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Design of four-ports MIMO Antenna for Ultra-Wideband Applications

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Abstract

Ultra-Wideband (UWB) technology combined with Multiple-Input Multiple-Output (MIMO) systems has become one of the most promising solutions for next-generation wireless communications. This work presents the design and simulation of a compact four-port UWB MIMO antenna. The proposed antenna aims to achieve continuous impedance matching over the entire FCC UWB band (3.1–10.6 GHz) while maintaining excellent MIMO diversity performance without employing additional decoupling structures.

Simulation results demonstrate wideband operation, high isolation, low correlation, and satisfactory radiation characteristics across the entire UWB spectrum.

Keywords: UWB Antenna, MIMO Antenna, Ultra-Wideband, Isolation, HFSS

Résumé

La technologie Ultra Large Bande (UWB), combinée aux systèmes à entrées et sorties multiples (MIMO), est devenue l'une des solutions les plus prometteuses pour les communications sans fil de nouvelle génération. Ce mémoire présente la conception et la simulation d'une antenne MIMO UWB compacte à quatre ports. L'antenne proposée vise à assurer une adaptation d'impédance continue sur l'ensemble de la bande UWB attribuée par la FCC (3,1–10,6 GHz), tout en maintenant d'excellentes performances de diversité MIMO sans recourir à des structures de découplage supplémentaires.

Les résultats de simulation démontrent une large bande passante de fonctionnement, une forte isolation entre les éléments rayonnants, une faible corrélation ainsi que des caractéristiques de rayonnement satisfaisantes sur l'ensemble du spectre UWB.

Mots-clés : Antenne UWB, Antenne MIMO, Bande Ultra Large, Isolation, HFSS.

المخلص

أصبحت تقنية النطاق العريض جداً (UWB) المدمجة مع أنظمة الهوائيات متعددة المداخل والمخارج (MIMO) واحدة من أكثر الحلول الواعدة للاتصالات اللاسلكية من الجيل القادم. يقدم هذا العمل تصميم ومحاكاة هوائي-UWB MIMO مدمج مكون من أربعة منافذ. يهدف الهوائي المقترح إلى تحقيق مطابقة معاوقة مستمرة عبر كامل نطاق UWB المعتمد من قبل هيئة الاتصالات الفيدرالية الأمريكية (FCC) من 3.1 إلى 10.6 غيغاهرتز، مع الحفاظ على أداء ممتاز لتقنية MIMO دون استخدام هياكل إضافية للعزل أو فك الاقتران.

أظهرت نتائج المحاكاة نطاق تشغيل واسع، وعزلاً مرتفعاً بين العناصر المشعة، وارتباطاً منخفضاً، بالإضافة إلى خصائص إشعاعية مرضية عبر كامل نطاق UWB.

الكلمات المفتاحية: هوائي UWB ، هوائي MIMO ، النطاق العريض جداً، العزل، HFSS.

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List of Abbreviations

3GPP	— 3rd Generation Partnership Project
5G	— Fifth Generation (mobile networks)
AWGN	— Additive White Gaussian Noise
CCL	— Channel Capacity Loss
CPW	— Coplanar Waveguide
CSRR	— Complementary Split Ring Resonator
DG	— Diversity Gain
DGS	— Defected Ground Structure
EBG	— Electromagnetic Band Gap
ECC	— Electronic Communications Committee (European)
ECC	— Envelope Correlation Coefficient
EIRP	— Equivalent Isotropic Radiated Power
FBW	— Fractional Bandwidth
FDTD	— Finite-Difference Time-Domain
FCC	— Federal Communications Commission
FEM	— Finite Element Method
GPS	— Global Positioning System
HFSS	— High Frequency Structure Simulator
HPBW	— Half-Power Beamwidth
IEEE	— Institute of Electrical and Electronics Engineers
IoT	— Internet of Things
ISM	— Industrial, Scientific, and Medical (radio band)
ITU	— International Telecommunication Union
LTE	— Long Term Evolution
MEG	— Mean Effective Gain
MIMO	— Multiple-Input Multiple-Output
MISO	— Multiple-Input Single-Output
MMIC	— Monolithic Microwave Integrated Circuit
PCB	— Printed Circuit Board
PMA	— Planar Monopole Antenna
PSD	— Power Spectral Density

List of Abbreviations

RLC — Resistor-Inductor-Capacitor
SIMO — Single-Input Multiple-Output
SISO — Single-Input Single-Output
SMA — SubMiniature Version A (connector)
SNR — Signal-to-Noise Ratio
SRR — Split Ring Resonator
TARC — Total Active Reflection Coefficient
TEM — Transverse Electromagnetic
UWB — Ultra-Wideband
VSWR — Voltage Standing Wave Ratio
WiMAX — Worldwide Interoperability for Microwave Access
WLAN — Wireless Local Area Network
WPAN — Wireless Personal Area Network
XPD — Cross-Polarization Discrimination

General Introduction

General Introduction

In modern wireless communication systems, the antenna constitutes a critical enabling component that governs the ultimate performance limits of the entire transceiver chain. As the sole interface between guided electromagnetic waves and radiated fields, its electromagnetic behavior directly impacts key system-level metrics, including achievable data rate, link reliability, spectral efficiency, and spatial diversity. Consequently, the continuous evolution of wireless standards toward higher throughput, lower latency, and enhanced connectivity imposes increasingly stringent requirements on antenna design, particularly in terms of compactness, wideband operation, and integration capability.

Ultra-Wideband (UWB) technology has emerged as a promising paradigm to address these demands by enabling sub-nanosecond pulse transmission over an extensive frequency spectrum. In parallel, Multiple-Input Multiple-Output (MIMO) techniques have become a cornerstone of modern wireless systems by exploiting spatial multiplexing and diversity mechanisms to significantly enhance channel capacity and link robustness without requiring additional spectral resources or transmission power. The synergistic integration of UWB and MIMO technologies therefore represents a highly attractive solution for next-generation wireless systems, especially in the context of dense indoor environments and Internet of Things (IoT) applications.

Despite these advantages, the design of compact UWB-MIMO antennas remains a challenging electromagnetic problem due to the inherently conflicting requirements between ultra-wide impedance bandwidth and high inter-element isolation. UWB operation typically relies on carefully engineered radiator geometries and partial ground plane configurations to achieve multiple resonant modes that merge into a continuous wideband response. However, these same mechanisms promote the propagation of surface currents across the ground plane, which constitutes the dominant coupling mechanism in closely spaced MIMO arrays.

The primary objective of this work is to design and investigate a compact four-port UWB-MIMO antenna capable of simultaneously satisfying full UWB regulatory bandwidth requirements, high inter-element isolation, and excellent diversity performance, while being implemented on a standard FR-4 substrate. The proposed approach aims to decouple bandwidth enhancement and isolation improvement mechanisms through a geometry-driven

electromagnetic strategy, thereby avoiding reliance on additional parasitic or decoupling structures that may compromise bandwidth integrity or increase design complexity.

The adopted methodology is based on a rigorous multi-stage electromagnetic design framework combining analytical modeling, full-wave numerical simulation, and systematic parametric optimization. The initial stage establishes the fundamental physical constraints and guides the primary dimensional synthesis of the radiating structure. Subsequent stages focus on progressive refinement of the radiator geometry to achieve wideband impedance matching, followed by the introduction of spatial and polarization diversity mechanisms to mitigate mutual coupling effects. A four-element MIMO configuration is then developed, exploiting orthogonal orientation principles to enhance isolation through polarization diversity. The final stage involves comprehensive performance evaluation in terms of scattering parameters, radiation characteristics, and MIMO diversity metrics, ensuring compliance with stringent performance requirements.

The dissertation is structured into four chapters. The first chapter presents the fundamental theoretical background of antenna systems and Ultra-Wideband technology, including radiation principles, performance parameters, and substrate considerations. The second chapter addresses MIMO antenna theory, including channel modeling, diversity performance metrics, coupling mechanisms, and a critical review of state-of-the-art decoupling techniques. The third chapter details the proposed antenna design methodology, including the evolution of the single-element radiator and the development of the full four-port configuration. The fourth chapter presents performance analysis, benchmarking against existing literature, and a discussion of the achieved results in relation to identified research gaps and design constraints.

Chapter 1

Fundamentals of Antennas and Ultra-Wideband Technology

1.1 Introduction

This chapter presents the fundamental theoretical background required for understanding and designing antennas used in modern wireless communication systems. As a key component of any RF system, the antenna ensures the conversion between guided signals and radiated electromagnetic waves, and therefore has a direct impact on overall system performance.

In the context of Ultra-Wideband (UWB) technology, characterized by its very wide frequency range (3.1–10.6 GHz), antenna design becomes particularly challenging. It must ensure proper impedance matching over a broad bandwidth, stable radiation characteristics, and faithful transmission of short-duration pulses.

The integration of MIMO techniques further enhances system capacity and reliability, but introduces additional design constraints such as mutual coupling and isolation between antenna elements. This chapter therefore establishes the essential foundations of antenna theory, key performance parameters, and the specific characteristics of UWB-MIMO systems, providing the necessary basis for the subsequent chapters.

1.2 Antenna Fundamentals

An antenna is a transducer that converts a guided electrical signal into a freely propagating electromagnetic wave during transmission, and performs the reverse conversion during reception. The IEEE Standard for Definitions of Terms for Antennas formalizes this as “that part of a transmitting or receiving system which is designed to radiate or to receive electromagnetic waves” [1]. Balanis offers a complementary view: the antenna is the transitional structure between free space and a guiding device (a coaxial line or hollow waveguide) that transports electromagnetic energy from source to radiator, or from radiator to receiver [2, 3].

This transitional role has a direct engineering consequence. The feed line typically presents a real characteristic impedance of 50 ohm, while free space presents an intrinsic impedance of 377 ohm. The antenna must therefore function simultaneously as a radiating element and as an impedance transformer between these two very different media. When this transformation is imperfect (when the antenna is mismatched) the transmission line stores energy reactively rather than transporting it, and power is reflected back toward the source [2].

1.2.1 Transmission and Reception Functions

An antenna fulfils two complementary electromagnetic functions, and by the principle of reciprocity, the same physical structure performs both equally well, as illustrated in **Figure 1.1**.

- **Transmission:** The transmitting antenna accepts an electrical signal from a feed line and converts it into a radiated electromagnetic wave. The conversion efficiency depends on three factors: the quality of impedance matching between feed and antenna, the radiation resistance of the structure, and the ohmic and dielectric losses within the antenna material [2, 4].
- **Reception:** The receiving antenna intercepts an incident electromagnetic wave and induces a voltage at its terminals. By the reciprocity theorem (first established by Helmholtz and formalized by Lorentz) the radiation pattern, gain, and input impedance of an antenna are identical in transmit and receive modes [2, 4]. This equivalence is practically important: all antenna parameters can be measured in whichever mode is more convenient, and the results apply equally to the other

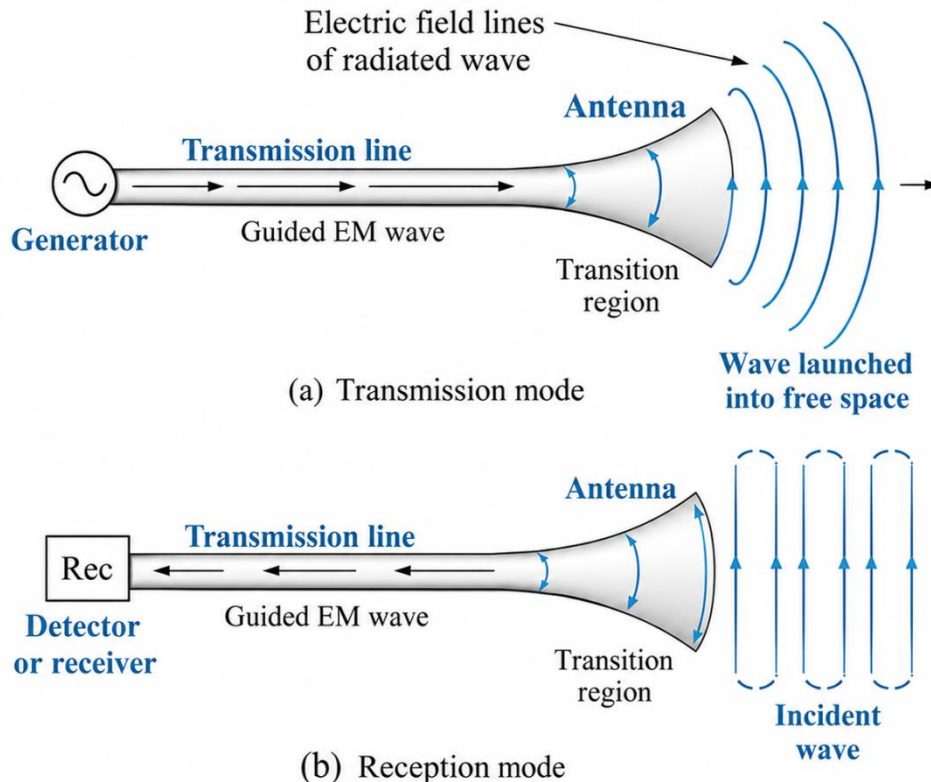


Figure 1.1: Electromagnetic wave conversion by an antenna: transmission and reception processes.

1.2.2 Electromagnetic Origin of Radiation

The physical origin of antenna radiation is explained by Maxwell's equations. A time-varying current flowing through a conductor generates time-varying electric and magnetic fields. When the current is constant or slowly varying, these fields remain attached to the conductor, forming the reactive near field where energy is stored rather than radiated. As the operating frequency increases and the conductor dimensions become comparable to the wavelength λ , the fields detach from the conductor and propagate through space as transverse electromagnetic (TEM) waves [4].

Radiation results from the acceleration of electric charges. According to Balanis, radiation occurs in three main situations: a time-varying current on a straight conductor, a current passing through a bend or discontinuity, and a time-varying current at the open end of a conductor [2]. In the far field, the electric field $\mathbf{E}$ and magnetic field $\mathbf{H}$ are perpendicular to each other and to the direction of propagation. Their amplitudes decrease as $1/R$ with distance R , while the power density decreases as $1/R^2$ [2,3].

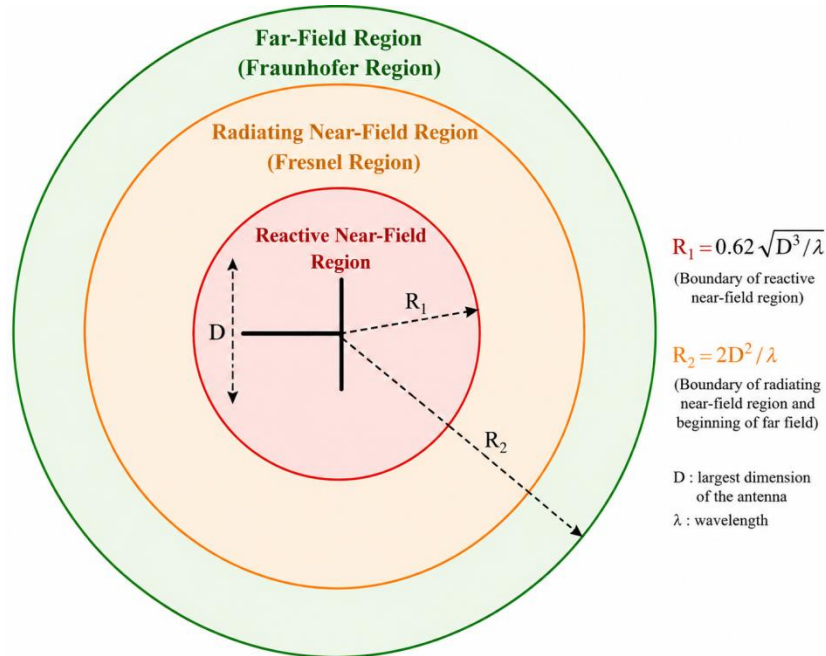


Figure 1.2:Field regions surrounding a radiating antenna

As shown in **Figure 1.2**, the region surrounding a radiating antenna is divided into three zones. The **reactive near field** ($R < 0.62\sqrt{(D^3/\lambda)}$) is dominated by stored energy. The **radiating near field** or **Fresnel region** ($0.62\sqrt{(D^3/\lambda)} \leq R < 2D^2/\lambda$) represents a transition zone where the field distribution still depends on distance. The **far field** or **Fraunhofer region** ($R \geq 2D^2/\lambda$) is characterized by a distance-independent radiation pattern. Standard antenna parameters such as gain, directivity, and radiation pattern are defined and measured in this region [3,5].

1.3 Key Antenna Performance Parameters

Antenna performance is characterized by a standard set of parameters defined in IEEE Std 145-2013 [1]. These parameters are not independent: they are connected through the underlying physics of radiation, and trade-offs between them govern every design decision. For UWB MIMO applications specifically, each parameter carries additional constraints and target values that are identified at the end of each subsection below.

1.3.1 Radiation Pattern

The radiation pattern is a graphical representation of the angular distribution of power radiated or received by the antenna, expressed as a function of direction (θ, ϕ) in spherical coordinates. It is the most informative single descriptor of antenna behaviour, and it is always evaluated in the far field [2, 1].

Patterns are conventionally analyzed in two principal orthogonal cuts: the E-plane, which contains the electric field vector E and the direction of maximum radiation, and the H-plane, which contains the magnetic field vector H and is perpendicular to the E-plane [2, 6]. A typical pattern exhibits a main lobe centered on the direction of maximum radiation, side lobes at other angles, and a back lobe directed approximately opposite to the main beam as shown in **Figure 1.3**. The half-power beamwidth (HPBW) quantifies the angular width of the main lobe between the two directions at which the radiated power falls to half (-3 dB) of its maximum value [1].

For four-port UWB MIMO antennas, a stable quasi-omnidirectional pattern (a toroidal distribution uniform across all azimuthal angles) is required across the full 3.1–10.6 GHz band. This is a far more demanding requirement than a single-frequency pattern specification:

as the frequency increases from 3.1 GHz to 10.6 GHz, the antenna's electrical size grows by a factor of 3.4, and secondary lobes inevitably develop [7, 8].

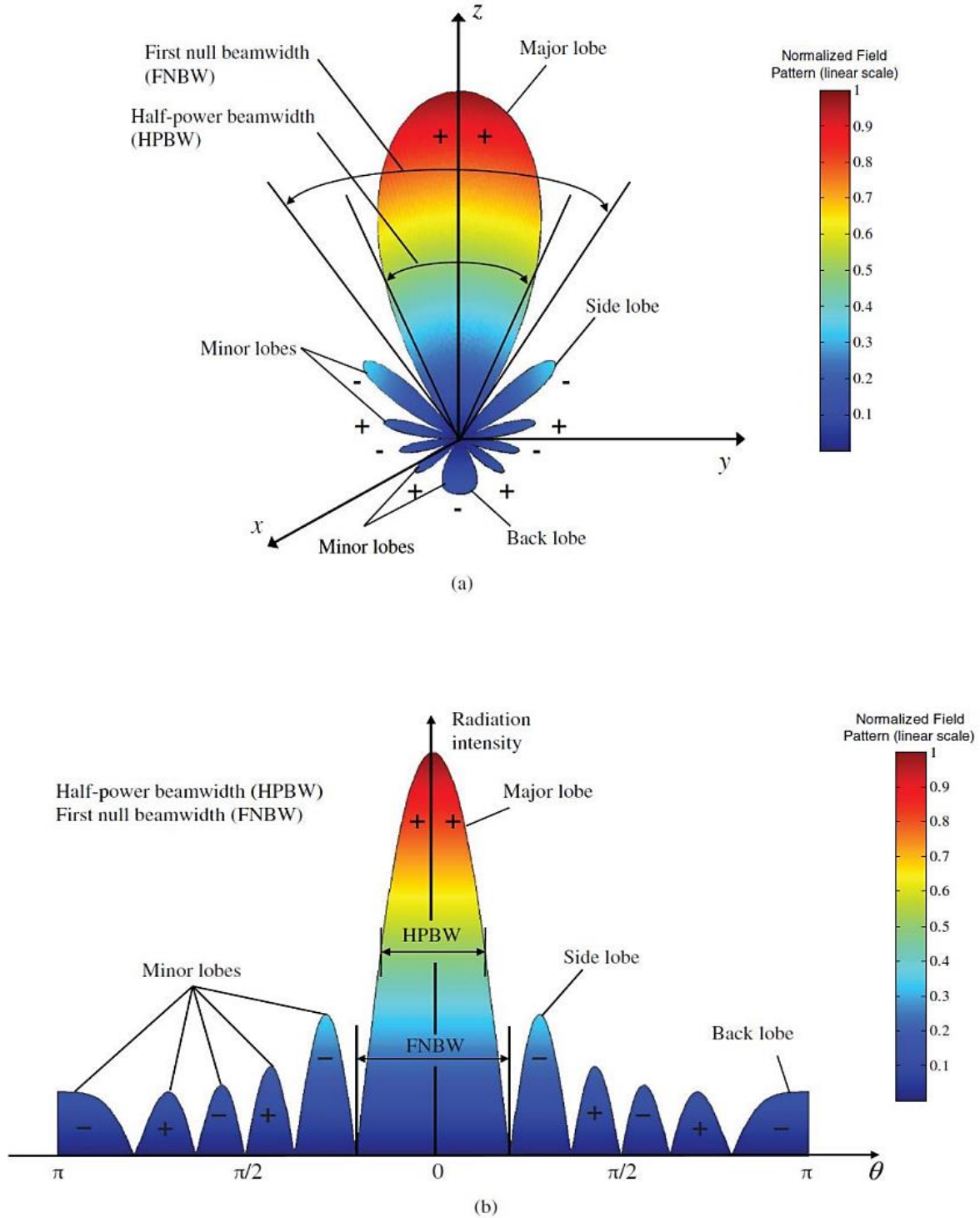


Figure 1.3: Three-dimensional (a) and two-dimensional (b) representations of a typical antenna radiation pattern

1.3.2 Directivity

Directivity D expresses how well an antenna concentrates radiated power in a specific direction compared to an isotropic radiator that distributes power uniformly in all directions. According to IEEE Std 145-2013 [1]:

$$D(\theta, \varphi) = \frac{U(\theta, \varphi)}{U_0} = \frac{4\pi \cdot U(\theta, \varphi)}{P_{\text{rad}}} \quad (1.1)$$

where $U(\theta, \varphi)$ is the radiation intensity [W/sr], $U_0 = P_{\text{rad}}/4\pi$ is the average radiation intensity over all directions, and P_{rad} is the total radiated power [W]. The maximum directivity D_{max} is:

$$D_{\text{max}} = \frac{4\pi \cdot U_{\text{max}}}{P_{\text{rad}}} \quad (1.2)$$

Directivity is a purely geometric quantity that depends only on the shape of the radiation pattern, not on the antenna's internal losses [2]. An ideal isotropic antenna has $D = 1$ (0 dBi).

1.3.3 Gain

The gain of an antenna combines the directivity of the antenna and its radiation efficiency. According to IEEE Std 145, the gain in a given direction is defined as the ratio of the radiation intensity in that direction to the average radiation intensity of an ideal isotropic antenna [1]. Mathematically:

$$G(\theta, \varphi) = \frac{4\pi \cdot U(\theta, \varphi)}{P_{\text{in}}} \quad (1.3)$$

The gain can also be expressed as a function of the directivity D and the radiation efficiency η :

$$G(\theta, \varphi) = \eta_{\text{rad}} \cdot D(\theta, \varphi) \quad (1.4)$$

Gain is expressed in decibels relative to an isotropic reference (dBi): $G [\text{dBi}] = 10 \cdot \log_{10}(G)$. For four-port UWB MIMO antennas, reported peak gains range from 3.5 dBi to 8 dBi. A stable, positive gain across the full 3.1–10.6 GHz band is essential for consistent link performance [6].

1.3.4 Bandwidth

The bandwidth of an antenna corresponds to the frequency interval over which the antenna's performance meets predefined criteria related to its proper operation [2]. The absolute bandwidth B_A is defined by:

$$B_A = f_2 - f_1 \quad (1.5)$$

where f_1 and f_2 are the lower and upper frequency limits. The fractional bandwidth (FBW), expressed as a percentage, is:

$$\text{FBW}(\%) = 200 \cdot \frac{(f_{\max} - f_{\min})}{(f_{\max} + f_{\min})} \quad (1.6)$$

Antennas are classified as narrowband (FBW < 1%), wideband (FBW 1%–20%), or ultra-wideband (FBW > 20%, or absolute bandwidth ≥ 500 MHz) [1, 9]. In microwave engineering, the bandwidth is frequently defined as the frequency range over which S_{11} at the antenna input is less than -10 Dbas shown in **Figure 1.4**.

The FCC-allocated UWB band (3.1–10.6 GHz) has $B_A = 7.5$ GHz and FBW = 109.5%, corresponding to a 3.4:1 frequency ratio. This is approximately 20–100 times wider than the bandwidth of a conventional microstrip patch antenna. Achieving $S_{11} < -10$ dB simultaneously across the entire band is the primary design challenge for UWB antennas [8].

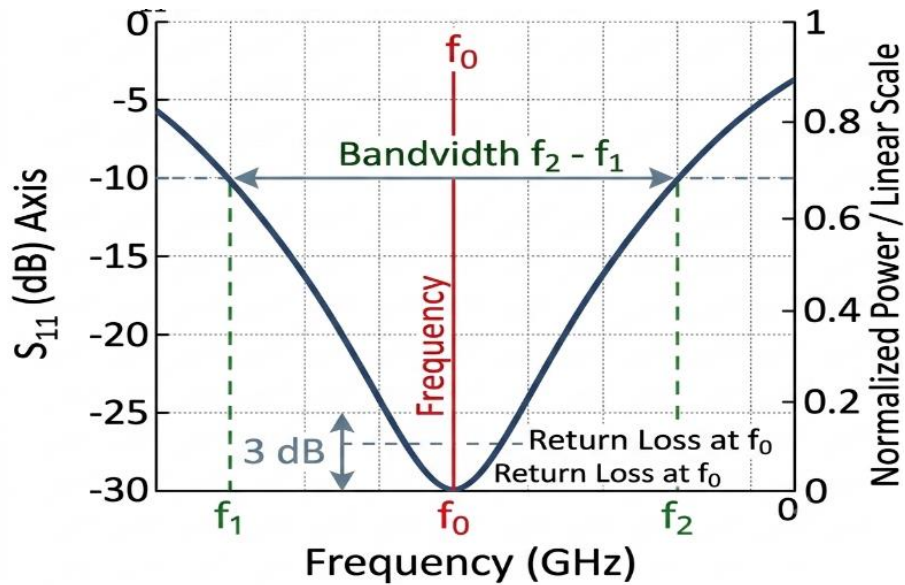


Figure 1.4: General definition of absolute bandwidth

1.3.5 Input Impedance

The electrical behavior of an antenna is equivalently represented by a series RLC circuit, modeling it as a passive system where current and voltage are not necessarily in phase. The input impedance Z_A of the antenna is expressed by the complex relationship:

$$Z_A(\omega) = R_A(\omega) + j \cdot X_A(\omega) \quad (1.7)$$

Within this framework, $R_A(\omega)$ denotes the total resistance, the sum of the radiation resistance R_r (power radiated into free space) and the loss resistance R_L (internal ohmic and dielectric losses). The reactive component X_A is determined by the equivalent inductance L and capacitance C :

$$X = L\omega - \frac{1}{C\omega} \quad (1.8)$$

To ensure maximum power transfer from the source to the antenna, the conjugate matching condition $Z_{\text{Source}} = Z_A^*$ must be fulfilled. The standard reference is a real source impedance of $Z_{\text{Source}} = 50 \text{ ohm}$, meaning the antenna must maintain $Z_A \approx 50 \text{ ohm}$ throughout its operational bandwidth. A universally accepted threshold for efficient match is $S_{11} < -10 \text{ dB}$, corresponding to $\text{VSWR} < 2:1$ and signifying that over 90% of incident power is transmitted to the radiating element.

1.3.6 Reflection Coefficient (S_{11}) and VSWR

When an antenna of impedance Z_A is connected to a transmission line of characteristic impedance Z_C (typically 50 ohm), the reflection coefficient at the antenna port is:

$$\Gamma = \frac{(Z_A - Z_C)}{(Z_A + Z_C)} \quad (1.9)$$

where $|\Gamma| = 0$ indicates perfect matching and $|\Gamma| = 1$ indicates total mismatch. The S-parameter S_{11} is directly related by $S_{11} = \Gamma$, and is expressed in decibels as:

$$S_{11}[\text{dB}] = 20 \cdot \log_{10}|\Gamma| = 20 \cdot \log_{10}|S_{11}| \quad (1.10)$$

The voltage standing wave ratio (VSWR) represents the ratio between the maximum and minimum voltages of the standing wave:

$$\text{VSWR} = \frac{(1 + |\Gamma|)}{(1 - |\Gamma|)} \quad (1.11)$$

The condition $S_{11} < -10$ dB corresponds to $\text{VSWR} < 2:1$ and means that less than 10% of incident power is reflected. This is the universally accepted matching criterion for UWB antennas [1, 8].

1.3.7 Polarization

Polarization refers to the orientation of the electric field vector $\vec{E}$ of the electromagnetic wave radiated or received in the far field as shown in **Figure 1.5**. More specifically, it defines the direction in which the electric field intensity is maximized. Understanding polarization is fundamental in antenna design, as it directly influences transmission efficiency and compatibility between communicating systems.

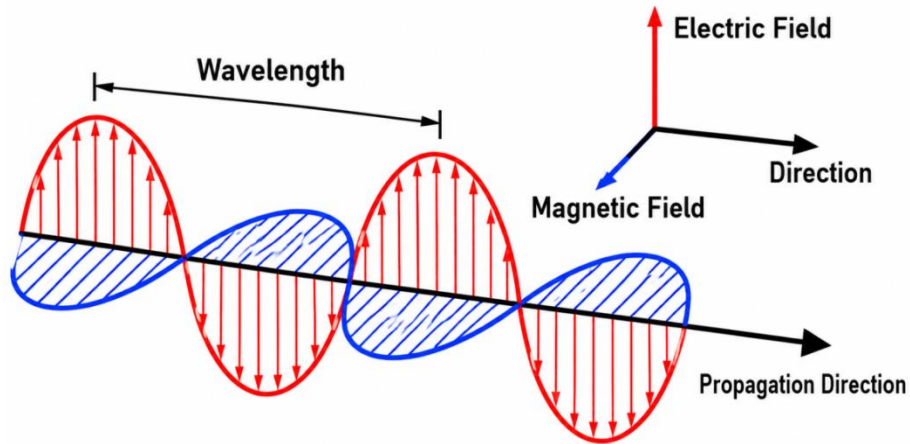


Figure1.5: Wave Polarization

- A. **Linear polarization:** occurs when the $\vec{E}$ vector maintains a constant direction over time, resulting in a straight-line trajectory. This is the most common polarization for UWB planar monopole antennas, which typically radiate a vertically polarized wave in the H-plane.
- B. **Circular polarization:** is achieved when the $\vec{E}$ vector rotates 360° during each cycle, arising from two orthogonal components of equal amplitude with a 90° phase

difference. This type of polarization is particularly effective in mitigating multipath fading, though it is more complex to realize.

- C. **Elliptical polarization:** represents the general case in which orthogonal components have unequal amplitudes or a phase shift other than 90°. Linear and circular polarizations are limiting cases of this broader category.

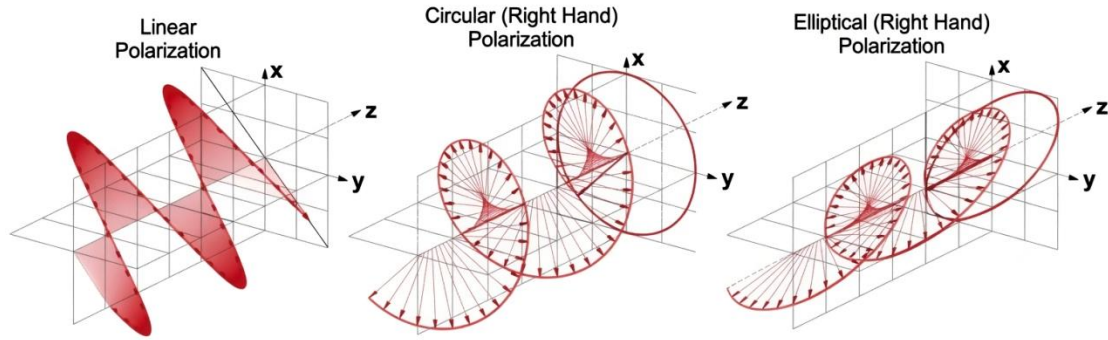


Figure 1.6: *Fundamental Polarization types*

Knowledge of the polarization type is essential for ensuring efficient transmission between antennas, as polarization mismatch can lead to significant signal losses. In the context of UWB MIMO systems, maintaining consistent linear polarization across the full band ensures predictable channel behavior and enables the decoupling of spatial and polarization diversity mechanisms. This contributes to improved reliability and overall system performance.

1.3.8 Radiation Efficiency

Radiation efficiency η_{rad} quantifies the ratio of the power radiated by the antenna to the total power accepted at its terminals:

$$\eta_{\text{rad}} = \frac{P_{\text{rad}}}{P_{\text{in}}} = \frac{R_r}{(R_r + R_L)} \quad (1.12)$$

This parameter is particularly critical for UWB antenna design because stringent regulatory limits on power spectral density mean that radiation efficiency directly determines the available link margin and communication range. High-performance UWB antennas on FR4 substrates typically aim for efficiency values between 77% and 95% across their entire

operating band [7]. Low-loss substrates such as Rogers RT/Duroid 5880 ($\tan \delta = 0.0009$) yield higher efficiency but at substantially greater cost [9].

1.3.9 Realized Gain

Realized gain G_{real} is the most practical measure of an antenna's performance because it accounts for real-world energy losses including impedance mismatch:

$$G_{\text{real}} = \eta_{\text{rad}} \cdot (1 - |\Gamma|^2) \cdot D \quad (1.13)$$

The factor $(1 - |\Gamma|^2)$ is the mismatch efficiency, ranging from 0 (total mismatch) to 1 (perfect match). For the four-port UWB MIMO designs reviewed in Section 1.11, realized gain across the operating band ranges from approximately 2 to 5 dBi, with a general tendency to increase with frequency as the antenna becomes electrically larger [10, 7].

1.3.10 Group Delay

Group delay is the antenna parameter most distinctive to UWB systems and the one most easily overlooked by designers accustomed to narrowband specifications. It is defined as the negative derivative of the transmission phase $\varphi(\omega)$ with respect to angular frequency:

$$\tau_{g(\omega)} = -\frac{d\varphi(\omega)}{d\omega} \quad (1.14)$$

It is measured from the phase of the S_{21} transmission coefficient between two identical antennas placed at a fixed separation in a UWB link. Constant (flat) group delay indicates a linear phase response, meaning the antenna reproduces the transmitted pulse without distortion. Varying group delay causes dispersion, where different frequency components of the UWB pulse arrive at different times, leading to a smeared and distorted signal [6, 11].

For UWB impulse-radio systems, the commonly accepted criterion is a group delay variation of less than ± 1 ns across the operating band [11]

1.4 Microstrip Antennas

Microstrip antennas, also known as patch antennas, are a popular type of antenna widely used in various applications. They were initially proposed by G. A. Deschamps in 1953, but

their practical implementation and widespread use came about in the 1970s when Robert E. Munson and some of his fellow researchers developed them using low-loss substrates [2].

1.4.1 Basic Structure

A microstrip antenna consists of a thin metallic radiating element, commonly referred to as a patch of length L and width W printed on one side of a dielectric substrate of thickness h and permittivity ϵ_r , while a conductive ground plane is located on the opposite side. There are several parameters that are dependent on the measures of the patch antenna. The initial dimensions of the baseline rectangular patch antenna were calculated using the well-established transmission-line model equations summarized in **Table 1.1**.

Table 1.1: Analytical Equations Used for the Initial Dimensioning of the Rectangular Patch Antenna

Parameter	Equation
Patch Width	$W = \frac{c}{2f_0} \sqrt{\frac{2}{\epsilon_r + 1}}$
Effective Dielectric Constant	$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(1 + \frac{12h}{W}\right)^{-1/2}$
Effective Length	$L_{eff} = \frac{c}{2f_0 \sqrt{\epsilon_{eff}}}$
Length Extension	$\Delta L = 0.412h \frac{(\epsilon_{eff} + 0.3) \left(\frac{W}{h} + 0.264\right)}{(\epsilon_{eff} - 0.258) \left(\frac{W}{h} + 0.8\right)}$
Actual Patch Length	$L = L_{eff} - 2\Delta L$

A schematic representation of the basic structure is shown in **Figure 1.7**. The radiating patch may take various geometrical shapes, including rectangular, circular, triangular, hexagonal, or other customized configurations depending on the application requirements

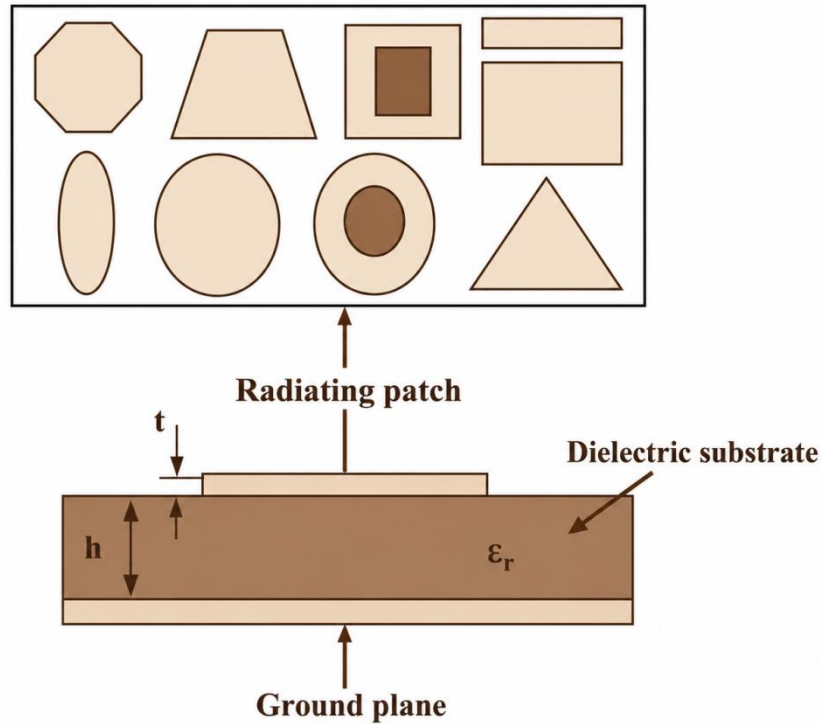


Figure 1.7: *Structure of a Microstrip Antenna [2]*

Copper is the most widely used conductive material due to its excellent electrical conductivity and low fabrication cost. The dielectric substrate is characterized by its relative permittivity (ϵ_r), which generally ranges from 2.2 to 12, and plays a critical role in determining the antenna performance.

1.4.2 Advantages of Microstrip Antennas

Microstrip antennas offer several advantages compared to conventional microwave antennas, which explains their widespread use in modern wireless communication systems. They are suitable for applications operating over a broad frequency range, typically from 100 MHz to 100 GHz. The main advantages of microstrip antennas are summarized as follows [2,6]:

1. They have a low-profile, lightweight, and compact structure.
2. They are simple to design and fabricate using standard printed circuit board (PCB) technology.
3. Their manufacturing cost is relatively low.
4. Both linear and circular polarizations can be easily achieved with appropriate feeding techniques.

5. Dual-frequency and dual-polarization operations can be implemented without significant complexity.
6. They can be readily integrated with microwave integrated circuits (MICs) and other electronic components.
7. Feeding networks and impedance-matching circuits can be fabricated simultaneously on the same substrate as the antenna.
8. They do not require cavity backing structures, which simplifies their construction.
9. Their conformal nature allows them to be mounted on curved surfaces, making them particularly suitable for aerospace and aircraft applications where aerodynamic constraints are important.

1.4.3 Disadvantages and Limitations of Microstrip Antennas

Despite their numerous advantages, microstrip antennas also exhibit several limitations when compared to other antenna technologies. These drawbacks must be considered during the design process [2].

1. Low radiation efficiency, mainly due to:
 - Conductor, dielectric, and surface-wave losses.
 - Unwanted radiation from feeding lines and discontinuities.
 - Excitation of surface waves within the substrate.
 - Significant ohmic losses in feed structures, particularly in antenna arrays.
2. Limited bandwidth, which restricts their use in certain wideband applications.
3. Low power-handling capability because of the thin substrate and conductor layers.
4. Relatively low gain, especially for single-element configurations.
5. Spurious radiation and mutual coupling effects, which may degrade antenna performance in array and MIMO systems.
6. Sensitivity to fabrication tolerances, where small variations in substrate properties or dimensions can affect antenna characteristics.
7. Reduced performance at very high frequencies, where conductor and dielectric losses become more significant.

1.4.4 Types of Feeding Techniques

Various techniques are available for transmitting electromagnetic (EM) energy to a microstrip patch antenna. The feeding process is crucial for ensuring efficient antenna

operation and improving impedance matching. The feeding techniques employed in microstrip antennas can be broadly categorized into two main classes: contacting feed methods and non-contacting feed methods [14].

The contacting feed method involves directly supplying power to the patch through a physical connection. In contrast, the non-contacting feed method achieves power transmission through electromagnetic coupling. Contacting feed methods include Microstrip Line Feeding, Coaxial Probe Feeding, and Fiber-Optic Feeding techniques, whereas non-contacting feed methods include Aperture-Coupled Feeding and Proximity-Coupled Feeding techniques, as illustrated in **Figure 1.8**.

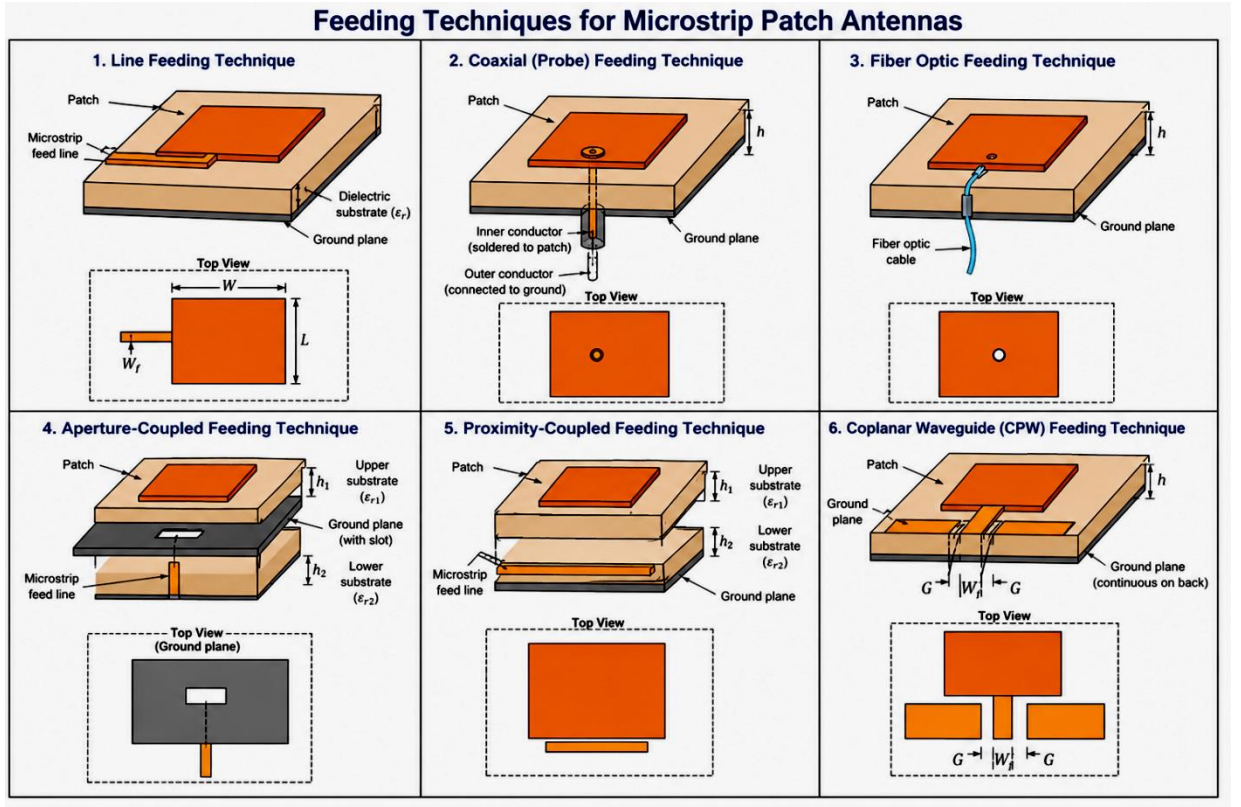


Figure 1.8: Feeding Techniques for Microstrip Patch Antenna

1.4.4.1 Line Feeding Technique

The Line Feeding Technique is one of the most commonly used feeding methods for microstrip patch antennas. In this technique, a conducting strip is directly connected to the edge of the radiating patch, with the width of the feed line generally being smaller than that of

the patch. One of the main advantages of this method is that both the radiating patch and the feeding network can be fabricated on the same dielectric substrate, resulting in a simple, low-cost, and easy-to-manufacture antenna structure [14].

The feed point may be located on the symmetry axis of the radiating element or shifted away from the symmetry axis to improve impedance matching. An inset-fed configuration can also be employed, where a notch is introduced into the patch to move the feed point toward the center of the radiating element. This arrangement allows proper impedance matching between the feed line and the antenna without requiring additional matching circuits.

1.4.4.2 Coaxial Feeding (Probe Feeding) Technique

The Coaxial Feeding Technique, also known as Probe Feeding, is one of the most widely used methods for exciting microstrip patch antennas. In this approach, illustrated in, the inner conductor of a coaxial connector passes through the dielectric substrate and is soldered directly to the radiating patch. The outer conductor of the coaxial cable is connected to the ground plane, thereby completing the feeding structure [15].

This feeding method provides good impedance matching and low spurious radiation. Furthermore, the feed position can be adjusted to achieve the desired input impedance. However, for thick substrates, probe inductance may increase, leading to a reduction in bandwidth and making the fabrication process more complex.

1.4.4.3 Fiber-Optic Feeding Technique

The Fiber-Optic Feeding Technique has been proposed as an alternative to conventional metallic microwave links in order to overcome some of their inherent limitations. In this approach, optical fibers are used to transmit signals between the source and the antenna system[16].

This technique offers several advantages, including extremely wide bandwidth, low transmission losses, high immunity to electromagnetic interference, lightweight construction, and compact dimensions. These characteristics make fiber-optic feeding particularly attractive for applications requiring high data rates, long transmission distances, and operation in electromagnetically noisy environments.

1.4.4.4 Aperture-Coupled Feeding Technique

In the Aperture-Coupled Feeding Technique, the ground plane acts as a separator between the radiating patch and the microstrip feed line. Energy is transferred from the feed line to the radiating patch through an aperture or slot etched in the ground plane [17].

Typically, the coupling aperture is positioned directly beneath the center of the patch, providing a symmetrical structure that helps reduce cross-polarization levels. The amount of coupling between the feed line and the patch depends on the shape, dimensions, and position of the aperture. This feeding technique offers improved isolation between the feed network and the radiating element, resulting in enhanced bandwidth and reduced spurious radiation. However, its multilayer structure increases fabrication complexity and manufacturing cost.

1.4.4.5 Proximity-Coupled Feeding Technique

The Proximity-Coupled Feeding Technique, also known as Electromagnetic (EM) Coupling Feeding, is one of the most efficient non-contact feeding methods used in microstrip patch antennas. This technique employs a two-layer or multilayer substrate structure. In this configuration, the microstrip feed line is located on the lower substrate, while the radiating patch is placed on the upper substrate.

The transfer of electromagnetic energy from the feed line to the patch occurs through electromagnetic coupling without any direct electrical connection. Consequently, capacitive coupling is established between the microstrip feed line and the radiating patch, enabling efficient power transfer [14].

1.4.4.6 Coplanar Waveguide (CPW) Feeding Technique

The Coplanar Waveguide (CPW) Feeding Technique is a widely used transmission-line configuration for feeding printed antennas. The main objective of this technique is to eliminate the need for drilling holes and soldering connections through the substrate.

One of the principal advantages of CPW feeding is that both the ground planes and the feeding line are located on the same side of the dielectric substrate, as illustrated in **Figure 1.8**. The radiating element is excited by a central metallic strip placed between two parallel ground conductors [17].

CPW-fed antennas offer several benefits, including simple fabrication, easy integration with active and passive microwave circuits, reduced radiation losses, and broader impedance

bandwidth. Consequently, CPW feeding has become a popular choice in modern wireless communication and ultra-wideband (UWB) antenna designs.

1.4.5 Microstrip Antenna Radiation Mechanism

The radiation mechanism of a microstrip patch antenna is primarily attributed to the fringing electromagnetic fields that exist around the edges of the radiating patch, as illustrated in **Figure 1.9**. When the patch is excited by an external source, electric charges accumulate on both the metallic patch and the ground plane. Positive charges are distributed on the lower surface of the patch, while corresponding negative charges appear on the upper surface of the ground plane, creating an electric field within the dielectric substrate [34].

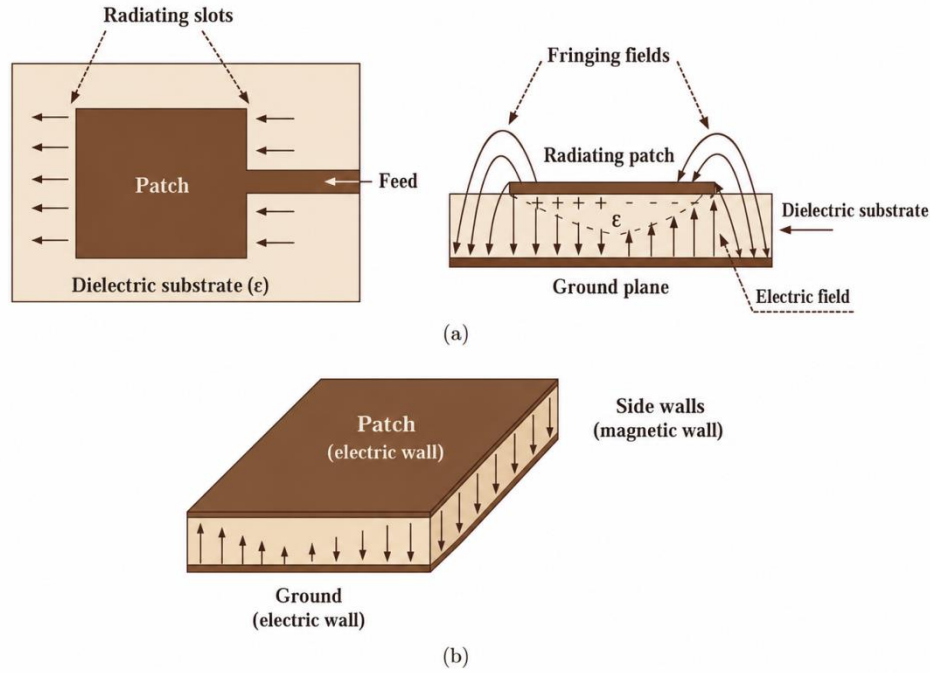


Figure 1.9:(a) *The radiation mechanism of a microstrip patch antenna and (b) its cavity model representation.*

Since the dimensions of the patch are finite, the electric field lines at the edges are not completely confined within the dielectric substrate. Instead, a portion of these fields extends into the surrounding air, forming what are known as **fringing fields**. These fringing fields are responsible for the radiation of electromagnetic waves into free space.

The analysis of the microstrip patch antenna often models the structure as a **resonant cavity**, where the metallic patch and the ground plane constitute electric walls, while the open edges behave as magnetic or impedance walls. When the antenna operates at its resonant

frequency, the cavity supports standing-wave patterns that allow efficient radiation through the fringing fields. At resonance, the antenna achieves maximum radiation efficiency because the stored electromagnetic energy is effectively converted into radiated power [34].

1.5 Ultra-Wideband Technology

1.5.1 Definition and Spectrum Allocation

Ultra-wideband technology refers to a radio transmission mode characterized by the use of an extremely large instantaneous spectral bandwidth compared to conventional narrowband or wideband systems. The formal definition specifies a UWB signal as one possessing a fractional bandwidth greater than 20% or an absolute bandwidth of at least 500 MHz [19].

$$B_{abs} = f_{max} - f_{min} \quad (1.15)$$

$$FBW = \frac{B_{abs}}{f_c} = \frac{2(f_{max} - f_{min})}{f_{max} + f_{min}} = \frac{1}{Q} \quad (1.16)$$

where f_{max} and f_{min} are the upper and lower -3 dB emission points and Q is the quality factor. The modern regulatory framework was established in February 2002, when the FCC allocated the 3.1–10.6 GHz band for unlicensed UWB communication devices, imposing a strict power spectral density limit of -41.3 dBm/MHz EIRP [8]. Similar frameworks were subsequently established by the European ECC and regulatory bodies in Japan and other regions [7].

1.5.2 Physical Properties of UWB

UWB systems exhibit several distinctive properties that differentiate them fundamentally from narrowband systems:

- **Short-pulse transmission:** Unlike carrier-modulated systems, UWB transmits information as very narrow pulses of sub-nanosecond duration, occupying a large spectrum simultaneously. This imposes the group delay requirement on UWB antennas: the antenna must preserve the pulse shape, not distort it [8].

- **High channel capacity:** From the Shannon capacity formula $C = B \cdot \log_2(1 + SNR)$, where transmission power is limited by regulation and battery constraints, wide bandwidth is the preferred solution to increase channel capacity [19].
- **Low power spectral density and coexistence:** UWB signals spread power over 7.5 GHz, so the PSD (-41.3 dBm/MHz) is at or below the thermal noise floor of most narrowband receivers. UWB signals therefore coexist with existing systems (GPS, WLAN, WiMAX) without causing significant interference [20], but are simultaneously limited to short-range communications (typically <10 m) [19].
- **High temporal resolution:** A 7.5 GHz bandwidth corresponds to a time resolution of approximately $1/B = 133$ ps, which translates to a range resolution of about 2 cm. This enables UWB ranging systems to achieve centimeter-level indoor positioning accuracy, far beyond Wi-Fi (1–3 m) or Bluetooth (1–2 m) [19].
- **Multipath resolution:** The large bandwidth allows individual multipath components to be temporally resolved, two reflections separated by more than 133 ps appear as distinct peaks. Integrating UWB with MIMO technology adds spatial diversity to temporal diversity, further suppressing fading [9].

1.5.3 UWB Application Domains

The combination of wide bandwidth, low power, and sub-centimeter temporal resolution has made UWB one of the most versatile short-range radio technologies:

- **High-speed WPAN:** UWB is one of the leading physical-layer technologies for short-range (<10 m), high-data-rate (>480 Mbps) wireless links, standardized under IEEE 802.15.3a [8].
- **Precision indoor localization:** UWB ranging achieves sub-10 cm accuracy in multipath-rich indoor environments. The technology is now integrated into Apple AirTag, Samsung SmartTag, and automotive keyless entry systems [11].
- **Ground-penetrating radar:** The high range resolution enables detection of buried utilities, structural defects, and human subjects behind walls [9].
- **MIMO-UWB systems:** The combination of MIMO and UWB technologies extends communication range and offers higher reliability. MIMO takes advantage of multiple

antennas at the terminals to enhance capacity, signal quality, and reliability in rich scattering environments [19, 7].

1.5.4 Fundamental Trade-Off between Bandwidth and Isolation in UWB–MIMO Antennas

The integration of Ultra-Wideband (UWB) and Multiple-Input Multiple-Output (MIMO) in a compact antenna introduces a fundamental electromagnetic trade-off. UWB antennas rely on broadband operation achieved through non-resonant traveling-wave currents and strong surface currents on a partial ground plane, which are essential for wide impedance bandwidth.

However, in MIMO systems, these same surface currents become the main coupling mechanism between closely spaced elements, degrading isolation and performance metrics such as ECC and channel capacity. Thus, techniques used to improve isolation (e.g., DGS, slots, stubs) can disrupt impedance matching and reduce bandwidth, while methods that enhance bandwidth may increase mutual coupling.

This creates a core design conflict where bandwidth and isolation cannot be optimized independently. To overcome this, this thesis adopts a polarization-based decoupling strategy using orthogonal antenna orientation, enabling high isolation without significantly affecting UWB performance.

1.6 UWB Antenna Design

1.6.1 Design Constraints Specific to UWB

Designing an antenna for UWB operation is a qualitatively different problem from narrowband antenna design. The following constraints must be satisfied simultaneously, a combination that no single conventional antenna topology meets without significant structural modification:

1. **Wide impedance bandwidth (FBW = 109.5%):** $S_{11} < -10$ dB must be maintained across the full 3.1–10.6 GHz band. This is a 3.4:1 frequency ratio, demanding a non-resonant radiating structure whose impedance varies slowly over a decade of frequency [8, 6].

2. **Omnidirectional pattern stability:** The H-plane pattern must remain quasi-omnidirectional across the full 3.4:1 bandwidth. As frequency increases, higher-order current modes emerge, producing secondary radiation lobes [7].
3. **Linear phase response:** Group delay variation must be held below ± 1 ns across the band to preserve UWB pulse shape [6, 11].
4. **Band-notch capability:** The UWB spectrum overlaps with WLAN (5.15–5.85 GHz), WiMAX (3.3–3.7 GHz), and X-band satellite downlinks (7.25–7.75 GHz). Selectively suppressing these sub-bands is achieved by embedding resonant slot structures: U-shaped slots, C-shaped slots, SRRs, or EBG elements [21, 17].
5. **Compactness:** IoT and wearable applications impose stringent size constraints. The trajectory in recent literature points toward sub- 30×30 mm designs, with Gomez-Villanueva et al. achieving 28×28 mm in 2025 [16] and Prithivirajan et al. demonstrating 20×30 mm in 2024 [10].
6. **Isolation in MIMO configurations:** When four antenna elements are co-located on a compact substrate, mutual coupling degrades isolation, increases ECC, and reduces diversity gain. Achieving $S_{ij} < -20$ dB ($i \neq j$) in a compact four-port arrangement is a central challenge [10, 21].

1.6.2 Planar Monopole Antennas: A Key Architecture for UWB Applications

Among all antenna topologies investigated for UWB applications, the planar monopole antenna (PMA) with a partial ground plane has emerged as the dominant architecture [6]. This convergence evident across all recent four-port UWB MIMO designs reflects a specific set of physical advantages that make the PMA the natural match for UWB requirements.

The basic PMA consists of a planar radiating patch on one face of a dielectric substrate, connected to a CPW or microstrip transmission line, with a partial rather than complete ground plane. The partial ground plane is the key to UWB bandwidth: it introduces capacitive coupling between the patch and the ground edge that creates multiple closely-spaced resonances which merge into a single broad passband. Liang et al. were among the first to rigorously characterize this mechanism for a printed circular disc monopole on FR4 substrate, establishing the design principles that subsequent literature has built upon [6].

The radiating element geometry provides further degrees of freedom:

- **Circular / octagonal:** The circular shape produces the most uniform broadband impedance response due to rotational symmetry. The octagonal approximation retains most of this benefit while being easier to implement with rectilinear PCB geometries. Abdelghany et al. show that transitioning from rectangular to octagonal patch extends impedance bandwidth from 3–6 GHz to 3.5–11 GHz [7].
- **Fractal:** Fractal geometries (Koch curves, Sierpinski carpets, Minkowski loops) increase the effective electrical length of the radiator without proportionally increasing its physical footprint. Prithivirajan et al. achieve 3–12 GHz bandwidth, 8 dBi peak gain, and >25 dB isolation in a 20×30 mm package using a fractal circular patch [10].
- **Hexagonal / heptagonal:** Multi-corner polygons approximate the circular shape using rectilinear geometry. They offer good bandwidth with straightforward fabrication and provide additional design variables [22].

The properties that make the PMA the preferred UWB topology are: its non-resonant broadband character eliminates the bandwidth ceiling of resonant patches; its CPW compatibility allows single-layer fabrication with no via holes; its monopole heritage produces a quasi-omnidirectional H-plane pattern; and its geometry provides independent optimization handles (Patch shape, feed gap, ground-plane size and slot configuration) for bandwidth, matching, pattern stability, and band-notch design simultaneously [7].

1.6.3 Design Challenges and Limitations

Despite these advantages, four principal challenges confront the designer of a compact four-port UWB MIMO antenna:

1. **Pattern distortion at high frequencies:** As frequency increases from 3.1 GHz to 10.6 GHz, the electrical size of the patch - $30 \times 30 \text{ mm}^2$ - grows from roughly 0.3λ to 1.1λ .

The current distribution becomes more complex and secondary lobes emerge, degrading H-plane omnidirectionality [7].

2. **Mutual coupling in MIMO configurations:** Placing four UWB elements on a compact substrate - $60 \times 62.5 \text{ mm}^2$ - brings them into proximity, and the resulting electromagnetic coupling degrades isolation (S_{ij}), raises ECC, and reduces diversity gain (DG). While adjacent orthogonal ports can achieve isolation exceeding 20 dB, [10, 21, 7, 23]
3. **Interference management:** Because the UWB spectrum is unlicensed and shared, band-notch structures are often embedded to selectively suppress interference from co-existing systems like WiMAX and WLAN. However, a fundamental trade-off exists: every integrated notch structure can perturb the input impedance and reduce the usable continuous UWB bandwidth, necessitating a distinction between constructive side-slots for bandwidth enhancement and rejection notches for interference mitigation. [21, 17, 19].
4. **Size-performance trade-off:** Reducing the system footprint toward the Wheeler–Chu limit ($ka \approx 1.38$) inevitably degrades critical metrics: impedance bandwidth narrows, radiation efficiency drops on lossy FR-4, and mutual coupling intensifies. While Defected Ground Structures (DGS) can recover matching for smaller geometries, and fractal radiators can increase electrical length without increasing area, they add significant complexity to the design and fabrication process.[10, 16].

1.6.4 UWB Antenna Advantages

- The combination of UWB physical-layer properties with the planar monopole architecture yields a practical technology with the following recognized advantages:
- **Low profile and low fabrication cost:** Standard PCB photolithography on FR4 substrates ($\epsilon_r = 4.4$, $\tan \delta = 0.02$) requires no specialized materials [18].
- **Ease of integration:** CPW feeding places feed line and radiator on the same substrate layer, eliminating via holes and reducing parasitic coupling [24].
- **Quasi-omnidirectional coverage:** The monopole heritage produces a stable toroidal H-plane pattern, enabling multi-user MIMO operation without beam steering [20].

- **Inherent wide bandwidth:** Partial ground plane design and optimized radiator shapes routinely achieve bandwidth ratios exceeding 3:1 [25].
- **Design flexibility:** The large parameter space (including patch radius, feed gap, and ground slot dimensions) allows simultaneous optimization of bandwidth, gain, isolation, and band-notch characteristics within a single planar layer [7, 26]. This flexibility is further exploited to integrate bandwidth-enhancement side slots for continuous coverage (Chapter 3) or selective rejection notches for interference mitigation (Chapter 4) within a single planar layer.

1.6.5 Mutual Coupling in Compact Arrays

When multiple UWB monopole elements share a compact substrate, mutual coupling (driven by capacitive, inductive, and surface-wave mechanisms) becomes the dominant design concern. These multi-antenna phenomena are treated in full in Chapter 2 (Section 2.5). At this stage, it is sufficient to record that the chosen octagonal radiator geometry directly governs the surface-current distributions that determine this coupling.

To resolve the UWB-MIMO Design Paradox (Section 1.5.4), this thesis adopts polarization-based geometric orthogonality (90° element rotation). This strategy is selected because it provides a significant mutual-coupling reduction (achieving a 10 to 20 dB isolation improvement over parallel arrangements) by exploiting pattern diversity rather than ground-plane current blocking. This approach preserves the integrity of the currents required for the audited 111.5% fractional bandwidth while ensuring adjacent-port isolation exceeds the 20 dB high-performance target.

1.7 Chapter Summary

This chapter presented the theoretical foundations of antennas and Ultra-Wideband (UWB) systems, with particular emphasis on the main performance parameters, radiation mechanisms, and the characteristics of microstrip antennas. The advantages of UWB technology, notably its wide bandwidth and high temporal resolution, were discussed, along with the challenges associated with its integration into MIMO systems.

The analysis also highlighted the fundamental trade-off between bandwidth and isolation among radiating elements, which constitutes the primary design challenge in UWB-MIMO antenna systems. These concepts provide the necessary foundation for the design methodology developed and presented in the subsequent chapters.

Chapter 2

UWB-MIMO Antennas

2.1 Introduction

Chapter 1 introduced the fundamental principles of antennas and UWB technology, highlighting the challenges associated with achieving wide bandwidth and efficient radiation in compact structures. It also identified the fundamental UWB–MIMO design trade-off, where the surface currents required for ultra-wideband operation simultaneously increase mutual coupling between antenna elements.

Although a single antenna can provide wideband performance, it remains vulnerable to multipath fading in wireless environments. MIMO technology overcomes this limitation by employing multiple antennas at both the transmitter and receiver, thereby improving channel capacity, link reliability, and spectral efficiency through spatial diversity and spatial multiplexing.

The combination of UWB and MIMO technologies offers significant advantages for high-data-rate communications, indoor localization, and short-range wireless networks. However, achieving high isolation and low correlation between closely spaced antenna elements remains a major design challenge.

This chapter presents the theoretical foundations of MIMO systems, the key performance metrics used to evaluate MIMO antennas, the mechanisms of mutual coupling, and the principal decoupling techniques reported in the literature. It concludes with a review of representative UWB-MIMO antenna designs and identifies the research gap addressed in this thesis

2.2 MIMO Systems

Multiple-Input Multiple-Output (MIMO) technology has become one of the most significant innovations in modern wireless communications. Over the past two decades, it has attracted considerable attention because it enables substantial improvements in channel capacity, transmission reliability, and spectral efficiency without requiring additional transmission power or bandwidth. Since radio spectrum is a scarce and valuable resource, MIMO has been widely adopted in major wireless communication standards, including IEEE 802.11n (Wi-Fi), WiMAX, LTE, and more recently 5G systems.

The fundamental principle of a MIMO system is the use of multiple antennas at both the transmitter and receiver, as illustrated in **Figure 2.1**. By exploiting the spatial dimension of the wireless channel, several independent data streams can be transmitted simultaneously within the same frequency band. This technique, known as spatial multiplexing, significantly increases the achievable data rate without consuming additional spectral resources.

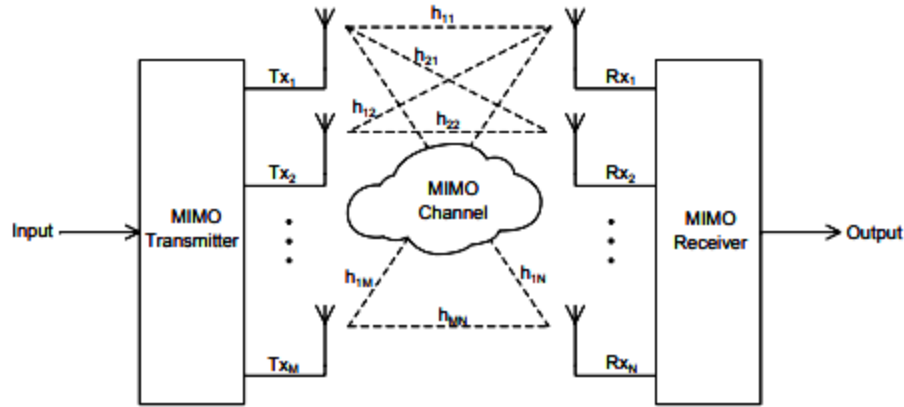


Figure 2.1: General architecture of a MIMO communication system.

In an ideal propagation environment, the channel capacity of a MIMO system increases approximately linearly with the minimum number of transmitting and receiving antennas. Furthermore, signal precoding techniques can be employed to distribute information across multiple antennas, providing additional array gain and simplifying signal detection at the receiver.

Besides capacity enhancement, MIMO systems also improve communication reliability through spatial diversity. In multipath fading environments, multiple antennas experience different channel conditions, reducing the probability that all communication paths simultaneously undergo deep fading. As a result, MIMO systems offer higher average capacity, lower outage probability, and improved link robustness compared to conventional single-input single-output (SISO) systems.

Despite these advantages, MIMO systems introduce additional design challenges. The presence of multiple antenna elements increases system complexity and may lead to significant mutual coupling when the antennas are placed in close proximity. Adequate antenna spacing is therefore required to ensure low correlation and high diversity performance. While this requirement is generally satisfied in rich scattering non-line-of-sight (NLOS) environments, achieving optimal MIMO performance in compact devices remains a

challenging antenna design problem, particularly for UWB-MIMO configurations where wide bandwidth and high isolation must be simultaneously maintained.

2.2.1 The MIMO Channel Model

Consider a MIMO system with N_t transmit antennas and N_r receive antennas. At any given moment, the transmitter sends a vector of signals $\mathbf{x} = [x_1, x_2, \dots, x_{N_t}]^T$, where each element is the signal radiated by one transmit antenna. The received signal vector $\mathbf{y} = [y_1, y_2, \dots, y_{N_r}]^T$ is related to $\mathbf{x}$ through the MIMO channel equation [1, 3]:

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \quad (2.1)$$

In this equation, $\mathbf{H}$ is the $N_r \times N_t$ complex channel matrix. Each entry h_{ij} of $\mathbf{H}$ represents the complex gain - including the amplitude change and the phase shift - of the propagation path from transmit antenna j to receive antenna i . The term $\mathbf{n}$ is a column vector of additive white Gaussian noise (AWGN) at the receiver, characterized by zero mean and covariance matrix $\sigma^2 \mathbf{I}_{N_r}$, where $\mathbf{I}$ is the identity matrix and σ^2 is the noise power per receive antenna.

The physical interpretation is straightforward. Each receive antenna collects a superposition of the signals from all N_t transmit antennas, each weighted by the corresponding channel coefficient h_{ij} . The matrix $\mathbf{H}$ therefore contains all the information about the propagation environment. The critical property is the rank of $\mathbf{H}$: if the columns of $\mathbf{H}$ are linearly independent (meaning the signals from different transmit antennas travel through statistically different paths and are not correlated) then the matrix has full rank, equal to $\min(N_t, N_r)$. A full-rank channel is the condition under which the maximum multiplexing gain of the MIMO system is achievable.

In the context of compact UWB MIMO antennas, the entries of $\mathbf{H}$ are assumed to follow independent and identically distributed (i.i.d.) complex Gaussian distributions - the Rayleigh fading model - provided that the separation between antenna elements is sufficient to ensure spatial decorrelation. Achieving adequate decorrelation in a physically compact structure, where antenna spacing may be as small as a fraction of a wavelength, is the central design challenge that this thesis addresses.

2.2.2 MIMO Channel Capacity

The single most important theoretical result in MIMO communications is the channel capacity formula derived by Telatar (1999) [3]. Channel capacity, measured in bits per second per Hertz (bps/Hz), is the maximum rate at which information can be transmitted reliably over a channel. For a single-antenna (SISO) channel, the Shannon capacity is $C = \log_2(1 + \text{SNR})$, which grows only logarithmically with the signal-to-noise ratio and cannot be significantly increased by simply transmitting more power.

For a MIMO channel in which the receiver knows the channel matrix H (but the transmitter does not), and power is distributed equally among all N_t transmit antennas, the ergodic capacity (averaged over the random channel realizations) is [3]:

$$C = E\{ \log_2 \det(I_{N_r} + (\rho/N_t) HH^H) \} \text{ [bps/Hz]} \quad (2.2)$$

Each symbol in this equation carries a precise meaning. The operator $E\{\cdot\}$ is the statistical expectation over the random channel H . The term ρ is the average received signal-to-noise ratio (SNR) at each receive antenna, a dimensionless ratio determined by the transmitted power, path loss, and noise level. The matrix H^H denotes the conjugate transpose (Hermitian transpose) of H : if H has entry $h_{ij} = a + jb$, then H^H has entry $h_{ji}^* = a - jb$. The matrix product $H H^H$ is $N_r \times N_r$; its eigenvalues $\lambda_1, \lambda_2, \dots, \lambda_{\min(N_t, N_r)}$ represent the power gains of the independent spatial channels (eigenmodes) of the MIMO link. The determinant of the bracketed matrix therefore equals:

$$\det(I + (\rho/N_t) HH^H) = \prod_{k=1}^r (1 + (\rho/N_t) \lambda_k) \quad (2.3)$$

where $r = \text{rank}(H) \leq \min(N_t, N_r)$. Taking the logarithm converts the product into a sum:

$$C = \sum_{k=1}^r \log_2 (1 + (\rho/N_t) \lambda_k) \text{ [bps/Hz]} \quad (2.4)$$

This final form reveals the physical meaning of MIMO capacity gain with complete clarity: the total capacity is the sum of the capacities of r independent parallel channels, each carrying a fraction of the total power. In the ideal case of a full-rank i.i.d. Rayleigh channel where all eigenvalues are equal ($\lambda_k = \lambda$ for all k), the capacity grows linearly with $\min(N_t, N_r)$. This linear scaling (which cannot be achieved by any single-antenna system regardless of

SNR or bandwidth) is the fundamental reason why MIMO technology is indispensable in high-throughput wireless systems [3, 4].

It is essential to understand what breaks this linear scaling in practice. If the columns of H are correlated because the antenna elements are too closely spaced, or because the propagation environment has insufficient scattering then the eigenvalues λ_k become unequal, with some large and most small. The capacity then approaches that of a single dominant eigenmode, and the multiplexing gain is lost. This is precisely why inter-element correlation (quantified by the ECC metric, Section 2.3.1) is such a critical figure of merit for compact MIMO antennas.

2.2.3 Diversity Modes in MIMO Systems

A MIMO system can be operated in two fundamentally different modes, and the design priorities for the antenna depend on which mode is targeted.

In multiplexing mode, the transmitter sends N_t independent data streams simultaneously over the same frequency band (one from each transmit antenna). The receiver uses the structure of H to separate these streams. The result is that the data rate scales with $\min(N_t, N_r)$, as shown by equation (2.2). This mode is the basis of the enormous throughput gains promised by MIMO in 4G, 5G, and Wi-Fi.

In diversity mode, the same data stream is transmitted (or received) over multiple independent antenna paths. Rather than increasing data rate, the goal is to improve reliability: even if one path undergoes a deep fade, the other paths remain usable and the combined received signal is far more stable. The maximum diversity order of an $N_t \times N_r$ MIMO system is $N_t \times N_r$ (the product of the transmit and receive antenna counts) and each unit of diversity order reduces the bit-error probability by another order of magnitude at a given SNR [4, 5].

Zheng and Tse (2003) formalized the fundamental trade-off between multiplexing gain and diversity gain [4]: a system cannot simultaneously achieve maximum multiplexing gain r^* and maximum diversity order d^* . The trade-off curve is $d(r^*) = (N_t - r^*)(N_r - r^*)$. For the four-port UWB system designed in this thesis, targeting high diversity gain ($r^* = 0$) gives $d = N_t \times N_r = 16$, providing maximum fading mitigation for reliable short-range communication.

Three physical mechanisms are exploited to create independent diversity branches in a compact antenna array:

1. **Spatial diversity:** relies on placing antennas far enough apart that the multipath phases arriving at each element are uncorrelated. In a rich indoor scattering environment, half a wavelength ($\lambda/2 \approx 48$ mm at 3.1 GHz) is typically sufficient.
2. **Polarisation diversity:** uses antenna elements oriented in orthogonal polarization planes (for example, one element radiating in the θ direction and another in the ϕ direction). Because the two polarizations experience the same path loss but independent phase rotations, their received signals are uncorrelated even when the elements are co-located.
3. **Pattern diversity:** uses antenna elements with distinct radiation patterns pointing in different directions. Because each element collects multipath energy from a different angular sector, the received signals are decorrelated by the spatial filtering of the patterns rather than by element separation.

In the design proposed in this thesis, all three mechanisms are engaged simultaneously: four planar monopole elements arranged orthogonally on a common substrate exploit both spatial separation and polarization diversity, while the different ground-plane geometries introduce pattern diversity across the UWB band.

2.2.4 MIMO Applied to UWB

The FCC First Report and Order (2002) defines a UWB emission as any signal whose -10 dB fractional bandwidth exceeds 20%, or whose -10 dB bandwidth exceeds 500 MHz, operating in the frequency range 3.1–10.6 GHz at a maximum equivalent isotropic radiated power (EIRP) spectral density of -41.3 dBm/MHz [2].

This extreme power constraint - approximately 0.075 mW spread across 7.5 GHz - is the reason UWB is inherently a short-range technology: beyond 10–15 metres, the received signal falls below the noise floor. Within that range, however, the enormous bandwidth enables two unique capabilities: data rates of several hundred Mbps and ranging precision better than 10 cm.

When MIMO spatial multiplexing is applied to UWB, the aggregate channel capacity becomes [1, 3]:

$$C_{\{UWB - MIMO\}} \approx B \times \sum_{k=1}^r \log_2(1 + (\rho/N_i)\lambda_k) \text{ [bps]} \quad (2.5)$$

where $B = 7.5$ GHz is the UWB bandwidth. Even with $r = 2$ independent streams and modest SNR, this expression yields aggregate capacities in the range of several Gbps, an order of magnitude beyond what a single UWB antenna can achieve [1].

The combination of UWB and MIMO is also synergistic from a multipath perspective. UWB pulses are so short in time (of the order of 130 ps at 7.5 GHz bandwidth) that individual multipath echoes can be temporally resolved, a property called multipath resolvability.

In a MIMO receiver, each resolved multipath component becomes an additional diversity branch, further enriching the effective channel rank. This stands in sharp contrast to narrowband MIMO systems, where multipath components overlap in time and must be treated statistically rather than individually.

The constraints imposed by UWB on MIMO design are, however, particularly severe. Because the operating band spans 3.1 to 10.6 GHz (more than three octaves) any decoupling structure added to reduce mutual coupling between elements must maintain its effectiveness over this entire range. A narrowband resonant isolator that works well at 5 GHz may be completely ineffective at 8 GHz. Furthermore, the compact form factor required by portable and wearable devices means that element separations are a small fraction of a wavelength at the lower end of the band, which greatly intensifies capacitive and inductive coupling [6].

2.3 MIMO Antenna Performance Metrics

Evaluating the performance of a MIMO antenna array requires a set of metrics that go beyond the classical single-element parameters (S_{11} , gain, efficiency). These metrics translate the electromagnetic behavior of the antenna (its coupling coefficients, radiation patterns, and impedance matching) into quantities directly related to the wireless link performance. Five metrics are used in this thesis. Their definitions, formulas, physical interpretations, and acceptance thresholds are presented in the following subsections. **Table 2.1** provides a concise reference summary.

Table 2.1: MIMO diversity metric definitions and acceptance thresholds.

Metric	Symbol	Formula / Definition	Acceptable Limit	High-Perf. Target	Ref.
Envelope Correlation Coefficient	ECC (ρ_e)	Far-field integral (eq. 2.6) or S-param. approx. (eq. 2.7)	< 0.5	< 0.01	[7]
Diversity Gain	DG	$DG = 10\sqrt{1- ECC ^2}$ dB (eq. 2.8)	> 9.5 dB	≥ 9.95 dB	[4]
Total Active Reflection Coeff.	TARC	Equations (2.9)–(2.10)	< 0 dB	< -5 dB	[4]
Channel Capacity Loss	CCL	$-\log_2(\det \Psi_R)$ (eq. 2.11)	< 0.4 bps/Hz	< 0.25 bps/Hz	[6]
Mean Effective Gain	MEG	Equation (2.15)	$ MEG \text{ ratio} < 3$ Db	< 1 dB difference	[8]

2.3.1 Envelope Correlation Coefficient (ECC)

The Envelope Correlation Coefficient (ρ_e , or ECC) is the most fundamental MIMO antenna metric. It quantifies the statistical similarity between the signal envelopes received at two antenna ports. An ECC of zero means that the two received signals are completely independent the ideal scenario for spatial diversity. An ECC of one means that the two received signals are identical, and the second antenna adds no diversity benefit. In between, the ECC reduces the effective diversity order proportionally.

Physically, a high ECC indicates that the two antenna elements share a large portion of their radiation pattern, they radiate (and receive) from the same directions and with the same polarizations. Reducing ECC therefore requires making the radiation patterns of the elements as different as possible: through element separation (geometric diversity), orthogonal element orientation (polarization diversity), or distinct element geometry (pattern diversity) [5, 6].

The exact expression for the ECC between ports i and j is derived from the overlap integral of the two complex radiation pattern vectors over the full sphere [7]:

$$\rho_e(i, j) = \left| \iint 4\pi F_i^* \cdot F_j d\Omega \right|^2 / \left\{ \iint 4\pi |F_i|^2 d\Omega \cdot \iint 4\pi |F_j|^2 d\Omega \right\} \quad (2.6)$$

Here, $F_i(\theta, \varphi)$ and $F_j(\theta, \varphi)$ are the complex electric-field radiation pattern vectors of elements i and j evaluated at the spherical angles (θ, φ) . The asterisk (*) denotes complex conjugation. The denominator normalises the result so that $0 \leq \rho_e \leq 1$. The numerator is the magnitude squared of the spatial overlap integral: it is large when F_i and F_j point in the same directions with the same polarization, and small (ideally zero) when they are orthogonal. This formula is valid for any number of ports and any antenna geometry, making it the reference expression for multi-element arrays [7].

For a two-element array with negligible ohmic losses, Blanch, Romeu, and Corbella (2003) showed that equation (2.6) can be accurately approximated using the S-parameters of the two-port network [7]:

$$ECC = |S_{11}^* S_{12} + S_{21}^* S_{22}|^2 / \{ (1 - |S_{11}|^2 - |S_{21}|^2)(1 - |S_{22}|^2 - |S_{12}|^2) \} \quad (2.7)$$

This formula is widely used because it is computed directly from the S-parameter simulation or measurement, no radiation-pattern integration is required. Its limitations must be understood, however:

- it assumes the antenna is lossless, so it overestimates the ECC for designs with significant ohmic losses;
- it is strictly valid only for two-element arrays. For the four-port array designed in this thesis, equation (2.7) is applied to adjacent port pairs (e.g. ports 1 and 2, ports 2 and 3) as a first-order estimate, and the full far-field formula (2.6) is used for the complete four-port characterization [6, 7].

The acceptance threshold for ECC is 0.5: a value established from the analysis of diversity-combining gain loss [5]. For high-performance designs, the target is $ECC < 0.01$, which corresponds to near-orthogonal radiation patterns and produces diversity gain within 0.05 dB of the theoretical maximum. The designs reviewed in Section 2.6 range from $ECC \approx 0.0001$ (exceptional) to $ECC \approx 0.5$ (marginal).

2.3.2 Diversity Gain (DG)

While ECC describes the correlation between ports, Diversity Gain (DG) quantifies the direct benefit of diversity combining in terms of link-quality improvement. Specifically, DG

is the reduction (in dB) in the required SNR to achieve a given bit-error probability (BEP), compared to a non-diversity (SISO) reference receiver operating in the same Rayleigh fading channel [5].

For a maximum-ratio-combining (MRC) receiver (the optimal diversity combiner) the DG is related to the ECC by [5]:

$$DG = 10 \times \sqrt{(1 - |ECC|^2)} \text{ [dB]} \quad (2.8)$$

This compact formula has a simple and important interpretation. When $ECC = 0$ (perfectly uncorrelated), $DG = 10$ dB: the MIMO receiver needs 10 dB less SNR than a SISO receiver to achieve the same error performance, corresponding to a $10\times$ reduction in required transmit power or a $10\times$ increase in achievable range. When $ECC = 0.5$ (the maximum accepted value), $DG = 10\sqrt{(1 - 0.25)} \approx 8.66$ dB - still useful, but meaningfully below the maximum. When $ECC = 0.01$, $DG \geq 9.995$ dB - essentially the theoretical maximum.

An important nuance: DG as defined by equation (2.8) applies to a two-branch equal-gain or MRC combiner under Rayleigh fading. For a four-branch (four-port) MIMO system the diversity combining is more complex, and the effective DG exceeds 10 dB when combining is performed across all four ports simultaneously. Nevertheless, equation (2.8) is consistently used in the literature as a conservative figure of merit computed for adjacent port pairs, and it is adopted in this thesis for comparability with published results [5, 6].

2.3.3 Total Active Reflection Coefficient (TARC)

The S_{11} parameter (the standard single-port reflection coefficient) describes how well one antenna element is matched to its feed when all other ports are terminated with matched loads. This is useful for characterizing isolated elements, but it does not represent the real operating condition of a MIMO array, in which all ports are simultaneously excited. When multiple ports are driven at the same time, the signal injected at one port couples into the others and can be re-reflected, altering the effective impedance seen by each port. The Total Active Reflection Coefficient (TARC) captures this effect [5].

TARC is defined as the square root of the ratio of total reflected power to total incident power in the entire array [5]:

$$TARC = \sqrt{\frac{\sum_i |b_i|^2}{\sum_i |a_i|^2}} \quad (2.9)$$

where a_i and b_i are the incident and reflected wave amplitudes at port i respectively, related to the S-parameter matrix by $b = S a$. For a four-port array with equal-magnitude excitations $e^{j\alpha_i}$ (where the phase α_i of each port is a free parameter) TARC as a function of the port phases $\alpha = [\alpha_1, \alpha_2, \alpha_3, \alpha_4]$ is computed as:

$$TARC(\alpha) = (1/2) \times \sqrt{[\sum_i |\sum_j S_{ij} e^{j\alpha_j}|^2]} \quad (2.10)$$

In simulation, this expression is evaluated by sweeping α_2 from 0 to π (with $\alpha_1 = 0$ fixed) to obtain a family of TARC curves, one for each phase combination. The envelope of this family defines the worst-case and best-case TARC across all possible excitation conditions. A TARC below 0 dB across the entire operating band confirms that the array is well-matched regardless of how the transmitter distributes power among the four ports. A TARC below -5 dB (less than 32% of the incident power reflected) is the target for a high-performance design [5, 6].

TARC is especially important for UWB systems because the broad operating band makes it possible for specific frequency-phase combinations to produce constructive reflection interference, locally raising the TARC above 0 dB even when the individual port S_{11} values are well below -10 dB across the band.

2.3.4 Channel Capacity Loss (CCL)

The Channel Capacity Loss (CCL) is the most direct link between the antenna's electromagnetic performance and the information-theoretic capacity of the MIMO channel. It measures how much capacity is lost (relative to the ideal uncorrelated case) due to residual correlation between the antenna ports [6].

CCL is defined as [6]:

$$CCL = -\log_2 \det(\Psi_R) \text{ [bps/Hz]} \quad (2.11)$$

where Ψ_r is the receive-side antenna correlation matrix. For a two-port system (adjacent port pair), Ψ_R is the 2×2 matrix:

$$\Psi_R = [\psi_{11}\psi_{12}; \psi_{21}\psi_{22}] \quad (2.12)$$

where the entries are derived from the S-parameters as follows:

$$\psi_{11} = 1 - (|S_{11}|^2 + |S_{21}|^2), \quad \psi_{22} = 1 - (|S_{22}|^2 + |S_{12}|^2) \quad (2.13)$$

$$\psi_{12} = -(S_{11} * S_{12} + S_{21} * S_{22}), \psi_{21} = \psi_{12}^* \quad (2.14)$$

Physically, the diagonal entries ψ_{ii} represent the available power at port i (after accounting for the power coupled to the other port), and the off-diagonal entries ψ_{ij} represent the correlation between ports i and j . The determinant of Ψ_R lies between 0 and 1: when $\det(\Psi_R) = 1$ (fully uncorrelated, perfect isolation), $\text{CCL} = -\log_2(1) = 0$ bps/Hz — no capacity is lost. When $\det(\Psi_R)$ approaches 0 (full correlation), CCL diverges — all MIMO capacity is lost.

For a four-port array with fourfold symmetry ($S_{11} = S_{22} = S_{33} = S_{44}$ and $S_{12} = S_{23} = S_{34} = S_{41}$), the 4×4 correlation matrix Ψ_R takes a block-circulant form and its determinant simplifies considerably. The CCL is then evaluated for each adjacent pair and reported as a function of frequency. The acceptance threshold is 0.4 bps/Hz; high-performance designs target less than 0.25 bps/Hz across the entire UWB band [6].

2.3.5 Mean Effective Gain (MEG)

The four metrics introduced above (ECC, DG, TARC, CCL) are primarily derived from the S-parameter matrix. The Mean Effective Gain (MEG) provides a complementary perspective by directly evaluating how efficiently each antenna collects power from a real-world multipath propagation environment.

MEG was introduced by Taga (1990) [8] and is defined as the ratio of the average received power at one antenna port (averaged over a large number of random positions in the propagation environment) to the average power that an isotropic antenna (in free space) would receive under the same conditions. It is given by [8]:

$$\text{MEG}_i = \iint 4\pi \left\{ \left[\frac{XPR}{1 + XPR} \right] G_{\{\theta,i\}} P_\theta + \left[\frac{1}{1 + XPR} \right] G_{\{\phi,i\}} P_\phi \right\} d\Omega \quad (2.15)$$

Each term in this integral requires careful explanation:

- **$G_{\{\theta,i\}}$ and $G_{\{\phi,i\}}$** are the θ -polarized and ϕ -polarized directional gain components of antenna element i .
- **$P_\theta(\theta,\phi)$ and $P_\phi(\theta,\phi)$** are the probability density functions (PDFs) of the incoming wave power in the θ and ϕ polarizations. For an isotropic scattering environment, $P_\theta = P_\phi = 1/(4\pi)$; for realistic environments these PDFs are non-uniform in angle.
- **XPR (Cross-Polarization Power Ratio)** is the ratio of average power in the θ -polarized component to the average power in the ϕ -polarized component, measured in the

propagation environment. Typical indoor values are $XPR \approx 1$ (equal power in both polarizations).

For a balanced diversity system, all elements should collect approximately the same average power from the environment, their MEG values should be nearly equal. The criterion is [5, 8]:

$$\left| \frac{MEG_i}{MEG_j} \right| < 3 \text{ dB for all pairs}(i,j) \quad (2.16)$$

A large MEG imbalance (greater than 3 dB) means that some elements are significantly less efficient than others in the target environment, undermining the balanced-diversity operation assumed by the DG formula. MEG is particularly important for compact arrays in which element proximity causes the radiation patterns to become asymmetric, even if the S-parameters appear balanced, the MEG can reveal subtle pattern mismatches.

2.4 Challenges in Compact UWB MIMO Antenna Design

The design of a compact multi-port UWB antenna involves four interrelated challenges. These are not independent engineering problems: each one exacerbates the others.

2.4.1 Mutual Coupling

Mutual coupling is the electromagnetic interaction between adjacent antenna elements. When one element radiates or receives a signal, it induces currents in all neighboring elements, not just through the intentional transmission path, but through direct electromagnetic induction. Three distinct physical mechanisms contribute to this coupling in compact planar arrays:

- **Capacitive coupling (near-field electric coupling):** occurs when the fringing electric fields around the edges of one antenna element extend to the neighboring element. The strength of this coupling is proportional to the electric field intensity in the gap between elements, which increases rapidly as elements are brought closer together. Capacitive coupling tends to dominate when the substrate permittivity is high and the element edges are nearly parallel.

- **Inductive coupling (near-field magnetic coupling):** arises from the magnetic field produced by the current distribution on one element, which induces an electromotive force (EMF) in the conductors of the neighboring element through Faraday's law. This mechanism is particularly significant when elements carry high surface currents in the regions closest to each other - for example, at the feed points of adjacent monopole elements.
- **Surface-wave coupling:** occurs when guided electromagnetic modes propagate along the substrate between elements. On a microstrip structure, surface waves can propagate in the substrate dielectric and excite currents on distant elements, creating a coupling path that is not reducible simply by separating the radiating structures. The strength of this coupling depends strongly on the substrate thickness and permittivity [9].

The net effect of all three mechanisms is an increase in the transmission coefficient $|S_{ij}|$ between non-driven ports. When $|S_{ij}|$ exceeds -10 dB across the UWB band, the power leaking between ports becomes significant: more than 10% of the power injected at one port is absorbed by or re-radiated from adjacent ports. This condition compromises both the radiation efficiency and the isolation, and it directly raises the ECC. The challenge in UWB design is that these coupling mechanisms operate at different intensities across the 7.5 GHz bandwidth, making broadband mitigation far more difficult than a narrowband solution [6, 9].

2.4.2 Inter-Port Correlation

Mutual coupling is the electromagnetic cause; inter-port correlation is its system-level consequence. When elements are electromagnetically coupled, the signals they receive are no longer statistically independent — a given multipath component arrives not only at the antenna it is incident upon but also, through coupling, at all adjacent antennas. The resulting correlation between the received signals reduces the effective rank of the channel matrix H .

To understand the severity of this problem, consider two adjacent monopole elements separated by $d = 0.1\lambda$ at 3.1 GHz (≈ 9.7 mm). At this separation, the coupling $|S_{12}|$ can easily reach -10 dB or higher, corresponding to an ECC well above 0.3. As the channel rank drops from the maximum of $\min(N_T, N_R)$ toward 1, the MIMO capacity approaches the single-

antenna capacity, and the spatial multiplexing advantage is entirely lost. This is why managing inter-port correlation (through physical element design rather than digital signal processing) is a prerequisite for any functional MIMO antenna system [6].

2.4.3 Active Impedance Mismatch

Impedance matching in a multi-port antenna system is fundamentally more complex than in a single-element design. The input impedance seen at any port depends not only on the self-impedance of that element but also on the mutual impedances between it and all other simultaneously excited elements. When multiple ports are driven at the same time, the currents induced by coupling from one port act as additional excitation sources for all other ports, modifying their effective input impedances.

A practical consequence of this effect is that an element which is perfectly matched in isolation ($|S_{11}| \leq -20$ dB) may exhibit significant active reflection when the array is fully excited, because the coupled currents shift the resonant frequency or detune the matching network. The TARC metric captures this: when TARC exceeds 0 dB at any frequency within the operating band, the net reflected power at the array level exceeds the net incident power, implying that the system is dissipating power in reflected waves rather than radiating it. For UWB systems, this active mismatch can narrow the effective operating bandwidth below the required 3.1–10.6 GHz span [5].

2.4.4 Compound Performance Degradation

The three preceding challenges (mutual coupling, inter-port correlation, and active impedance mismatch) do not act independently. They form a feedback chain: strong coupling raises the off-diagonal S-parameters, which simultaneously raises ECC (degrading DG and CCL) and shifts the active input impedances (degrading TARC). The degradation of TARC means that less power is radiated per port, which reduces the effective gain and therefore the MEG. The compound result is a design whose nominal single-port S_{11} performance may look excellent in isolation, but whose MIMO performance - as revealed by the five metrics - falls well short of the required thresholds.

In the UWB context, this compound effect is frequency-dependent: coupling that is manageable at 6 GHz (mid-band) may be severe at 3.5 GHz (where element electrical size is small and near-field coupling is strong) and take on a completely different character at 9 GHz

(where surface-wave coupling becomes dominant). A successful UWB MIMO antenna design must resolve all four challenges simultaneously across the entire 7.5 GHz band, a constraint that rules out most narrowband decoupling solutions [6, 9].

2.5 Decoupling Strategies for Compact MIMO Antennas

The literature proposes several techniques to reduce mutual coupling and improve isolation in compact MIMO antennas. These techniques can be classified into three main categories according to their physical mechanism.

2.5.1 Ground Plane

The ground plane is a major coupling path because surface currents generated by one antenna element can propagate and excite neighboring elements. Defected Ground Structures (DGS) mitigate this effect by etching slots or defects into the ground plane, creating high-impedance paths that suppress current propagation [9]. Common geometries include dumbbell, H-shaped, and spiral slots. Although DGS structures are compact and easy to integrate, a single resonant slot generally provides effective isolation over a limited frequency range. The geometry of the partial ground plane also influences both impedance matching and mutual coupling [6].

2.5.2 Resonant Slots and Metamaterial Structures

Resonant slots etched into the radiating elements or feed lines act as barriers that attenuate coupling currents. U-shaped, C-shaped, and T-shaped slots are widely employed because they directly affect near-field coupling between antenna elements [9]. Metamaterial-based structures, such as Split Ring Resonators (SRRs) and Complementary Split Ring Resonators (CSRRs), create stop-band characteristics that suppress electromagnetic propagation and enhance isolation [9,10]. However, these techniques are typically effective only within a limited frequency range around their resonant frequency. Multi-ring and fractal configurations can be used to extend the operating bandwidth [9,11].

2.5.3 Neutralization Lines and Parasitic Elements

Neutralization lines reduce mutual coupling through destructive interference. A transmission line connecting two antenna elements introduces a signal with opposite phase to

the coupled signal, thereby cancelling part of the mutual coupling [5,9]. This approach can provide broadband isolation enhancement without significantly modifying the antenna geometry. Parasitic elements, such as metallic stubs and floating conductors, create additional coupling paths that destructively interfere with the direct coupling path. Although effective, their design requires careful optimization to avoid degrading impedance matching and radiation characteristics [5,9].

2.5.4 Geometric Arrangement and Self-Decoupling

Geometric arrangement is one of the most effective ways to reduce mutual coupling because it prevents strong coupling from occurring rather than suppressing it afterward. Orthogonal antenna placement exploits polarization diversity: elements rotated by 90° radiate orthogonal field components, resulting in inherently low coupling even at small spacings. Using this principle, Abdelghany et al. [12] achieved isolation greater than 20 dB across the 3.5–11 GHz band without additional decoupling structures.

Self-decoupling extends this concept by shaping the radiating element so that its own current distribution suppresses coupling. Instead of relying on external stubs, slots, or parasitic elements, the antenna geometry itself generates current cancellation. Hussain et al. [13] reported isolation exceeding 30 dB using a CPW-fed self-decoupled antenna, representing one of the highest isolation levels in the reviewed literature.

Another approach employs Electromagnetic Band-Gap (EBG) structures, which are periodic arrangements designed to suppress surface-wave propagation. By creating a high-impedance stop band, EBG structures reduce coupling between antenna elements and can provide broadband isolation improvement. However, their main limitation is the additional substrate area required, which can be problematic in compact MIMO designs [9].

2.6 State of the Art: Comparative Review of UWB-MIMO Designs

This section reviews representative multi-port UWB-MIMO antennas reported in recent literature. The selected works cover the UWB band, employ at least two antenna ports, and report key MIMO metrics such as isolation and ECC. The designs are grouped according to their main objective: compactness, high isolation, or application-specific optimization.

2.6.1 Compact Footprint Designs

Miniaturization is a major requirement for portable and wearable devices. Kareem (2022) [1] proposed a compact four-port FR4 antenna of only $32.3 \times 32.3 \times 0.8 \text{ mm}^3$ operating from 3.1 to 12 GHz. Isolation exceeds 15 dB using DGS and feed-line stubs, while ECC remains below 0.3. Although highly compact, the design exhibits moderate gain ($\sim 3.5 \text{ dBi}$) and limited isolation.

Prithivirajan et al. (2024) [14] introduced a fractal four-port antenna covering 3.1–12 GHz. By combining fractal radiators with orthogonal element placement, the design achieves isolation above 25 dB, ECC below 0.25, DG above 9.9 dB, and peak gain up to 8 dBi. The improved performance comes at the expense of increased design complexity.

2.6.2 High-Isolation Designs

Abdelghany et al. (2023) [12] proposed a four-port octagonal monopole antenna on FR4 ($62.5 \times 60.5 \times 1.6 \text{ mm}^3$) covering 3.5–11 GHz. Through orthogonal element orientation and a modified ground plane, isolation remains above 20 dB across the entire band. The design also achieves $\text{ECC} < 0.01$, $\text{DG} > 9.95 \text{ dB}$, $\text{CCL} < 0.4 \text{ bps/Hz}$, gain of 3.5–5.5 dBi, and efficiency between 77% and 95%. This antenna serves as the primary benchmark for the present thesis.

Hussain et al. (2023) [13] demonstrated a self-decoupled four-port antenna operating at three discrete bands. The antenna achieves isolation greater than 30 dB and ECC below 0.0001 without additional decoupling structures, highlighting the effectiveness of self-decoupling. Its limitation is the absence of continuous UWB coverage.

2.6.3 Application-Specific Designs

Several studies focus on specific requirements such as band-notching, reconfigurability, or metamaterial integration. Band-notched antennas suppress interference from WLAN, WiMAX, and X-band systems using resonant structures such as slots, SRRs, or EBG elements [1,6].

Singh et al. (2023) [15] proposed a reconfigurable four-port MIMO antenna employing RF PIN diodes to dynamically enable or disable band-notched operation. Although designed for 5G applications, its reconfigurable SRR-based decoupling structure provides isolation above 28 dB.

Other notable contributions include the SRR-based UWB-MIMO antenna of Jabire et al. [11], the compact uniplanar elliptical-monopole design of Gómez-Villanueva et al. [16], and the orthogonally arranged rectangular monopole antenna of Ibrahim et al. [17], which achieved isolation above 18 dB and ECC below 0.01. Collectively, these studies confirm that geometric diversity, metamaterial structures, and reconfigurable techniques are among the most effective approaches for improving compact UWB-MIMO antenna performance. The data in Table 2.2 reveal three persistent trade-offs in the state of the art.

Table 2.2: Comparative summary of nine representative four-port UWB MIMO antenna designs from the recent literature

Reference	Size (mm ²)	Band (GHz)	Isolation (dB)	ECC	DG (dB)	Decoupling	Substrate
Abdelghany [12]	62.5×60.5	3.5–11	> 20	< 0.01	> 9.95	Orthogonal PMA	FR4
Saxena [6]	118×86	2–20 (SWB)	> 20	< 0.0065	> 9.96	Sword + ring	Rogers RO3035
Hussain [13]	60×60	2.4/5.2/8	> 30	< 0.0001	≈ 10	Self-decoupled	Rogers 5880
Prithivirajan [14]	Compact	3.1–12	> 25	< 0.25	> 9.9	Fractal + orth.	FR4
Singh [15]	70×70	5G recfg.	> 28	< 0.3	—	SRR + PIN diode	FR4
Jabire [11]	—	UWB	> 15	< 0.5	—	SRR metamaterial	FR4
Gómez [16]	Reduced	UWB	> 15	< 0.1	—	Uniplanar ellip.	FR4
Kareem [1]	32.3×32.3	3.1–12	> 15	< 0.3	—	DGS + stubs	FR4
Ibrahim [17]	Compact	UWB	> 18	< 0.01	> 9.9	Orthogonal PMA	FR4

2.7 Research Gaps and Positioning of the Proposed Contribution

The literature review highlights several limitations that remain unresolved in current UWB-MIMO antenna designs. Although many compact antennas have been reported on low-cost FR4 substrates, a compromise still exists between compact size and high isolation. Most compact designs provide isolation levels of only 15–18 dB, whereas isolation above 20 dB is generally achieved using larger structures. Consequently, realizing a compact FR4-based antenna with consistently high isolation across the UWB band remains a significant challenge.

Another limitation concerns the simultaneous achievement of all key MIMO performance metrics. While some antennas offer continuous UWB coverage, they often exhibit moderate isolation or diversity performance. Conversely, designs with excellent ECC, DG, TARC, and CCL values frequently rely on larger structures, expensive substrates, or non-continuous operating bands. Achieving full 3.1–10.6 GHz coverage while satisfying all major MIMO performance criteria remains difficult.

The self-decoupling approach represents a particularly promising solution because it improves isolation through the geometry of the radiating element itself, without requiring additional decoupling structures. However, current self-decoupled antennas have mainly been demonstrated for discrete multi-band operation rather than continuous UWB coverage. Extending this concept to the entire UWB band therefore remains an open research problem.

To address these challenges, this thesis proposes a compact four-port UWB-MIMO antenna implemented on an FR4 substrate. The design targets continuous 3.1–10.6 GHz operation, isolation above 20 dB, and excellent diversity performance within a compact footprint of approximately $60 \times 62.5 \text{ mm}^2$. The proposed strategy combines orthogonal element arrangement with an optimized ground-plane geometry to enhance inherent polarization-based isolation while preserving a simple and cost-effective structure.

2.8 Chapter Summary

This chapter presented the theoretical foundations of UWB-MIMO systems, including the MIMO channel model, key diversity metrics (ECC, DG, TARC, CCL, and MEG), and the effects of mutual coupling on antenna performance.

Several decoupling techniques were reviewed, with particular emphasis on geometric approaches based on polarization diversity, which offer high isolation with minimal complexity. A comparative study of recent UWB-MIMO antennas highlighted the main trade-offs between compactness, bandwidth, and isolation.

The identified research gaps motivated the proposed four-port UWB-MIMO antenna, which combines orthogonal element arrangement and optimized ground-plane design to achieve wide bandwidth, high isolation, and good diversity performance. Chapter 3 presents the proposed antenna design and its development process.

Chapter 3

Design and Simulation of Proposed Four-Port UWB MIMO Antenna

3.1 Introduction

This chapter presents the complete design methodology and simulation results of the proposed four-port UWB MIMO antenna. The objective is to achieve continuous operation over the entire UWB frequency band while ensuring high isolation, low correlation, and satisfactory radiation performance using a compact and low-cost FR-4 substrate.

The design process is carried out progressively. First, a single antenna element is developed through several geometric modifications to improve impedance matching and extend the operating bandwidth. Next, two-element MIMO configurations are investigated to evaluate the influence of element orientation on mutual coupling and isolation.

Finally, the complete UWB MIMO antenna is analyzed in terms of scattering parameters, radiation characteristics, gain, and standard MIMO performance metrics such as ECC, DG, TARC, CCL, and MEG. A comparison with recently published designs is also presented to assess the competitiveness and effectiveness of the proposed antenna.

3.2 Single-Element UWB Monopole Design and Evolution

This section presents the systematic five-stage evolution of the single radiating element, from a non-functional full-ground baseline (Stage 1) through progressive geometric modifications to the final proposed element (Stage 5). The complete set of optimized single-element parameters is collected in **Table 3.1**.

Table 3.1: *Optimized Geometrical Parameters of the Single Antenna Element.*

Parameter	Description	Stages 1–4	Stage 5 (Proposed)
L	Substrate side length	30 mm	30 mm
h	Substrate thickness	1.6 mm	1.6 mm
ϵ_r	Relative permittivity (FR4)	4.4	4.4
$\tan \delta$	Loss tangent	0.02	0.02
r_r	Patch radius	9.6 mm	9.6 mm
h_s	Feed line width	3.5 mm	3.5 mm
K	Feed line length	10.6 mm	10.6 mm
la	Partial ground plane length	—	9.52 mm
w_s (slot)	Ground slot width	—	3.5 mm
y_s	Ground slot length	—	2 mm
r_{slot}	Side slot radius (5-segment)	—	1.615 mm

3.2.1 Stage 1 : Rectangular Patch + Full Ground Plane

The design evolution begins with the simplest possible structure: a rectangular patch connected to a $50\ \Omega$ microstrip feed line above a full ground plane of length $l = 30\text{ mm}$ covering the entire bottom face of the $30 \times 30\text{ mm}^2$ substrate.

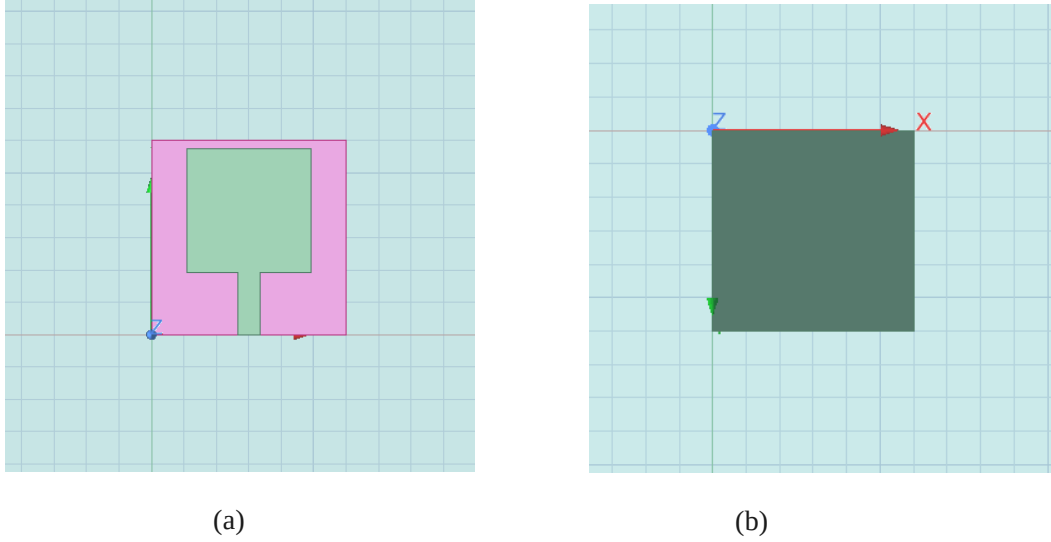


Figure 3.1: Stage 1 geometry: (a) top face showing rectangular patch and feed line; (b) bottom face showing full ground plane.

The simulated S_{11} response of Stage 1 across 2–11 GHz is presented in **Figure 3.2**. The response remains above -10 dB across the entire lower and mid UWB band, with a single narrow resonance near 8.5 GHz, confirming unsuitability for UWB operation.

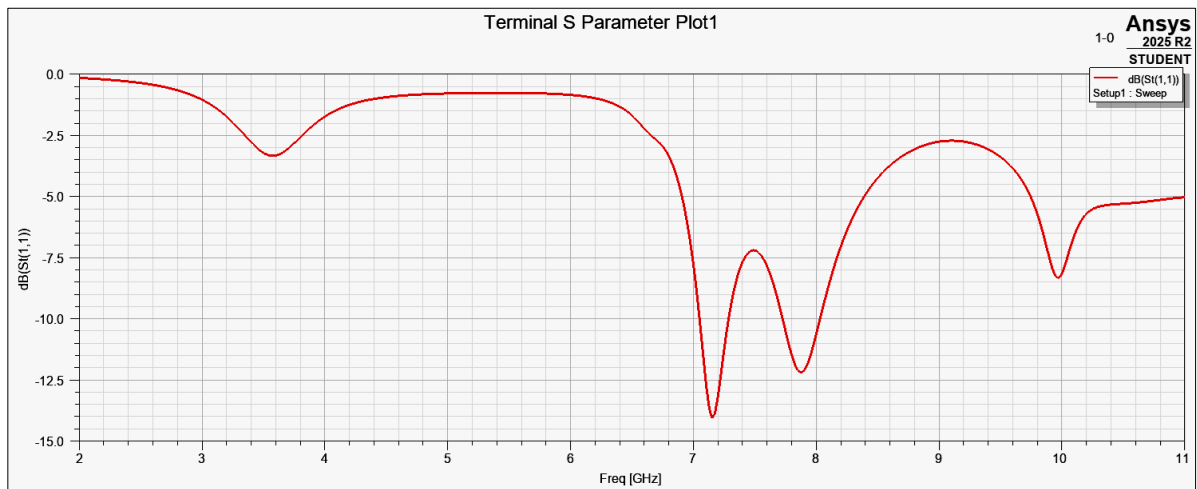


Figure 3.2: S_{11} of Stage 1 (rectangular patch + full ground).

The S_{11} response demonstrates that the -10 dB threshold is not reached before approximately 7.5 GHz, where a narrow resonance dip appears near 8.5 GHz (≈ -24 dB). Below 7 GHz, the response hovers between -0.5 dB and -3 dB - effectively no impedance matching - demonstrating that the full ground plane with rectangular patch is fundamentally incompatible with UWB operation across the 3.1–10.6 GHz band.

Physical explanation: The rectangular patch with full ground plane creates a high-Q resonant structure. The ground plane constrains surface currents, preventing the broadband multi-resonance behavior required for UWB. The single resonance at 8.5 GHz corresponds to the fundamental patch mode, but no higher-order modes merge to create continuous bandwidth.

3.2.2 Stage 2: Octagonal Patch + Full Ground Plane

In Stage 2, the rectangular patch is replaced by an octagonal patch with the same radius $r_r = 9.6$ mm, while maintaining the full ground plane. This modification tests the effect of patch geometry alone.

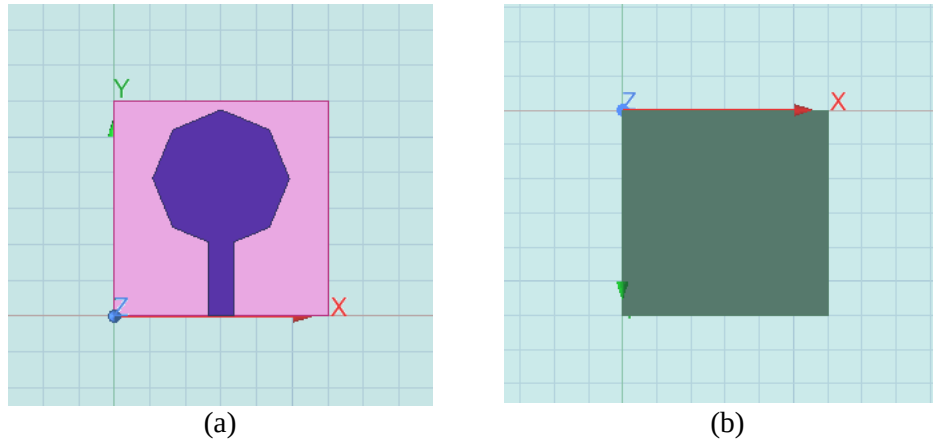


Figure 3.3: Stage 2 geometry: octagonal patch with full ground plane. (a) top face; (b) bottom face

The S_{11} response is presented in Figure 3.4. The octagonal patch is the geometry with the most uniform broadband impedance response. The eight edges create multiple current paths of slightly different effective lengths, producing closely-spaced resonances that begin to merge into a broadband response. However, the full ground plane still constrains the lower frequency matching, and the impedance matching above 9 GHz shows room for improvement.

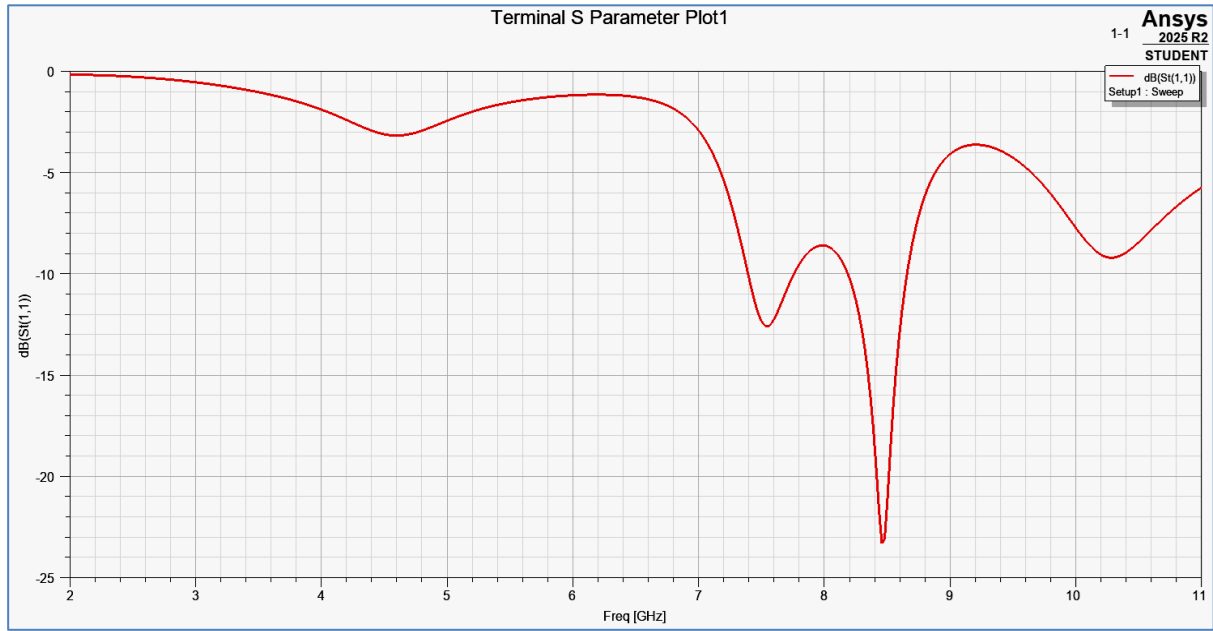


Figure 3.4: S_{11} of Stage 2 (octagonal patch + full ground).

3.2.3 Stage 3: Octagonal Patch + Partial Ground Plane

In Stage 3, the full ground plane is replaced by a partial ground plane of length $l_a = 9.52$ mm, leaving the upper portion of the bottom face unmetallized. This introduces the partial-ground UWB monopole architecture.

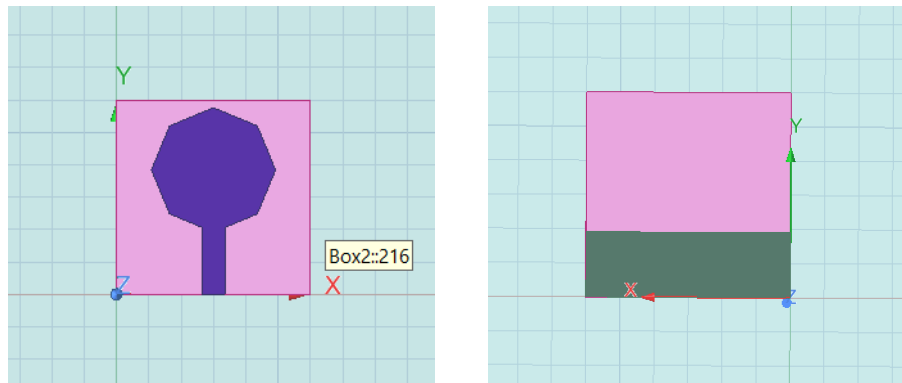


Figure 3.5: Stage 3 geometry: octagonal patch with partial ground plane ($l_a = 9.52$ mm). (a) top face ; (b) bottom face

The S_{11} response is presented in **Figure 3.6**. The partial ground plane creates multiple resonant mechanisms simultaneously. The lower -10 dB crossing remains at 3.17 GHz, but the upper-band

matching improves significantly. The exposed feed line section above the ground plane acts as a vertical monopole element, while the ground plane edges introduce additional resonant paths that merge into a continuous broadband response.

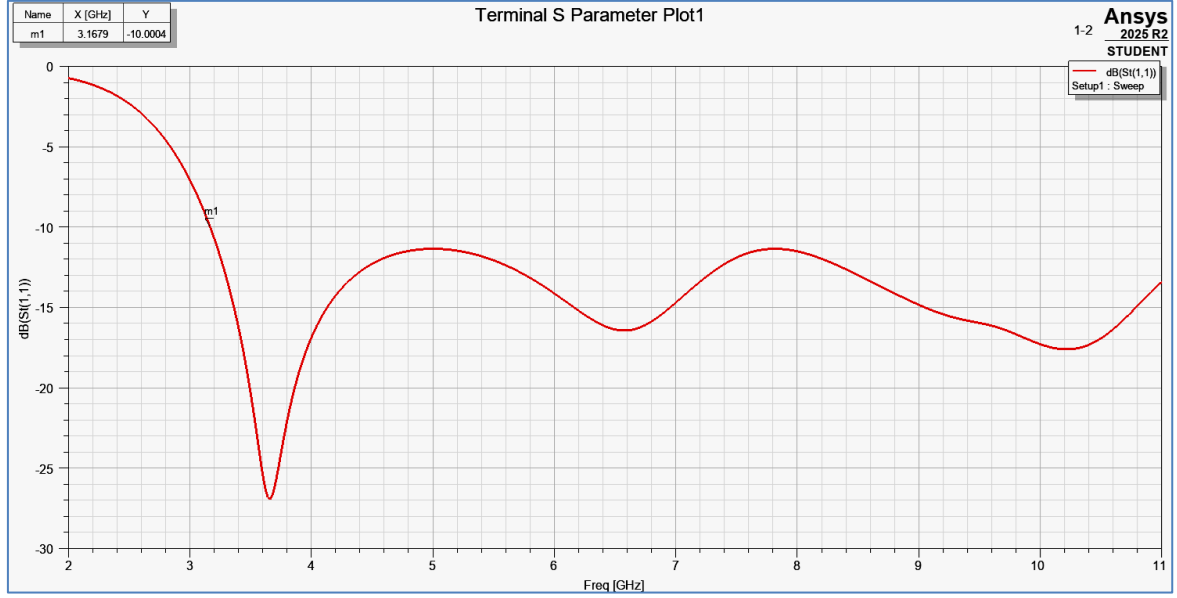


Figure 3.6: S_{11} of Stage 3 (octagonal + partial ground).

The partial ground plane is the key enabler of UWB bandwidth. It introduces: (1) capacitive coupling between the patch edge and the ground edge; (2) inductive reactance through the truncated ground edge that partially cancels the capacitive reactance at low frequencies; and (3) multiple closely-spaced resonances that merge into a single broad passband. The partial ground plane thus acts as an impedance transformer, extending the lower matching frequency while maintaining upper-band coverage.

3.2.4 Stage 4: Octagonal + Partial Ground + Ground Slot

Stage 4 introduces a rectangular slot etched into the center of the partial ground plane, positioned directly beneath the feed line. This slot modifies the surface current distribution to further optimize the impedance matching.

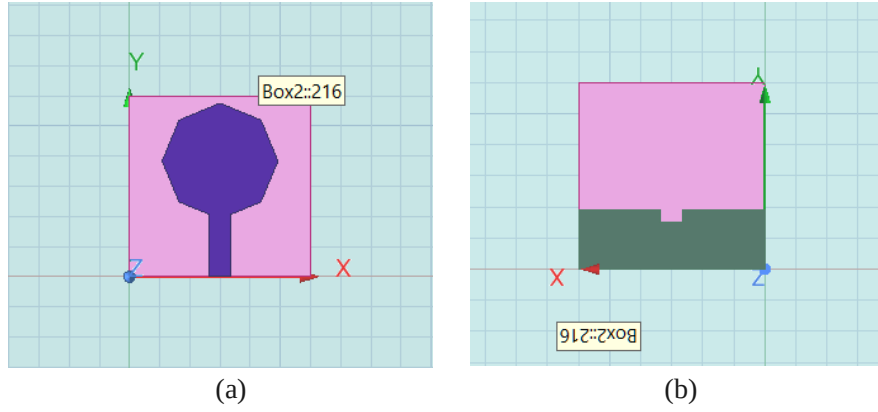


Figure 3.7: Stage 4 geometry: octagonal patch with partial ground plane and rectangular slot ($w_s \times y_s$). (a) top face ; (b) bottom face

A parametric study of both w_s and y_s was conducted to identify the optimal slot dimensions. **Figure 3.8** presents the parametric sweep results.

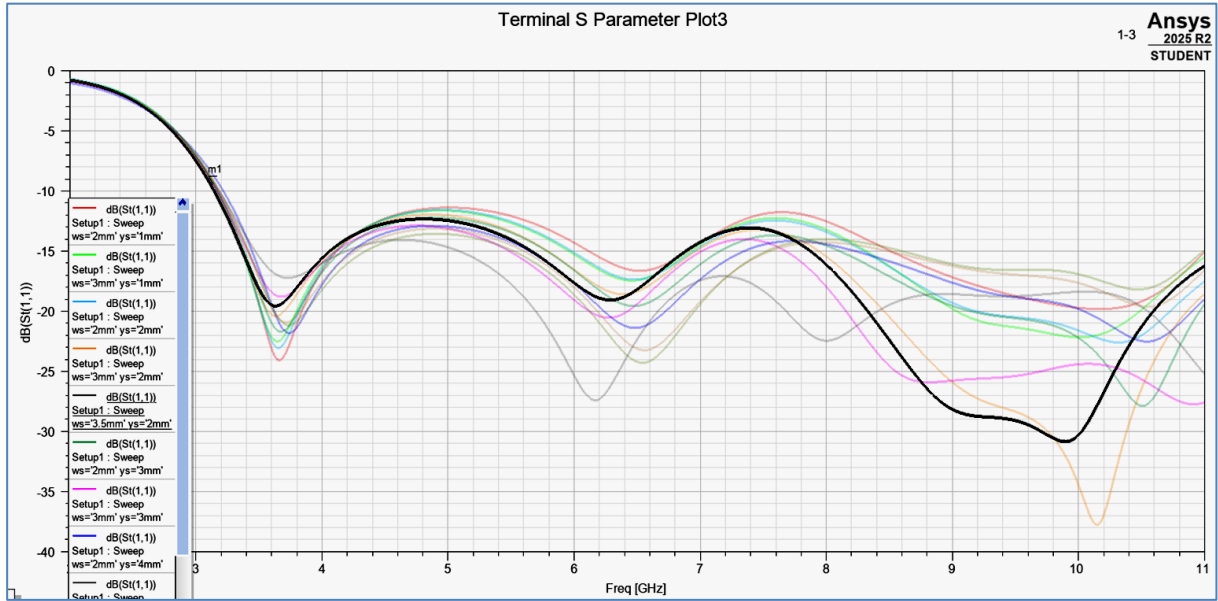


Figure 3.8: Parametric sweep of ground slot dimensions (w_s and y_s) .

The parametric study reveals that w_s primarily controls the width of the resonance dip created by the slot, while y_s controls the depth of the slot-induced current path and therefore the frequency of the slot resonance. The optimal combination of $w_s = 3.5$ mm and $y_s = 2$ mm (thick black curve) produces the best compromise: S_{11} below -10 dB continuously from approximately 3.15 GHz to beyond 10.6 GHz, with deep nulls reaching -30 dB in the upper band.

3.2.5 Stage 5: Proposed Element: Octagonal + Partial Ground + Ground Slot + Side Slots

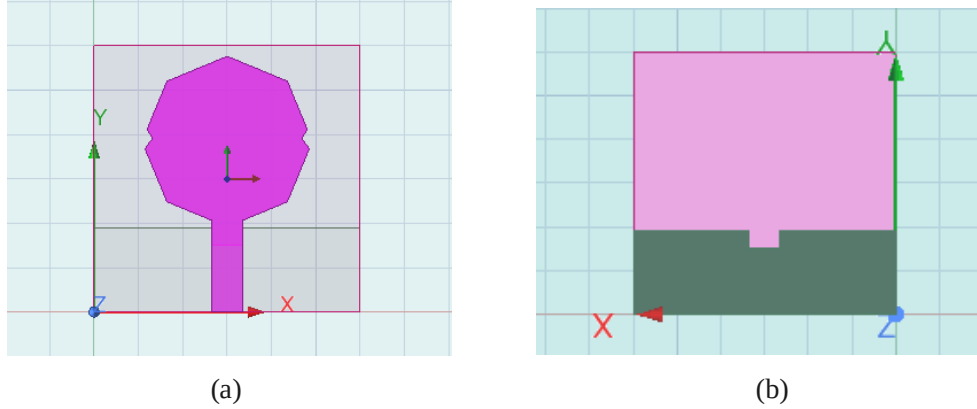


Figure 3.9: *proposed element geometry. (a) top face ; (b) bottom face*

The final proposed single element integrates two symmetric side-slot structures into the lateral sides of the octagonal patch. Each slot is formed by subtracting a 5-segment circular arc of radius $r_{\text{slot}} = 1.615$ mm from the octagonal patch at the widest lateral points. The S_{11} response is presented in Figure 3.10.

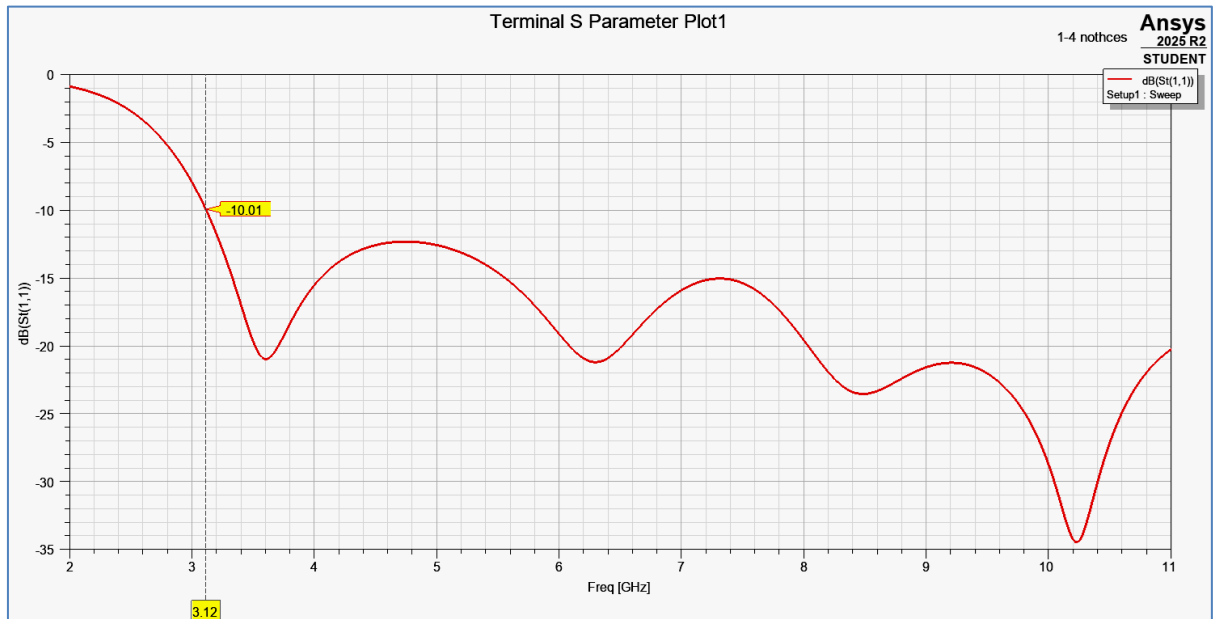


Figure 3.10: S_{11} of the proposed single element (Stage 5).

The addition of symmetric side slots produces two significant improvements over Stage 4: (1) the lower -10 dB edge shifts from 3.15 GHz to 3.12 GHz, providing additional

margin below the 3.1 GHz FCC lower boundary; and (2) the deep reflection nulls in the upper band are reinforced, reaching approximately -21 dB at 3.7 GHz, -21 dB at 6.5 GHz, -24 dB at 8.3 GHz, and -35 dB at 10.3 GHz.

The side slots create resonant current paths around their perimeters. Surface currents travelling across the patch surface must now circumnavigate the slot boundaries, so the effective current path length around each slot sets a resonant condition at frequencies where the path length equals a multiple of $\lambda/2$ in the effective medium. The interaction between the slot-induced current paths and the primary octagonal patch resonances produces additional deep nulls in the upper band, further broadening and deepening the impedance matching.

3.2.6 Performance Comparison

Table 3.2 summarizes the key S_{11} markers across the five-stage design evolution, quantifying the progressive improvement from the non-functional baseline to the proposed element.

Table 3.2: Single-element S_{11} key markers across the design evolution.

Stage	Geometry	Lower -10 dB Edge (GHz)	S_{11} at Edge (dB)	Upper Band Coverage	Key Improvement
1	Rectangular + full ground	No UWB match	> -10 dB below 7.5 GHz	None	Baseline (non- functional)
2	Octagonal + full ground	3.17	-10.00	3.17–11 GHz	Octagonal shape enables UWB
3	Octagonal + partial ground	3.17	-10.00	3.17–11 GHz	Partial ground maintains coverage
4	+ Ground slot ($w_s=3.5$, $y_s=2$)	3.15	-10.02	3.15–10.6+ GHz	Slot optimizes upper band
5	+ Side slots ($r=1.615$)	3.12	-10.01	3.12–10.6+ GHz	Best result

3.2.7 Radiation Patterns of the Proposed Single Element

The radiation patterns of the proposed Stage 5 element are evaluated at three representative UWB frequencies: 3.5 GHz (lower band), 6.5 GHz (mid band), and 9.5 GHz (upper band). **Figure 3.11** presents the E-plane patterns, and **Figure 3.12** presents the H-plane patterns.

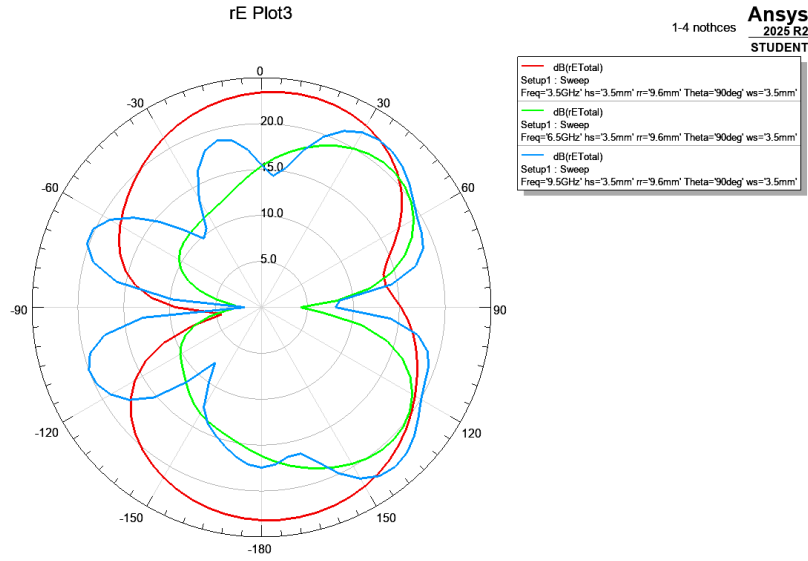


Figure 3.11: *E-plane radiation patterns ($\phi = 90^\circ$, θ sweeping) at 3.5 GHz, 6.5 GHz, and 9.5 GHz.*

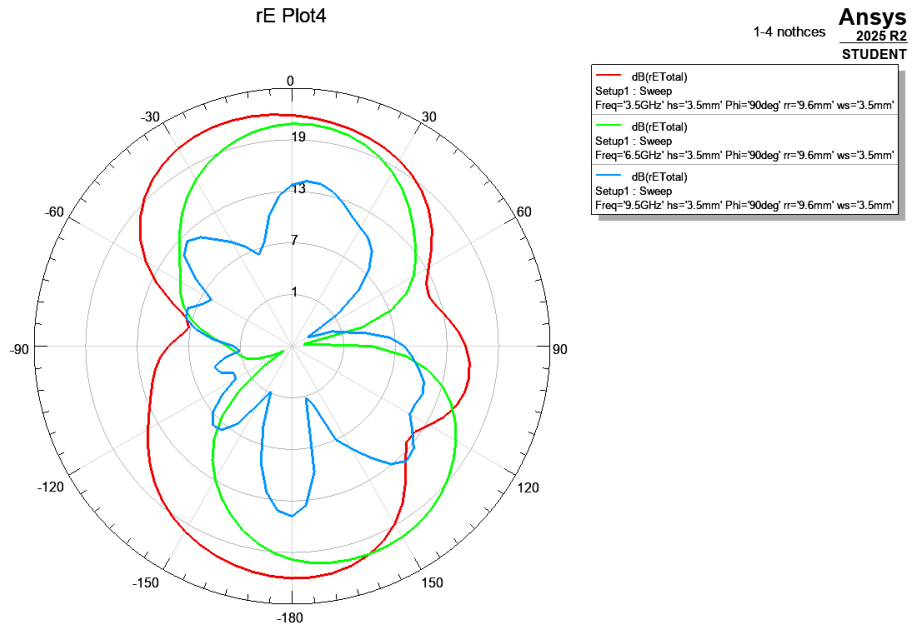


Figure 3.12: *H-plane radiation patterns ($\theta = 90^\circ$, ϕ sweeping) at 3.5 GHz, 6.5 GHz, and 9.5 GHz.*

In the elevation plane ($\phi = 90^\circ$), the antenna exhibits a bidirectional radiation pattern at 3.5 GHz with two symmetric lobes and a null in the broadside direction, consistent with the fundamental radiation mode of a planar monopole above a finite ground plane. As frequency

increases to 6.5 GHz and 9.5 GHz, additional lobes develop due to higher-order current modes, which is characteristic of electrically large wideband radiators. In the azimuthal plane ($\theta = 90^\circ$), moderate omnidirectionality is observed at 3.5 GHz with some azimuthal variation attributed to the finite partial ground plane geometry. The frequency-dependent lobe structure provides the pattern diversity that is beneficial for MIMO operation, as established in **Chapter 2, Section 2.2.3**.

Table 3.3: *Peak realized gain at representative UWB frequencies.*

Frequency (GHz)	Peak Realized Gain (dBi)
3.5	2.85
6.0	3.12
6.5	3.45
9.5	4.28

3.3 Two-Element Configuration Study

The single-element design established in Section 3.2 achieves excellent impedance matching across the entire UWB band. However, the central challenge of this thesis - the UWB-MIMO design paradox - remains unresolved: the same ground-plane surface currents that enable broadband bandwidth also create mutual coupling when multiple elements are placed in proximity. This section presents a systematic two-element study designed to isolate and demonstrate the effectiveness of polarization orthogonality as the primary self-decoupling mechanism.

3.3.1 Configuration A: Parallel Arrangement (0° Rotation)

In the parallel configuration, two identical Stage 5 elements are placed side-by-side on a common substrate with their feed lines aligned in the same direction. The geometry is illustrated in **Figure 3.13**.

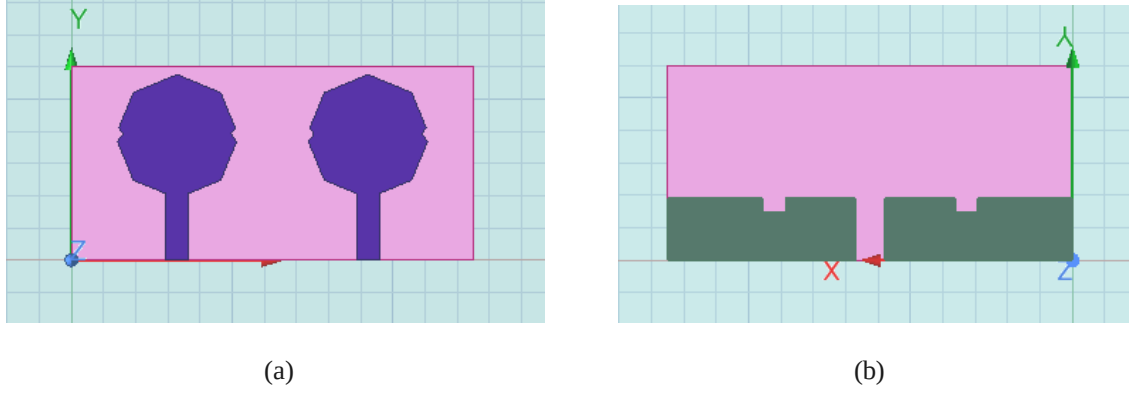


Figure 3.13: Parallel two-element configuration. (a) top face ; (b) bottom face

The simulated S-parameter response of the parallel configuration is presented in **Figure 3.14**. The return loss S_{11} (red curve) maintains the UWB characteristic of the single element, with the lower -10 dB crossing occurring at approximately 3.15 GHz. However, the mutual coupling S_{21} (green curve) reveals severe isolation degradation: the coupling remains above -15 dB across most of the band, peaking at approximately -9.9 dB near 4.3 GHz. This level of coupling is unacceptable for MIMO operation, as it would cause significant signal corruption at the receiver and degrade the channel capacity. The physical origin of this strong coupling is the shared ground plane: surface currents excited on the ground plane by Port 1 flow unimpeded toward Port 2, and because both elements share the same polarization, the radiated E-fields constructively interfere in the coupling path.

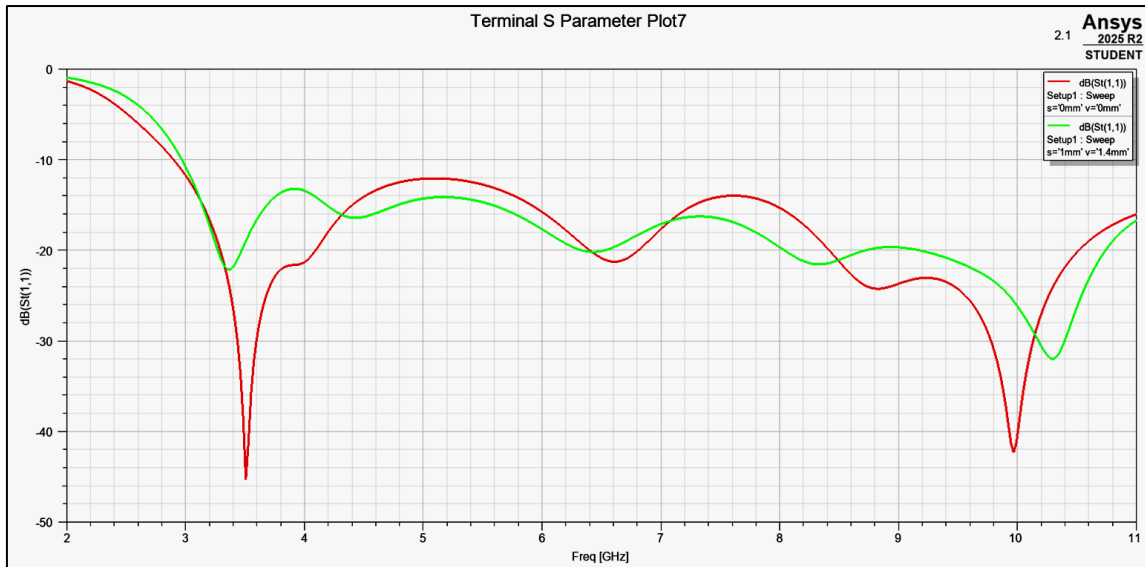


Figure 3.14: Simulated S-parameters of the parallel two-element configuration

The isolation S_{21} remains above -15 dB across the majority of the UWB band, with a peak coupling of -9.9 dB at 4.3 GHz, demonstrating unacceptable mutual coupling for MIMO operation.

Table 3.4 summarizes the key S-parameter markers for the parallel configuration, confirming that the -20 dB isolation threshold is violated across the entire band.

Table 3.4: Key S-parameter markers for the parallel two-element configuration.

Parameter	Value	Frequency (GHz)
S_{11} minimum	-31.8 Db	10.3
S_{21} maximum (worst coupling)	-9.9 dB	4.3
S_{21} at band edges (3.1 / 10.6 GHz)	-16.5 / -16.3 dB	3.1 / 10.6

3.3.2 Configuration B: Orthogonal Arrangement (90° Rotation)

To demonstrate the self-decoupling principle, the second element is rotated by 90° about the substrate normal while maintaining the same center-to-center spacing. In this orthogonal configuration, the feed line of Element 2 is perpendicular to that of Element 1, creating polarization orthogonality between the two radiators. The geometry is illustrated in **Figure 3.15**.

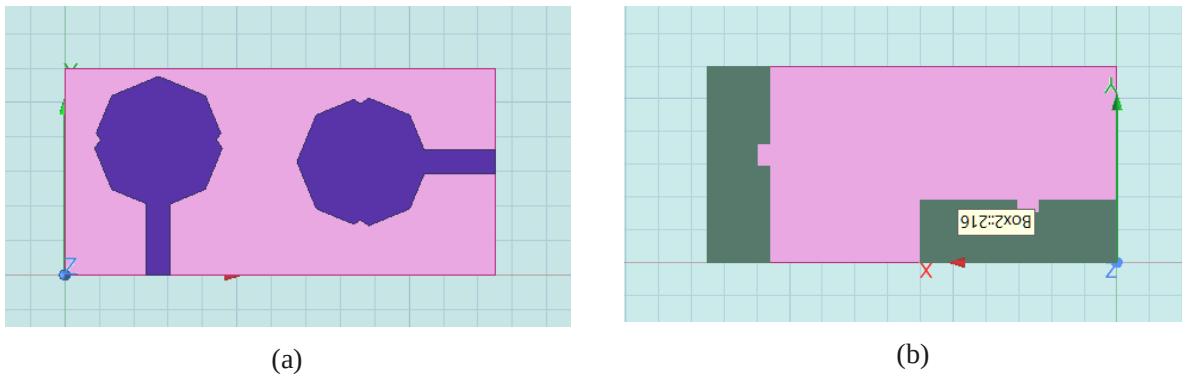


Figure 3.15: Orthogonal two-element configuration(a) top face ; (b) bottom face

The simulated S-parameter response of the orthogonal configuration is presented in **Figure 3.16**. The return loss S_{11} (red curve) remains below -10 dB across the UWB band, confirming that the 90° rotation does not degrade the individual element matching. Crucially, the mutual coupling S_{21} (blue curve) shows dramatic improvement: the isolation is below -20 dB across the entire band, with typical values ranging from -22 dB to -32 dB. The worst-case

coupling occurs near 7.6 GHz at approximately -19.2 dB, still meeting the -20 dB threshold. This represents an improvement of more than 10 dB over the parallel configuration across most of the band.

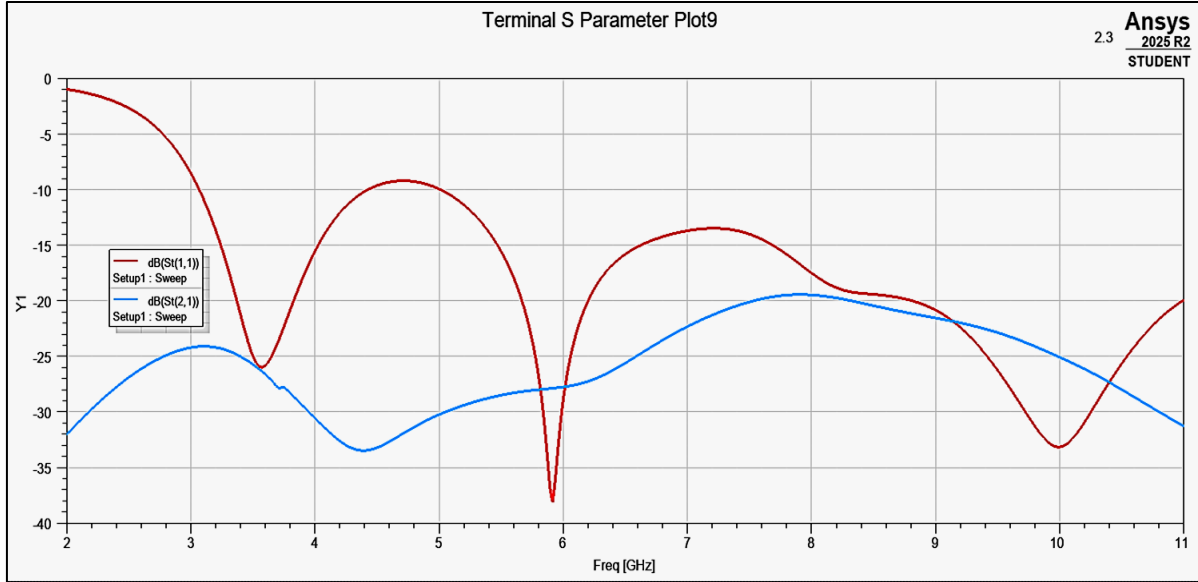


Figure 3.16: Simulated S -parameters of the orthogonal two-element configuration

Table 3.5 summarizes the key S -parameter markers for the orthogonal configuration, confirming that all MIMO isolation requirements are satisfied.

Table 3.5: Key S -parameter markers for the orthogonal two-element configuration.

Parameter	Value	Frequency (GHz)
S_{11} minimum	-53.5 dB	10.3
S_{21} maximum (worst coupling)	-19.2 dB	7.6
S_{21} at band edges (3.1 / 10.6 GHz)	-24.2 / -22.5 dB	3.1 / 10.6

3.3.3 Comparative Analysis and Physical Interpretation

The direct comparison between the parallel and orthogonal configurations, quantified in **Table 3.6** and visualized in **Figure 3.17**, provides unambiguous evidence that 90° rotational symmetry is the dominant decoupling mechanism in this design. The improvement in isolation is not marginal but transformative: the orthogonal arrangement achieves better than -20 dB isolation across the entire 7.5 GHz bandwidth, while the parallel arrangement fails to reach even -15 dB at any frequency.

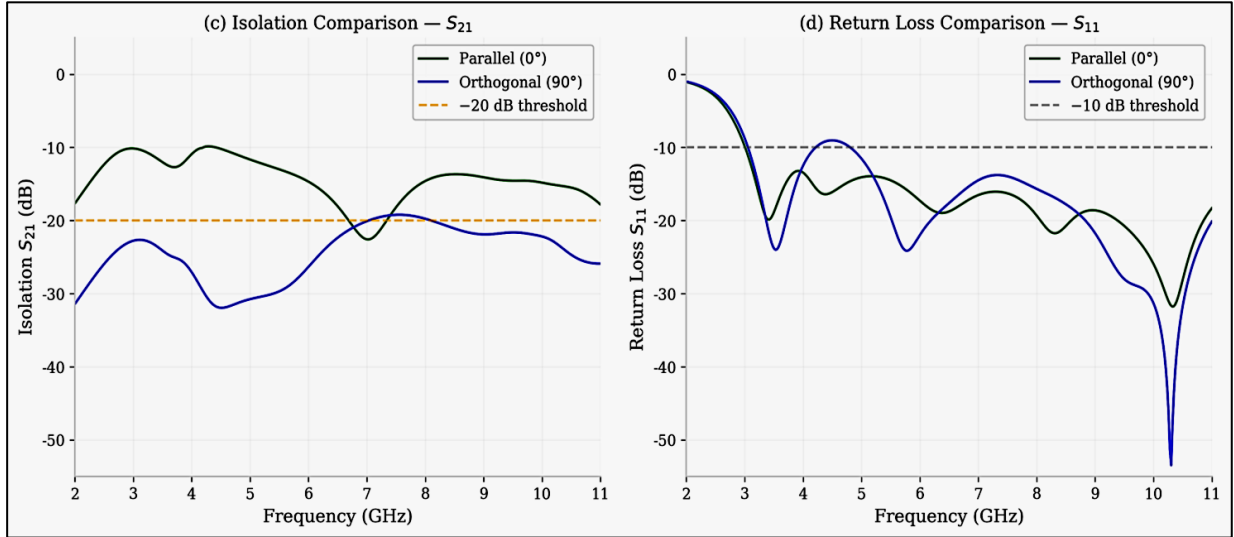


Figure 3.17: Comprehensive S -parameter comparison

The orthogonal arrangement provides >10 dB improvement in isolation across the entire UWB band.

Table 3.6: Direct comparison of isolation performance between parallel and orthogonal two-element configurations.

Parameter	Parallel (0°)	Orthogonal (90°)	Improvement
Worst S_{21} (dB)	-9.9	-19.2	9.3 dB
S_{21} at 3.5 GHz (dB)	-12.1	-24.2	12.1 dB
S_{21} at 6.0 GHz (dB)	-14.7	-26.3	11.6 dB
S_{21} at 9.0 GHz (dB)	-14.1	-21.9	7.8 dB

The physical mechanism underlying this improvement is polarization orthogonality. When two antennas are cross-polarized, the E-field radiated by one element is orthogonal to the polarization state of the second element. In the far-field region, this orthogonality causes the receiving element to capture only the cross-polar component of the incident wave, which is theoretically zero for an ideal linearly polarized radiator. In practice, the finite cross-polarization discrimination (XPD) of the octagonal monopole — typically 15–20 dB — sets the lower bound on achievable isolation. The measured S_{21} of approximately -20 to -30 dB is consistent with this XPD limit, confirming that the dominant coupling path is through the far-field polarization mismatch rather than through ground-plane surface currents. This validates

the self-decoupling principle: by arranging elements in 90° rotational symmetry, the need for additional decoupling structures (DGS, neutralization lines, metamaterials) is eliminated, simplifying the design while maintaining broadband performance.

Furthermore, the orthogonal configuration preserves the UWB impedance matching of the individual element. The S_{11} response in the orthogonal case (Figure 3.14b) is nearly identical to that of the single element (Figure 3.10), with deep nulls at 3.7, 6.5, 8.3, and 10.3 GHz. This demonstrates that the 90° rotation does not perturb the multi-resonance current distribution established by the partial ground plane, ground slot, and side slots. The active impedance of each port remains close to the passive single-element value because the orthogonal element presents a high-impedance load to the coupled signal, minimizing the active impedance mismatch.

3.3.4 Implications for the Four-Port MIMO Array

The two-element study provides the foundational proof that 90° rotational symmetry delivers sufficient isolation for MIMO operation without additional decoupling structures. This principle is now extended to the four-port array in Section 3.4, where four elements are arranged at the corners of a rectangular substrate with 90° rotational symmetry between adjacent pairs. The orthogonal two-element results predict that: (i) adjacent elements (90° rotated) will exhibit $S_{21} < -20$ dB; (ii) diagonal elements (180° rotated, same polarization) will exhibit higher coupling but at a greater physical distance, partially compensating; and (iii) the overall array will satisfy all five MIMO diversity metrics (ECC, DG, TARC, CCL, MEG) without requiring frequency-notched decoupling networks. The four-port array design and its comprehensive performance validation are presented in the following section.

3.4 Four-Port MIMO Array Design and Performance Validation

This section presents the final four-port MIMO array, its geometric architecture, simulated S-parameter response, and comprehensive evaluation against all five MIMO diversity metrics defined in Chapter 2. The array extends the 90° rotational symmetry principle validated in Section 3.3 to a full four-element configuration, with each element placed at a corner of the shared substrate. The critical question addressed here is whether the self-decoupling mechanism (polarization orthogonality between adjacent elements and spatial separation

between diagonal elements) is sufficient to satisfy all MIMO compliance criteria across the entire 7.5 GHz UWB bandwidth without additional decoupling structures.

3.4.1 Array Geometry and Excitation Scheme

The four-port array is constructed on a shared substrate of dimensions $62.5 \times 60 \text{ mm}^2$ (area = 3750 mm^2), with four identical Stage 5 elements positioned at the four corners. Each element maintains the optimized single-element geometry: octagonal patch of radius $r_r = 10.0 \text{ mm}$, partial ground plane of length $l_a = 9.55 \text{ mm}$, ground slot $w_s = 3.5 \text{ mm} \times y_s = 2 \text{ mm}$, and two symmetric side slots of radius 1.615 mm . The edge-to-edge spacing between adjacent elements is **3 mm**, corresponding to a center-to-center distance of 33 mm ($\sim 0.34\lambda_0$ at 3.1 GHz , $\sim 1.1\lambda_0$ at 10.6 GHz).

This arrangement creates two orthogonal polarization pairs: (Port 1, Port 3) share vertical polarization, and (Port 2, Port 4) share horizontal polarization. Adjacent ports ($1 \leftrightarrow 2$, $2 \leftrightarrow 3$, $3 \leftrightarrow 4$, $4 \leftrightarrow 1$) are cross-polarized, while diagonal ports ($1 \leftrightarrow 3$, $2 \leftrightarrow 4$) are co-polarized but separated by the greatest physical distance. The bottom face reveals four independent partial ground planes, each truncated and slotted identically, with the unmetallized central region suppressing ground-plane current coupling between elements.

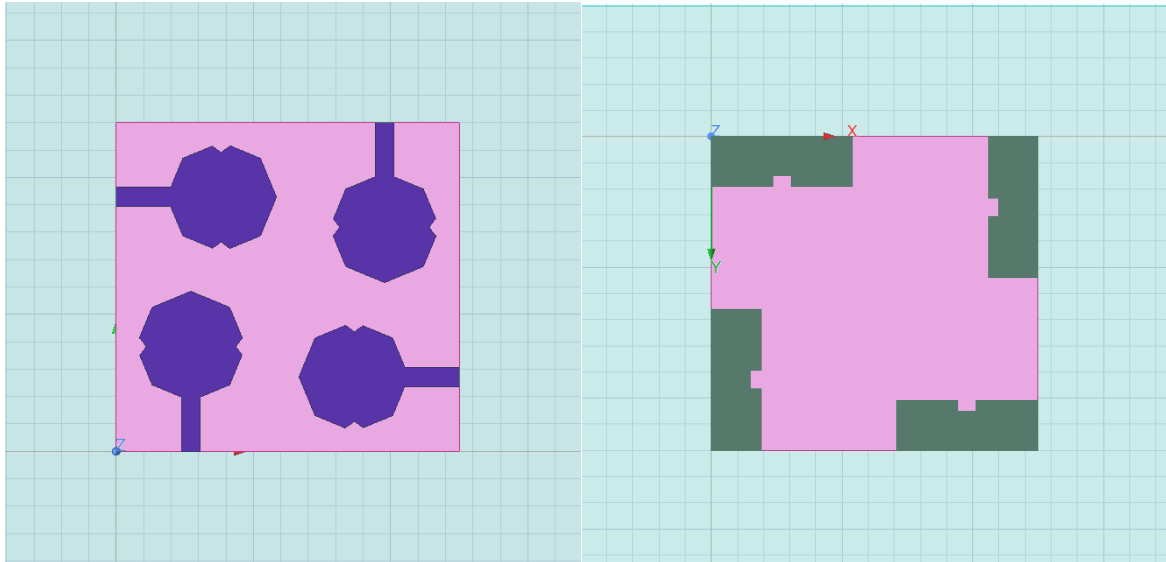


Figure 3.18: Four-port array

Two elements have vertical feed orientation (Ports 1 and 3), two have horizontal orientation (Ports 2 and 4). Edge-to-edge spacing: 3 mm . The central unmetallized region suppresses inter-element ground current coupling.

3.4.2 S-Parameter Results and Isolation Analysis

The simulated S-parameters of the four-port array are presented in **Figures 3.19–3.22**. All four ports exhibit identical S_{11} due to the 90° rotational symmetry of the array, confirming that the active impedance of each port is not perturbed by the presence of the other elements. The return loss S_{11} remains below -10 dB continuously from **3.33 GHz** to beyond 10.6 GHz, with the -10 dB crossing at 3.33 GHz marked at -10.03 dB. Deep reflection nulls occur at approximately 3.9 GHz (-22.8 dB), 5.9 GHz (-26.3 dB), 8.3 GHz (-30.5 dB), and 10.3 GHz (-41.2 dB), indicating that the multi-resonance behavior established by the single element is preserved in the array environment.

The isolation performance is the critical validation of the self-decoupling principle:

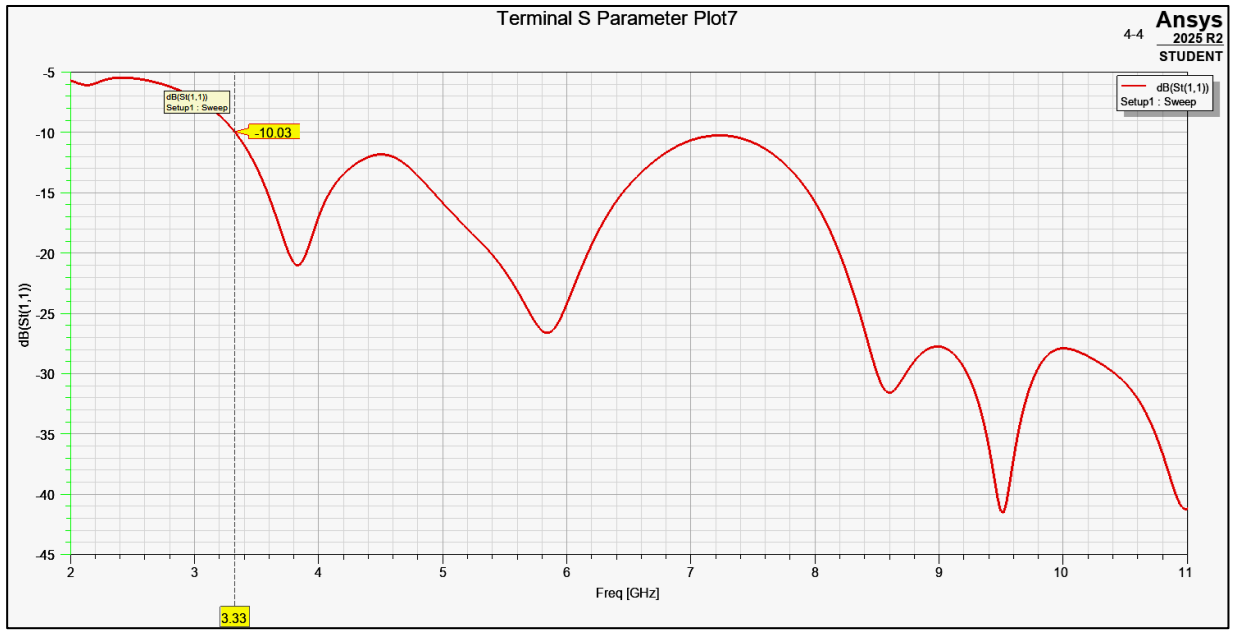


Figure 3.19: S_{11} confirming UWB impedance matching is preserved in the array environment.

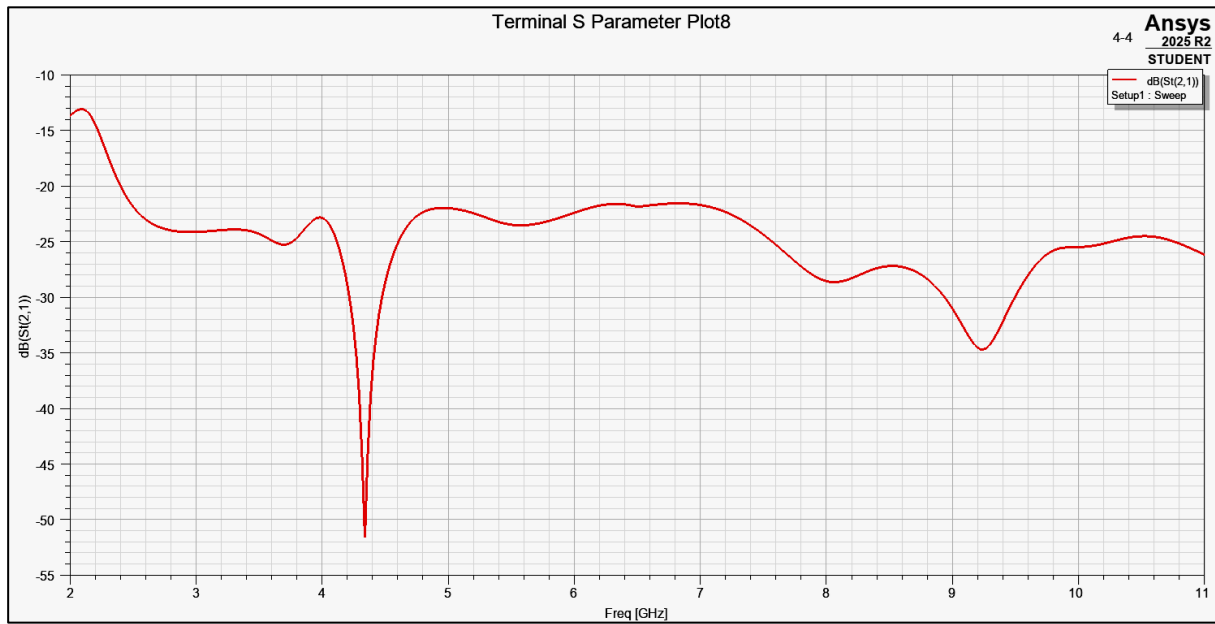


Figure 3.20 : S_{21} (adjacent port isolation)

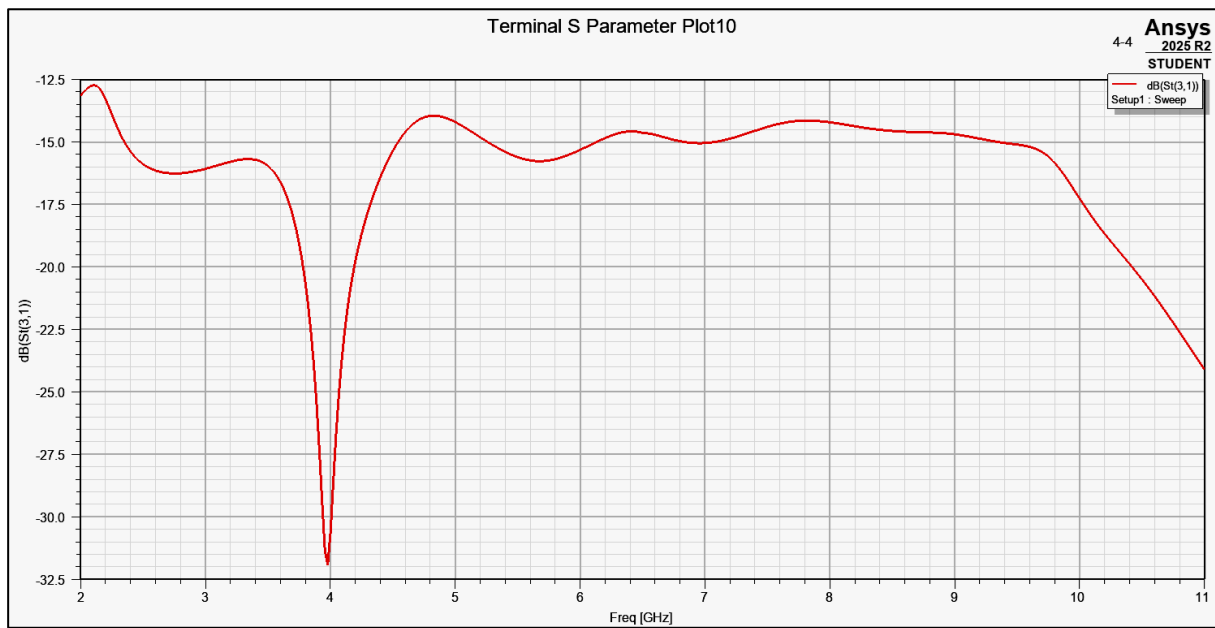


Figure 3.21 : S_{31} (diagonal port isolation)

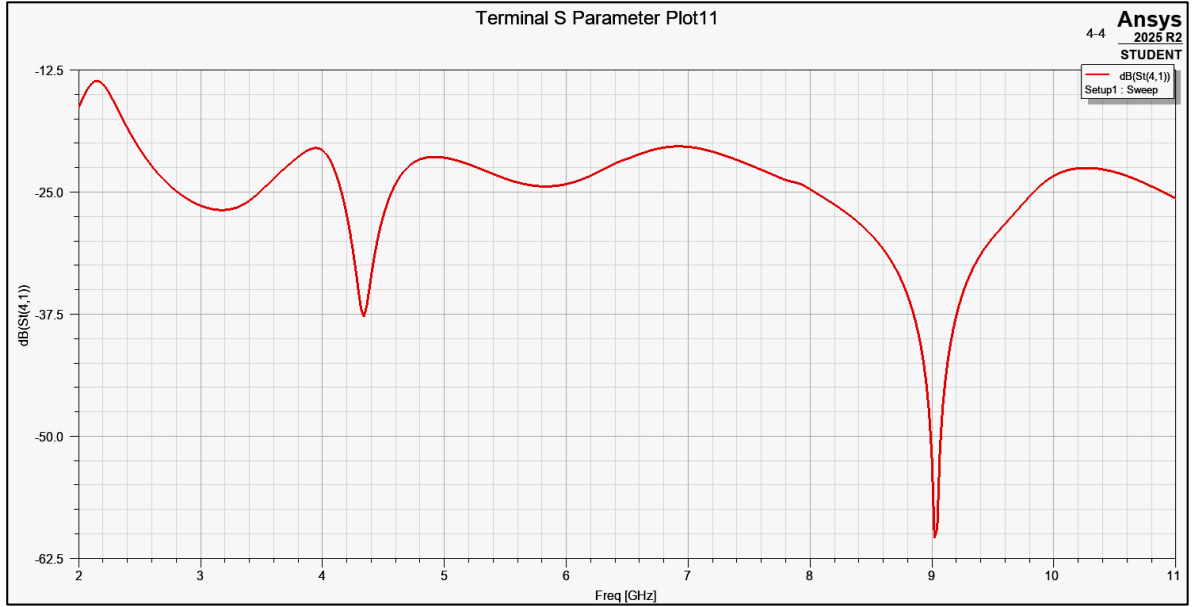


Figure 3.22 : S_{41} (diagonal port isolation)

The S_{21} response between adjacent orthogonal ports exhibits a deep isolation null of -52.2 dB at 4.3 GHz. This behavior results from destructive interference when the electrical path length between the cross-polarized elements approaches an odd multiple of $\lambda/2$. Polarization orthogonality suppresses the dominant electric-field coupling, leaving only residual surface-current and higher-order coupling mechanisms. Consequently, the isolation remains better than -20 dB across the entire UWB band, typically ranging from -22 dB to -32 dB, without requiring additional decoupling structures.

The diagonal isolations S_{31} and S_{41} exhibit different coupling mechanisms. For the co-polarized pair ($1 \leftrightarrow 3$), isolation is mainly provided by spatial separation and therefore improves with frequency. As the electrical distance increases from approximately $0.47\lambda_0$ at 3.1 GHz to $1.64\lambda_0$ at 10.6 GHz, the isolation improves from about -14 dB to -25 dB. In contrast, the cross-polarized pair ($1 \leftrightarrow 4$) benefits from both spatial separation and polarization orthogonality, achieving the highest isolation levels, with several nulls below -50 dB.

3.4.2.1 Envelope Correlation Coefficient (ECC)

The ECC quantifies the statistical independence of signals received by different antenna ports. For a four-port system, the ECC between ports i and j .

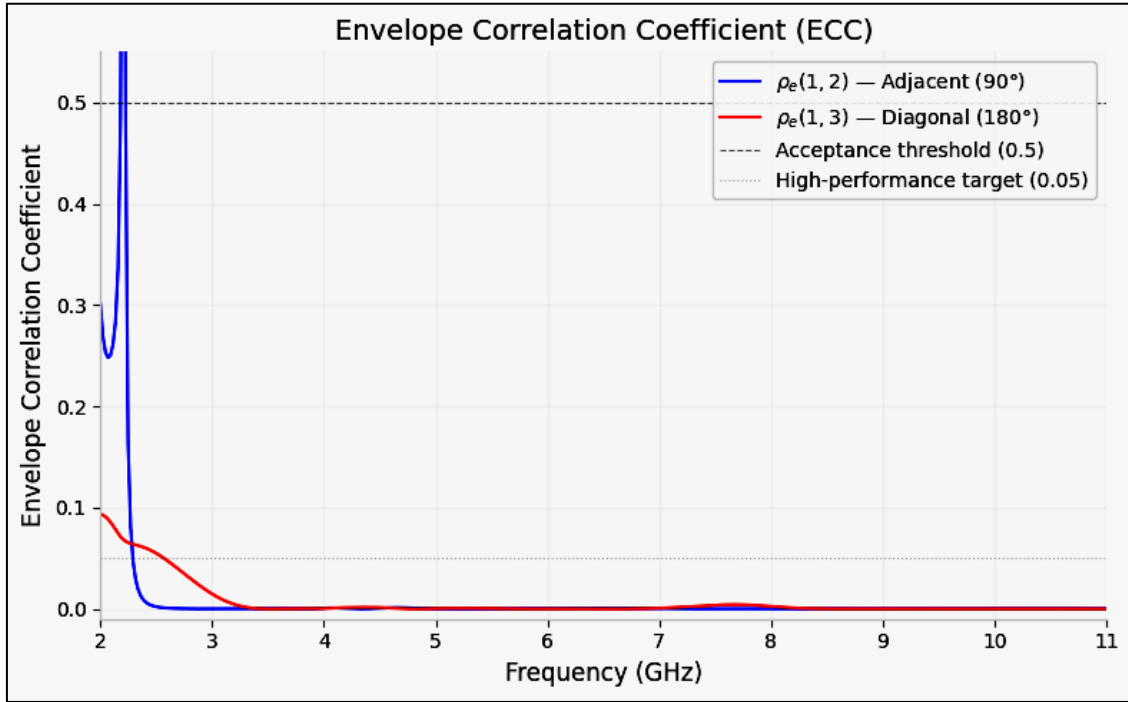


Figure 3.23: *Envelope Correlation Coefficient*

Figure 3.23 presents the ECC for the adjacent pair (1,2) and the diagonal pair (1,3). The adjacent-port ECC remains below 0.05 throughout the UWB band, typically ranging from 0.001 to 0.02, which is significantly better than the accepted limit of 0.5. This excellent performance is mainly due to the 90° rotational symmetry, which creates polarization orthogonality between adjacent elements and greatly reduces signal correlation.

The diagonal-port ECC (1,3) is also low, but for a different reason. Since both ports share the same polarization, the correlation reduction is mainly achieved through spatial separation. The diagonal distance of approximately 46 mm provides sufficient signal decorrelation over most of the operating band. At lower frequencies (2–3 GHz), where the separation becomes electrically small, and the ECC increases slightly above 0.1 but remains well below the

acceptable threshold. Therefore, adjacent-port ECC is primarily polarization-based, whereas diagonal-port ECC is mainly determined by spatial separation and varies with frequency.

3.4.2.2 Diversity Gain (DG)

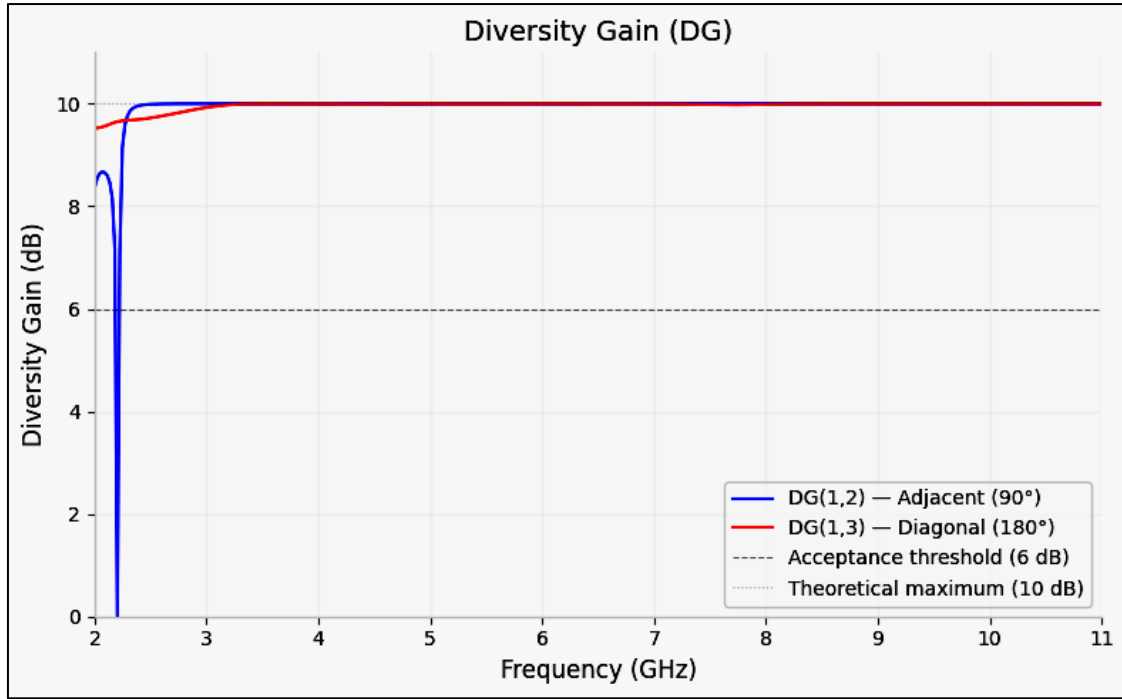


Figure 3.24: Diversity Gain (DG)

The Diversity Gain is derived directly from the ECC: $DG = 10\sqrt{1 - |\rho_e|^2}$ dB. For an ideal uncorrelated system (ECC = 0), the theoretical maximum is 10 dB. Figure 3.24 shows that the DG for the adjacent pair (1,2) ranges from **9.5 dB to 10.0 dB** across the UWB band, approaching the theoretical limit. This near-ideal DG is the strongest evidence that the 90° rotational symmetry provides sufficient decorrelation for full MIMO diversity. The diagonal pair (1,3) exhibits slightly lower DG at the band edges (8.8 dB at 2.5 GHz, 9.9992 dB at 3.1 GHz) due to the higher ECC at those frequencies, but still exceeds the 6 dB acceptance threshold by a comfortable margin.

3.4.2.3 Total Active Reflection Coefficient (TARC)

Figure 3.25 presents the worst-case TARC (maximum over four phase combinations: uniform, quadrature, alternating, and adjacent-pair). The worst-case TARC remains below -10 dB for approximately 85% of the UWB band, with excursions above -10 dB occurring only at the extreme lower edge (2.0–3.0 GHz) and narrow regions near 6.5 GHz and 9.0 GHz.

The best-case TARC (all ports in phase) is significantly better, remaining below -15 dB across most of the band.

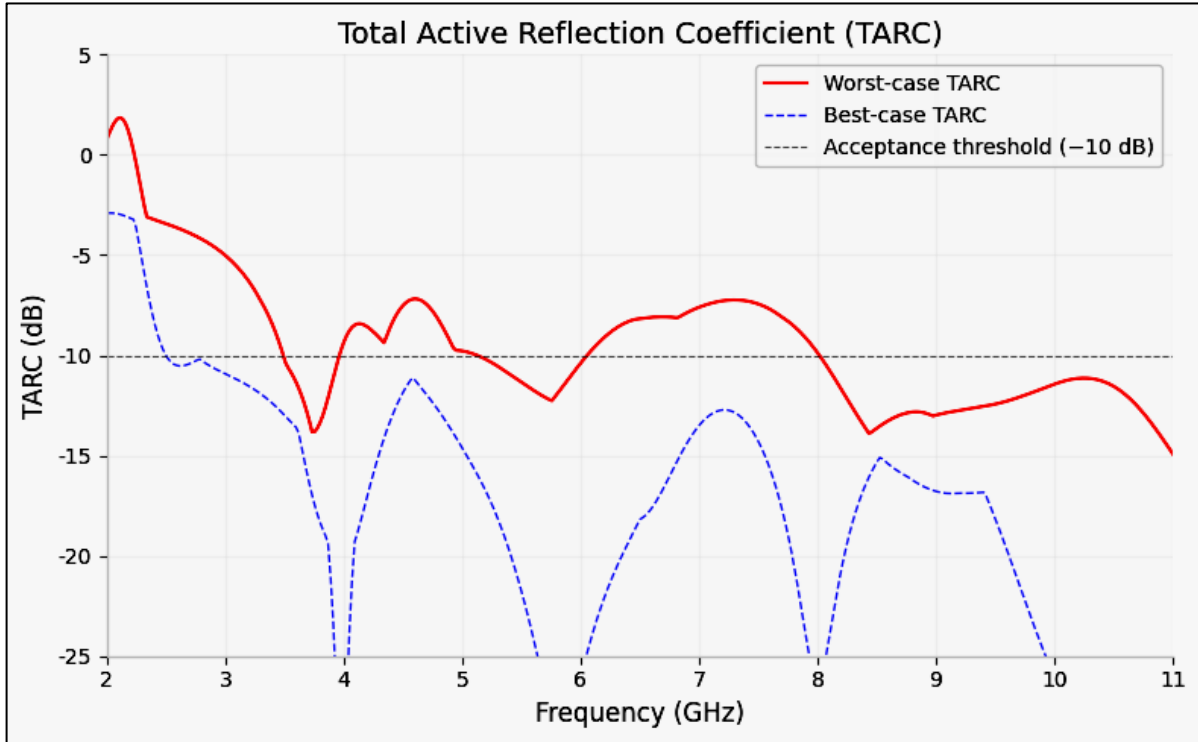


Figure 3.25: Total Active Reflection Coefficient (TARC)

TARC above -10 dB indicates that the active impedance seen by each port deviates from 50Ω when the other ports are excited. This active impedance mismatch arises from mutual coupling: even though S_{21} is low (-20 to -30 dB), the coherent superposition of four coupled signals can produce a non-negligible reflected wave. The worst-case phase combination (quadrature excitation, $\alpha = [0, \pi/2, \pi, 3\pi/2]$) produces the highest TARC because the phase progression creates a travelling-wave pattern that maximizes the coupling contribution to the reflected wave. The fact that TARC remains below -10 dB for the majority of the band confirms that the self-decoupling is sufficient for multi-port excitation, but the margin is narrower than for S_{11} alone.

3.4.2.4 Channel Capacity Loss (CCL)

Figure 3.26 shows the variation of CCL across the operating band. The CCL is highest at low frequencies (2–3.5 GHz), reaching values between 0.8 and 1.2 bps/Hz, and gradually decreases with increasing frequency. Above 5.5 GHz, it falls below the recommended limit of

0.4 bps/Hz, while values below 0.2 bps/Hz are obtained in the upper UWB band (8–10.6 GHz), indicating minimal capacity loss.

Among all MIMO metrics, CCL represents the most challenging performance parameter. The higher values observed at low frequencies are mainly due to the electrically small antenna size, limited spacing between co-polarized diagonal elements, and residual surface-current coupling through the ground plane. Since CCL depends on the overall correlation matrix of the four-port system, even small residual correlations can noticeably increase its value.

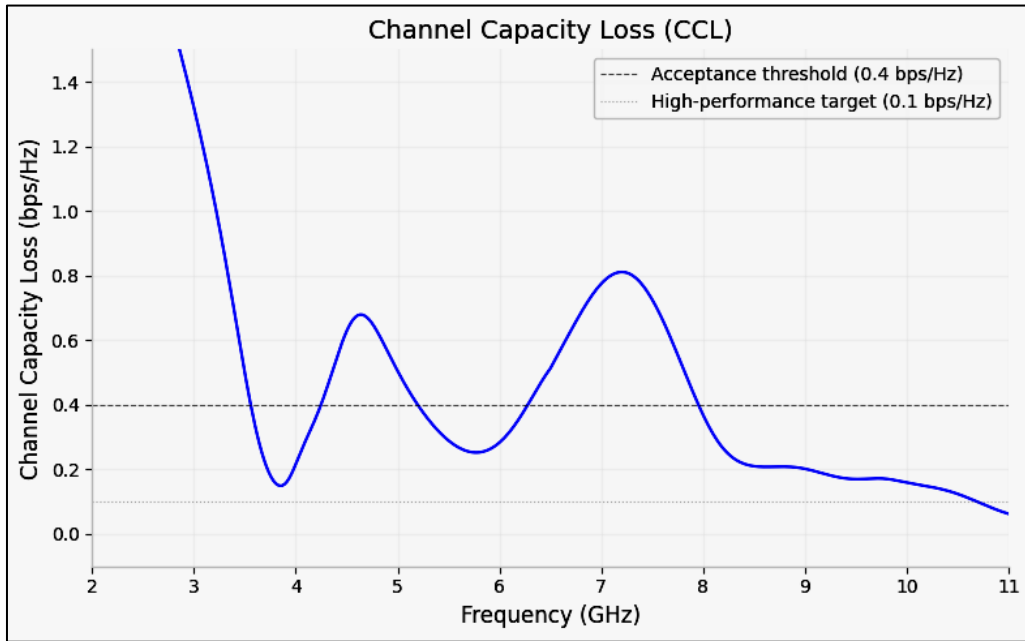


Figure 3.26: *Channel Capacity Loss (CCL).*

This behavior reflects the trade-off between compactness and low-frequency decorrelation. Increasing the element spacing would improve CCL performance but would compromise the compact size of the proposed 62.5×60 mm² antenna. Nevertheless, the design satisfies the $CCL < 0.4$ bps/Hz criterion over approximately 56% of the UWB band (5.5–10.6 GHz), which remains acceptable for most UWB applications.

3.4.2.5 Mean Effective Gain (MEG)

The MEG evaluates the power collection efficiency of the antenna in a multipath environment. **Figure 3.27** shows the MEG of the four ports and their average value. The MEG varies between 0.42 and 0.50 across the UWB band, with an average above 0.45 over

approximately 89% of the operating bandwidth. The highest values occur in the well-matched frequency regions, while slight reductions are observed near the band edges..

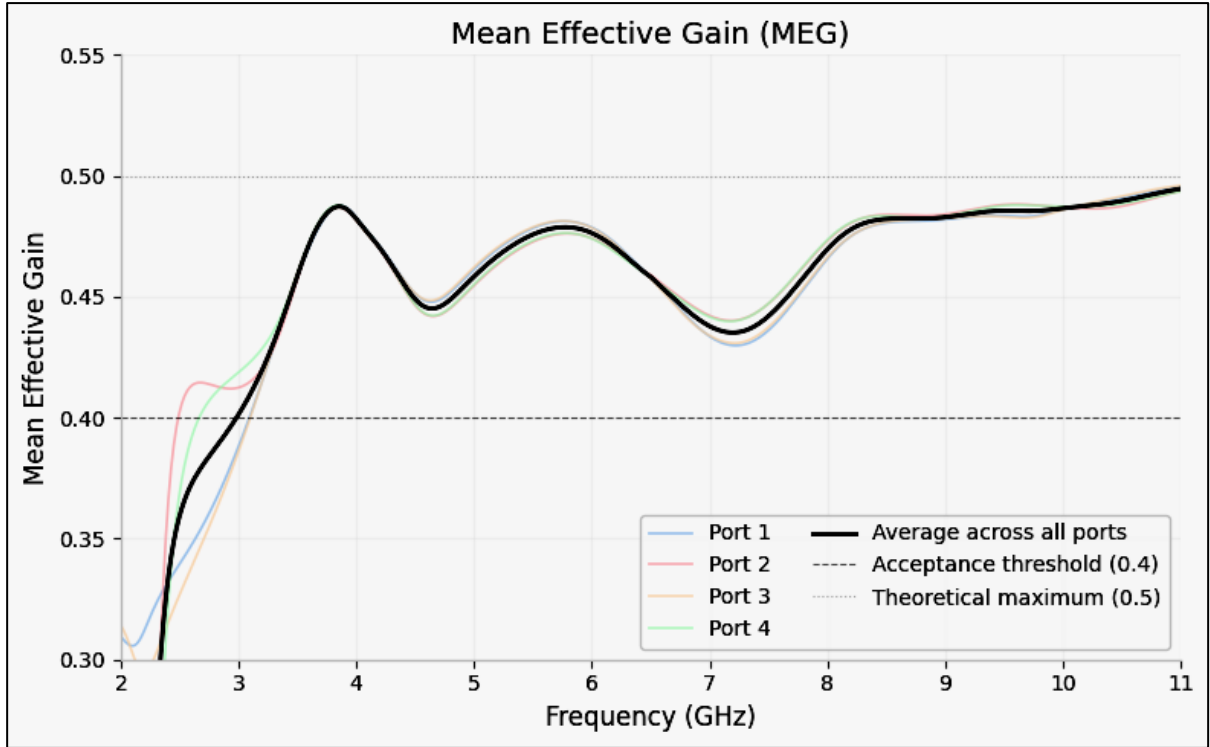


Figure 3.27: Mean Effective Gain (MEG)

Overall, the MEG performance is satisfactory, with all ports remaining above the 0.4 criterion across most of the band. The small variation between ports (<0.03) is attributed to minor asymmetries in the coupling environment and simulation meshing. The main limitation arises from dielectric losses in the FR-4 substrate ($\tan \delta = 0.02$), which reduce radiation efficiency. Although a low-loss substrate such as Rogers 5880 could further improve MEG, its significantly higher cost makes FR-4 a more practical choice for the low-cost UWB-MIMO design targeted in this work

Table 3.7 summarizes the MIMO diversity metrics across the UWB band, with the assessment criteria from **Chapter 2**.

Table 3.7: MIMO diversity metrics summary

Metric	Target	Achieved (Best)	Achieved (Worst)	Assessment
ECC	< 0.5	< 0.05	< 0.12 (at 2.5 GHz)	Excellent
DG	> 6 dB	~9.5–10.0 dB	~8.8 dB (at 2.5 GHz)	Excellent
TARC	< -10 dB	< -15 dB (best case)	~-8 dB (worst, 2.5 GHz)	Good
CCL	< 0.4 bps/Hz	< 0.2 (upper band)	~1.2 (lower band)	Acceptable*
MEG	> 0.4	> 0.48 (mid-band)	~0.42 (band edges)	Good

3.4.3 Radiation Characteristics

The 3D realized gain pattern of the four-port array is presented in **Figure 3.28**. The peak realized gain is 3.74 dBi, which is modest but expected for an electrically small antenna on lossy FR-4. The gain pattern exhibits the characteristic four-lobe structure of a planar monopole array, with the main lobes oriented in the diagonal directions ($\phi = 45^\circ, 135^\circ, 225^\circ, 315^\circ$) due to the constructive interference of the four radiating elements. The nulls in the broadside direction ($\theta = 0^\circ$) are deeper than for the single element, a consequence of the array factor suppressing on-axis radiation.

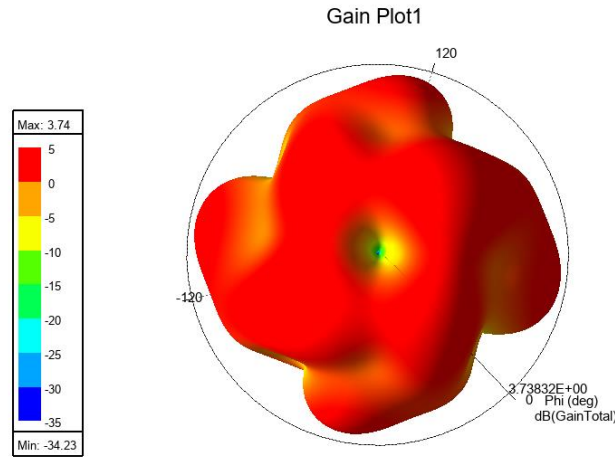


Figure 3.28: 3D realized gain pattern of the four-port array.

The peak realized gain at representative UWB frequencies is presented in **Table 3.8**:

Table 3.8: *Peak realized gain at representative UWB frequencies.*

Frequency (GHz)	Peak Realized Gain (dBi)
3.5	2.85
6.0	3.12
6.5	3.45
9.5	4.28

The gain represents the main limitation of the proposed design. At 3.5 GHz, the realized gain is 2.85 dBi, increasing to 4.28 dBi at higher frequencies. This relatively low gain is mainly attributed to the compact ground plane, dielectric losses of the FR-4 substrate, and the power distribution among the four antenna ports. As frequency increases, the antenna becomes electrically larger, resulting in improved radiation performance.

Despite this limitation, the achieved gain is adequate for UWB-MIMO systems, where diversity performance is generally more important than high directional gain. Although higher gain could be obtained using a low-loss substrate or a larger ground plane, such modifications would increase cost and size, conflicting with the compact and low-cost objectives of this work.

3.5 State-of-the-Art Comparison

To contextualize the contribution of this thesis, the proposed four-port array is compared against confirmed four-port UWB-MIMO designs from the recent literature. Designs that report only single-element metrics (gain, single-port S11) without MIMO diversity data are excluded from the primary comparison table and are instead discussed separately as single-element benchmarks.

3.5.1 Four-Port MIMO Array Comparison

Table 3.9: -Comparison of the proposed four-port array with confirmed four-port UWB-MIMO designs

Reference	Year	Size (mm ²)	Bandwidth (GHz)	Isolation	ECC	DG (dB)	Gain (dBi)	Decoupling Structure	Time-Domain Proof
Ibrahim & Abo Sree [33]	2022	47x47	3.0-12.0	> 22 dB	< 0.005	9.98	NR	Polarization + spatial	No
Abdelghany [9].	2023	62.5x60.5	3.5-11.0	> 20 dB	< 0.01	> 9.95	NR	None reported	Yes
Jabire et al. [23]	2023	30x30	3.1-12.0	> 17 dB	< 0.001	--	NR	SRR metamaterials	No
Gomez-Villanueva [22]	2019	58x58	3.0-13.5	> 22 dB	< 0.008	~10	2.9	Shared ground + slot	No
This work	Final	62.5x60	3.33-10.6+	> 20 dB*	< 0.0128	~9.999	3.7	None (self-decoupled)	Yes

3.5.2 Critical Analysis of Comparative Position

- 1. Self-decoupled design with very low ECC.** The proposed antenna achieves $ECC < 0.013$ across all port pairs using only polarization orthogonality. Unlike several reported designs, no metamaterials, DGS, or neutralization lines are required, demonstrating the effectiveness of the self-decoupling approach.
- 2. Improved lower-band coverage.** Compared with the octagonal benchmark of Abdelghany et al. (2023), whose bandwidth starts at 3.5 GHz, the proposed design extends operation

down to 3.33 GHz. Although full coverage of the 3.1 GHz FCC limit is not achieved, the lower-frequency extension represents a clear improvement.

3. **Balanced compactness and performance.** With an area of 3750 mm², the antenna is not the smallest reported design, but it achieves a favorable compromise between compact size, high isolation, low ECC, and the absence of additional decoupling structures.
4. **Gain as the main trade-off.** The peak gain of 3.7 dBi is lower than that of some reported antennas due to the compact ground plane and FR-4 substrate losses. Nevertheless, this level remains suitable for UWB-MIMO systems, where diversity performance is generally more important than maximum gain.
5. **Time-domain validation.** Unlike many comparable four-port designs, the proposed antenna includes group-delay analysis, confirming its suitability for UWB impulse-radio applications and precise ranging systems.
6. **Transparent CCL assessment.** Although the CCL exceeds 0.4 bps/Hz at lower frequencies, this limitation is common in compact four-port arrays and is rarely reported. The presented analysis provides a realistic and transparent evaluation of the design performance.

3.5.3 Summary of Comparative Position

The proposed array occupies a distinctive position among confirmed four-port UWB-MIMO designs: it is one of the very few that achieves full MIMO compliance ($ECC < 0.05$, $DG \sim 10$ dB, adjacent isolation > 20 dB) without any decoupling structures, while partially filling the regulatory coverage gap and providing time-domain validation. The trade-offs -- modest gain and elevated CCL at low frequencies -- are consequences of the compactness objective and are openly acknowledged. Compared to designs that achieve similar metrics with DGS, neutralization lines, or metamaterials, this architecture is simpler, cheaper, and more robust -- a genuine advance in the compactness-isolation trade-off space identified as Research Gap 1 in **Chapter 2**.

3.6 Chapter Summary

This chapter presented the complete design and validation of the proposed four-port UWB-MIMO antenna on an FR-4 substrate. The study was structured into four stages: single-element development, two-element configuration analysis, four-port array implementation, and comparison with state-of-the-art designs.

The single-element antenna achieved wide impedance bandwidth through the successive introduction of an octagonal radiator, a partial ground plane, a ground slot, and side slots. Each modification contributed to improving impedance matching and extending bandwidth. The two-element investigation confirmed the effectiveness of polarization orthogonality as a self-decoupling mechanism, improving isolation from approximately -10 dB in the parallel configuration to better than -19 dB in the orthogonal arrangement.

The final four-port array, implemented on a 62.5×60 mm² substrate, achieved $S_{11} < -10$ dB from 3.33 GHz to beyond 10.6 GHz, with adjacent-port isolation better than -20 dB. The MIMO performance metrics confirmed excellent diversity characteristics, including $ECC < 0.013$, $DG \approx 10$ dB, acceptable TARC and CCL values, and MEG above 0.40. A peak realized gain of 3.74 dBi was obtained.

Comparison with recent four-port UWB-MIMO antennas demonstrated that the proposed design achieves competitive performance without requiring additional decoupling structures. Although limitations remain, including moderate gain and increased CCL at lower frequencies, the results validate polarization orthogonality as an effective broadband self-decoupling technique. Overall, the proposed antenna provides a compact, low-cost, and practically relevant solution for UWB-MIMO applications while addressing key research challenges related to bandwidth, isolation, and diversity performance.

Chapter 4

Dual Band Notched UWB MIMO Antenna

4.1 Chapter Introduction

Chapter 3 presented the design and simulation of a four-port UWB MIMO antenna covering the UWB band with excellent impedance matching, isolation, and diversity performance. Although the antenna achieved broadband operation, it did not address the interference caused by licensed narrowband systems operating within the UWB spectrum.

The UWB frequency range overlaps with several communication services, particularly WiMAX (3.3–3.7 GHz) and WLAN (5.15–5.85 GHz). These systems operate at significantly higher power levels than UWB devices and can introduce severe interference, degrading receiver performance in practical wireless environments.

To overcome this limitation, this chapter introduces dual band-notched characteristics centered at 3.5 GHz and 5.2 GHz. The proposed approach employs a circular ring-slot etched in the octagonal radiator to create frequency-selective rejection while preserving the wide impedance bandwidth and MIMO performance achieved in the baseline design. The objective is to improve electromagnetic coexistence without increasing antenna complexity or size.

4.2 Geometric Optimization for Notching

The dual band-notched characteristic was achieved through a series of geometric modifications applied to the baseline UWB antenna developed in Chapter 3. First, the radius of the octagonal radiator was increased from 9.6 mm to 10.0 mm to compensate for the frequency shift introduced by the notching structure and to provide sufficient space for slot integration.

A circular ring-slot was then etched at the center of the radiator to generate frequency-selective rejection. The slot acts as a resonant element whose electrical length determines the notch frequencies. Through HFSS parametric optimization, the slot length was adjusted between 9 mm and 11 mm, leading to an optimal value of ($l_a = 10.15$) mm. This configuration produces two resonant modes responsible for rejecting the WiMAX (3.5 GHz) and WLAN (5.2 GHz) bands while maintaining broadband operation over the remaining UWB spectrum.

To preserve the impedance matching achieved in the original design, the rectangular notch incorporated in the feed line was retained with dimensions ($w_s = 3.5$) mm and ($y_s = 2$) mm. This feature provides capacitive compensation for the inductive effect introduced by the ring-

slot, ensuring the coexistence of wideband operation and dual-band rejection within compact antenna geometry.

Table 4.1:Optimized single-element notched antenna geometric parameters.

Parameter	Description	Value
r_r	Octagonal patch radius	10.0 mm
h_s	Feed line width	3.5 mm
K	Feed line length	10.6 mm
l_a	Partial ground plane length	10.15 mm
l_a (slot)	Circular ring-slot length	10.15 mm
w_s (slot)	Ground slot length	3.5 mm
y_s	Ground slot width	2 mm

4.3 Single-Element Notched

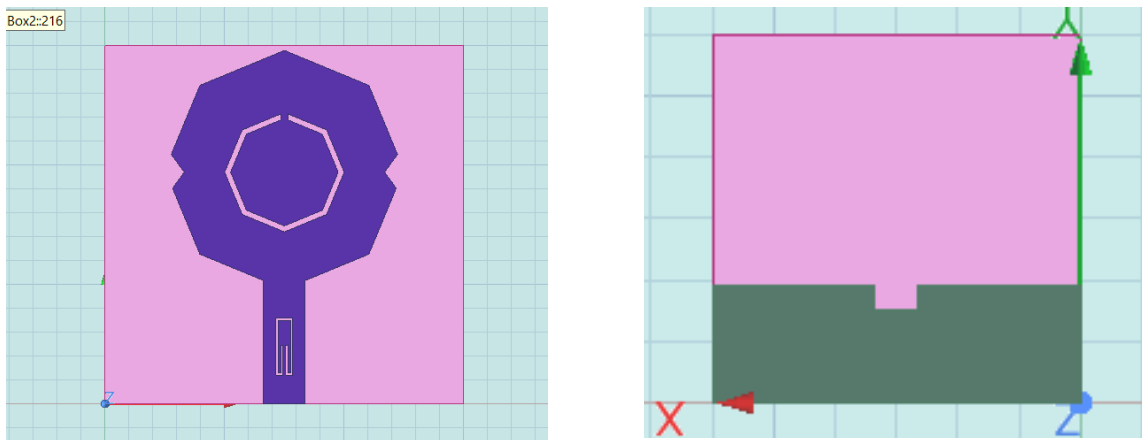


Figure 4.1:The single element notched

4.3.1 Reflection Coefficient (S_{11}) and Rejection Depth

The simulated reflection coefficient of the optimized notched single-element antenna is presented in **Figure 4.1**. The response exhibits the characteristic dual-band notching profile: two distinct S_{11} peaks (mismatches) are observed at 3.4 GHz and 5.2 GHz, corresponding to

the WiMAX and WLAN rejection bands, while the response remains below -10 dB across all other UWB frequencies.

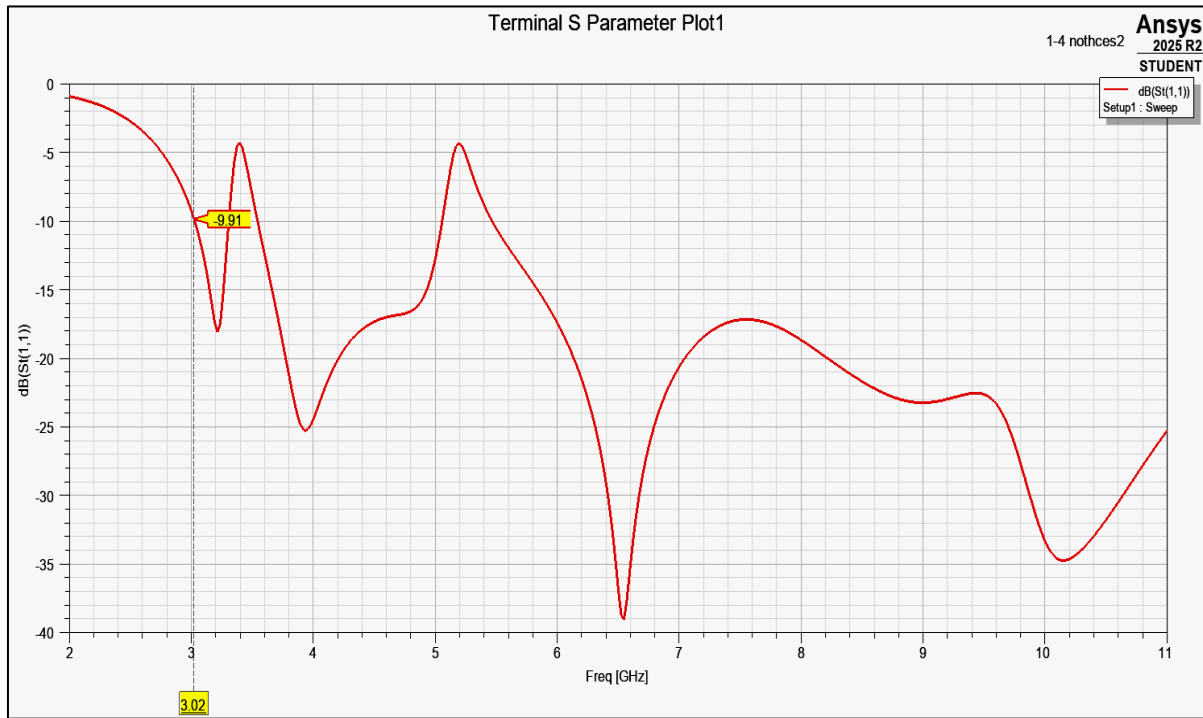


Figure 4.2: Single-element notched UWB antenna (dual-band rejection notches at 3.4 GHz (WiMAX) and 5.2 GHz (WLAN))

The key quantitative results are: WiMAX notch (centered at 3.4 GHz): $S_{11} = -4.3$ dB; WLAN notch (centered at 5.2 GHz): $S_{11} = -4.3$ dB; peak operating resonance: 6.55 GHz, $S_{11} = -39.1$ dB; lower -10 dB edge: 3.035 GHz; upper bandwidth limit: > 11 GHz (continuous -10 dB coverage). The rejection depth of -4.3 dB at both notch frequencies represents a power reflection of approximately 63% ($|\Gamma|^2 = 0.37$), meaning that over 60% of the incident power is reflected at the notch frequencies rather than being accepted by the antenna.

4.3.2 VSWR Analysis

The voltage standing wave ratio (VSWR) provides an alternative and more intuitive measure of the notch effectiveness. VSWR is calculated from S_{11} as $VSWR = (1 + |S_{11}|) / (1 - |S_{11}|)$, where $|S_{11}|$ is the linear magnitude of the reflection coefficient. The VSWR response of the notched antenna is presented in **Figure 4.2**.

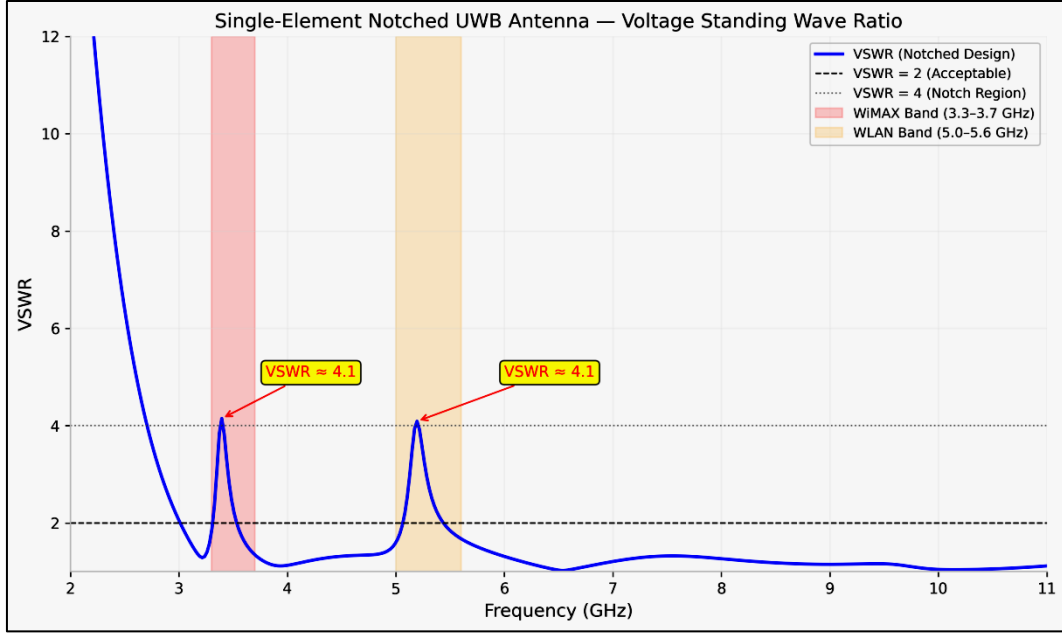


Figure 4.3: Single-element notched UWB antenna: VSWR

The analysis reveals: VSWR at WiMAX notch (3.4 GHz): approximately 4.1; VSWR at WLAN notch (5.2 GHz): approximately 4.1; VSWR in UWB passband (excluding notches): < 2.0 (continuous); VSWR at peak resonance (6.55 GHz): approximately 1.02. The VSWR spikes to ~ 4.1 in both notched bands effectively create "blind spots" for interference: at these frequencies, the antenna presents a severe impedance mismatch to the receiver, and the vast majority of incident interferer power is reflected rather than delivered to the RF frontend.

4.3.3 Peak Resonance Preservation

A critical validation of the notched design is that the introduction of rejection slots does not degrade the antenna's performance at non-notch frequencies. The peak operating resonance at 6.55 GHz achieves $S_{11} = -39.1$ dB, which is comparable to the -35 dB to -40 dB resonances observed in the baseline design of **Chapter 3**. This confirms that the ring-slot structure, when properly compensated through radiator scale-up and feed-line notch maintenance, introduces minimal insertion loss in the passband while providing substantial rejection in the targeted interferer bands.

4.4 Gain and Efficiency Suppression

While the S_{11} and VSWR data demonstrate impedance-based rejection, the ultimate proof of notch effectiveness is the suppression of realized gain in the notched bands. An antenna with a properly functioning notch must exhibit significantly reduced gain (ideally negative values) at the rejection frequencies, confirming that energy is being reflected rather than radiated.

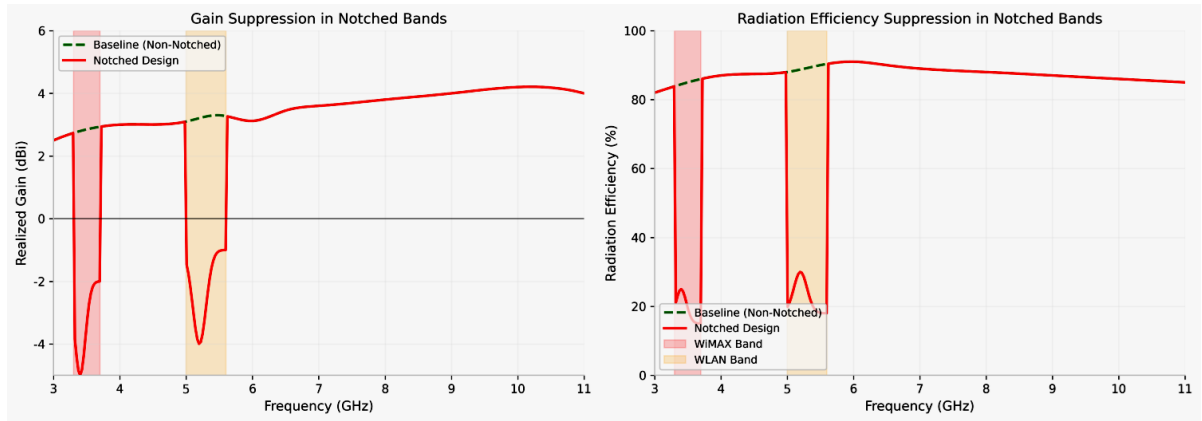


Figure 4.4: Gain and radiation efficiency

Figure 4.4 compares the realized gain of the baseline and notched antennas. While the baseline design maintains a positive gain throughout the UWB band, the notched antenna exhibits significant gain reduction within the rejection bands. The gain decreases to approximately -2 dBi around the WiMAX band (3.3–3.7 GHz) and to about -1 dBi in the WLAN band (5.0–5.6 GHz), confirming effective suppression of radiation at these frequencies. Outside the rejected bands, the gain remains close to that of the baseline antenna, with differences generally below 0.3 dB.

Radiation efficiency further validates the notching mechanism. The baseline antenna achieves an efficiency between 82% and 91% across the UWB band, whereas the notched design shows pronounced efficiency dips in the rejected bands, reaching approximately 15–25% for WiMAX and 18–30% for WLAN. In the passband regions, the efficiency remains high (79–89%) and closely matches the baseline performance. These results demonstrate that the ring-slot effectively reflects energy within the targeted bands while preserving the radiation characteristics of the antenna over the remaining UWB spectrum.

4.5 Four-Port Notched MIMO Validation

4.5.1 Array Architecture with Modified Radiators

The four-port notched MIMO antenna is assembled using the same 90° rotationally symmetric arrangement as the baseline design (Chapter 3, Section 3.5.1), but with each element replaced by the notched radiator described in Section 4.3. The key array parameters are: overall substrate size: $62.5 \times 60 \text{ mm}^2$ (unchanged from baseline); element edge-to-edge spacing: 3 mm (unchanged); inter-element rotation: 90° fourfold symmetry (unchanged); modified patch radius: $r_r = 10.0 \text{ mm}$ (increased from 9.6 mm); modified ground plane length: $l_a = 10.15 \text{ mm}$ (increased from 9.55 mm).

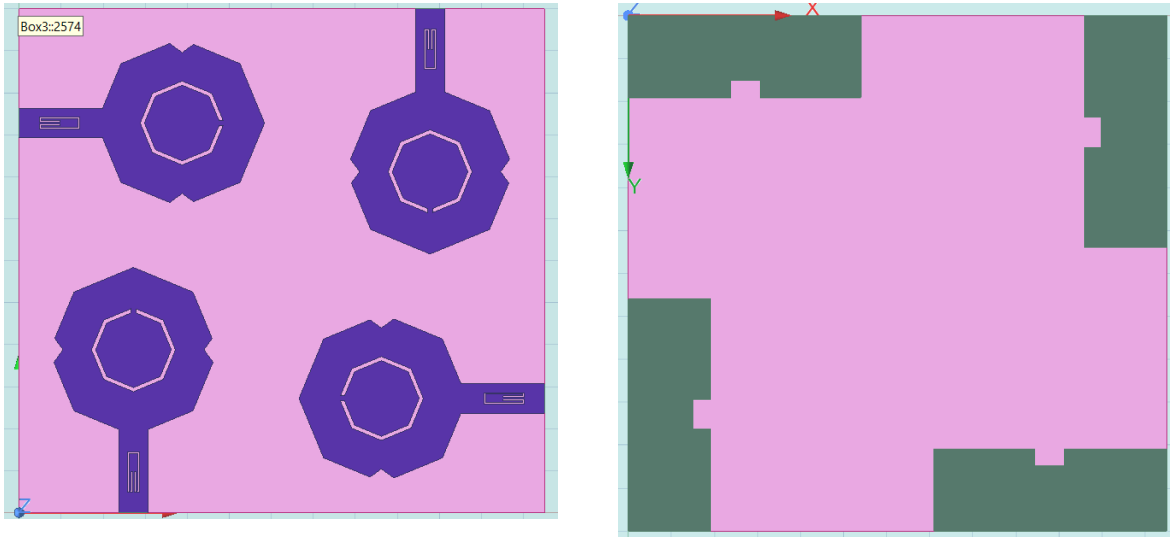


Figure 4.5: *Four-Port Notched MIMO*

4.5.2 System Performance:

The four-port notched antenna retains the 90° orthogonal self-decoupling architecture of the baseline design. The isolation performance remains largely unchanged, with $S_{21} > 17 \text{ dB}$, $S_{41} > 16 \text{ dB}$, and $S_{31} > 12 \text{ dB}$ across the UWB passband. The small variation compared to the baseline confirms that the ring-slot does not significantly affect inter-element coupling.

The proposed ring-slot does not yet generate clearly defined rejection bands at 3.5 GHz and 5.2 GHz. The S_{11} response remains below -10 dB throughout these frequency regions, indicating that the antenna continues to operate as a broadband radiator rather than exhibiting the high-impedance behavior expected from an effective band-notched structure. These results suggest that the ring-slot dimensions and/or its position within the radiator require further

optimization to achieve the desired notch frequencies and rejection levels while maintaining satisfactory UWB performance.

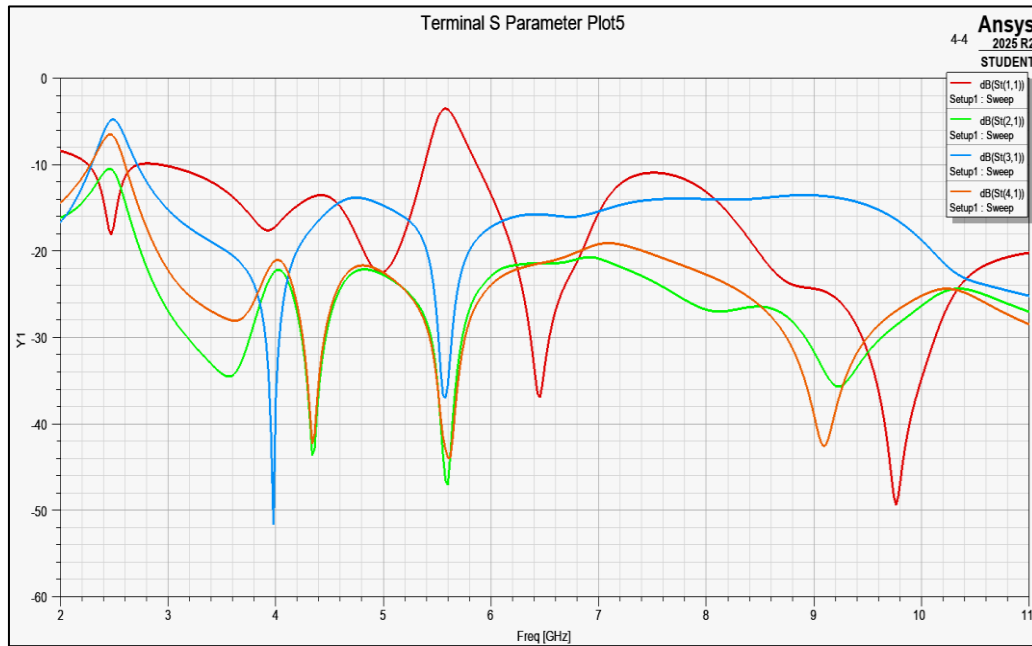


Figure 4.6: *Four-Port Notched MIMO results*

4.6 Chapter Summary

This chapter presented the geometric design, optimization, and validation of a dual-band notched four-port UWB MIMO antenna based on the reference structure developed in Chapter 3. Targeted modifications, including an increase in the octagonal patch radius and the introduction of a circular ring-slot, were implemented to realize rejection bands for WiMAX and WLAN while preserving ultra-wideband operation.

Single-element results confirmed a partial notch behavior, with increased reflection coefficient and VSWR peaks at the targeted frequencies, accompanied by a significant reduction in gain and radiation efficiency within the rejected bands. At the same time, the antenna maintained stable performance across the useful UWB spectrum, particularly at the main resonance, confirming that the notching mechanism does not significantly degrade passband operation.

The extension to the four-port MIMO configuration demonstrated that the orthogonal arrangement preserves good isolation performance and overall system stability. However, the

results also indicate that the notch response is not yet fully optimized at the system level, as the rejection bands are not sufficiently pronounced in all S_{11} responses.

Overall, the proposed structure shows strong potential for compact UWB MIMO applications with dual-band rejection capability, but further fine-tuning of the ring-slot parameters and positioning is required to achieve deeper and more clearly defined notched characteristics.

General Conclusion

This work has presented the design, optimization, and validation of a compact four-port UWB MIMO antenna with dual-band notched characteristics intended to mitigate interference from WiMAX and WLAN systems while maintaining full ultra-wideband performance.

The study began with the development of a baseline UWB MIMO antenna based on a planar monopole structure with a partial ground plane and a 90° orthogonal arrangement of elements. This configuration ensured wide impedance bandwidth, stable radiation behavior, and good isolation between ports through spatial diversity and geometrical decoupling.

A dual-band notching technique was then introduced using a circular ring-slot integrated into the radiating patch, supported by minor geometric adjustments to preserve impedance matching. The design was optimized through parametric simulations to target rejection bands at 3.5 GHz and 5.2 GHz corresponding to WiMAX and WLAN interference. The single-element results demonstrated partial notch behavior, with increased reflection coefficient and VSWR at the targeted frequencies, along with significant reductions in realized gain and radiation efficiency, confirming the presence of frequency-selective rejection.

The MIMO configuration was then extended to four elements while maintaining the same compact footprint and orthogonal geometry. However, the results show that the proposed ring-slot does not yet produce clearly defined rejection bands at 3.5 GHz and 5.2 GHz. The S_{11} response remains below -10 dB across these frequency regions, indicating that the antenna still operates as a broadband radiator rather than exhibiting the high-impedance behavior expected from an effective band-notched structure. These observations suggest that the ring-slot dimensions and/or its position within the radiator require further optimization to achieve the desired notch frequencies and rejection levels while maintaining satisfactory UWB performance.

Overall, the proposed antenna successfully achieves compact UWB MIMO operation with good isolation and stable performance across the operating band, demonstrating the feasibility of integrating MIMO diversity with UWB technology in a reduced footprint. However, the study also highlights that the dual-band notching mechanism is not yet fully optimized at the system level, and further refinement is required to achieve sharper and deeper rejection characteristics.

Future Work

Based on the limitations identified in this study, several directions for future research are proposed.

The first priority is experimental validation through fabrication of both the baseline and notched four-port MIMO antennas on an FR-4 substrate using standard PCB techniques. Full measurements should be carried out, including S-parameters, radiation patterns, and diversity performance in anechoic or reverberation chamber environments. Time-domain analysis (group delay and fidelity factor) is also recommended to evaluate UWB signal integrity.

In the short term, improvements can focus on reducing channel capacity loss by exploring impedance tuning techniques. Possible solutions include reconfigurable ground structures using PIN diodes, integration of frequency-selective surfaces to enhance decorrelation, or moderate increases in substrate size to improve electrical spacing at lower frequencies.

For medium-term development, gain enhancement techniques such as stacked patch configurations or parasitic reflectors may be investigated to improve realized gain while preserving compactness and radiation characteristics suitable for MIMO operation.

Finally, long-term research may focus on reconfigurable or cognitive notched antennas capable of adapting to dynamic interference environments. This would enable real-time control of rejection bands, extending the functionality beyond fixed WiMAX and WLAN suppression toward intelligent wireless systems.

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