmitted signal vector composed of \( M \) statistically independent equal power components with Gaussian distribution and that the RX has no prior knowledge of the channel, the MIMO channel capacity is given by
\[
C = \log_2 \left( \det \left( I_N + \frac{\text{SNR}}{N} H H^H \right) \right)
\] (A-2)

where \( I_N \) is the \( M \times N \) identity matrix and \( H^H \) denotes the conjugate transpose.

From (2), it can be seen that the capacity is maximised when \( H H^H = NI_N \), whereby in this case, all eigenvalues in \( H H^H \) are equal and the determinant takes its maximum value. Therefore, all MIMO sub-channels are orthogonal. The MIMO system can then be seen as \( N \) parallel SISO links.

A 2x2 MIMO example is given as follows:
\[
H H^H = \begin{bmatrix}
e^{-jk(r_{11}-r_{21})} & e^{-jk(r_{12}-r_{22})} \\
2 & 2
\end{bmatrix}
\] (A-3)

Where \( k \) is a wavenumber to the corresponding carrier frequency and \( r_{11}, r_{12}, r_{22} \) and \( r_{21} \) are the distances between TX 1 and RX 1, TX 1 and RX 2, TX 2 and RX 2 and TX 2 and RX 1. Using \( H H^H = 2I_2 \), where \( I_2 \) is a 2x2 identity matrix, Equation (A-3) is simplified to:
\[
H H^H = 2 \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}
\] (A-4)
\[ \therefore e^{jk(r_{11} - r_{21})} + e^{jk(r_{12} - r_{22})} = 0 \]

and

\[ e^{jk(r_{21} - r_{11})} + e^{jk(r_{22} - r_{12})} = 0 \]

Equation (A-4) is true when the phases of the two arguments in each phasor term are 180° out of phase. This will come up with the case where:

\[ k(r_{11} - r_{21}) = k(r_{12} - r_{22}) + \pi \quad (A-5) \]

And

\[ k(r_{21} - r_{11}) = k(r_{22} - r_{12}) + \pi \quad (A-6) \]

Hence after taking both Equation (A-5) and (A-6) as absolute and equating them:

\[ k(r_{11} - r_{21} - r_{22} + r_{12}) = (2p + 1)a \pi \quad (A-7) \]

where \( p \in 0, 1, 2, \ldots \). Thus it can be seen that for odd multiples of \( \pi \), this translates to the next equation in terms of every odd multiples of half wavelength:

\[ |r_{11} - r_{12} + r_{22} - r_{21}| = (2p + 1) \frac{\lambda_{\text{wavelength}}}{2} \quad (A-8) \]

where \( \lambda_{\text{wavelength}} \) is the wavelength.
Appendix B

Figure B-1: The connection of hybrid coupler when attached to the STAR dipole. The black dotted line represents the cable connection from coupler to STAR dipole. The blue and red lines represent the reflections that might have occur at the input ports of hybrid coupler.

Figure B-2: Measured S-parameter of Coupler 1

Figure B-3: Measured S-parameter of Coupler 2
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