

**HYBRID BEAMFORMING TECHNIQUES FOR MILLIMETER-WAVE
MIMO SYSTEMS**

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HYBRID BEAMFORMING TECHNIQUES FOR MILLIMETER-WAVE MIMO SYSTEMS

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CONTENTS

	<u>Page</u>
ACKNOWLEDGEMENT.....	iv
CONTENTS	v
LIST OF FIGURES	vii
LIST OF TABLES	x
LIST OF SYMBOLS	xi
ABBREVIATIONS.....	xiii
ÖZET	xv
ABSTRACT.....	xvii
1. INTRODUCTION	1
1.1. Thesis Contributions	2
2. THEORETICAL PART.....	6
2.1. Introduction To MIMO and XL-MIMO Systems.....	6
2.1.1. Beamspace MIMO channel	8
2.1.2. MIMO design issues	10
2.1.2.1. Spatial-wideband effect	10
2.1.2.2. Near-field propagation.....	11
2.2. Study I: Lens Antenna Subarray Design.....	12
2.3. Study II & III: Spatial-Wideband Effect in Wideband Massive MIMO	14
2.4. Physical Layer Authentication (PLA) in Beamspace MIMO	17
2.4.1. Study IV: PLA scheme in beamspace MIMO systems.....	19
2.4.2. Study V: Robust tracking-based PLA in MIMO systems.....	19
3. EXPERIMENTAL PART	21
3.1. Study I: Lens Antenna Subarray Design in Multi-User MIMO Systems	21
3.1.1. System model	22
3.1.2. Problem formulation	24
3.1.3. Proposed sub-grouped LAS transceiver design	25
3.1.3.1. Transceiver design	26
3.1.3.2. Proposed sub-grouped-based precoding	26
3.1.3.3. Complexity evaluation.....	28
3.1.3.4. Energy efficiency	30
3.2. Study II: Spatial-Wideband Effect in UWB Massive MIMO.....	30
3.2.1. System model and problem analysis	30
3.2.2. The proposed LAS-based system design	34
3.2.2.1. Proposed transceiver design.....	34
3.2.2.2. Proposed threshold-based precoding	39
3.2.2.3. Extension to multiple users scenario	41
3.2.3. System analysis	42
3.2.3.1. Capacity	42
3.2.3.2. Power consumption.....	43
3.2.3.3. Practical implementation	46
3.3. Study III: Spatial-Wideband Effect in UWB XL-MIMO Systems.....	47
3.3.1. System model and problem formula	47
3.3.2. Proposed beam gain analysis under beam squint effect	48
3.3.2.1. Beam gain in far-field propagation.....	49
3.3.2.2. Beam gain in near-field propagation	49
3.3.2.3. Beam squint interference region	51
3.3.3. Proposed inter-beam interference analysis	54

3.3.3.1.	Inter-beam interference	54
3.3.3.2.	System capacity	55
3.3.4.	Proposed beam scheduling technique for IBI mitigation.....	58
3.3.4.1.	Proposed beam scheduling technique	58
3.3.4.2.	Complexity evaluation	61
3.4.	Study IV: PLA in BeamSpace MIMO.....	61
3.4.1.	System model and assumptions	61
3.4.2.	Proposed PLA scheme	63
3.4.2.1.	Proposed distance signature.....	63
3.4.2.2.	Proposed hardware signature	64
3.4.2.3.	Proposed hybrid signature	65
3.4.2.4.	Detection model	66
3.5.	Study V: Tracking-Based PLA in BeamSpace Massive MIMO	68
3.5.1.	System model and assumptions	68
3.5.2.	Proposed narrowband tracking-based PLA approach.....	70
3.5.2.1.	ED signature extraction in narrowband transmission	70
3.5.2.2.	Detection model	72
3.5.2.3.	Tracking process for the ED signature in narrowband transmission..	73
3.5.3.	Proposed wideband tracking-based PLA approach	76
3.5.3.1.	Beam squint-based ED signature in wideband transmission	77
3.5.3.2.	Tracking process for the ED signature in wideband transmission.....	79
4.	RESULTS AND DISCUSSION	81
4.1.	Results of Study I.....	81
4.2.	Results of Study II	87
4.2.1.	Proposed precoder design performance	88
4.2.2.	Energy efficiency performance.....	90
4.2.3.	Impact of the design parameters selection	92
4.2.4.	Performance in multiuser case	93
4.3.	Results of Study III	94
4.3.1.	IBI analysis	95
4.3.2.	Capacity analysis	97
4.4.	Results of Study IV	99
4.5.	Results of Study V	102
4.5.1.	Narrowband transmission	103
4.5.2.	Wideband transmission	104
5.	CONCLUSIONS AND FUTURE WORK.....	109
	BIBLIOGRAPHY.....	113
	CURRICULUM VITAE.....	122

LIST OF FIGURES

Figure 2.1: a) Spatial MIMO channel representation, and b) its beamspace MIMO channel representation.	9
Figure 2.2: General PLA system operation.	18
Figure 3.1: The LAS-MIMO system model for a) the fully-connected and b) the proposed sub-grouped architectures.	23
Figure 3.2: Normalized beam gain vs beam angle analysis at different antenna array size with $f_c = 60$ GHz.	33
Figure 3.3: Normalized beam gain vs beam angle analysis at different beam direction with $f_c = 60$ GHz.	33
Figure 3.4: Normalized beam gain vs beam angle analysis at different fractional bandwidth with $f_c = 60$ GHz.	34
Figure 3.5: The proposed LAS analog transceiver design for a multicarrier signal in mMIMO systems.	35
Figure 3.6: Sub-grouping the subcarriers in the subband filters.	36
Figure 3.7: Squinting amount of a given subband for any desired direction within LAS's scanning range is within the tolerable range ($\Omega_{3dB}/8$). Here, (3.35) is used to calculate N_F , with $W = 5$ GHz and $f_c = 60$ GHz.	37
Figure 3.8: The signal flow chart in the proposed LAS design.	41
Figure 3.9: Power consumption vs P analysis. Here, both the conventional mMIMO and the proposed LAS-based designs follow parameters in Table 4.2	45
Figure 3.10: Relationship between subcarrier frequencies and the resultant beam direction with respect to the ideally desired direction. It is noted that the angle of the transmitted beam deviates from the exact direction when the frequency shifts from the center frequency f_c . This is due to the effect of squinting of the beam in UWB mMIMO systems. Here we assume $M = 128$, $f_c = 28$ GHz, the system bandwidth $W = 3$ GHz, the number of subcarrier frequencies $N = 1024$, and $\hat{\theta} = 2\pi \frac{(m-1)}{M}$ with $m = 20$	48
Figure 3.11: System model in near-field propagation.	50
Figure 3.12: Multi-beam transmission with respect to the subcarrier frequencies. Here we assume $M = 128$, $f_c = 28$ GHz, the system bandwidth $W = 3$ GHz, and the number of subcarrier frequencies $N = 1024$	52
Figure 3.13: The normalized IBI pattern in the space (in dB). Four scenarios are considered as follows: a) narrowband signaling in far-field propagation, b) narrowband signaling in near-field propagation, c) UWB signaling in far-field propagation, and d) UWB signaling in near-field propagation. Here, it is assumed that the system operates at $f_c = 28$ GHz with bandwidth $W = 5$ GHz, $M = 256$, and $K = 20$ users distributed randomly in the service area with $M' = K$ beams where each user is served with one beam, while the desired user is located at $\hat{\theta} = \pi/6$	53
Figure 3.14: Capacity vs beam angle analysis. The solid line represents the beam capacity in the far-field propagation, while the dashed line represents the beam capacity in the near-field propagation. Here, it is assumed that the system operates at $f_c = 28$ GHz, $W = 5$ GHz, $K = 20$ users, and the desired user is located at d_F for far-field analysis and at $d_F/32$ for near-field analysis.	57
Figure 3.15: The concept of subcarrier grouping in the proposed sub-grouped beam-based scheme.	59
Figure 3.16: The mmWave MIMO system model.	62

Figure 3.17: An example of extracting the proposed distance signature for the beamspace channel representation with $L_p = 6$ and $N = 64$.	64
Figure 3.18: The distribution of η_L for the normal case K_0 and spoofing case K_1 for different channel sparsity levels. The distribution is obtained by simulating 500000 realizations generated from a geometric-based channel model with channel gain of complex normal distribution $\mathcal{CN}(0, 1)$ and the angles follow uniform distribution $\sim \mathcal{U}(-\pi/2, \pi/2)$. Then, the proposed distance signature is extracted from the channel. Lastly, fitting tools in [1] are utilized to find the best matching distribution for the fading amplitude of the actual hypothesis.	66
Figure 3.19: System model for a mmWave MIMO scenario with an attacker.	69
Figure 3.20: The beamspace channel representation a) without considering beam squint effect, and b) with beam squint effect.	71
Figure 3.21: The proposed tracking-based PLA flowchart.	76
Figure 4.1: Power consumption vs N_{RF} for different N_L in the fully-connected LAS-MIMO system architecture.	82
Figure 4.2: Power consumption vs N_L for different sub grouping at $N_{RF} = 4$ (dash lines) and $N_{RF} = 8$ (solid lines) in the proposed sub-grouped LAS-MIMO architecture.	83
Figure 4.3: Spectral efficiency vs SNR for $N_{RF} = 4$ in the proposed sub-grouped LAS-MIMO architecture.	84
Figure 4.4: Energy efficiency vs SNR for $N_{RF} = 4$ in the proposed sub-grouped LAS-MIMO architecture.	84
Figure 4.5: Spectral efficiency vs SNR for $N_{RF} = 8$ in the proposed sub-grouped LAS-MIMO architecture.	85
Figure 4.6: Energy efficiency vs transmitted power P_x for $N_{RF} = 8$ in the proposed sub-grouped LAS-MIMO architecture.	86
Figure 4.7: Beam gain vs normalized frequency.	88
Figure 4.8: Complexity analysis.	89
Figure 4.9: Capacity vs bandwidth analysis.	90
Figure 4.10: Power consumption vs M analysis.	90
Figure 4.11: Energy efficiency vs bandwidth analysis.	91
Figure 4.12: Capacity vs SNR analysis in a single user scenario.	92
Figure 4.13: a) Capacity and b) energy efficiency vs number of subband filters N_F analysis.	93
Figure 4.14: Capacity vs SNR analysis in multi-user scenario.	94
Figure 4.15: The normalized IBI pattern in the space (in dB) for Scenario I for: a) far-field propagation without beam squinting, b) near-field propagation without beam squinting, c) far-field propagation under beam squint effect, and d) near-field propagation under beam squint effect.	96
Figure 4.16: The normalized IBI pattern in the space (in dB) for Scenario II for: a) far-field propagation without beam squinting, b) near-field propagation without beam squinting, c) far-field propagation under beam squint effect, and d) near-field propagation under beam squint effect.	97
Figure 4.17: System capacity vs number of users analysis.	98
Figure 4.18: System capacity vs SNR analysis.	99
Figure 4.19: Probability of detection vs probability of false alarm at $L_p = 4$.	100
Figure 4.20: Probability of detection vs SNR analysis at $\alpha = 0.5$ and $N = 64$ for $L_p = 4$ and $L_p = 20$. Solid lines represent the signatures with perfect channel estimation, while the dash lines represent the signatures with imperfect channel estimation.	101

Figure 4.21: Probability of false alarm vs SNR analysis for correlated and uncorrelated channels at $\gamma = 0.5$, $N = 64$, and $L_p = 4$	101
Figure 4.22: Probability of detection vs probability of false alarm analysis for a) static, b) low-dynamic, and c) high-dynamic environment in narrowband transmission.	103
Figure 4.23: Probability of detection vs probability of false alarm analysis for a) static, b) low-dynamic, and c) high-dynamic environment in wideband transmission.	105
Figure 4.24: Probability of misdetection vs system bandwidth analysis.	106
Figure 4.25: Probability of detection vs channel estimation error analysis.	107
Figure 4.26: The normalized signature variation vs time analysis.	107



LIST OF TABLES

Table 4.1: Simulation parameters for the proposed LAS-MIMO design	82
Table 4.2: Simulation parameters of the proposed LAS design	87
Table 4.3: Simulation parameters for the IBI-based scenarios	95
Table 4.4: Simulation parameters for the tracking-based PLA scheme	102



LIST OF SYMBOLS

λ	: Wavelength
α	: Probability of False Alarm
γ	: Probability of Detection
F_{RF}	: PS Precoder Matrix at all RF Chains
$f_{RF}^{(k)}$	: PS Precoder Vector at k -th RF Chain
F_{LAS}	: LAS Precoder Matrix
F_{BB}	: Baseband Precoder Matrix
$s_t^{(l)}$	: Beam Selection Vector at Lens l
h_k	: Narrowband Channel Vector of User k
H	: Channel Matrix
$\hat{H}$	: Estimated Channel Matrix
y_k	: Received Signal at User k
n	: AWGN
a	: Steering Vector
β	: Channel Coefficient
F	: Total Precoder Matrix
d	: Inter-Element Spacing
$ \cdot $	: Absolute Value
I_N	: $N \times N$ Identity Matrix
R_{full}	: Spectral Efficiency in Conventional MIMO Design
R_g	: Spectral Efficiency in Proposed Sub-Grouped Design
A_t	: Sub-Grouped PS Precoder Matrix
EE	: Energy Efficiency
C	: Capacity
BW	: System Bandwidth
R	: Spectral Efficiency
P_c	: Power Consumption
$\hat{\theta}$	: AoA/AoD
τ	: Propagation Delay
$\Delta\tau_m$	: Delay Across Array Elements
g	: Beam Gain Function
BL	: Beam Loss
Ω_{3dB}	: 3 dB Beamwidth
N_F	: Number of Analog Subband Filters
IL_{SW}	: Insertion Loss of Switches
w	: Analog Beamforming Vector
AF	: Array Factor
$r_m^{(q)}$	: Distance Between the q -th User and the m -th Antenna Element
r_q	: Distance Between the First Antenna Element and the q -th User
IBI	: Inter-Beam Interference
P_r	: Received Signal Power
PL	: Path Loss
P_I	: Interference Power
C_A	: Mutual Coupling Matrix
K_0	: Hypothesis Test of Legitimate Node Sent the Received Signal
K_1	: Hypothesis Test of Legitimate Node did not Send the Received Signal

η_{th}	: Threshold to Separate Between K_0 and K_1
$\mathbf{U}$	: DFT Unitary Matrix
δ	: Variation of Normalized Enhanced Signature
$(\cdot)^H$	: Hermitian (Conjugate Transpose)
$(\cdot)^T$	: Transpose
$tr(\cdot)$	: Trace of Matrix $\mathbf{A}$
σ	: Noise Standard Deviation



ABBREVIATIONS

3GPP	: Third Generation Partnership Project
4G	: Fourth Generation
5G	: Fifth Generation
6G	: Sixth Generation
ANS	: Array Non-Stationarity
AoA	: Angle of Arrival
AoD	: Angle of Departure
AWGN	: Additive White Gaussian Noise
BAW	: Bulk Acoustic Wave
BL	: Beam Loss
BS	: Base Station
CFR	: Channel Frequency Response
CIR	: Channel Impulse Response
CSI	: Channel State Information
DAC	: Digital-to-Analog Converter
DFT	: Discrete Fourier Transform
ED	: Enhanced Distance
EE	: Energy Efficiency
EIRP	: Effective Isotropic Radiated Power
EKF	: Extended Kalman Filter
EM	: Electromagnetic
FDD	: Frequency Division Duplexing
FFT	: Fast Fourier Transform
DFT	: Discrete Fourier Transform
IA	: Initial Access
IBI	: Inter-Beam Interference
IoT	: Internet-of-Things
LAA	: Lens Antenna Array
LAS	: Lens Antenna Subarray
LoS	: Line-of-Sight
MCCM	: Mean Channel Covariance Matrix
MCM	: Mean Channel Matrix
MIMO	: Multiple-Input Multiple-Output
mMIMO	: Massive MIMO
mmWave	: Millimeter-Wave
MPMT	: Multiple Poles Multiple Throws
NR	: New Radio
OFDM	: Orthogonal Frequency Division Multiplexing
PLA	: Physical Layer Authentication
PS	: Phase Shifter
PW	: Planar Wavefront
RACH	: Random Access Channel
RF	: Radio-Frequency
ROC	: Receiver Operating Characteristics
RS	: Reference Signal
RSRP	: Reference Signal Received Power
RSS	: Received Signal Strength
SAW	: Surface Acoustic Wave
SIC	: Successive Interference Cancellation

SINR	: Signal to Interference and Noise Ratio
SLL	: SideLobe Level
SLS	: Sector-Level Sweep
SNR	: Signal-to-Noise Ratio
SPMT	: Single Pole Multiple Throws
SVD	: Single Value Decomposition
SW	: Spherical Wavefront
TA	: Traditional Array
TDD	: Time Division Duplexing
TTD	: True-Time-Delay
UCA	: Uniform Circular Array
UE	: User Equipment
ULA	: Uniform Linear Array
UPA	: Uniform Planar Array
VR	: Visibility Region
UWB	: Ultra WideBand
XL	: Extra Large-scale
ZF	: Zero-Forcing

MİLİMETRE DALGA MIMO SİSTEMLERİ İÇİN HİBRİT HÜZME OLUŞTURMA TEKNİKLERİ

ÖZET

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Araştırmamız, gelecekteki kablosuz ağlarda iyileştirilmiş enerji verimliliği, spektral verimlilik ve güvenlik ihtiyacından kaynaklanmaktadır. Mevcut hibrit çoklu giriş çoklu çıkış (MIMO) sistemleri, yüksek maliyet ve karmaşıklık, mekansal geniş bant etkisi (ışın saçılımı olarak bilinir), ışınlar arası girişim, ilk erişim gecikmeleri ve sahtecilik saldırılarına karşı savunmasızlık gibi önemli zorluklarla karşı karşıyadır. Bu tez, yeni nesil iletişim sistemleri için özel olarak tasarlanmış yenilikçi çözümlerle bu sorunları ele almaktadır. İlk olarak, büyük anten dizilerini ve birden fazla kullanıcıyı verimli bir şekilde yönetmek için hibrit ön kodlama algoritmasıyla birleştirilmiş alt gruplu lens anten alt dizisi (LAS-MIMO) mimarisi öneriyoruz. Bu tasarım, donanım yükünü azaltır ve toplam oranı en üst düzeye çıkarırken kullanıcılar arası girişimi en aza indirir. Büyük MIMO uygulamasına sahip ultra geniş bant sinyalizasyonu (UWB) için, mekansal geniş bant etkisini azaltmak, ışın kazanımını ve sistem kapasitesini artırmak için LAS tasarımını ve analog alt bant filtrelerini entegre eden bir alıcı-verici tasarımı sunulmaktadır. Özellikle, önerilen tasarım faz kaydırıcılardan gelen geniş bant sinyalini dar bant sinyal gruplarına ayırır ve mercekleme altında kapsamlı bir arama tabanlı anahtarlama/ön kodlama mekanizması aracılığıyla saçılıklarını kontrol eder. Ayrıca, karmaşıklığı önemli ölçüde en aza indirirken iyi performans gösteren basitleştirilmiş, eşikli arama tabanlı ön kodlama algoritması önerilir. Daha sonra, hem uzak alan hem de yakın alan yayılımında ışın saçılımından etkilenen UWB masif MIMO sistemlerinde ışınlar arası girişimi (IBI) ele alınmaktadır ve burada IBI modelinin sistem kapasitesi üzerindeki etkisi değerlendirilmektedir. IBI etkisi, alt gruplandırılmış ışın tabanlı bir zamanlama şemasıyla hafifletilir ve sistem performansı önemli ölçüde iyileştirilir. Güvenlik için, hem dar bant hem de geniş bant iletimlerde ışın alanı MIMO kanalının özelliklerini kullanarak, ışın alanı MIMO kanal gösterimindeki ana bileşenlerin konumlarından benzersiz bir mesafe imza vektörünün çıkarıldığı yeni bir fiziksel katman kimlik doğrulama (PLA) şeması sunulmaktadır. Dinamik ortamlardaki hareketlilik endişelerini gidermek için bir izleme filtresi tanıtılmaktadır. Bu filtre, kimlik doğrulama sisteminin depolanan benzersiz imzayı sürekli olarak izlemesini ve güncellemesini sağlayarak genel kimlik doğrulama performansını artırmaktadır. Kapsamlı simülasyonlar ve analizlerle doğrulanan çözümlerimiz, enerji verimliliği, spektral verimlilik, sistem kapasitesi ve güvenlikte önemli iyileştirmeler göstermektedir. Bu il-lerlemeler toplu olarak kritik zorlukları ele alarak, yeni nesil kablosuz iletişim sistemleri için sağlam, verimli ve güvenli çerçeveler sunmaktadır. Önerilen yöntemler potansiyel

olarak genelleştirilebilir olup, alanda daha fazla araştırma ve geliştirme için sağlam bir temel sağlamaktadır.



Anahtar sözcükler: Işın şaşı etkisi, ışın alanı MIMO, ışınlar arası girişim (IBI), mMMO, dar bant, PLA, UWB, XL-MIMO.

HYBRID BEAMFORMING TECHNIQUES FOR MILLIMETER-WAVE MIMO SYSTEMS

ABSTRACT

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Our research is driven by the need for improved energy efficiency, spectral efficiency, and security in future wireless networks. Current hybrid multiple-input multiple-output (MIMO) systems face significant challenges, including high cost and complexity, spatial wideband effect (known as beam squint), inter-beam interference, initial access delays, and vulnerability to spoofing attacks. This thesis addresses these problems through innovative solutions tailored for next-generation communication systems. First, we propose a sub-grouped lens antenna subarray (LAS-MIMO) architecture coupled with a hybrid precoding algorithm to efficiently manage large antenna arrays and multiple users. This design reduces hardware overhead and maximizes the sum rate while minimizing inter-user interference. For ultra-wideband signaling (UWB) with massive MIMO implementation, we introduce a transceiver design integrating LAS design and analog sub-band filters to mitigate the spatial wideband effect, enhancing beam gain and system capacity. Specifically, the proposed design chunks the wideband signal from the phase shifters into groups of narrowband signals and controls their squints through an exhaustive search-based switching/precoding mechanism under the lenses. Furthermore, a simplified, thresholded search-based precoding algorithm is proposed, demonstrating good performance while significantly minimizing complexity. Then, we tackle inter-beam interference (IBI) in the UWB massive MIMO systems affected by beam squint in both far-field and near-field propagation, where the impact of the IBI model on system capacity is evaluated. The IBI effect is alleviated with a sub-grouped beam-based scheduling scheme, significantly improving system performance. For security, utilizing properties of the beamspace MIMO channel in both narrowband and wideband transmissions, we present a novel physical layer authentication (PLA) scheme where a unique distance signature vector is extracted from the positions of the principal components within the beamspace MIMO channel representation. A tracking filter has been introduced to address mobility concerns in dynamic settings. This filter allows the authentication system to continuously track and update the stored unique signature, enhancing overall authentication performance. Our solutions also incorporate advanced optimization techniques to further enhance the scalability and adaptability of the proposed systems under varying network conditions. By considering realistic environmental factors such as mobility, channel fluctuations, and network topology, the proposed designs can dynamically adjust, ensuring consistent performance and robustness in diverse operational scenarios.

Our solutions, validated through extensive simulations and analysis, demonstrate substantial improvements in energy efficiency, spectral efficiency, system capacity, and security. These advancements collectively address critical challenges, offering robust, efficient, and secure frameworks for next-generation wireless communication systems. The proposed methods are potentially generalizable, providing a solid foundation for further research and development in the field.



Keywords: Beam squint effect, beamspace MIMO, inter-beam interference (IBI), mMIMO, narrowband, PLA, UWB, XL-MIMO.

CHAPTER 1

1. INTRODUCTION

Multiple-Input Multiple-Output (MIMO) technology has revolutionized wireless communications by exploiting the spatial dimension of the wireless channel to enhance both capacity and reliability. By employing multiple antennas at the transmitter, receiver, or both, MIMO systems capitalize on the natural multipath propagation of wireless signals. Rather than treating multipath as a limitation, MIMO transforms it into an advantage by creating parallel data streams, thereby significantly increasing throughput and spectral efficiency. This capability makes MIMO a cornerstone of advanced communication standards such as Fourth Generation (4G) LTE, Fifth Generation (5G) New Radio (NR), and emerging Sixth Generation (6G) systems.

The key benefits of MIMO include increased data rates through spatial multiplexing, improved reliability via diversity techniques, and enhanced coverage and signal strength with beamforming. Spatial multiplexing enables the simultaneous transmission of independent data streams over the same frequency band, multiplying the effective data rate. Diversity techniques mitigate fading by combining multiple versions of the same signal received through different paths, ensuring robust communication even in challenging environments. Beamforming, on the other hand, focuses signal energy in specific directions, improving coverage while reducing interference.

The importance of MIMO systems lies in their ability to efficiently utilize limited spectral resources while meeting the growing demand for high data rates and reliable connectivity. They are critical to enabling advanced applications in 5G and beyond. Moreover, with the integration of higher frequency bands such as Millimeter-Wave (mmWave) into wireless networks, MIMO systems play a pivotal role in mitigating increased path loss and enhancing spatial resolution.

Despite these advantages, designing and implementing MIMO systems poses several challenges. In massive MIMO and Ultra Wideband (UWB) systems, the spatial-wideband effect and beam squinting occur, causing signals at different frequencies or subcarriers to deviate in direction, degrading system performance. Similarly, near-field propagation in Extra Large-scale (XL)-MIMO introduces Array Non-Stationarity (ANS), where channel characteristics vary across the antenna array due to Spherical Wavefront (SW) effects and localized Visibility Regions (VRs). Addressing these phenomena requires advanced channel modeling, efficient signal processing algorithms, and robust beam alignment techniques.

Integrating Physical Layer Authentication (PLA) into MIMO systems is essential to harness the unique channel characteristics of next-generation networks. In 5G and beyond, MIMO technology employs highly directional beam patterns and sparse channel representations, enabling improved discrimination between legitimate and illegitimate users through attributes such as Angle of Arrival (AoA) and Angle of Departure (AoD). These features provide a robust foundation for advanced authentication mechanisms. However, adopting PLA in Beamspace MIMO systems introduces new challenges. The high-dimensional channel matrix and dynamic nature of beamspace representations demand efficient processing to maintain real-time performance. Additionally, environmental factors such as mobility and interference can affect the sparsity and distinguishability of beamspace attributes, potentially reducing their reliability as authentication metrics. Overcoming these challenges requires advanced signal processing techniques and adaptive models capable of handling the variability and complexity of Beamspace MIMO systems while ensuring seamless integration with existing PLA frameworks.

1.1. Thesis Contributions

- Achieving high spectral efficiency and energy efficiency is critical, particularly in mmWave and massive MIMO networks, where the dense deployment of antenna elements significantly increases system complexity and power consumption. Traditional hybrid beamforming architectures, while reducing the number of Radio-Frequency (RF) chains, often face challenges such as high computational complexity and excessive power usage due to extensive phase shifter networks. To address

these limitations, innovative architectures like lens antenna subarrays (Lens Antenna Subarray (LAS)) have been proposed, leveraging the sparsity of beamspace channels to enhance efficiency. However, fully-connected LAS designs still pose significant challenges in terms of scalability, hardware feasibility, and computational demands for real-time operation in multi-user scenarios. Building on this context, the work in [2] presents a sub-grouped LAS-MIMO architecture with a low-complexity hybrid precoding design tailored for mmWave multi-user MIMO systems. By dividing the RF paths into subgroups, each linked to a portion of the lens subarray, this architecture significantly reduces the implementation complexity and RF power consumption of fully-connected LAS structures. The study proposes a beam selection-based sum-rate maximization algorithm that partitions the optimization process into manageable sub-rate problems, integrating a partially exhaustive beam selection approach for efficiency. To further enhance performance, user interference within each subgroup is mitigated using Successive Interference Cancellation (SIC). Simulation results confirm the effectiveness of the proposed architecture and algorithm, demonstrating a comparable sum-rate to fully-connected hybrid Traditional Array (TA)-MIMO systems, along with substantial energy efficiency gains—nearly 400% and 45% improvements at high Signal-to-Noise Ratio (SNR) for eight users. These contributions mark a significant advancement in balancing energy efficiency and complexity in LAS-MIMO systems, addressing critical challenges in next-generation mmWave networks.

- To address the challenges posed by the beam squint problem in UWB massive MIMO (mMIMO) systems, where conventional solutions often compromise energy efficiency or involve impractical hardware complexity, the work in [3] introduces a novel analog transceiver design leveraging a modified LAS structure combined with analog subband filters. This design effectively compensates for beam squinting by dividing the wideband signal into subbands, aligning the resultant beams more precisely with the intended direction, and employing an energy-efficient switching network. The system, which incorporates a threshold-based precoder to reduce computational complexity, is adaptable to worst-case scenarios without requiring hardware modifications, making it suitable for green communication networks. Furthermore,

the framework supports a beam squint-based multiple access scheme, enabling flexible beam steering to serve multiple users, as demonstrated in [4].

- To address the challenges of Inter-Beam Interference (IBI) in multi-user MIMO systems, particularly in the context of UWB XL-MIMO networks, [5] introduces an accurate model that considers the impact of beam squint in both near-field and far-field propagation scenarios. This model not only defines the interference region in the angle-frequency domain but also provides a comprehensive analysis of IBI behavior in relation to distance, angle, and frequency. The work also proposes a sub-grouped beam-based scheduling mechanism that categorizes subcarrier frequencies based on their beam squint to reduce interference and optimize system capacity. Through simulation results, it is shown that the proposed scheduling method achieves significant improvements in system capacity while effectively mitigating IBI compared to traditional methods.
- In [6], a novel PLA scheme is proposed to address spoofing attacks in mmWave MIMO networks, leveraging the unique characteristics of 5G MIMO channels. The scheme uses a combination of distance, hardware, and hybrid signatures derived from beamspace MIMO channel representation and mutual coupling effects. Specifically, the distance signature is calculated by analyzing the principal components of the beamspace channel, while the hardware signature incorporates the antenna array's mutual coupling effects to enhance security. A hybrid signature is then constructed by combining these two methods. The proposed scheme is evaluated based on simulation results, which show its effectiveness in improving authentication performance while considering factors like channel sparsity and the number of antenna elements. This work offers an innovative approach to enhancing security in mmWave networks, distinguishing itself from traditional PLA methods by utilizing the rich characteristics of the MIMO channel and device hardware imperfections.
- In [7], a novel tracking-based PLA scheme is proposed to address the challenges of channel variations and user mobility in wideband mmWave MIMO systems, particularly in both Time Division Duplexing (TDD) and Frequency Division Duplexing (FDD) transmission modes. Unlike previous PLA methods that are tailored

for narrowband transmission and are limited to TDD systems, this approach introduces a versatile solution that works effectively in both static and dynamic environments. The scheme is based on beamspace parameters, such as AoA and AoD, which exhibit stability and reliability even in the presence of channel fluctuations over time and frequency. The proposed PLA method extends the distance-based signature introduced in previous work by focusing on the principal components of the beamspace channel, creating a unique authentication signature that remains robust against channel variations. This new method is independent of path gain, making it particularly reliable even with changes in the strength of individual signal paths. The signature also benefits from angle reciprocity, which ensures its applicability in both TDD and FDD systems. To handle mobility scenarios, the scheme incorporates a filtering-based tracking algorithm that allows for continuous updates of the authentication signature, reducing mismatches in the signature database and minimizing the authentication process's overhead. This approach enhances the scheme's ability to maintain authorized connections in dynamic environments, ensuring longer connection durations. The proposal is further extended to wideband mmWave MIMO systems by integrating the proposed signature with the beam squint effect, improving the robustness of the authentication process. Comprehensive simulation analyses are conducted to validate the scheme's effectiveness, demonstrating its strong performance in various environments with different levels of channel sparsity and mobility.

CHAPTER 2

2. THEORETICAL PART

2.1. Introduction To MIMO And XL-MIMO Systems

Employing multiple antenna elements at the transmitter, receiver, or both can significantly enhance system performance by effectively utilizing the multipath effects inherent in communication channels. These multipath effects, which result from signal reflections, diffraction, and scattering, are typically viewed as sources of interference. However, employing multiple antennas can harness these effects to transmit additional data, thus improving data throughput and the SNR.

MIMO technology leverages the principle of utilizing multiple antennas to separate signals traveling through different paths and exploits their capabilities to improve data throughput and SNR. This technological advancement significantly improves overall system performance by effectively managing and utilizing the multipath effects. MIMO technology is pivotal in modern wireless communication systems, including 5G and beyond, due to its ability to increase spectral efficiency and data rates.

MIMO technologies improve system capacity by enhancing energy and spectral efficiency through spatial dimensions, allowing more data to be transmitted within the same bandwidth. They ensure a reliable connection by providing robustness against fading, shadowing, and path loss, which are common issues in wireless communication channels. Multiple antennas also allow for better Channel State Information (CSI) and exploitation, crucial for adaptive transmission techniques. They also help align and cancel interference, improving overall system performance.

Multiple antenna modes include diversity, multiplexing, and beamforming, offering distinct advantages and applications. In diversity mode, the same data stream is transmitted

from all antennas simultaneously but independently, which improves signal coverage and increases diversity gain. This mode is particularly effective at low SNR levels to counteract fading, especially in channels with rich scatterers, thereby enhancing reliability. Different data streams are transmitted from each antenna independently and simultaneously in multiplexing mode. This approach is ideal for high SNR levels and channels with rich scatterers, maximizing the data transmission rate and significantly boosting overall throughput. In beamforming mode, the same data stream is transmitted from all antennas with a slight time delay between antenna elements. This technique improves signal coverage and provides antenna gain. Beamforming is particularly effective in sparse channels as it reduces interference. It is implemented using smart antennas, such as phased or adaptive arrays, which direct the signal toward the intended receiver, enhancing signal quality and coverage.

In practice, using multiple antennas involves a trade-off between throughput and coverage for a fixed array size or fixed capacity. When aiming to maximize throughput, the system may focus on spatial multiplexing, which increases data rates by transmitting multiple data streams simultaneously. However, this approach can lead to reduced coverage, as the adequate signal strength at the edges of the coverage area may decrease. Conversely, when prioritizing coverage, techniques like beamforming and diversity are employed to enhance signal strength and reliability over a larger area. These techniques ensure that even users at the edge of the coverage area receive a strong and reliable signal. Still, they may limit the overall data rate achievable by the system.

The Third Generation Partnership Project (3GPP) standards, which include technologies like 4G, 5G NR, and recently goes towards 6G, use multicarrier waveforms such as Orthogonal Frequency Division Multiplexing (OFDM) for data transmission in wireless networks to use the available spectrum and provide robust communication links efficiently. MIMO technology is highly compatible with these multicarrier signals, leveraging their characteristics to enhance system performance. Combining MIMO with multicarrier signals offers several significant advantages. The integration of MIMO and OFDM enhances spectral efficiency by allowing multiple data streams to be transmitted simultaneously over the same frequency band. This combination also improves robustness, as OFDM's resilience to frequency-selective fading complements MIMO's diversity techniques, ensuring a robust communication link even in challenging environments. Furthermore,

beamforming and interference alignment techniques in MIMO systems, when paired with OFDM, lead to better interference management and improved signal quality. Additionally, the granularity of OFDM subcarriers facilitates flexible resource allocation, optimizing the performance of MIMO systems.

2.1.1. Beamspace MIMO channel

As the operating frequency shifts to higher bands, the MIMO channel characteristics undergo significant changes, necessitating a more detailed investigation of the channel and its modeling. Higher frequency bands, such as mmWave, introduce new challenges, including increased path loss, higher susceptibility to atmospheric absorption, and more pronounced effects of multipath propagation. These changes can affect signal propagation, coverage, and overall system performance, requiring more advanced and precise channel models to represent and predict behavior in these frequency ranges accurately. Additionally, the increased spatial resolution afforded by higher frequencies may lead to more complex channel structures that need to be characterized and managed effectively. Therefore, developing and refining channel models that account for these frequency-dependent effects becomes crucial as frequencies rise to ensure robust and efficient system design and operation.

Due to the sparse nature of mmWave MIMO channels, these channels can be effectively represented in the angle domain. In this representation, the MIMO channel appears sparse, allowing for a more efficient and precise analysis of the channel characteristics. The angle domain channel representation bridges digital signal processing and antenna signal processing, enabling better integration and optimization of these technologies. As the number of antenna elements in a MIMO system increases, the resolution in the angle domain also increases, allowing for finer differentiation between incoming signal paths. This improved resolution enhances the system's ability to capture and process signals, leading to better performance accurately.

A specific case of the angle domain is the beamspace domain introduced in [8]. The beamspace domain provides a discretized approximation for the spatial channel by uniformly sampling the angle-delay-Doppler space at the Nyquist rate [9]. This discretization simplifies the analysis and processing of the signals but can introduce a power leakage problem. Power leakage occurs when energy from one beam spills into adjacent beams,

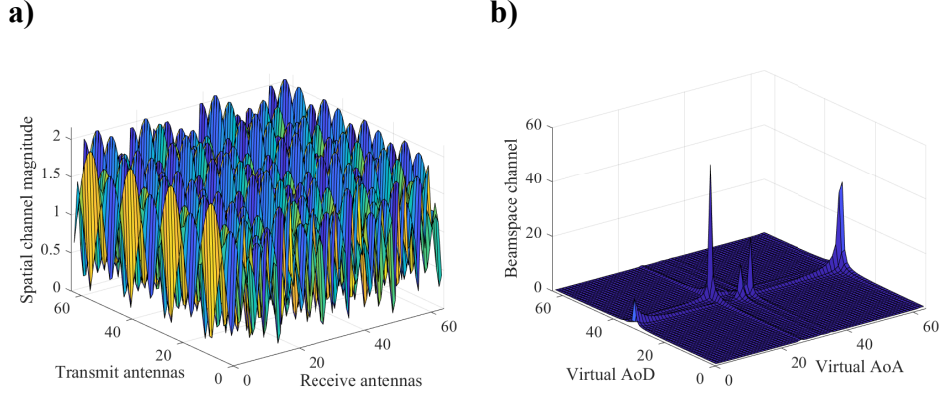


Figure 2.1: a) Spatial MIMO channel representation, and b) its beamspace MIMO channel representation.

which can degrade system performance by causing interference and reducing the clarity of the received signals. Despite this issue, beamspace representation remains a valuable tool for understanding and optimizing mmWave MIMO channels, particularly when combined with advanced signal processing techniques that can mitigate power leakage effects.

MmWave MIMO channel can be transformed from its spatial representation, seen in **Figure 2.1a**, to the beamspace one, seen in **Figure 2.1b**, using a Discrete Fourier Transform (DFT) approach. Beamspace channel representation is sparse as the number of antenna elements is large enough with fewer paths contributing to each spatial bin in the beamspace channel, where this sparsity property can reflect the number of transmitters [10]. Using beamspace channel representation can reduce the hardware transceiver complexity, channel matrix dimension, and precoder design [11], leading to a precise channel estimation approach and a faster beam alignment process. Note that the properties of the beamspace channel representation (principal components with the corresponding group of virtual AoA/AoD bins, as shown in **Figure 2.1b**) are robust to noise and can be accurately estimated bins/angles with low overhead [12]. By focusing on key parameters such as AoAs, AoDs, path gains, and delays, the system can achieve accurate channel characterization while managing computational demands effectively.

The sparsity or rank of the channel in this domain can be leveraged to simplify the channel estimation process. Instead of estimating the entire channel matrix, channel parameters can be calculated under certain conditions: a very large antenna array for better spatial

resolution, operation in high-frequency bands to ensure channel sparsity, and using a half-wavelength array to avoid angle estimation ambiguity.

Presenting the MIMO channel in the beamspace domain facilitates more efficient channel estimation, enhances beamforming, reduces hardware complexity, improves interference management, speeds up beam alignment, and contributes to more accurate localization, better spectral efficiency, and adaptive resource allocation. These benefits collectively lead to improved system performance and more effective utilization of resources in modern wireless communication systems. Some of these advantages are going to be exploited in the following chapters.

2.1.2. MIMO design issues

The design of MIMO systems involves addressing several critical challenges that impact their performance and effectiveness. Among these challenges are spatial-wideband effects and ANS, which can significantly influence the system's behavior and efficiency.

2.1.2.1. Spatial-wideband effect

MMIMO and UWB transmission have been considered among the candidate technological enablers for enhancing the performance and efficiency of the next-generation wireless networks [13], [14]. mMIMO can improve spectrum and energy efficiency, combat small-scale fading through channel hardening, and extend network coverage by overcoming the Path Loss (PL) problem [15]. UWB transmission not only facilitates high data rate communication but also provides resilience against multipath fading and covertness against jamming attacks [16]. Advantages of both mMIMO and UWB also span to the RF sensing aspect of the wireless networks that have been recently considered under the Joint Radar and Communication (JRC) framework [17]. They, respectively, provide acceptable spatial and temporal multipath resolution, thereby facilitating accurate localization and ranging of the target objects.

Despite these desirable advantages, it has been shown that mMIMO systems implementing UWB signaling suffer from spatial-wideband effect, recapitulated hereunder. The Huygens-Fresnel wave propagation principle dictates that, in the antenna array systems, the received signal at different array elements is a slightly delayed version of the original

signal unless the incident signal is perpendicular to the array. The delay incurred across the elements depends on the inter-element spacing and the signal's AoA/AoD. For a system with a relatively small number of antennas, as it is in the conventional small-scale MIMO systems, the maximum delay across the antenna aperture can be much smaller than the symbol duration, and thus its effect can be ignored. However, with high-dimensional antenna arrays, i.e., mMIMO, and with UWB signaling, this delay can be in the order or even more significant than the symbol duration, leading to delay squinting effect in the spatial-delay domain, i.e., considerable delay spread is observed across the array even in the pure Line-of-Sight (LoS) propagation condition. The delay squinting problem renders the steering vector frequency-dependent in the angular-frequency domain. In multicarrier systems like OFDM, signals at different subcarriers will point to different physical directions. Signals pertained to such derailed subcarriers might not arrive at the intended receiver or align with sidelobes or nulls of the receiver's radiation pattern, thereby degrading the performance. This phenomenon is referred to as beam squinting [18], [19].

2.1.2.2. Near-field propagation

The next generation of wireless networks is set to incorporate much larger antenna arrays, known as XL-MIMO, which will feature significantly larger aperture sizes compared to current mMIMO systems. These XL-MIMO arrays, with an extremely high number of antenna elements, promise to enhance significantly spectral and energy efficiency, offer superior spatial resolution, and support reliable massive access. However, XL-MIMO represents more than just a simple extension of conventional mMIMO systems; it introduces new challenges to the MIMO channel model that require advanced solutions.

A significant challenge in XL-MIMO systems is the phenomenon of ANS [20], where channel characteristics differ markedly across various antenna elements. This variation arises from two primary propagation conditions: SW and clustered VR.

The curvature of SW leads to variations in the AoAs and propagation distances to each element in an XL-MIMO array for the direct path. Depending on its position, this results in distinct phase shifts, delays, path losses, and polarization mismatch losses for each array element. Consequently, there are significant differences in received power, path gains, and even AoA across the array [21]. Moreover, because the curvature of SW changes with distance, these channel variations depend not only on the path angles but also on the

spatial distance between the signal source and the array. This angle-distance-dependent variation further accentuates the spatial uniqueness of the channel, introducing additional complexity from both angular and distance perspectives.

In complex propagation environments where most scatterers are within the near-field region, various segments of an XL-MIMO array may encounter different blockages and clusters of scatterers, leading to the concept of VR. Specifically, a communication node in different locations interacts with distinct clusters, and clusters at varying positions are visible to other parts of the array. This interaction defines the node's VR within the XL-MIMO system. The characteristics of these VRs—such as their size, delay spread, Doppler spread, and inter-element channel correlations—can vary widely depending on the clusters' size, mobility, and spatial arrangement relative to the connected nodes. These variations significantly impact the ANS observed in XL-MIMO systems.

These propagation conditions necessitate the development of new techniques for channel estimation, signal design, beamforming, and overall signal processing to effectively manage the complex and varying characteristics of XL-MIMO systems. Advanced algorithms and approaches must be designed to account for the non-stationary nature of the array and the diverse visibility regions, ensuring that the benefits of XL-MIMO, such as improved efficiency and resolution, can be fully realized. On the other hand, SW and VR characteristics can be exploited to enhance the system's performance.

2.2. Study I: Lens Antenna Subarray Design

MIMO systems can provide directional transmission with high gain; however, each antenna element requires its own RF chain in digital beamforming, resulting in high cost and power consumption [22]. To overcome this problem, various hybrid beamforming architectures have been developed with fewer RF chains compared to fully digital beamforming systems [23]. Hybrid precoding algorithms are developed in these systems to optimize beamforming in analog and digital domains. They are based either on spatially sparse precoding techniques as in [24] or on a codebook with an iterative search procedure as in [25] to calibrate a small number of RF chains with a large number of antenna elements in the array.

Many wireless systems using hybrid beamforming architectures require a large number of

Phase Shifters (PSs) to provide uniform data transmission, resulting in power consumption that outweighs the energy-saving benefits of using fewer RF chains, as discussed in [23]. Therefore, alternative designs such as Lens Antenna Array (LAA) [26] along with beamspace MIMO channel representation [9] are introduced to make the hybrid beamforming MIMO systems more applicable in practice and replace the bulky PS network with a simple switching network. The LAA enables the representation of the mmWave MIMO channel in a sparse beamspace domain by using a unitary DFT matrix. In addition, LAA can focus the signal power beam on different antenna elements. Due to the sparseness of the beamspace channel in mmWave systems, the number of concentrated energy beams is much less than that of the number of antenna elements, leading to the use of beam selection techniques [27]–[29]. The design of a beamforming system for a large antenna array, as in massive MIMO [30]–[32], with only a single-lens array presents some difficulties, including power leakage problem [33], large focal depth of the lens [34], and lack of scalability, since a larger/smaller lens must be redesigned when the number of antenna elements changes. Therefore, dividing the aperture area occupied by a large single lens into smaller lenses is proposed in [35]–[37] under the name of LAS to obtain a competitive system in terms of RF design feasibility and power consumption reduction with minimal impact on per-chain capacity.

In order to investigate the performance of the hybrid LAS-MIMO architecture, an analog/digital precoding technique was used based on an exhaustive search in [37]. The results in [37] show that the hybrid LAS architecture remarkably improves spectral efficiency and energy efficiency compared to a single-lens system under various mmWave channel conditions. Compared to a MIMO system without a lens array, i.e., hybrid TA-MIMO architecture [24], the LAS design shows a significant improvement in energy efficiency at sufficient data rate levels. However, the practical use of a fully-connected hybrid LAS-MIMO architecture with a precoding algorithm based on exhaustive search is challenging due to its high computational complexity. Therefore, a sub-grouped LAS-MIMO architecture with a low-complexity hybrid precoding design is proposed for mmWave multi-user MIMO systems. This study was published in IEEE Access [2].

2.3. Study II & III: Spatial-Wideband Effect In Wideband Massive MIMO

Initial investigation of the spatial-wideband effect, also known as beam squint effect, can be found in the literature on radar technology, as large antenna array systems have been deployed there ever since [38]–[41]. In wireless communication networks, the beam squint problem emerged recently after introducing the mMIMO and UWB transceiver concepts. Nevertheless, several studies have already been dedicated to tackling the beam squinting problem in wireless systems.

The proposed approaches can be roughly categorized into the digital domain and analog domain-based beam squint minimization approaches. The ideal and straightforward way of dealing with the beam squint problem in the digital domain is by designing a compensation matrix for each subcarrier that eliminates the squint-inducing phase term introduced by the spatial-wideband effect. However, this is deemed impractical as the dimension of the required compensation matrix is prohibitively large [42]. Accordingly, digital domain-based hybrid precoding approaches have been proposed in several studies [42]–[49]. Specifically, the digital precoder is divided into two phases: performing a conventional digital precoding process for each subcarrier and then compensating for the phase drifting caused by each frequency-dependent analog precoder. Some authors have opted for radically different approaches to tackle the problem, such as the works in [50], [51]. Although they provided promising results by approaching the situation from a codebook design perspective, complexity issues stemming from hardware requirements or prohibitively large codebook sizes seem to haunt the proposed approaches. For example, authors of [52] proposed an Alamouti-based beamforming scheme that minimizes beam gain variation of all subcarriers within the operational bandwidth. Since the proposed beam pattern optimization involves Eigenvalues and Eigenvalues vector computation, the computational complexity of this scheme grows with the number of antennas. In a nutshell, the digital precoding-based solutions may only offer minimal beam gain improvement while exacerbating system complexity owing to the fact that the beam squint phenomenon originates from the analog domain of the beamforming process.

The analog domain-based beam squint minimization approaches are discussed in [45], [46], [53]–[57]. In [53], [55], [56], authors proposed an analog precoder that leverages the True-Time-Delay (TTD) provided by Rotman lens array to reduce the beam squint

problem. The rationale behind the approach is that the TTD provided by the Rotman lens inherently circumvents the extra transition time delay across the array (which is the main cause of the squint problem as discussed above), which reduces the beam squint effect. [46] designs a sparse antenna structure where each analog output signal is controlled by two PSs at each RF chain. The PS-based proposed design can adjust the phase and amplitude of the analog precoder simultaneously while avoiding the constant-amplified precoding constraint and improving the transmission rate. However, the PSs addition causes a huge power consumption, which leads to an incompetent energy-efficient system. [57] and its extension [54] employ both bandpass filters and extra PSs to facilitate subband-based beam squinting compensation. The main idea behind this approach is to chunk a wideband signal into relatively narrowband signals using a bank of bandpass filters, followed by an additional phase shift with respect to the center frequency of each filter. While the approach sounds viable, the overall analog design still faces some challenges that must be addressed. Specifically, the adoption of an extra layer of PSs network has an obvious drawback of reducing the system's energy efficiency.

In order to compensate for the beam squint effect in the analog domain and enhance system performance while maintaining an energy-efficient system without the use of extra PS network as the aforementioned works, an analog antenna design is proposed in UWB mMIMO systems [3]. This study was published in IEEE Transactions on Communications [3].

In multi-user MIMO scenarios, the beams generated from a substantial antenna array experience a relatively significant sidelobe level (SLL), leading to the occurrence of IBI in the angular domain. IBI manifests when the sidelobes of one or more beams interfere with the main lobes of other beams. This phenomenon arises because the angular separation between individual beams is sufficient, allowing each beam to utilize the same time-frequency resources. The IBI imposes constraints on the Signal to Interference and Noise Ratio (SINR), consequently limiting the data rate achievable per beam.

The modeling of IBI is a critical aspect of network design and has been the focus of several noteworthy research efforts. In recent investigations, such as the work presented in [58], an IBI model is introduced for a Uniform Planar Array (UPA) within 5G cellular systems, employing mMIMO and beamforming technology. This model serves as a foundation

for analyzing key network metrics, including cell capacity and outage probability. Another significant contribution to the field of IBI modeling can be found in the research conducted in [59], which introduces an innovative approach based on the multi-elliptical propagation model. This approach aids in determining the necessary minimum separation angle between co-channel beams, which is instrumental in addressing IBI challenges in multi-beam antenna systems.

While modeling IBI is crucial, practical implementation of interference reduction methods is equally vital to ensure optimal network performance. These methods encompass intelligent resource allocation, which optimizes power and bandwidth allocation, as seen in research such as [60]–[62]. Additionally, techniques like amplitude tapering, as discussed in [63], offer control over beam shape but require comprehensive channel information and may have limitations in terms of Effective Isotropic Radiated Power (EIRP). Zero-forcing methods, explored in works like [64], [65], involve adjustments in amplitude excitation across antennas for beam shaping, but they also require comprehensive channel information and may impact EIRP. Subarray beamforming and beam combining strategies, as studied in research such as [66], [67], aim to optimize beamforming gain; however, it may reduce the gain achieved using a massive number of antennas.

The ultimate objective of these interference reduction methods is to ensure sufficient angular separation between multiple beams, thereby controlling the level of IBI within the network. Through the combination of advanced modeling and the practical implementation of these interference reduction strategies, it becomes possible to enhance the overall performance and efficiency of massive MIMO networks within the context of 5G and beyond.

The XL-MIMO represents the next stage in the evolution of contemporary mMIMO technology, which increases the number of antennas by at least an order of magnitude to several hundred or even thousands. This dramatic increase results in unparalleled improvements in both spectral efficiency and spatial resolution for wireless networks. However, the transition from mMIMO to XL-MIMO is not merely an increase in antenna count or size; it fundamentally changes the channel characteristics. Precisely, it shifts from conventional far-field Planar Wavefront (PW) propagation to near-field SW propagation, resulting in a channel that transitions from stationarity to non-stationarity over the antenna elements.

This leads to changes in the IBI effect on the system performance. The research in [68], [69] introduces a wideband near-field beam training mechanism to optimize beam angles in a frequency-division approach and beam distances in a time-division approach. It is essential to underscore that this training mechanism is tailored for a single-beam transmission system and does not account for the presence of IBI.

Therefore, an accurate IBI model is introduced in [5], taking into account the impact of beam squint in both near-field and far-field propagation scenarios within UWB XL-MIMO networks.

2.4. Physical Layer Authentication (PLA) In Beamspace MIMO

PLA is a process that uses physical-layer attributes to build secure wireless communication networks in the analog domain, where the legitimate users are verified and identified from the illegitimate ones. Compared to the other conventional security mechanisms, PLA provides the following advantages:

- The PLA protects the physical layer features with the noisy observation only where the receiver can measure the attributes instantaneously. This allows for low computational requirements, processing delay, network overhead, and modest energy consumption.
- PLA is an analog-based mechanism that depends on either device imperfections and/or its environment.
- The device hardware imperfections and its environments offer unpredictable and unique attributes due to the diverse combinations available, which provide natural refresh mechanisms and multi-dimensional protections.
- Unlike traditional methods, which require an additional trusted third party in the system, PLA techniques use only available knowledge of physical attribute statistics to provide security.

It is useful to note that the PLA mechanisms do not replace the upper-layer authentication methods but are considered support, providing a higher security level in the system.

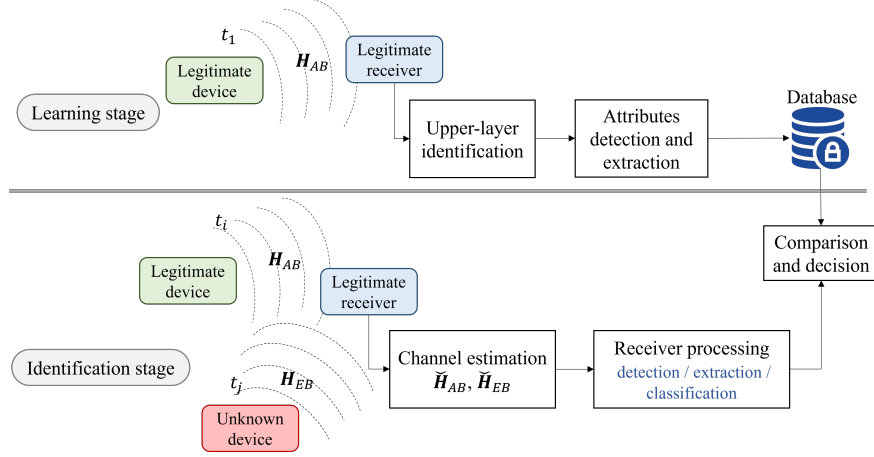


Figure 2.2: General PLA system operation.

PLA techniques are categorized into RF/hardware-based and channel-based attributes. The channel-based PLA techniques employ the channel features between the legitimate transmitter and receiver, such as Received Signal Strength (RSS), Channel Impulse Response (CIR), and Channel Frequency Response (CFR) to authenticate the sender by observing the unique instantaneous/average characteristics' of the estimated CSI then comparing the current CSI characteristics to the authentic CSI that previously reserved during the past transmission as shown in **Figure 2.2**. Although channel-based attributes do not require high-end signal analyzers to extract the features, they rely on the assumption that the channel features of different transmitted devices have a strong spatial decorrelation.

Different from the existing security schemes based channel attributes discussed in 4G MIMO wireless networks, the new channel representation of MIMO in 5G communication networks present unique features, including uniqueness of beam patterns, distinguishability of Sector-Level Sweep (SLS) SNR trances, and sparsity of angle/virtual channel representation.

Beamspace channel representation simplifies the 5G MIMO channel to a sparse model, reducing hardware complexity, channel matrix size, and precoder design [11]. This approach enhances channel estimation accuracy and speeds up beam alignment. Additionally, the beamspace representation's principal components and virtual AoA/AoD angles are robust to noise, enabling precise angle estimation with low overhead [12]. Therefore, several PLA techniques are set up using beamspace properties in the following subsections.

2.4.1. Study IV: PLA scheme in beamspace MIMO systems

Channel-based authentication systems exploit the phenomenon of spatial decorrelation between devices, which ensures a unique location-specific signature for each node. The work in [70] uses MIMO channel coefficients to build a multi-feature-based authentication method. Similarly, [71] exploits the location-specific property of wireless channels for PLA in a mMIMO system, also incorporating hardware impairment into the authentication scheme. Utilizing the 5G MIMO channel characteristics, the studies in [72] and [73] introduce the use of the beamspace channel representation in mmWave MIMO systems to detect spoofing attacks. However, storing the beamspace channel matrix as a signature in the database causes storage overhead due to the significant increase in the number of elements in the antenna array.

The MIMO antenna array can also be exploited to provide efficient PLA schemes. For example, The work in [74] exploits perturbed antenna array geometries for a robust PLA network. Similarly, the work in [75] utilizes the spatial beam patterns during the beam searching process of mmWave multiple antenna devices for RF fingerprinting by exploiting the inherent impairments in antenna array structure and PSs. [76] exploits the mmWave MIMO channel features, AoAs, AoDs, and channel gain, with the combination of hardware phase noise features to provide a novel PLA scheme.

Motivated by the MIMO channel characteristics in the 5G networks, a novel PLA system against spoofing attacks in mmWave networks is proposed in [6] using distance, hardware, and hybrid signatures based on beamspace MIMO channel representation and mutual coupling effect.

2.4.2. Study V: Robust tracking-based PLA in MIMO systems

The PLA studies, including those discussed in the previous section, are susceptible to channel variations and user mobility. Given that they are primarily tailored for narrow-band transmission, they cannot be applied to wideband mmWave MIMO systems, which are crucial for achieving high data rates. Furthermore, these methods can only be implemented in the TDD transmission and are not viable in the FDD mode. Therefore, a novel

tracking-based PLA scheme is proposed in [7] that takes advantage of the beamspace parameters for the mmWave MIMO channels, offering versatility in static and dynamic environments and robustness against channel variations in both time and frequency domains. The work builds upon the stability of beamspace parameters, such as AoAs and AoDs, in dynamic settings, contrasting their reliability with counterparts in time and frequency domains [77]. The proposed PLA framework employs a tracking-based approach, creating a unique authentication signature by extracting principal components from the beamspace representation of the mmWave MIMO channel. This scheme is designed to operate effectively in static and dynamic environments, covering narrowband and wideband mmWave MIMO systems.



CHAPTER 3

3. EXPERIMENTAL PART

3.1. Study I: Lens Antenna Subarray Design In Multi-User MIMO Systems

The performance of a MIMO system depends primarily on signal processing algorithms, channel propagation characteristics, and the configuration of the antenna array. An antenna array refers to a set of antenna elements positioned with specific inter-element spacing. These elements are connected to the RF analog part of the transceiver design through PSs or switches, facilitating individual connections to each antenna element. The antenna array can be designed along one dimension, such as Uniform Linear Array (ULA), or two dimensions, such as UPA and Uniform Circular Array (UCA). The choice of the array depends on the specific application requirements, including desired beamforming capabilities, spatial resolution, and coverage area.

One of the recent interesting antenna arrays is the LAA. The use of an Electromagnetic (EM) lens as an antenna array geometry in MIMO systems provides several significant benefits, along with some notable challenges. The lens enables beamspace representation for the MIMO channel, allowing for the concentration of beams from different directions on different antenna elements, which reduces the need for complex signal processing at each antenna element. This capability not only enhances directional signal reception but also reduces system power consumption and hardware costs. However, the insertion of LAA introduces a power leakage problem and suffers from a lack of scalability, as larger-size lenses need to be redesigned when the number of antenna elements changes, as in the case of massive and XL-MIMO systems. Additionally, the lens design has limited beam shaping, which prevents implementing a precoder/combiner over a multipath environment, leading to suboptimal performance in such scenarios.

To mitigate the challenges associated with large LAAs, the concept of LAS has been introduced. LAS utilizes multiple smaller lenses, each covering a subset of the antenna elements. This new antenna array design offers several advantages:

- **Improved Scalability:** LAS can be more easily scaled than a single large lens array. Adding or removing subarrays is simpler and less costly than redesigning a large lens, providing greater flexibility in system design and deployment.
- **Enhanced Performance in Multipath Environments:** By dividing the antenna array into smaller subarrays, LAS can better manage and mitigate multipath effects. Each subarray can perform localized beamforming and signal processing, leading to improved performance in environments where signals arrive from multiple directions.
- **Flexible Beamforming:** LAS allows for more sophisticated beamforming and signal processing techniques. Each subarray can implement its own precoding and combining strategies, optimizing the overall beam shape and directionality. This flexibility leads to better utilization of the spatial dimensions and improved system performance.
- **Reduced Power Leakage:** Smaller lenses are more efficient in focusing power, reducing the overall power leakage compared to a single large lens. This efficiency enhances the system's energy usage and minimizes interference with neighboring channels.

3.1.1. System model

Consider a downlink mmWave MIMO system using LAS. The Base Station (BS) uses N_t antenna elements connected to N_L lenses. Each lens has M antenna elements, so $M = N_t/N_L$ where M is an integer, as shown in **Figure 3.1a**. Note that all M lens sizes are much smaller than a typical single-lens [26] system; hence, the architecture is referred to as LAS. In addition, the BS has N_{RF} RF chains to simultaneously serve K single-antenna users. Since the total number of active users can be arbitrarily larger than N_{RF} in practice, we assume that the BS first selects N_{RF} users from the total active user pool before transmission to achieve full multiplexing gains [78], [79].

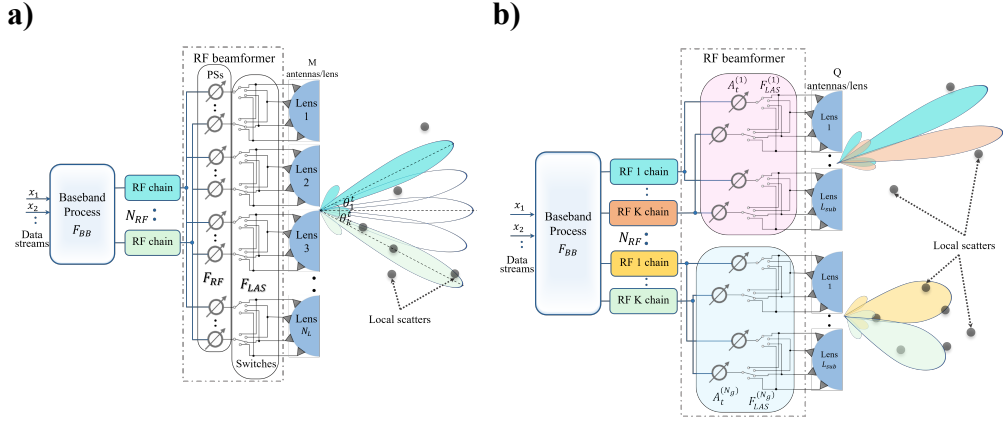


Figure 3.1: The LAS-MIMO system model for **a)** the fully-connected and **b)** the proposed sub-grouped architectures.

As in **Figure 3.1a**, the data stream is first passed through a baseband precoder $\mathbf{F}_{BB} \in \mathbb{C}^{N_{RF} \times N_{RF}}$. Then, the output of each RF chain is passed through a B -bit PS network with $\mathbf{f}_{RF}$ precoding vector. The PS precoder of the k -th RF chain is

$$\mathbf{f}_{RF}^{(k)} = \frac{1}{\sqrt{N_L}} [e^{j\phi_{1,k}}, e^{j\phi_{2,k}}, \dots, e^{j\phi_{N_L,k}}]^T, \quad (3.1)$$

where $\phi_l = \frac{2\pi b}{2^B}$, $l = 1, 2, \dots, N_L$. b can take any value from $b = \{b; b = 0, 1, \dots, 2^B - 1\}$ [80]. The total PS precoder at all RF chains is $\mathbf{F}_{RF} = [\mathbf{f}_{RF}^{(1)}, \mathbf{f}_{RF}^{(2)}, \dots, \mathbf{f}_{RF}^{(N_{RF})}]$. After that, the analog signal is routed through a switching mechanism modeled by a $\mathbf{F}_{LAS}$ precoder to the lenses for transmission, given as follows

$$\mathbf{F}_{LAS} = \begin{bmatrix} \mathbf{s}_t^{(1)} & \mathbf{0}_{M \times 1} & \cdots & \mathbf{0}_{M \times 1} \\ \mathbf{0}_{M \times 1} & \mathbf{s}_t^{(2)} & & \mathbf{0}_{M \times 1} \\ \vdots & & \ddots & \vdots \\ \mathbf{0}_{M \times 1} & \mathbf{0}_{M \times 1} & \cdots & \mathbf{s}_t^{(N_L)} \end{bmatrix}, \quad (3.2)$$

where $\mathbf{s}_t^{(l)}$ is the $M \times 1$ beam selection vector at lens l overlaid by all RF chains. For x transmitted streams, the received signal at user k is given as

$$y_k = \mathbf{h}_k \mathbf{F}_{LAS} \mathbf{F}_{RF} \mathbf{F}_{BB} \mathbf{x} + n_k, \quad (3.3)$$

where $\mathbf{h}_k \in \mathbb{C}^{1 \times N_t}$ is the narrowband channel of user k and $n \sim \mathcal{CN}(0, \sigma_o^2)$ is an additive white Gaussian noise. Here we use the clustered geometric channel representation, which

is useful for modeling mmWave propagation, i.e., the k -th user channel is given as [81]

$$\mathbf{h}_k = \sum_{i=1}^{P_k} \beta_{i,k} \mathbf{a}^H(\theta_{i,k}^t), \quad (3.4)$$

where P_k is the number of channel paths of user k , $\beta_{i,k}$ is the channel coefficient with complex normal distribution $\mathcal{CN}(0, 1)$ on the i -th path for the k -th user, $\theta_{i,k}^t$ is the AoD, and $\mathbf{a}$ is the response vector of the transmit antenna array. For a uniformly spaced antenna array, the steering vector is given as

$$\mathbf{a}(\theta) = \left[1, e^{-j2\pi \frac{d}{\lambda} \sin \theta}, \dots, e^{-j2\pi \frac{d}{\lambda} (N_t-1) \sin \theta} \right]^T, \quad (3.5)$$

where $d = \lambda/2$ being inter-element spacing and λ is the carrier wavelength.

3.1.2. Problem formulation

Assume that the total design precoding $\mathbf{F} = \mathbf{F}_{\text{LAS}} \mathbf{F}_{\text{RF}} \mathbf{F}_{\text{BB}}$, the achievable spectral efficiency for all users is given by [24]

$$R_{\text{full}} = \log_2 \left(\left| \mathbf{I}_{N_{\text{RF}}} + \frac{\gamma}{N_{\text{RF}}} \mathbf{H} \mathbf{F} \mathbf{F}^H \mathbf{H}^H \right| \right), \quad (3.6)$$

where γ is the SNR value, and $\mathbf{H} = [\mathbf{h}_1^T, \mathbf{h}_2^T, \dots, \mathbf{h}_{N_{\text{RF}}}^T]^T$ is the channel matrix of all users.

In general, the precoding design aims to maximize the spectral efficiency of overall users, taking into account the transmit power constraint and the constant modulus constraint of the analog beamformers. Therefore, the problem is formulated as follows

$$\begin{aligned} (\mathbf{F}_{\text{LAS}}^{\text{opt}}, \mathbf{F}_{\text{RF}}^{\text{opt}}, \mathbf{F}_{\text{BB}}^{\text{opt}}) &= \arg \max_{\mathbf{F}_{\text{LAS}}, \mathbf{F}_{\text{RF}}, \mathbf{F}_{\text{BB}}} R_{\text{full}}, \\ \text{s.t. } \mathbf{F}_{\text{RF}} &\in \mathcal{F}_{\text{RF}}, \\ \mathbf{F}_{\text{LAS}} &\in \mathcal{F}_{\text{LAS}}, \\ \text{tr} \left(\mathbf{F}_{\text{BB}}^H \mathbf{F}_{\text{RF}}^H \mathbf{F}_{\text{RF}} \mathbf{F}_{\text{BB}} \right) &= N_{\text{RF}}, \\ \|\mathbf{F}_{\text{RF}}\|_F^2 &= 1, \end{aligned} \quad (3.7)$$

given that $\mathcal{F}_{\text{LAS}}$ is the set containing all feasible LAS selected beams [37] while $\mathcal{F}_{\text{RF}}$ is the set of feasible RF precoders, i.e., the set of $N_t \times N_{\text{RF}}$ matrices for $\mathbf{F}_{\text{RF}}$ matrix with constant magnitude entries [24].

It is assumed that the BS perfectly estimates the channel matrix $\mathbf{H}$ using an effective channel estimation scheme such as the techniques presented in [82]–[88] to obtain the estimated

channel $\hat{\mathbf{H}}$.

The problem in (3.7) is ideally solved by an exhaustive search over all possible sets for the precoder $\mathbf{F}_{\text{LAS}}$ and $\mathbf{F}_{\text{RF}}$, while the digital precoder $\mathbf{F}_{\text{BB}}$ can be computed by using zero-forcing technique according to [89] as follows

$$\mathbf{F}_{\text{BB}} = \left(\mathbf{F}_{\text{RF}}^H \mathbf{F}_{\text{LAS}}^H \hat{\mathbf{H}}^H \hat{\mathbf{H}} \mathbf{F}_{\text{LAS}} \mathbf{F}_{\text{RF}} \right)^{-1} \left(\hat{\mathbf{H}} \mathbf{F}_{\text{LAS}} \mathbf{F}_{\text{RF}} \right)^H. \quad (3.8)$$

However, it should be noted that due to the non-convex feasibility constraint $\mathbf{F}_{\text{RF}} \in \mathcal{F}_{\text{RF}}$, there are no general solutions to (3.7) [24], and the optimization over $\mathbf{F}_{\text{LAS}}$ and $\mathbf{F}_{\text{RF}}$ jointly is unfeasible for the following reasons:

1. To design $\mathbf{F}_{\text{RF}}$, an exhaustive search from a predefined set is required to find an optimal solution. However, the possible choices in the set increase exponentially with the increase in lenses and PSs resolutions.
2. In beam selection matrix $\mathbf{F}_{\text{LAS}}$, as defined in [37], the complexity is limited to the number of lenses N_L and antenna elements associated with each lens M . For a fixed M , as the number of N_L in the system increases, the number of search angles increases exponentially since each lens has its own angle that must be set optimally.

Consequently, we propose a MIMO system using a sub-grouped hybrid LAS architecture (i.e., sub-grouped LAS-MIMO system) to reduce the complexity of the precoder and to further improve the energy efficiency of the system.

3.1.3. Proposed sub-grouped LAS transceiver design

In this section, we first present the proposed sub-grouped LAS transceiver design, where the conventional LAS-MIMO system is grouped into sub-grouped LAS-MIMO. Then, a sub-grouped-based precoding is proposed for the presented LAS design. After that, the complexity of the proposed algorithm is calculated, and the model for estimating the energy efficiency of the fully connected and the proposed sub-grouped LAS architectures is presented.

3.1.3.1. Transceiver design

In conventional fully-connected LAS-MIMO design, as shown in **Figure 3.1a**, the signal out of each RF chain is passed from all PSs to the activated antenna element by a switching network under each lens in the LAS structure. Hence, the transmitted beam is launched from all N_L lenses in the design. This results in producing similar beam gains as in conventional fully-connected TA-MIMO system with significantly less hardware requirement and thus less power consumption [37]. However, as the size of the antenna array increases, the channel realization varies in each part of the antenna array. Therefore, different users served by this system face diverse environments. For instance, a user at 30° direction experiences a high-sparse channel with only one LoS path. This user would prefer to be served with the maximum beamforming gain that the system can provide. Another user, who is in the 130° direction, may be faced with multiple scatterers that create a multipath channel. To serve this user, an optimized beamforming gain is required based on the realization of his channel, which is different from that of the first user.

In order to serve multiple users with different requirements and further improve the energy efficiency of the conventional fully-connected LAS-MIMO system, a sub-grouped LAS design is proposed in this work by dividing the multiple lenses structure in fully-connected LAS-MIMO design into N_g sub-groups. In each sub group, there are $L_{\text{sub}} = \frac{N_L}{N_g}$ lenses with $K_{\text{sub}} = \frac{N_{\text{RF}}}{N_g}$ RF chains connected to all L_{sub} lenses as shown in **Figure 3.1b**, given that K_{sub} is chosen to be an integer and there are Q antenna elements under each lens. Unlike the conventional fully-connected LAS-MIMO design, in the proposed sub-grouped LAS-MIMO structure, the signal generated from one RF chain belongs to only one sub-group. It does not connect to other lenses in the other subgroups. **Figure 3.1b** illustrates the transmitter design of the proposed sub-grouped LAS-MIMO architecture, while the receiver obeys the inverse architecture. The following subsection focuses on the hybrid precoding design to control the proposed sub-grouped LAS architecture.

3.1.3.2. Proposed sub-grouped-based precoding

In the proposed sub-grouped architecture, the hybrid precoder turns to be a spatial block diagonal such as $\mathbf{F} = \text{diag}\{\mathbf{F}_{(1)}, \mathbf{F}_{(2)}, \dots, \mathbf{F}_{(n)}, \dots, \mathbf{F}_{(N_g)}\} = [\mathbf{F}'_{(1)}, \mathbf{F}'_{(2)}, \dots, \mathbf{F}'_{(N_g)}]$. This

is due to the simplified structure of $\mathbf{F}_{\text{LAS}}$, $\mathbf{F}_{\text{RF}}$, and $\mathbf{F}_{\text{BB}}$, given as follows

$$\mathbf{F}_{\text{LAS}} = \begin{bmatrix} \mathbf{F}_{\text{LAS}}^{(1)} & \mathbf{0}_{Q L_{\text{sub}} \times L_{\text{sub}}} & \cdots & \mathbf{0}_{Q L_{\text{sub}} \times L_{\text{sub}}} \\ \mathbf{0}_{Q L_{\text{sub}} \times L_{\text{sub}}} & \mathbf{F}_{\text{LAS}}^{(2)} & & \mathbf{0}_{Q L_{\text{sub}} \times L_{\text{sub}}} \\ \vdots & & \ddots & \vdots \\ \mathbf{0}_{Q N_{L_{\text{sub}}} \times N_{L_{\text{sub}}}} & \mathbf{0}_{Q L_{\text{sub}} \times L_{\text{sub}}} & \cdots & \mathbf{F}_{\text{LAS}}^{(N_g)} \end{bmatrix}, \quad (3.9)$$

$$\mathbf{F}_{\text{RF}} = \begin{bmatrix} \mathbf{A}_t^{(1)} & \mathbf{0}_{L_{\text{sub}} \times K_{\text{sub}}} & \cdots & \mathbf{0}_{L_{\text{sub}} \times K_{\text{sub}}} \\ \mathbf{0}_{L_{\text{sub}} \times K_{\text{sub}}} & \mathbf{A}_t^{(2)} & & \mathbf{0}_{L_{\text{sub}} \times K_{\text{sub}}} \\ \vdots & & \ddots & \vdots \\ \mathbf{0}_{L_{\text{sub}} \times K_{\text{sub}}} & \mathbf{0}_{L_{\text{sub}} \times K_{\text{sub}}} & \cdots & \mathbf{A}_t^{(N_g)} \end{bmatrix}, \quad (3.10)$$

$$\mathbf{F}_{\text{BB}} = \begin{bmatrix} \mathbf{F}_{\text{BB}}^{(1)} & \mathbf{0}_{K_{\text{sub}} \times K_{\text{sub}}} & \cdots & \mathbf{0}_{K_{\text{sub}} \times K_{\text{sub}}} \\ \mathbf{0}_{K_{\text{sub}} \times K_{\text{sub}}} & \mathbf{F}_{\text{BB}}^{(2)} & & \mathbf{0}_{K_{\text{sub}} \times K_{\text{sub}}} \\ \vdots & & \ddots & \vdots \\ \mathbf{0}_{K_{\text{sub}} \times K_{\text{sub}}} & \mathbf{0}_{K_{\text{sub}} \times K_{\text{sub}}} & \cdots & \mathbf{F}_{\text{BB}}^{(N_g)} \end{bmatrix}, \quad (3.11)$$

where $\mathbf{A}_t$ is a $L_{\text{sub}} \times K_{\text{sub}}$ matrix where its vector follows (3.1) with L_{sub} elements.

In addition to the above beamformer constraints, the constraints in (3.7) for the optimization problem are also considered in the proposed sub-grouped scheme. Considering that the precoding on different sub-grouped in the LAS architecture is independent, the optimization problem can be divided into a set of suboptimization problems as stated in [90] using the SIC concept. Therefore, based on the proposed model for sub-grouped systems, (3.6) can be modified as follows

$$R_g = \sum_{n=1}^{N_g} \log_2 \left(\left| \mathbf{I}_{K_{\text{sub}}} + \frac{\gamma}{N_{\text{RF}}} (\mathbf{F}'_{(n)})^H \hat{\mathbf{H}}^H \mathbf{V}_{n-1}^{-1} \hat{\mathbf{H}} \mathbf{F}'_{(n)} \right| \right), \quad (3.12)$$

where $\mathbf{F}'_{(n)}$ is $N_t \times K_{\text{sub}}$ sub-matrix of matrix $\mathbf{F}$, and $\mathbf{V}_n = \mathbf{I}_{N_{\text{RF}}} + \frac{\gamma}{N_{\text{RF}}} \hat{\mathbf{H}} \mathbf{F}_n \mathbf{F}_n^H \hat{\mathbf{H}}^H$ is the auxiliary matrix for interference cancellation between the sub-groups with $\mathbf{V}_0 = \mathbf{I}_{N_{\text{RF}}}$ where $\mathbf{F}_n$ is $N_t \times nK_{\text{sub}}$ matrix contain the first nK_{sub} columns of $\mathbf{F}$. Thus, since $\mathbf{F}'_{(n)}$ has non-zero values at $\mathbf{F}_{(n)}$ only, the sub-optimization problem can be equivalently written as

$$\mathbf{F}_{(n)}^{\text{opt}} = \arg \max_{\mathbf{F}_{(n)}} \log_2 \left(\left| \mathbf{I}_{K_{\text{sub}}} + \frac{\gamma}{N_{\text{RF}}} (\mathbf{F}'_{(n)})^H \hat{\mathbf{H}}^H \mathbf{V}_{n-1}^{-1} \hat{\mathbf{H}} \mathbf{F}_{(n)} \right| \right), \quad (3.13)$$

which follows the constraints stated in (3.7). The optimization problem starts by optimizing the achievable spectral efficiency of the first sub-grouped array $\mathbf{F}_{(1)}$ by excessively

searching all possible switches' and PS's set for the $\mathbf{F}_{\text{LAS}}^{(1)}$ and $\mathbf{A}_t^{(1)}$ optimization, respectively. To relax the optimization problem, the ℓ -th column in the sub-grouped PS precoder $\mathbf{A}_t^{(1)}$ can be modeled by a beamforming vector as follows

$$\mathbf{a}_t^\ell = \frac{1}{\sqrt{L_{\text{sub}}}} \left[1, e^{j2\pi \frac{dL_{\text{sub}}}{\lambda} \sin \theta_\ell}, \dots, e^{j2\pi \frac{dL_{\text{sub}}}{\lambda} (L_{\text{sub}}-1) \sin \theta_\ell} \right]^T, \quad (3.14)$$

where $\sin \theta_\ell = \Phi_\ell$ with $\Phi_\ell = b/2^B$ and b is chosen so that θ_ℓ is approximately equal to the AoD of the user ℓ (i.e., $\theta_\ell \approx \hat{\theta}_\ell$). After selecting $\mathbf{F}_{\text{LAS}}^{(1)}$ from the predefined set $\mathcal{F}_{\text{LAS}}$, $\mathbf{F}_{\text{BB}}^{(1)}$ is calculated by modifying (3.8) as follows

$$\mathbf{F}_{\text{BB}}^{(1)} = \left(\left(\mathbf{H} \mathbf{F}_{\text{LAS}}^{(1)} \mathbf{A}_t^{(1)} \right)^H \mathbf{H} \mathbf{F}_{\text{LAS}}^{(1)} \mathbf{A}_t^{(1)} \right)^{-1} \left(\mathbf{H} \mathbf{F}_{\text{LAS}}^{(1)} \mathbf{A}_t^{(1)} \right)^H, \quad (3.15)$$

where $\mathbf{H}$ is $K_{\text{sub}} \times QL_{\text{sub}}$ sub matrix of matrix $\hat{\mathbf{H}}$ that is corresponding to the given sub-grouped. Then, the matrix $\mathbf{V}_1$ is updated. The same procedures are accomplished for the other sub-grouped antennas until the last sub-grouped array. The proposed sub-grouped precoding is summarized in **Algorithm 3.1**.

It is worth mentioning that in the sub-grouped LAS-MIMO system, each sub-group has a lower antenna gain compared to the fully-connected architecture, which may prevent the coverage of the distant users that the fully-connected architecture can cover. However, the LAS-MIMO system can provide less interference to all nearby users close enough to a particular subgroup. This is due to the ability of the proposed precoding algorithm to optimize each subgroup independently before updating the auxiliary matrix $\mathbf{V}_n$.

3.1.3.3. Complexity evaluation

A complexity evaluation is performed to show that the search procedure for the optimal precoder design is simplified with the proposed sub-grouped LAS-MIMO system as opposed to the fully-connected LAS-MIMO system. To obtain the optimal precoder, we first need to select a candidate $\mathbf{F}_{\text{LAS}}$ and the $\mathbf{F}_{\text{RF}}$, and then calculate the $\mathbf{F}_{\text{BB}}$ according to the (3.8). Since the constraint (3.7) for the fully-connected LAS-MIMO system contains $M \times N_L$ nonzero elements, the search process requires $2^{M \times N_L}$ possible $\mathbf{F}_{\text{LAS}}$, while $(2^B)^{N_L \times N_{\text{RF}}}$ requires possible $\mathbf{F}_{\text{RF}}$ quantities for $\mathcal{F}_{\text{LAS}}$, where B represents the bit resolution of the PSs. Therefore, $(2^B)^{N_L \times N_{\text{RF}}} \times 2^{M \times N_L}$ possible search can be obtained to design optimal $\mathbf{F}_{\text{LAS}}$, $\mathbf{F}_{\text{RF}}$, and $\mathbf{F}_{\text{BB}}$ which makes the computational cost prohibitive due to the large N_t in mmWave massive MIMO systems (i.e., for $M = 8$, $N_t = 64$, $N_L = 8$, and $B = 4$

Algorithm 3.1 Proposed Hybrid Precoding Algorithm of the Proposed Sub-grouped LAS-MIMO System

Require: $N_g, N_{\text{RF}}, \hat{\mathbf{H}}, \gamma, N_L, B$

▶ Input

Ensure: $R_{g,\max}$

▶ Output

```

1: Initialize  $\mathbf{V}_0 = \mathbf{I}_{N_{\text{RF}}}$ 
2: Build all possible sets for  $\mathcal{F}_{\text{LAS}}$  for the proposed sub-grouped LAS-MIMO design
3: for  $n = 1, 2, \dots, N_g$  do
4:   for  $\ell = 1, 2, \dots, K_{\text{sub}}$  do
5:     Define  $\theta_\ell \approx \hat{\theta}_\ell$  based on resolution value  $B$ 
6:     Define  $\mathbf{a}_t^\ell$  as in (3.14)
7:   end for
8:   for  $i = 1, 2, \dots$ , all possible set in  $\mathcal{F}_{\text{LAS}}$  do
9:     Select the beams in  $\mathbf{F}_{\text{LAS}}^{(n,i)}$  from  $\mathcal{F}_{\text{LAS}}$  beams set,
10:    Calculate  $\mathbf{F}_{\text{BB}}^{(n,i)}$  using (3.15),
11:    Calculate  $\mathbf{F}_{(n,i)} = \mathbf{F}_{\text{LAS}}^{(n,i)} \mathbf{A}_t^{(n,i)} \mathbf{F}_{\text{BB}}^{(n,i)}$ ,
12:    Calculate  $R_{(n,i)} = \log_2 \left( \left| \mathbf{I}_{K_{\text{sub}}} + \frac{\gamma}{N_{\text{RF}}} (\mathbf{F}_{(n,i)})^H \hat{\mathbf{H}}^H \mathbf{V}_{n-1}^{-1} \hat{\mathbf{H}} \mathbf{F}_{(n,i)} \right| \right)$ .
13:   end for
14:    $\mathbf{F}_{(n)}^{\text{opt}} = \arg \max R_{(n)}$ ,
15:    $R_n^{\max} = \max R_{(n)}$ ,
16:   Calculate  $\mathbf{V}_n = \mathbf{I}_{N_{\text{RF}}} + \frac{\gamma}{N_{\text{RF}}} \hat{\mathbf{H}} \mathbf{F}_n \mathbf{F}_n^H \hat{\mathbf{H}}^H$ .
17: end for
18:  $R_{g,\max} = \sum_{n=1}^{N_g} R_n^{\max}$ 

```

the number of searches required is $(2^4)^8 \times 2^{8 \times 8} \approx 8 \times 10^{28}$ for each RF chain).

On the other hand, in the proposed sub-grouped LAS-MIMO system, the number of searches required is significantly reduced because the precoders in the different subgroups are independent. Therefore, the search process is performed independently in each group. Moreover, the PS precoder is fixed and is set based on the direction of the users. Thus, for a given subgroup, the number of searches required for $\mathbf{F}_{\text{LAS}}^{(n)}$ is $2^{Q \times L_{\text{sub}}}$. Therefore, only $N_g \times 2^{Q \times L_{\text{sub}}}$ needs to be searched to find the appropriate precoder for the proposed sub-grouped LAS-MIMO design (i.e., considering $N_t = 64$, $N_L = 8$, $N_g = 4$ subgroups, and $Q = 8$ antennas per lens, the number of searches required is $4 \times 2^{8 \times 2} \approx 2.6 \times 10^5$). According to the numerical results regarding the number of required searches, the precoding algorithm proposed in **Algorithm 3.1** is suitable and affordable for the proposed sub-grouped LAS-MIMO system.

3.1.3.4. Energy efficiency

Energy efficiency can be defined as the number of bits that can be transmitted per unit of energy or as the ratio between the achievable sum rate and the total power consumed by the system, as in [37], [91], which can be expressed as follows

$$EE = \frac{C}{P_c + \frac{P_x}{\rho}} = \frac{\text{Bw } R}{P_c + \frac{P_x}{\eta_{\text{PA}} \eta_{\text{SW}}}}, \quad (3.16)$$

where R is the spectral efficiency of a given system. C , Bw , P_c , and P_x represent the system capacity, the available system bandwidth, the total circuit power consumed, and the total RF power to be transmitted by the antenna elements, respectively. In addition, η_{PA} is the power amplifier efficiency and η_{SW} is the efficiency of the switches, which can be expressed as $\eta_{\text{SW}} = 10^{-\zeta \text{IL}_{\text{SW}}/10}$, where ζ is the number of series switches needed to implement the architecture under each lens and IL_{SW} is the insertion loss for the switch.

In order to calculate the energy efficiency of the LAS-MIMO systems, a power model is proposed for **Figure 3.1a** in [37] where the power consumption is defined as

$$P_c = N_{\text{RF}} N_L P_{\text{PS}} + \zeta N_{\text{RF}} N_L P_{\text{SW}} + N_{\text{RF}} P_{\text{RF}}, \quad (3.17)$$

where P_{PS} , P_{SW} , and P_{RF} are the power consumption of PS, switches, and RF chain, respectively. For the proposed sub-grouped LAS-MIMO architecture shown in **Figure 3.1b**, the power consumption model of the fully-connected LAS-MIMO architecture given in (3.17) needs to be modified as

$$P'_c = N_{\text{RF}} \frac{N_L}{N_g} P_{\text{PS}} + \zeta N_{\text{RF}} \frac{N_L}{N_g} P_{\text{SW}} + N_{\text{RF}} P_{\text{RF}}. \quad (3.18)$$

3.2. Study II: Spatial-Wideband Effect In UWB Massive MIMO

3.2.1. System model and problem analysis

A mMIMO system with a BS equipped with M -antennas ULA is considered. Suppose there are L_p channel paths arriving at the BS where each path $\ell \in \{1, 2, \dots, L_p\}$ is associated with a passband complex gain α_ℓ , AoA/AoD $\hat{\theta}_\ell \in [-\pi/2, \pi/2)$, and a propagation delay $\tau_\ell \in [0, \tau_{\text{max}}]$, where τ_{max} is the channel's maximum delay spread. For a mMIMO system implementing very wideband signaling, there exists a non-trivial amount of delay

across the array elements (with respect to the first element), given by [19]

$$\Delta\tau_{m,\ell} = (m-1) \frac{d \sin \hat{\theta}_\ell}{c} \quad (3.19)$$

with c is speed of light, $d = \lambda_c/2$ being inter-element spacing where λ_c is the carrier wavelength, and $m \in \{1, 2, \dots, M\}$. Consequently, the total delay of an ℓ^{th} path observed by m^{th} element can be given as

$$\tau_{m,\ell} = \tau_\ell + \Delta\tau_{m,\ell}. \quad (3.20)$$

Accordingly, the CIR observed by the m^{th} element is given as

$$h_m(t) = \sum_{\ell=1}^{L_p} \beta_\ell e^{-j2\pi f_c \tau_{m,\ell}} \delta(t - \tau_{m,\ell}), \quad (3.21)$$

where β is the complex path gain, f_c is the carrier frequency given as $f_c = c/\lambda_c$ where c is the speed of light.

Taking the Fourier transform of (3.21) and with some simplification, the CFR of the considered link is obtained as [14]

$$h_m(f) = \sum_{\ell=1}^{L_p} \tilde{\beta}_\ell e^{-j2\pi(m-1) \frac{d \sin \hat{\theta}_\ell}{\lambda_c}} \underbrace{e^{-j2\pi(m-1) \frac{f d \sin \hat{\theta}_\ell}{c}}}_{\text{squint-inducing term}} e^{-j2\pi f \tau_\ell}, \quad (3.22)$$

where $\tilde{\beta}_\ell = \beta_\ell e^{-j2\pi f_c \tau_\ell}$ is the equivalent baseband path gain. Stacking the channels from all M antennas, the CFR vector over the whole array can be expressed as

$$\mathbf{h}(f) = \sum_{\ell=1}^{L_p} \tilde{\beta}_\ell \underbrace{\mathbf{a}_{\text{ideal}}(\hat{\theta}_\ell) \odot \mathbf{a}_{\text{squint}}(\hat{\theta}_\ell, f)}_{= \mathbf{a}_{\text{eff}}}(\hat{\theta}_\ell, f) e^{-j2\pi f \tau_\ell}, \quad (3.23)$$

where $\mathbf{a}_{\text{ideal}}(\hat{\theta}_\ell) = [1, e^{-j2\pi \frac{d \sin \hat{\theta}_\ell}{\lambda_c}}, \dots, e^{-j2\pi(M-1) \frac{d \sin \hat{\theta}_\ell}{\lambda_c}}]$ is the perfect spatial-domain steering vector and $\mathbf{a}_{\text{squint}}(\hat{\theta}_\ell, f) = [1, e^{-j2\pi \frac{f d \sin \hat{\theta}_\ell}{c}}, \dots, e^{-j2\pi(M-1) \frac{f d \sin \hat{\theta}_\ell}{c}}]$ is a frequency-dependent vector that induces beam-squinting problem in the system. As such, $\mathbf{a}_{\text{squint}}$ modifies $\mathbf{a}_{\text{ideal}}$ into a frequency-dependent effective steering vector $\mathbf{a}_{\text{eff}}$

$$\mathbf{a}_{\text{eff}}(\hat{\theta}_\ell, f) = \left[1, e^{-j2\pi(f+f_c) \frac{d \sin \hat{\theta}_\ell}{c}}, \dots, e^{-j2\pi(M-1)(f+f_c) \frac{d \sin \hat{\theta}_\ell}{c}} \right]^T. \quad (3.24)$$

Based on the array response $\mathbf{a}_{\text{eff}}(\theta, f)$ and a pre-designed beamforming vector $\mathbf{w} =$

$\mathbf{a}_{\text{ideal}}(\hat{\theta})$, the beam gain function is give by

$$\begin{aligned} g^{(\text{BS})}(\theta, f) &= \frac{1}{\sqrt{M}} \mathbf{w}^H \mathbf{a}_{\text{eff}}(\theta, f), \\ &= \frac{1}{\sqrt{M}} \sum_{m=1}^M e^{-j2\pi \frac{d}{c}(m-1)[(f+f_c) \sin \theta - f_c \sin \hat{\theta}]}. \end{aligned} \quad (3.25)$$

Assuming that $x(\theta, f) = \frac{d}{c}[(f + f_c) \sin \theta - f_c \sin \hat{\theta}]$, the beam gain can be written as

$$g^{(\text{BS})}(\theta, f) = \frac{1}{\sqrt{M}} \frac{\sin(\pi x(\theta, f)M)}{\sin(\pi x(\theta, f))} e^{-j\pi x(\theta, f)(M-1)}. \quad (3.26)$$

In multicarrier signaling, $\mathbf{a}_{\text{eff}}$, being frequency-dependent, directs signal pertaining to different subcarriers to different directions away from the desired direction. From (3.23), the effective direction of the signal at subcarrier k centered at $f_k = ((k - 1) - (N - 1)/2) \times \Delta f + f_c$, where $k = 1, 2, \dots, N$ with N and Δf being the Fast Fourier Transform (FFT) size and subcarrier spacing, respectively, can be found as [50], [92]

$$\theta(f_k) = \sin^{-1} \left(\frac{\sin \hat{\theta}}{1 + \frac{f_k - f_c}{f_c}} \right), \quad (3.27)$$

which does not align with the desired direction. The amount of incurred beam deviation due to k^{th} subcarrier can be given as

$$\Delta\theta(f_k) = |\hat{\theta} - \theta(f_k)|. \quad (3.28)$$

Likewise, the Beam Loss (BL) in the desired direction due to squinting caused by k^{th} subcarrier can be quantified by

$$\text{BL}(f_k) = \left| \frac{g(\hat{\theta}) - g^{(\text{BS})}(\hat{\theta}, f)}{g(\hat{\theta})} \right|_{f=f_k-f_c}, \quad (3.29)$$

where $g(\hat{\theta}) = g^{(\text{BS})}(\hat{\theta}, f)|_{f=f_k-f_c}$ and $f_k = f_c$ is the ideal beam gain in the desired direction in the absence of beam squinting which is equal to $\sqrt{M}$. This beam loss directly translates into the system capacity degradation. For a system with N_{RF} RF chains operating with bandwidth W under LoS channel and SNR γ , the system capacity is given by

$$C_{\text{TA}} = \frac{W}{N} \sum_{r=1}^{N_{\text{RF}}} \sum_{k=1}^N \log_2 \left[1 + \frac{\gamma}{N_{\text{RF}} \frac{W}{N}} |g^{(\text{BS})}(\hat{\theta}_r, f_{k,r})|^2 \right]. \quad (3.30)$$

From the above analysis, it can be deduced that the effect of beam squint on the system performance is determined by three main factors: the number of antenna elements M , system bandwidth with respect to the operating frequency W/f_c , and the desired beam

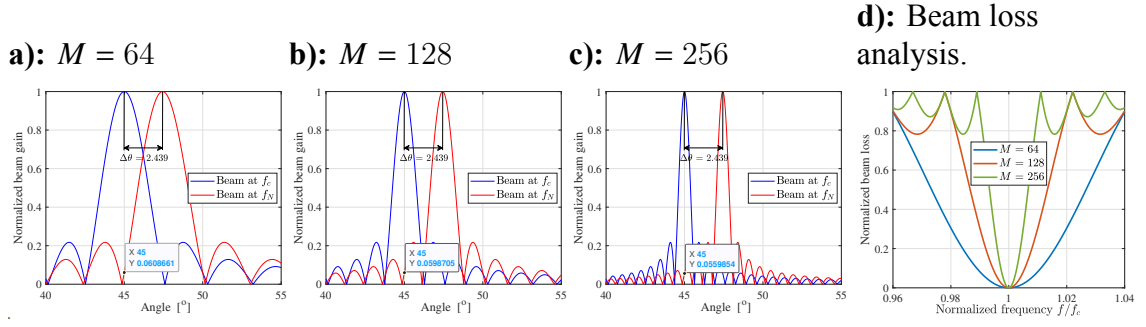


Figure 3.2: Normalized beam gain vs beam angle analysis at different antenna array size with $f_c = 60$ GHz.

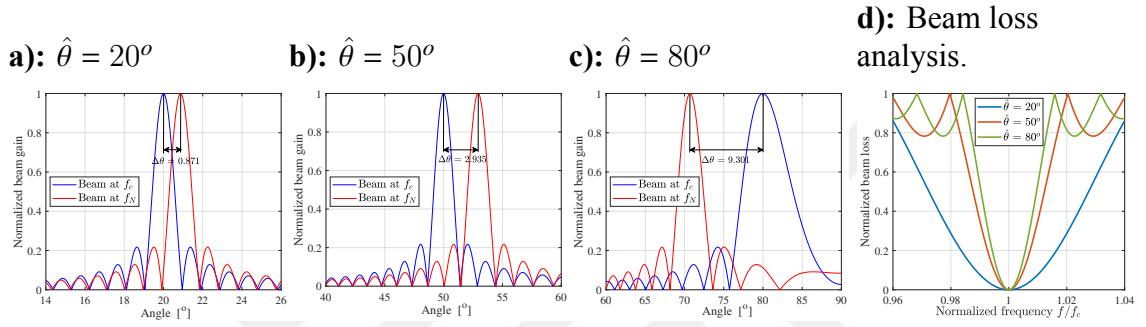


Figure 3.3: Normalized beam gain vs beam angle analysis at different beam direction with $f_c = 60$ GHz.

direction $\hat{\theta}$. **Figure 3.2**, **Figure 3.3**, and **Figure 3.4** illustrate the impacts of these factors on the beam directions across the subcarriers and the consequent beam loss. For clarity, only beams due to the middle subcarriers ($f_k = f_c$) and the furthest subcarrier ($f_k = f_N$) are plotted.

Figure 3.2a, **Figure 3.2b**, and **Figure 3.2c** show that as the array size increases, the maximum beam deviation from the desired direction remains the same. However, since the beams become narrower with large array sizes, the squinted beams tend to split from each other. The narrow squinted beams contribute less gain toward the desired direction as most of their energy is focused toward the squinted direction. This leads to higher beam loss as depicted in **Figure 3.2d**. Therefore, larger array dimensions worsen the effect of the beam squint phenomenon on the system performance.

In **Figure 3.3a**, **Figure 3.3b**, and **Figure 3.3c**, the impact of the $\hat{\theta}$ is illustrated. It is clear that the squinting phenomenon intensifies as the desired direction $\hat{\theta}$ deviates from the array's boresight. **Figure 3.3a** shows that when $\hat{\theta}$ aligns with arrays boresight (i.e., $\hat{\theta} = 0$) no squinting effect is observed regardless of the array size and the used bandwidth.

However, as the $\hat{\theta}$ deviates to the right or left of the array's boresight, beam squint comes into the picture, and its effect increases as $\hat{\theta}$ increases, as illustrated by **Figure 3.3b** and **Figure 3.3c**. Consequently, the beam loss also grows with $\hat{\theta}$ as shown in **Figure 3.3d**.

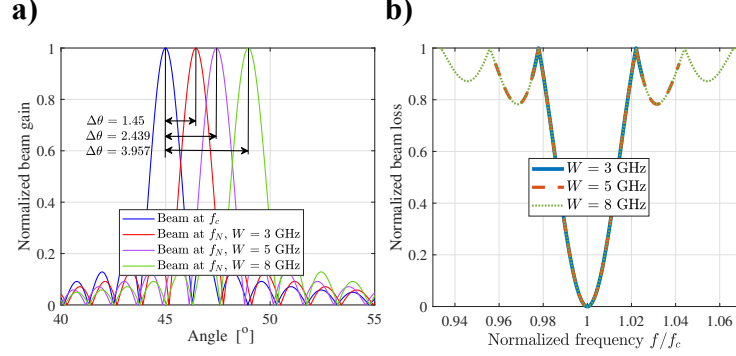


Figure 3.4: Normalized beam gain vs beam angle analysis at different fractional bandwidth with $f_c = 60$ GHz.

Figure 3.4 shows the effect of bandwidth W on the beam squint. The analysis is done for $W = 3$ GHz, 5 GHz, and 8 GHz. As the bandwidth increases, edge subcarriers incur more frequency deviations from the carrier frequency f_c . This causes them to incur a larger squinting effect as well as more beam loss which directly degrades the system's throughput.

Since the use of massive antenna arrays and UWB signaling is inevitable in the future wireless system, techniques to deal with the beam squint problem are highly necessary in order to ensure reliable performances in these systems. The transceiver design presented in the subsequent section is dedicated to minimizing the squinting problem taking the complexity and energy efficiency issues into account.

3.2.2. The proposed LAS-based system design

This section discusses the proposed transceiver design and its novel analog domain pre-coding technique to minimize the beam squint effect.

3.2.2.1. Proposed transceiver design

This work leverages the unique design of the LAS transceiver presented in our earlier work [37] that gives the inherent ability to control the signal in the analog domain.

The conventional LAS design consists of N_L discrete lenses with P antenna elements

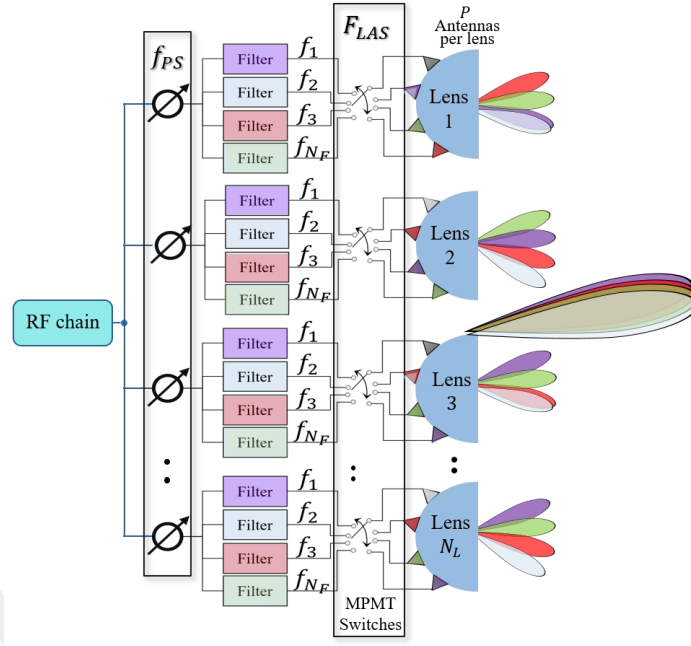


Figure 3.5: The proposed LAS analog transceiver design for a multicarrier signal in mMIMO systems.

connected to the focal surface of each lens. The total number of antenna elements under all lenses is equal M , placed in a way that the resultant antenna gain of the array is similar to half-wave spaced M antenna elements set in a ULA. The lenses are connected to the PS network by Single Pole Multiple Throws (SPMT) switches. One antenna element is activated under each lens by controlling the SPMT switching network to provide a sub-beam with a specific direction and low gain. As a result, one narrower beam with higher beamforming gain is produced for transmission from the combination of these sub-beams, which is approximately equal to the beam generated by a half-wave spaced M antenna elements placed in a ULA.

In this work, a modified LAS structure is proposed to specifically account for the beam squinting effect. The modified design features analog subband filters inserted between the PSs and the switching elements. The role of these filters is to chunk the wideband signal into groups of narrowband signals such that the subcarriers within each group experience a relatively similar amount of squinting. It also features the Multiple Poles Multiple Throws (MPMT) instead of the SPMT switching network for relaying all the narrowband groups to their respective antenna elements (based on the degree of their squinting) under the lenses, as shown in **Figure 3.5**.

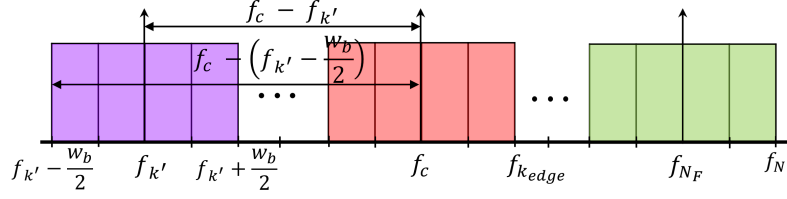


Figure 3.6: Sub-grouping the subcarriers in the subband filters.

At the transmitter, at each RF chain, the UWB signal passes through the PS network where each PS adds a fixed delay to the incoming RF signal based on the desired direction $\hat{\theta}$ ¹. As such, the PS precoder design is determined by $\mathbf{a}_{\text{ideal}}$ and can be given as

$$\mathbf{f}_{\text{PS}} = \frac{1}{\sqrt{N_L}} \left[1, e^{j\frac{2\pi d_L}{\lambda_c} \sin \hat{\theta}}, \dots, e^{j\frac{2\pi d_L}{\lambda_c} (N_L-1) \sin \hat{\theta}} \right]^T, \quad (3.31)$$

where d_L is the distance between the lenses assumed to be $d_L = P \times d$. The output of each PS is then passed through N_L groups of analog subband filters (each group has N_F parallel analog subband filters) to chunk the wideband signal into N_F narrowband signals of bandwidth W_b where $W_b = W/N_F$ and W is the system bandwidth. N_F is decided based on the amount of squint that the system can tolerate.

The extreme case for the design is to compensate the squint due to each individual subcarrier, which would require $N_L \times N$ filters for the whole design. To relax the design requirement, it is assumed that the squinting within the quarter of 3 dB beamwidth $\Omega_{3\text{dB}}$ is tolerable inside each subband, i.e.,

$$|\theta(f_{k_{\text{edge}}}) - \theta(f_{k'})| \leq \frac{\Omega_{3\text{dB}}}{8}, \quad (3.32)$$

where $\Omega_{3\text{dB}} \approx (0.89\lambda)/(Md)$ for ULA [93], $f_{k'} = ((k' - 1) - (N_F - 1)/2) \Delta f + f_c$ is the center frequency of the filter, with $k' = 1, 2, \dots, N_F$, and $f_{k_{\text{edge}}} = f_{k'} \pm W_b/2$ is the carrier frequency of a subcarrier at the edge of k' -th subband, see **Figure 3.6**.

¹Note that the multipath index ℓ is dropped here and in the subsequent equations just for the sake of notational simplicity.

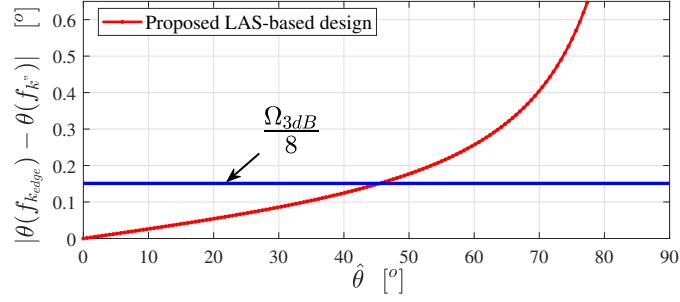


Figure 3.7: Squinting amount of a given subband for any desired direction within LAS's scanning range is within the tolerable range ($\Omega_{3dB}/8$). Here, (3.35) is used to calculate N_F , with $W = 5$ GHz and $f_c = 60$ GHz.

Using (3.27), (3.32) can be simplified to

$$\left| \sin^{-1} \left(\frac{\sin \hat{\theta}}{1 + \frac{\Delta f}{2f_c} (2k' - N_F - \frac{N}{N_F} - 1)} \right) - \sin^{-1} \left(\frac{\sin \hat{\theta}}{1 + \frac{\Delta f}{2f_c} (2k' - N_F - 1)} \right) \right| \leq \frac{\Omega_{3dB}}{8}, \quad (3.33)$$

s.t. $N \bmod N_F = 0$.

For simplicity, we consider the subband centered at f_c such that $f_{k'} = f_c$ and $k' = (N_F + 1)/2$. Consequently, (3.33) reduces to

$$\left| \hat{\theta} - \sin^{-1} \left(\frac{\sin \hat{\theta}}{1 + \frac{W}{2f_c N_F}} \right) \right| \leq \frac{\Omega_{3dB}}{8}, \quad (3.34)$$

s.t. $N \bmod N_F = 0$.

Considering that the severity of the beam squint increases as $\hat{\theta}$ grows (equation (3.19)), N_F is optimized based on the maximum scanning angle of the LAS system (i.e., $\hat{\theta} = \pi/4$ [37]) such that it is applicable for all users within the scanning range (See **Figure 3.7**). This leads to

$$N_F \geq \frac{W \sin(\pi/4 - \Omega_{3dB}/8)}{\sqrt{2}f_c(1 - \sqrt{2} \sin(\pi/4 - \Omega_{3dB}/8))}. \quad (3.35)$$

Here it is pertinent to mention that W in (3.35) should be the maximum bandwidth to be supported by the system such that the designed N_F is also applicable for arbitrary signal bandwidths below W . Note that when the signal bandwidth is less than the specified maximum bandwidth, some subbands (edge subbands in particular) may be turned off as they may contain no subcarrier.

Each narrowband signal passes through the switching mechanism using MPMT switches is modeled by a precoder F_{LAS} to the lenses for transmission. The MPMT switches allow simultaneous activation of N_F (one for each subband) antennas under each lens. For the n^{th} lens (where $n = 0, 1, \dots, N_L - 1$), the switching mechanism is represented by a selection vector $\mathbf{s}^{(n)} \in B^{P \times 1}$. If an antenna element is activated, its location in the selection vector is represented by logic 1. The activated p -th antenna element produces a beam at direction $\theta^{(n)}(p) = \frac{\pi}{4} - \frac{p}{(P-1)}\frac{\pi}{2}$, where P is the total number of antenna elements under each lens and $p = 0, 1, \dots, P - 1$ as explained in [37]. On the other hand, the inactivated elements' locations are represented by logic 0 in the selection vector $\mathbf{s}^{(n)}$. Therefore, the switch precoder matrix $\mathbf{F}_{\text{LAS}}$ is given as

$$\mathbf{F}_{\text{LAS}} = \begin{bmatrix} \mathbf{s}^{(0)} & \mathbf{0} & \dots & \mathbf{0} \\ \mathbf{0} & \mathbf{s}^{(1)} & & \mathbf{0} \\ \vdots & & \ddots & \vdots \\ \mathbf{0} & \mathbf{0} & \dots & \mathbf{s}^{(N_L-1)} \end{bmatrix}, \quad (3.36)$$

where $\mathbf{0}$ is a vector of size $P \times 1$ with all elements set to zeros. Note that the zeros in the $\mathbf{F}_{\text{LAS}}$ matrix indicate that there is no contribution between subarrays to the individual beams that they generate. For the k' -th subband filter, one antenna element is activated under each lens.

Since the lens generates a beam toward a specific direction based on the selected antenna to be activated under it, the binary selection vector $\mathbf{s}^{(n)}$ can be equivalent to [37]

$$\mathbf{s}^{(n)} = \frac{1}{\sqrt{P}} \left[e^{-j2\pi \frac{d}{c} f_c \frac{P-1}{2} \sin \theta^{(n)}}, \dots, 1, \dots, e^{-j2\pi \frac{d}{c} f_c \frac{-(P-1)}{2} \sin \theta^{(n)}} \right]^T. \quad (3.37)$$

Eventually, the effective RF precoder, $\mathbf{f}_{\text{RF}}$, for the proposed design is given as

$$\begin{aligned} \mathbf{f}_{\text{RF}} &= \mathbf{F}_{\text{LAS}} \mathbf{f}_{\text{PS}}, \\ &= \frac{1}{\sqrt{N_L}} \begin{bmatrix} 1 \times \mathbf{s}^{(0)} \\ e^{j2\pi \frac{d}{c} P f_c \sin \hat{\theta}} \times \mathbf{s}^{(1)} \\ \vdots \\ e^{j2\pi \frac{d}{c} P f_c (N_L-1) \sin \hat{\theta}} \times \mathbf{s}^{(N_L-1)} \end{bmatrix}. \end{aligned} \quad (3.38)$$

Therefore, the $(Pn + p)$ -th element in $\mathbf{f}_{\text{RF}}$ vector is given as

$$\{\mathbf{f}_{\text{RF}}\}_{(Pn+p)} = \frac{1}{\sqrt{N_L P}} e^{j2\pi \frac{d}{\lambda_c} P n \sin \hat{\theta}} e^{-j2\pi \frac{d}{\lambda_c} (\frac{(P-1)}{2} - p) \sin \theta^{(n)}}. \quad (3.39)$$

Let $\theta_{LAS}(k')$ be the k' -th subband's effective beam direction given by the proposed LAS design with arbitrary values of $\theta^{(n)}$'s across the lenses, then the maximum beam gain $g_{\max}$ of the design at k' -th subband is obtained as

$$\begin{aligned} g_{\max}(\theta_{LAS}(k'), f_{k'}) &= \mathbf{f}_{\text{RF}}^H \mathbf{a}_{\text{eff}}(\theta_{LAS}(k'), f_{k'}), \\ &= \frac{1}{\sqrt{N_L P}} \sum_{n=0}^{N_L-1} \sum_{p=0}^{P-1} e^{-j2\pi \frac{d}{c} (Pn+p)(f_{k'}+f_c) \sin \theta_{LAS}(k')} \\ &\quad \times e^{-j2\pi \frac{d}{c} f_c P n \sin \hat{\theta}} e^{-j2\pi \frac{d}{c} f_c (\frac{P-1}{2}-p) \sin \theta^{(n)}}. \end{aligned} \quad (3.40)$$

For each subband, the proposed design seeks to find the set $\Phi(k') = \{\theta^{(1)}, \theta^{(2)}, \dots, \theta^{(n)}, \dots, \theta^{(N_L)}\}$ that gives optimum $\theta_{LAS}(k')$, i.e.,

$$\theta_{LAS}^{\text{opt}}(k') = \min_{\Phi(k')} |\hat{\theta} - \theta_{LAS}(k')|. \quad (3.41)$$

This can be done by the exhaustive search approach whose complexity for each subband can be quantified as $O(P^{N_L} - k' + 1)$. Note that increasing the number of lenses in the design achieves better angle resolution in (3.41) to reach exact $\hat{\theta}$, which enhances the beam gain and thus system capacity. However, this may lead to an exponential growth of the precoding complexity. Therefore, an optimum number of lenses needs to be considered in the system design that achieves acceptable angle resolution with affordable complexity overhead.

3.2.2.2. Proposed threshold-based precoding

In order to reduce the complexity of the exhaustive search approach, a threshold-based search mechanism is proposed. In this case, a threshold is specified (based on the desired performance) to allow the selection of a sub-optimal set of antennas over the lenses for a given k' -th subband. Once the sub-optimal set that satisfies the specified threshold is found, the search process continues with the next subband. This avoids the searching overall $(P^{N_L} - k' + 1)$ possible combinations of the antenna under all lenses for a given subband, thereby reducing the complexity.

In the proposed precoder, it is assumed that, for the middle subband (i.e., $f_{k'}|_{k'=N_F/2}$), the beam deviation within a specific threshold (i.e., $\Omega_{3\text{dB}}$ of the desired beam direction) is acceptable. Since different subbands experience different amounts of squint, the threshold is scaled based on $f_{k'}$ with respect to f_c . Hence, the search seeks to find a sub-optimal set

Algorithm 3.2 The proposed LAS framework for beam squint compensation.

Require: $f_c, W, M, \hat{\theta}$. ▷ Input

Ensure: $\mathbf{F}_{\text{LAS}}^{(\text{opt})}$. ▷ Output

- 1: Calculate the optimum number of analog subband filters N_F as specified in (3.35).
 - 2: Build the proposed LAS design as illustrated in **Figure 3.5**.
 - 3: Set the PS precoder as in (3.31).
 - 4: Define all possible sets for $\mathbf{s}_t^{(n)}$ to build a possible $\mathbf{F}_{\text{LAS}}$.
 - 5: Define the half-power beamwidth $\Omega_{3\text{dB}}$ of the design.
 - 6: **for** $k' = N_F/2, N_F/2 \pm 1, \dots$ **do**
 - 7: Select the initial set of searching as $\Phi_{\text{init}}(k') = \{\hat{\theta}, \hat{\theta}, \dots, \hat{\theta}\}$.
 - 8: Calculate $\theta_{\text{LAS}}(k')$ that satisfies (3.40).
 - 9: **while** $|\hat{\theta} - \theta_{\text{LAS}}(k')| > \frac{\Omega_{3\text{dB}}}{2} \times \left(1 + \left|\frac{f_{k'} - f_c}{f_c}\right|\right)$ **do**
 - 10: Go to the next set $\Phi(k')$.
 - 11: Repeat step 8.
 - 12: **end while**
 - 13: Assign the sub-optimum set to the given subband and remove this set from the possible sets for the next subband.
 - 14: **end for**
 - 15: Build $\mathbf{F}_{\text{LAS}}$ from the optimum selected sets.
-

of antennas for the k' -th subband that satisfies

$$|\hat{\theta} - \theta_{\text{LAS}}(k')| \leq \frac{\Omega_{3\text{dB}}}{2} \times \left(1 + \left|\frac{f_{k'} - f_c}{f_c}\right|\right). \quad (3.42)$$

Since the squint causes beam deviation from the desired direction $\hat{\theta}$, the search process for a given subband begins around $\hat{\theta}$ by setting the initial beam direction under the lenses as $\Phi_{\text{init}}(k') = \{\hat{\theta}, \hat{\theta}, \dots, \hat{\theta}\}$. Note that each subband has a different amount of squinting, thus the combination of $\theta^{(n)}$'s that makes up the optimum set $\Phi(k')$ is unique for each subband. Therefore, such a combination can be removed from the search sets of the next subbands, thereby reducing the search complexity even further. Note that, with the $\Phi(k')$ initialization approach proposed above, the subband with the least squinting converges faster. As such, the searching process is initialized from the middle to the edge subbands to reduce the size of the search sets at the edge subbands.

The proposed LAS-based design with its precoding is summarized in **Algorithm 3.2**. Additionally, **Figure 3.8** provides a quick recap of how the proposed system operates. In a nutshell, for each subband, one antenna element is selected under each lens. Since the aim of the design is to minimize the beam squint for a given subband, the selected antenna combinations across the lenses should result in one best effective beam that has the

minimum deviation from the desired beam direction $\hat{\theta}$. For a design with N_L lenses and P antennas under each lens, there are P^{N_L} possible ways of selecting the antennas under the lenses for each given subband. To find the best effective beam, the exhaustive search algorithm can be employed to search over all P^{N_L} possible combinations, and the one that results in the best beam is selected for a given subband. Note that since different subbands experience different amounts of squinting, the optimum selected antenna combination across the lenses can differ for each subband. The beam search complexity can be reduced by employing the proposed threshold-based precoding algorithm instead of the full exhaustive search algorithm.

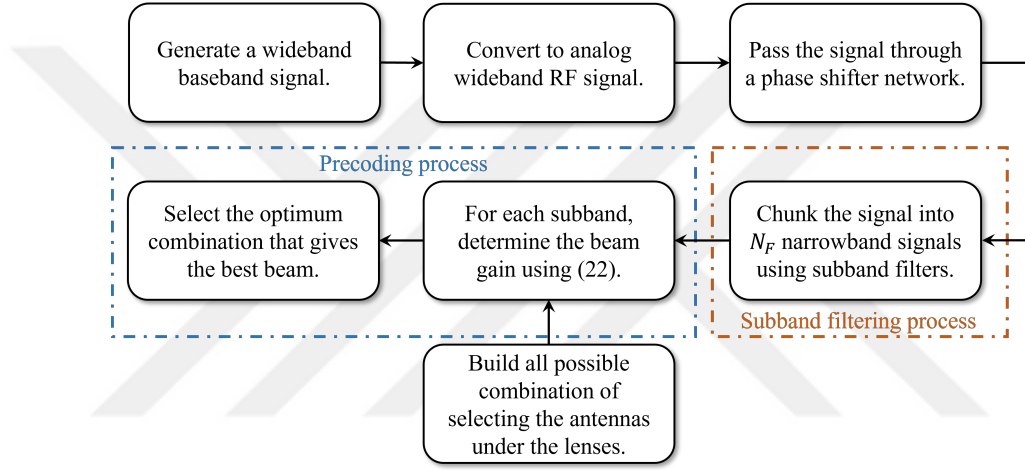


Figure 3.8: The signal flow chart in the proposed LAS design.

It is important to mention that, although the proposed design is described from the transmitter perspective, it can also be applied on the receiver side as long as the receiving node implements the mMIMO system and the beam squint problem exists.

3.2.2.3. Extension to multiple users scenario

Considering a multi-user UWB mMIMO system in which the BS makes use of the M antennas and N_{RF} RF chains to communicate with U single-antenna users where $U \leq N_{\text{RF}}$. The system is adjusted for transmission/reception based on its own structure as follows:

- In fully-connected hybrid mMIMO designs [24], the array size under each RF chain can be considered massive, and therefore the proposed design should be implemented in each RF chain. At this point, the RF precoder in (3.38) is extended

to multiple RF chains connection, and a digital precoder is needed in the system. Hence, the overall hybrid analog-digital precoder is given as

$$\mathbf{F} = \mathbf{F}_{\text{RF}}\mathbf{F}_{\text{BB}}, \quad (3.43)$$

where $\mathbf{F}_{\text{BB}} \in \mathbb{C}^{N_{\text{RF}} \times U}$ is the digital baseband precoder and the size of $\mathbf{F}_{\text{RF}}$ is $M \times N_{\text{RF}}$ where each column in $\mathbf{F}_{\text{RF}}$ represents the precoder design of the signal at one RF chain controlled in the analog domain as given in (3.38). Note that the proposed precoding in this work should be applied here to control the design and compensate for the beam squint effect.

- In a sub-connected hybrid mMIMO designs [94], if the size of the sub-connected part of each RF chain experiences a remarkable amount of squinting, the proposed design must be implemented on each subarray. Otherwise, the given sub-connected part is implemented as conventional MIMO systems.

At this point, it should also be mentioned that the conventional digital precoder $\mathbf{F}_{\text{BB}}$ based on singular value decomposition for multi-RF LAS structure [37] can be implemented with the proposed design when multiple RF chains are used. For example, a linear Zero-Forcing (ZF) precoder can be applied as a digital baseband precoder as $\mathbf{F}_{\text{BB}}^{\text{ZF}} = \mathbf{H}_{\text{eff}}^H (\mathbf{H}_{\text{eff}} \mathbf{H}_{\text{eff}}^H)^{-1}$ where $\mathbf{H}_{\text{eff}} = \mathbf{H}\mathbf{F}_{\text{RF}}$ and $\mathbf{H}$ is the channel matrix.

3.2.3. System analysis

In this section, the proposed design is analyzed in terms of capacity and power consumption.

3.2.3.1. Capacity

Once the beam squint effect is compensated, the lost beam gain is reclaimed, thus the capacity is improved toward the desired direction. With the proposed design, for a single RF chain, the capacity is given as

$$C_{\text{LAS}} = \sum_{k'=1}^{N_F} \frac{W_b}{N'} \sum_{k''=0}^{N'-1} \log_2 \left[1 + \frac{\gamma N'}{W_b} \left| g \left(\theta_{\text{LAS}}^{\text{opt}}(k'), f_{k''|k'} \right) \right|^2 \right], \quad (3.44)$$

where $\theta_{\text{LAS}}^{\text{opt}}(k')$, $f_{k''|k'}$ is the effective beam direction due to all lens elements for k'^{th} subband. $N' = W_b/\Delta f$ is the number of subcarriers within each subband, and $f_{k''|k'}$ is the

frequency of the k''^{th} subcarrier within k''^{th} subband. Here, it is important to reiterate that $\theta_{LAS}^{opt}(k')$ for k''^{th} subband is obtained based on the subband's center frequency $f_{k'}$, thus other subcarriers at $f_{k''} \neq f_{k'}$ within the subband may still cause some residual squinting. This is taken care of during the design of N_F in (3.35).

In order to compare the capacity of our proposed design with the capacity of (3.30) under beam squint effect, given that the ratio W_b/N' is fixed over all subbands which are equal to the ratio W/N in the conventional capacity formula, it is noticed that the capacity is mainly decided by the beam gain obtained in different physical directions at different subcarriers. This indicates that the beam gain loss caused by the beam squint effect is vital to the system capacity. Therefore, the capacity comparison can be simplified to beam gain comparison.

Based on (3.26) and (3.40), the ratio between the beam gain of the proposed design and the beam gain of the conventional mMIMO system under beam squinting effect at each subband frequency is given in

$$\frac{g(\theta_{LAS}^{opt}(k'), f_{k'})}{g^{(BS)}(\hat{\theta}, f_{k'})} = \frac{\frac{1}{\sqrt{N_L P}} \sum_{n=0}^{N_L-1} \sum_{p=0}^{P-1} e^{-j2\pi \frac{d}{c} (Pn+p)(f_{k'}+f_c) \sin \theta_{LAS}^{opt}(k')} e^{-j2\pi \frac{d}{c} Pn \sin \hat{\theta}} e^{-j2\pi \frac{d}{c} f_c (\frac{P-1}{2}-p) \sin \theta_{opt}^{(n)}}}{\frac{1}{\sqrt{M}} \sum_{m=1}^M e^{-j2\pi \frac{d}{c} (m-1) [(f_{k'}+f_c) \sin \hat{\theta} - f_c \sin \hat{\theta}]}} , \quad (3.45)$$

where $\theta_{opt}^{(n)}$ belong to the optimum set $\Phi_{opt}(k') = \{\theta_{opt}^{(1)}, \theta_{opt}^{(2)}, \dots, \theta_{opt}^{(n)}, \dots, \theta_{opt}^{(N_L)}\}$ that gives $\theta_{LAS}^{opt}(k')$.

Since the proposed design depends on several factors to achieve the best performance such as the number of subband filters, number of lenses, and the desired direction, the beam ratio in (3.45) varies at each subband to be less or larger than one. Therefore, a careful design with an optimum number of subband filters and lenses needs to be taken into account to achieve the best capacity performance. This has been numerically analyzed in **Figure 4.12** in the numerical results Section. Additionally, the effect of changing the number of lenses is discussed in the next subsection.

3.2.3.2. Power consumption

Due to the large antenna array structure in mMIMO systems, the power consumption of the circuits can account for a significant portion of the total system power consumption.

Although the transmit power is the main consumer in the transmitter design, the power consumption of the circuits in each transmit antenna is also a non-negligible element in relation to the number of transmit antennas due to other analog devices such as Digital-to-Analog Converter (DAC) as the main power consumers in the RF chain, PSs, and power amplifiers. The power consumption model of the conventional mMIMO system is analyzed in [37] and is given as

$$\mathcal{P}_c^{\text{TA}} = \underbrace{\frac{\mathcal{P}_{\text{tx}}}{\eta_{\text{TA}}\eta_{\text{PA}}}}_{\text{Effective Tx power term, } \mathcal{P}^{\text{TA}}} + \underbrace{N_{\text{RF}}M\mathcal{P}_{\text{PS}}}_{\text{PS network term}} + \underbrace{N_{\text{RF}}\mathcal{P}_{\text{RF}}}_{\text{RF term}}, \quad (3.46)$$

where $\mathcal{P}_{\text{tx}}$, η_{TA} and η_{PA} are the system's transmit power, array aperture efficiency and power amplifier's efficiency, respectively. For the conventional mMIMO, η_{TA} is usually assumed to be 90% [37]. $\mathcal{P}_{\text{PS}}$ and $\mathcal{P}_{\text{RF}}$ represent the amount of power consumed by a PS and an RF chain, respectively.

Similarly, the total power consumed by the proposed LAS design can be divided into four parts, given as

$$\mathcal{P}_c^{\text{LAS}} = \underbrace{\frac{\mathcal{P}_{\text{tx}}}{\eta_{\text{LAS}}\eta_{\text{PA}}\eta_{\text{SW}}\eta_{\text{filter}}}}_{\text{Effective Tx power term, } \mathcal{P}^{\text{LAS}}} + \underbrace{N_{\text{RF}}N_L\mathcal{P}_{\text{PS}}}_{\text{PS network term}} + \underbrace{N_{\text{RF}}N_L\mathcal{P}_{\text{SW}}}_{\text{switching network term}} + \underbrace{N_{\text{RF}}\mathcal{P}_{\text{RF}}}_{\text{RF term}}, \quad (3.47)$$

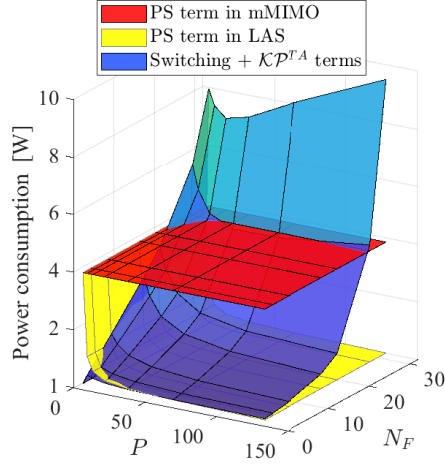
where η_{LAS} , $\eta_{\text{SW}} = 10^{-\text{IL}_{\text{SW}}/10}$ and $\eta_{\text{filter}} = 10^{-N_F\text{IL}_{\text{filter}}/10}$ are LAS aperture, switch, and subband filters power efficiency, respectively, with IL_{SW} and $\text{IL}_{\text{filter}}$ being insertion losses of the switch and the filter. The LAS aperture efficiency is generally assumed to be 80% due to lens/air mismatch [37].

In order to analyze the total power consumption by the conventional mMIMO and the proposed LAS design, we compare the corresponding terms in (3.46) and (3.47). Note that the power consumed by RF chains (i.e., the RF terms in (3.46) and (3.47)) are the same in both cases, thus, can be disregarded in the comparison. The effective power terms $\mathcal{P}^{\text{TA}}$ and $\mathcal{P}^{\text{LAS}}$ can be related by $\mathcal{P}^{\text{LAS}} = \mathcal{K}\mathcal{P}^{\text{TA}}$ where

$$\mathcal{K} = \frac{\eta_{\text{TA}}}{\eta_{\text{LAS}}\eta_{\text{SW}}\eta_{\text{filter}}}. \quad (3.48)$$

Since $\eta_{\text{LAS}} < \eta_{\text{TA}}$ and considering the extra insertion losses due to the filters and the switches, $\mathcal{K}$ is always greater than one, which makes $\mathcal{P}^{\text{LAS}} > \mathcal{P}^{\text{TA}}$. This means that the

a): Power consumption of each term



b): Total power consumption

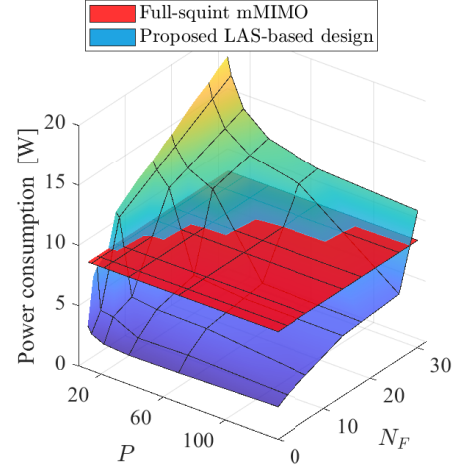


Figure 3.9: Power consumption vs P analysis. Here, both the conventional mMIMO and the proposed LAS-based designs follow parameters in **Table 4.2**.

LAS design requires more power to achieve the same effective isotropic radiated power as the conventional mMIMO design. Furthermore, the presence of switches (which are active circuitry elements) introduces an extra power consumption in the LAS design. Nevertheless, this extra power consumption in the LAS design is counteracted by a significant reduction in the number of PSs. The number of PSs in the LAS design is reduced by the factor of P (i.e., $M = N_L \times P$), which significantly reduces the PS term in (3.47) as compared to (3.46). The power consumed by these individual terms is illustrated in **Figure 3.9a** as a function of P and N_F . Apparently, the PS term in the conventional mMIMO is independent of both P and N_F , while the PS term in the proposed design only depends on P . This is plausible since, for a given M , the number of PS ($= N_L$ for $N_L > 1$) required in the LAS design decreases as P increases. Therefore, increasing P leads to the reduction of the total power consumed by the PS term. As the extreme case where $P = M$, i.e., single-lens design in which PS network is no longer required², the PS term vanishes from (3.47). For the switching and the effective power terms, their total power consumption varies with both P and N_F . For low N_F values, the total power consumed by these two terms decreases as P increases. On the other hand, for high N_F values, the total power varies convexly with P . Note that, with high N_F , the total power consumed by the proposed LAS design surpasses that of the conventional mMIMO, as shown by **Figure 3.9b**.

²In the LAS design, PS network is responsible for connecting the lenses by introducing the delays to the signal passed to each lens. As such, in the case of a single lens system, PS network is no longer required

In order to ensure low power consumption in the proposed design, the choice of P and N_F needs to be done carefully. Note that the choice of N_F is governed by (3.35). In order to facilitate power efficiency, the minimum N_F that satisfies (3.35) can be chosen. The choice of P , which is more flexible, can follow after choosing N_F . Note that, as it can be seen from **Figure 3.9b**, although higher values of P are advantageous in terms of power consumption, it compromises the controllability of the beam squint effect through the PS network due to the reduced number of lenses and PSs. On the other hand, lower values of P cause relatively high power consumption and it also leads to the loss of the beam squint controllability through the switching network. Therefore the best practice is to heuristically select a moderate value of P that facilitates both adequate controllability of the beam squint and low power consumption.

It is clear from the above analysis that the choice of P and N_F are critical for both beam squint controllability (which has a direct impact on the system capacity) and the total power consumption. Therefore, it is also pertinent to analyze the system's Energy Efficiency (EE) as a function of these two design parameters. The EE is generally defined as the ratio of the system capacity and its average power consumption, given as

$$EE = \frac{C_{\text{LAS}}}{P_c^{\text{LAS}}}. \quad (3.49)$$

3.2.3.3. Practical implementation

Since the proposed LAS architecture features some hardware design, it is important that we shed light on the practical feasibility and availability of the additional hardware components required to implement the architecture. In fact, there has been some immense effort invested in realizing the practical implementation of LAS systems similar to the one exploited in this work. For instance, the works of G. Mumcu in [36] and [95] have developed some prototypes of LAS architectures with 1D and 2D lenses, respectively, and provided detailed design steps. Apparently, the practical realization of the modified LAS architecture proposed in our work may follow similar design steps only with the insertion of the subband filter as an extra design step.

It is also important to highlight that, for the proper functioning of the proposed architecture, the required narrowband filters should be of high quality and operate at high-frequency bands. In the current state-of-the-art, such filters are bulky and expensive. However, there

has been active research on the realization of such filters based on Surface Acoustic Wave (SAW) and Bulk Acoustic Wave (BAW) technologies, which is expected to address both the bulkiness and cost issues [96].

Furthermore, our proposed LAS architecture replaces the SPMT switches in the traditional LAS with MPMT switches. Both SPMT and MPMT switch configurations have been widely studied and adopted in various practical systems [97], [98]. However, the limited number of poles in most of the commercially available MPMT switches may pose some difficulty in the realization of the proposed system, especially when a number of subbands N_F is huge. Nevertheless, higher-order MPMT switching configurations can also be achieved by using multiple SPMT switches arranged serially. High quality SPMT switches operating in a wide range of frequencies are widely available on the market at very low cost [99].

3.3. Study III: Spatial-Wideband Effect In UWB XL-MIMO Systems

3.3.1. System model and problem formula

An UWB multi-user mMIMO system is assumed, with a BS equipped with M antenna elements allocated in a ULA, serving K single-antenna users. For a far-field propagation assumption, at the ℓ -th channel path, the received uplink signal from the k -th user to the BS is given as [18]

$$\mathbf{y}_k(t) = \left[\alpha_\ell s_k(t - \tau_{m,\ell,k}) e^{j2\pi f_c(t - \tau_{m,\ell,k})} \right]_{m \in \mathcal{I}(M)}, \quad (3.50)$$

where α_ℓ is the complex channel coefficient at path ℓ , $s_k(t)$ is the transmitted signal at time t with carrier frequency f_c , and $\tau_{m,\ell,k}$ is the time delay of the signal that is sent to the m -th antenna element from path ℓ of user k given as $\tau_{m,\ell,k} = \tau_{\ell,k} + \Delta\tau_{m,\ell,k}$ where $\Delta\tau_{m,\ell,k} = (m - 1) \frac{d \sin \hat{\theta}_\ell^{(k)}}{c}$ is the delay across the array elements with respect to the first element with $d = \lambda_c/2$ is the antenna spacing, λ_c is the carrier wavelength, $\hat{\theta}_\ell^{(k)}$ is the desired AoA received from user k , c is the speed of light, $m \in \mathcal{I}(M) = \{m : m = 1, 2, \dots, M\}$. Removing the carrier frequency term $e^{j2\pi f_c t}$, the frequency domain of the equivalent baseband received signal at all paths is

$$\mathbf{y}_k(f) = \sum_{\ell=1}^{L_{p,k}} \left[\alpha_{\ell,k} s_k(f) e^{-j2\pi(f+f_c)\tau_{m,\ell,k}} \right]_{m \in \mathcal{I}(M)}. \quad (3.51)$$

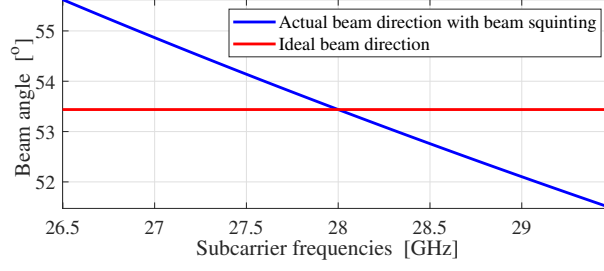


Figure 3.10: Relationship between subcarrier frequencies and the resultant beam direction with respect to the ideally desired direction. It is noted that the angle of the transmitted beam deviates from the exact direction when the frequency shifts from the center frequency f_c . This is due to the effect of squinting of the beam in UWB mMIMO systems. Here we assume $M = 128$, $f_c = 28$ GHz, the system bandwidth $W = 3$ GHz, the number of subcarrier frequencies $N = 1024$, and $\hat{\theta} = 2\pi \frac{(m-1)}{M}$ with $m = 20$.

Assuming $\mathbf{a}(\hat{\theta}_\ell^{(k)}, f)$ is the $M \times 1$ wideband array steering vector equal to

$$\mathbf{a}(\hat{\theta}_\ell^{(k)}, f) = \left[1, e^{-j2\pi(1+\frac{f}{f_c})\frac{d \sin \hat{\theta}_\ell^{(k)}}{\lambda_c}}, \dots, e^{-j2\pi(1+\frac{f}{f_c})\frac{(M-1)d \sin \hat{\theta}_\ell^{(k)}}{\lambda_c}} \right]^T. \quad (3.52)$$

Then, the frequency domain channel response vector of the system can be given as

$$\mathbf{h}_k(f) = \sum_{\ell=1}^{L_k} \alpha_{\ell,k} e^{-j2\pi f \tau_{\ell,k}} \mathbf{a}(\hat{\theta}_\ell^{(k)}, f), \quad (3.53)$$

where L_k is the number of channel paths of user k .

Observations reveal that delays in the received symbols induce a phase shift in the frequency domain, a phenomenon elucidated in [14], resulting in a departure from the intended direction. When transmitting multicarrier signals, such as OFDM with N subcarrier frequencies, each subcarrier, as governed by (3.52), exhibits a beam direction that deviates from the desired one, refer to **Figure 3.10**, primarily due to the influence of the beam squint effect, as discussed in [18]. It's noteworthy that the channel described in (3.53) is inherently frequency-dependent, with the magnitude and location of non-zero channel elements varying in accordance with the subcarrier index. Consequently, implementing analog beamforming based on this channel poses a significant design challenge.

3.3.2. Proposed beam gain analysis under beam squint effect

In this section, we first analyze the beam gain in far-field propagation given the system introduced in the previous section and then in the near-field propagation under the beam squint effect. After that, the beam squint interference region is defined for a specific

subcarrier frequency.

3.3.2.1. Beam gain in far-field propagation

For single beam generation in a phased antenna array, analog beamforming is implemented by a frequency-independent phase shifter network. In order to point the beam direction from these phase shifters to a specific angle $\hat{\theta}$, the analog beamforming vector of beam q is set to

$$\mathbf{w}_q = \mathbf{a}^H(\hat{\theta}^{(q)}), \quad (3.54)$$

where $\hat{\theta}^{(q)}$ is the unique spatial direction of beam q (i.e., LoS channel without beam squint effect)³. The beam pattern of beam q can be calculated from the array factor formula as [63]

$$\begin{aligned} \text{AF}_q(\theta, f) &= \mathbf{w}_q \mathbf{a}(\theta, f), \\ &= \sum_{m=1}^M e^{-j2\pi \frac{d}{\lambda_c} (m-1) \left[\left(1 + \frac{f}{f_c}\right) \sin \theta - \sin \hat{\theta}^{(q)} \right]}. \end{aligned} \quad (3.55)$$

Note that the maximum value of $\text{AF}_q(\theta, f)$ is obtained when $(1 + f/f_c) \sin \theta - \sin \hat{\theta}^{(q)} = 0$. Hence, the corresponding beam direction at a given frequency is

$$\theta(f) = \sin^{-1} \left(\frac{\sin \hat{\theta}^{(q)}}{1 + \frac{f}{f_c}} \right), \quad (3.56)$$

which indicates that for a different frequency, the beam is directed to a different angle.

3.3.2.2. Beam gain in near-field propagation

For an extremely large antenna array, the potential served users are located in the near-field region, which is within a distance less than Fraunhofer distance ($d_F \leq \frac{2D^2}{\lambda_c}$ where D is the array aperture). Hence, the delay $\Delta\tau_m$ across the antenna element to a given user is not only an angle-dependent but also a distance-dependent delay, i.e.,

$$\Delta\tau_m = \frac{r_m^{(q)}}{c}, \quad (3.57)$$

where $r_m^{(q)} = \sqrt{r_q^2 + (m-1)^2 d^2 - 2r_q(m-1)d \sin(\hat{\theta}^{(q)})}$ is the distance between the q -th user and the m -th antenna element by taking the distance r_q (the distance between the first

³Note that the multipath index ℓ is dropped here and in the subsequent equations due to the assumption that the system has only one path between the transmitter and receiver which is the LoS path.

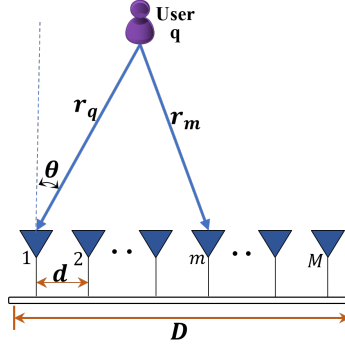


Figure 3.11: System model in near-field propagation.

antenna element and the q -th user) as a reference point, as illustrated in **Figure 3.11**.

Given that the magnitude of each element is a distance-dependent [100]; thus, the m -th antenna element in the channel steering vector for the q -th user is given as

$$a_m^{\text{NF}}(\hat{\theta}, f, r_q) = \frac{\sqrt{\beta_o}}{r_m} e^{-j \frac{2\pi}{\lambda_c} r_m (1 + \frac{f}{f_c})}, \quad (3.58)$$

where β_o is the channel power at a reference distance equal to 1 m [101]. The analog beamforming vector of beam q in the near-field propagation is given as

$$\mathbf{w}_q^{\text{NF}} = \left(\mathbf{a}^{\text{NF}}(\hat{\theta}^{(q)}, r_q) \right)^H. \quad (3.59)$$

The array factor formula in near-field propagation is given as

$$\begin{aligned} \text{AF}_q^{\text{NF}}(\theta, f, r_q) &= \mathbf{w}_q^{\text{NF}} \mathbf{a}(\theta, f, r_q), \\ &= \sum_{m=1}^M \frac{\beta_o}{r_m r_m^{(q)}} e^{-j \frac{2\pi}{\lambda_c} \left[\left(1 + \frac{f}{f_c} \right) r_m - r_m^{(q)} \right]}. \end{aligned} \quad (3.60)$$

Based on (3.60), the beam gain is maximized when $(1 + f/f_c)r_m - r_m^{(q)} = 0$. Hence, the corresponding beam pointing in the angular-distance domain at a specific frequency is given by

$$r_m(f) = \frac{r_m^{(q)}}{1 + \frac{f}{f_c}}, \quad (3.61)$$

observing that $r_m(f)$ is a function of the distance and angle given as $r_m(f) = \sqrt{r^2(f) + (m-1)^2 d^2 - 2r(f)(m-1)d \sin(\theta(f))}$. The deviation in the focusing point angle can be calculated under far-field assumption as in (3.56) while the deviation in the

distance can be calculated with respect to the first antenna element $m = 1$ as

$$r(f) = \frac{r_q}{1 + \frac{f}{f_c}}. \quad (3.62)$$

This indicates that the beam deviates in both angle and distance; thus, the beam focusing point is turned away from the targeted one.

3.3.2.3. Beam squint interference region

In multi-beam transmission, the estimation of desired beam angles in a sparsely populated environment can be classified into three categories: on-grid, off-grid, and gridless. The off-grid approach allows for more flexibility in choosing desired angles but still requires a grid. Off-grid and gridless methods can improve estimation performance and reduce error but also increase computational complexity and mutual coherence. On-grid sparse methods assume that the desired angles lie on a fixed set of grid points taken from $\Omega = \{0, \frac{2\pi}{M}, \dots, \frac{2\pi(M-1)}{M}\}$, which is suitable for applying existing approaches to sparse representation. However, this assumption only holds when the number of grid points is huge (i.e., M is massive). This work focuses solely on the on-grid approach due to its simplicity and suitability for mMIMO systems.

In far-field propagation, the on-grid method samples the grid points from a finite set to maintain orthogonality between generated beams. Hence, each far-field path is represented by only one grid point. However, since the amount of beam squinting is both angle-dependent and frequency-dependent, it is impossible to sample the grid points in a way that entirely avoids beam squinting. As a result, the beam direction may be displaced outside the grid, even if the desired direction $\hat{\theta}_\ell$ was selected from one of the on-grid points. Severe squinting amounts, such as at the edge grid points, can cause interference between neighboring beams within their main lobes. For example, given the beams in **Figure 3.12**, the interfered subcarrier f_I is appeared when the squinted angle of beam q is met with the squinted angle of beam k at the last subcarrier index f_N as

$$\begin{aligned} \theta^{(k)}(f)|_{f=f_N-f_c} &= \theta^{(q)}(f)|_{f=f_I-f_c}, \\ \sin^{-1}\left(\frac{\sin \hat{\theta}^{(k)}}{1 + \frac{f_N-f_c}{f_c}}\right) &= \sin^{-1}\left(\frac{\sin \hat{\theta}^{(q)}}{1 + \frac{f_I-f_c}{f_c}}\right). \end{aligned} \quad (3.63)$$

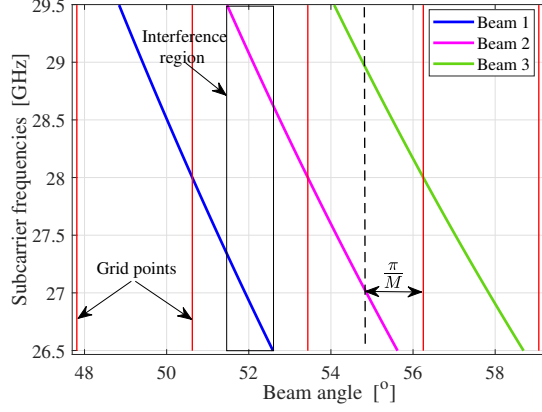


Figure 3.12: Multi-beam transmission with respect to the subcarrier frequencies. Here we assume $M = 128$, $f_c = 28$ GHz, the system bandwidth $W = 3$ GHz, and the number of subcarrier frequencies $N = 1024$.

Removing $\sin^{-1}$ from both side, (3.63) becomes

$$\frac{\sin \hat{\theta}^{(k)}}{1 + \frac{f_N - f_c}{f_c}} = \frac{\sin \hat{\theta}^{(q)}}{1 + \frac{f_I - f_c}{f_c}}, \quad (3.64)$$

$$\left(1 + \frac{f_I - f_c}{f_c}\right) \sin \hat{\theta}^{(k)} = \left(1 + \frac{f_N - f_c}{f_c}\right) \sin \hat{\theta}^{(q)}.$$

Then, the interference subcarrier point of beam q to the edge of beam k can be given as

$$f_I = \frac{\sin \hat{\theta}^{(q)}}{\sin \hat{\theta}^{(k)}} f_N, \quad (3.65)$$

$$\text{s.t. } f_1 \leq f_I \leq f_N.$$

The area in beam q where the frequency is equal to or greater than the interfered subcarrier f_I ($f \geq f_I$) is called the beam interference region. The interfered subcarrier f_I is the subcarrier where adjacent beams begin to interfere with each other. One solution to interference is to turn off the subcarriers located within the interference region for both beams. However, due to the non-linear behavior of the IBI at the edge frequencies for different angles, a linear subcarrier scheduling for each user is not practical. Later in this work, an appropriate technique will be considered to reduce the IBI.

In near-field propagation, the beam is considered to be distance and angle-dependent [102]. Hence, as the beam is generated based on joint angle-distance information of each near-field path [103], the distances and angles are sampled from the entire angular-distance domain extending the on-grid approach from a 1D domain (i.e., angular domain) to a 2D

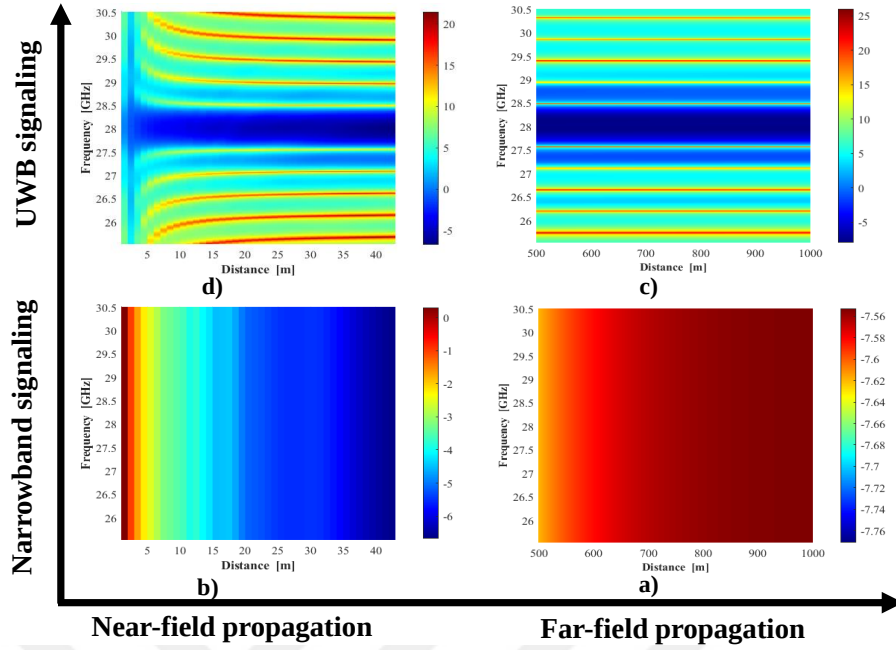


Figure 3.13: The normalized IBI pattern in the space (in dB). Four scenarios are considered as follows: **a)** narrowband signaling in far-field propagation, **b)** narrowband signaling in near-field propagation, **c)** UWB signaling in far-field propagation, and **d)** UWB signaling in near-field propagation. Here, it is assumed that the system operates at $f_c = 28$ GHz with bandwidth $W = 5$ GHz, $M = 256$, and $K = 20$ users distributed randomly in the service area with $M' = K$ beams where each user is served with one beam, while the desired user is located at $\hat{\theta} = \pi/6$.

domain (i.e., angular-distance domain, which is also called the polar domain).

Under the beam squint effect, the interference region of beam q to its neighboring beams is calculated based on $r_m^{(q)}$ in which the deviation in the beam focusing point is distance and angle dependent. Two beams are interfering at a specific subcarrier frequency f_I^{NF} when the phase of edge subcarrier f_N of beam k at antenna m is equal to the phase of beam q at the given subcarrier f_I^{NF} , i.e.,

$$r_m^{(k)}(f)|_{f=f_N-f_c} = r_m^{(q)}(f)|_{f=f_I^{\text{NF}}-f_c} ,$$

$$\frac{r_m^{(k)}}{1 + \frac{f_N-f_c}{f_c}} = \frac{r_m^{(q)}}{1 + \frac{f_I^{\text{NF}}-f_c}{f_c}} . \quad (3.66)$$

Hence, the interfering subcarrier point is given as

$$\begin{aligned}
f_I^{\text{NF}} &= f_N \frac{r_m^{(q)}}{r_m^{(k)}} , \\
&= f_N \sqrt{\frac{r_q^2 + (m-1)^2 d^2 - 2r_q(m-1)d \sin(\hat{\theta}^{(q)})}{r_k^2 + (m-1)^2 d^2 - 2r_k(m-1)d \sin(\hat{\theta}^{(k)})}} , \\
\text{s.t. } f_1 &\leq f_I^{\text{NF}} \leq f_N .
\end{aligned} \tag{3.67}$$

3.3.3. Proposed inter-beam interference analysis

In this section, the IBI model under the effect of beam squint is derived, and consequently, the system capacity is presented in near-field and far-field propagation.

3.3.3.1. Inter-beam interference

Due to the frequency-dependence property of the beam pattern, given in (3.55) and (3.60), in multi-beam transmission, the radiation of a beam at a certain frequency index interferes with the radiation of other beams (especially for the adjacent ones) causing IBI which can limit the achievable data rate of the system. The IBI from the k -th beam to the q -th beam can be defined as the radiation of beam k at a specific frequency f towards the radiation of beam q over the radiation of beam q .

Considering near-field propagation, the IBI from the m -th beam to the q -th beam is given as

$$\text{IBI}_{kq}^{\text{NF}}(\hat{\theta}^{(q)}, f, r_q) = \frac{\text{AF}_k^{\text{NF}}(\hat{\theta}^{(q)}, f, r_q)}{\text{AF}_q^{\text{NF}}(\hat{\theta}^{(q)}, f, r_q)}. \tag{3.68}$$

For a system with M' beams serving K users ($K \leq M'$), the total beam interference at beam q focused at a specific location for a given frequency is given as

$$\text{IBI}_q^{\text{NF}}(\hat{\theta}^{(q)}, f, r_q) = \sum_{\substack{k=1 \\ k \neq q}}^{M'} \text{IBI}_{kq}^{\text{NF}}(\hat{\theta}^{(q)}, f, r_q). \tag{3.69}$$

If the far-field assumption holds, the IBI becomes angle and frequency-dependent as

$$\text{IBI}_{kq}(\hat{\theta}^{(q)}, f) = \frac{\text{AF}_k(\hat{\theta}^{(q)}, f)}{\text{AF}_q(\hat{\theta}^{(q)}, f)}. \tag{3.70}$$

The total beam interference at beam q for a specific frequency in the far-field propagation

is given as

$$\text{IBI}_q(\hat{\theta}^{(q)}, f) = \sum_{\substack{k=1 \\ k \neq q}}^{M'} \text{IBI}_{kq}(\hat{\theta}^{(q)}, f). \quad (3.71)$$

Assuming a narrowband transmission with no beam squinting and far-field propagation, the IBI becomes only angle dependent as discussed in [66].

Figure 3.13 shows the IBI behaviour of a beam directed toward $\pi/6$ along different distances. In narrowband transmission, the IBI behaves almost the same over distances in the far-field propagation (see **Figure 3.13a**) while decreases as moving away from the antenna array in the near-field propagation as shown in **Figure 3.13b**. This is due to the complicated relationship between the electric and magnetic fields in the near-field region, which causes out-of-phase reactive components that return to the antenna elements. Hence, the interference between the transmitted beams increases abnormally, leading to a high IBI level. In wideband transmission, considering UWB signaling, the beam squint phenomena becomes dominant in the system and significantly affects the IBI behavior in both near-field and far-field propagation as shown in **Figure 3.13c** and **Figure 3.13d**. The beam squint effect causes beams at different frequencies to focus on different locations rather than the desired one, which causes interference to be at its maximum at specific subcarriers while minimized at other ones. This leads to the importance of having an appropriate mechanism to control the IBI level in the system based on the squinting amount.

3.3.3.2. System capacity

Considering the multi-beam mMIMO system design presented in this work, where the beams are generated and transmitted using an on-grid approach, the Shannon bounded capacity for beam q at subcarrier frequency f_n can be expressed as

$$\begin{aligned} C_q(f_n) &= W_q \log_2 (1 + \text{SINR}_q(f_n)), \\ &= W_q \log_2 \left(1 + \frac{P_r^{(q)}(f_n)}{P_I^{(q)}(f_n) + W_q N_o} \right), \end{aligned} \quad (3.72)$$

where N_o is noise power spectral density, W_q is the bandwidth allocated to beam q at the n -th subcarrier which is equal to $W_q = W/N$ where W is the allocated system bandwidth and N is the total number of subcarriers in the system. $P_r^{(q)}(f_n)$ is the received signal

power of beam q at subcarrier frequency f_n given as

$$P_r^{(q)}(f_n) = \frac{P_t^{(q)} G_t^{(q)}(f_n) G_r}{\text{PL}_q}, \quad (3.73)$$

where $P_t^{(q)}$ is the transmitted power allocated to beam q , G_t is the transmitted antenna array gain of beam q at the n -th subcarrier, G_r is the received antenna gain, and PL is the path loss model. The lossless passive antenna array gain at a given subcarrier is given as [104]

$$G_t^{(q)}(f_n) = \frac{2\pi |\text{AF}_{qq}|^2}{\int_0^{2\pi} |\text{AF}_q(\theta)|^2 d\theta}, \quad (3.74)$$

where AF_{qq} represents the gain of beam q at its desired direction $\hat{\theta}^{(q)}$ which follows (3.55) if the beam is propagated in far-field region or follows (3.60) if the beam is propagated in near-field region.

The path loss model adopted here is based on measurements made in [105] and given as

$$\text{PL}_q [\text{dB}] = \gamma + 10\beta \log_{10}(\rho_q) + \chi_\sigma, \quad (3.75)$$

where γ is the intercept in dB, β is the slope, ρ_q is the distance that the q -th beam is traveling to reach the receiver, and the shadowing factor χ_σ is a zero-mean Gaussian random variable with a standard deviation σ in dB.

The interference power P_I for the q -th beam at subcarrier frequency f_n is given by

$$P_I^{(q)}(f_n) = \sum_{\substack{k=1 \\ k \neq q}}^{M'} \frac{P_t^{(k)} \text{IBI}_{kq} G_t^{(q)}(f_n) G_r}{\text{PL}_k}. \quad (3.76)$$

where IBI_{kq} is the interference of beam k towards beam q follows either (3.70) for beam generated in the far-field or (3.68) for beams generated in the near-field region.

For M' beams in the system with N subcarriers in each beam, the total system capacity is given through the following expression

$$\begin{aligned} C_{\text{tot}} &= \sum_{k=1}^{M'} \sum_{n=1}^N C_k(f_n), \\ &= \sum_{k=1}^{M'} C_k, \end{aligned} \quad (3.77)$$

where $C_k = \sum_{n=1}^N C_k(f_n)$ is the beam capacity over all subcarriers.

The beam capacity C_q is analyzed in **Figure 3.14** where the beams are orthogonally

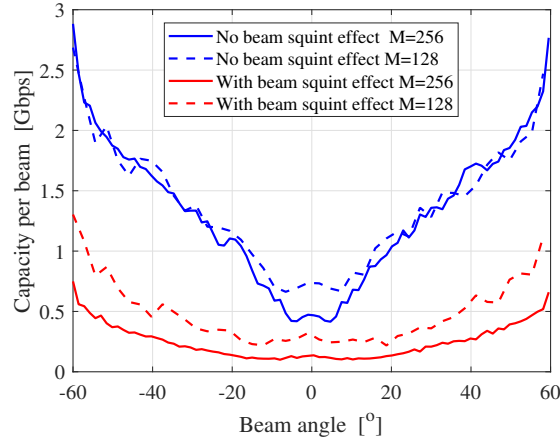


Figure 3.14: Capacity vs beam angle analysis. The solid line represents the beam capacity in the far-field propagation, while the dashed line represents the beam capacity in the near-field propagation. Here, it is assumed that the system operates at $f_c = 28$ GHz, $W = 5$ GHz, $K = 20$ users, and the desired user is located at d_F for far-field analysis and at $d_F/32$ for near-field analysis.

spread on fixed grid points taken from Ω given that the antenna array scans the range of $[-\pi/3, \pi/3]$. The figure shows that the beam capacity decreases in the near-field propagation compared to the far-field one due to the SW effect in increasing the IBI level as discussed in **Figure 3.13** by comparing the interference level between **Figure 3.13a** and **Figure 3.13b**. As the beam squint effect is considered in the system, the capacity drops for the beams pointing away from the 0° beam. This is because the beam squinting amount depends on three factors; system bandwidth with respect to the operating carrier frequency, antenna array size, and beam direction. By fixing the first two factors, the effect of beam direction distinctly appears in the capacity analysis, as presented in the figure.

The observation from **Figure 3.14** indicates a decline in beam capacity, notably a 12.4% reduction attributed to the influence of spherical wavefront phenomena in near-field propagation. Additionally, there is a 12.1% drop in capacity due to the beam squinting effect. When both the near-field and squinting effects are considered, the cumulative impact leads to a substantial 24.5% decrease from the maximum attainable beam capacity. This marked reduction significantly impairs the overall system capacity. Consequently, it becomes imperative to devise a mechanism to mitigate this degradation and effectively address the challenges posed by the beam squint effect, especially in the context of extremely large antenna arrays.

3.3.4. Proposed beam scheduling technique for IBI mitigation

Since beam squinting has a major effect on the system capacity compared to the non-stationarity of near-field propagation, in this Section, we propose a grouping mechanism to reduce the IBI and enhance the system capacity by considering the model of a transmitted beam with distance and angle-dependent under beam squint effect.

3.3.4.1. Proposed beam scheduling technique

Our proposed grouping method is rooted in evaluating the squinting amount of each beam within the system. This involves a meticulous assessment of subcarrier frequencies, which are intelligently divided into distinct subgroups according to their respective squinting characteristics. As depicted in **Figure 3.15**, these subgroups are dynamically toggled between active and inactive modes. This strategic modulation of subcarrier subgroups plays a pivotal role in minimizing interference between beams. By strategically organizing the subcarriers in this manner, we effectively mitigate the impact of interference, improving the overall performance and reliability of the system. This dynamic grouping mechanism is designed to adapt to changing conditions and optimize resource utilization, ensuring efficient and robust operation in various scenarios.

The proposed grouping mechanism operates as follows:

- At a given beam angle $\hat{\theta}$, considering a specific system bandwidth W , operating frequency f_c , and antenna array size M , and assuming a multicarrier system design, the first step involves calculating the squinting amount. This is defined as the extent to which the beam deviates from its desired direction at a given frequency, i.e.,

$$\text{BS} = |\hat{\theta} - \theta(f)|. \quad (3.78)$$

To divide the total subcarriers in the system into N_g groups, the assumption is made that subcarriers within one group experience almost the same squinting amount as the center subcarrier in that group. This assumption is based on considering that the squinting amount between the edge subcarrier $f_{k_{\text{edge}}}$ and the center one $f_{k'}$ within the group is $\text{BS} \leq \Omega_{3\text{dB}}/8$ [4], where $\Omega_{3\text{dB}}$ is the 3 dB beamwidth, approximately

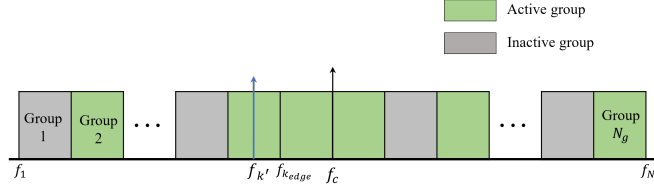


Figure 3.15: The concept of subcarrier grouping in the proposed sub-grouped beam-based scheme.

$\Omega_{3\text{dB}} \approx (0.9\lambda_c)/(Md)$ for a ULA design [93]. In other words,

$$|\theta(f_{k_{\text{edge}}}) - \theta(f_{k'})| \leq \frac{\Omega_{3\text{dB}}}{8}, \quad (3.79)$$

where $f_{k'} = ((k' - 1) - (N - 1)/2) \times \Delta f + f_c$ with N and Δf being the FFT size and subcarrier spacing, respectively.

- Continuing with the assumption of $f_{k'} = f_c$ and $k' = (N_g + 1)/2$, the number of groups N_g for beam k can be calculated by satisfying the condition

$$\left| \hat{\theta}^{(k)} - \sin^{-1} \left(\frac{\sin \hat{\theta}^{(k)}}{1 + \frac{W}{2f_c N_g^{(k)}}} \right) \right| \leq \frac{\Omega_{3\text{dB}}}{8}, \quad (3.80)$$

s.t. $N \bmod N_g^{(k)} = 0$.

Hence, $N_g^{(k)}$ is calculated as

$$N_g^{(k)} \geq \frac{W \sin(\hat{\theta}^{(k)} - \Omega_{3\text{dB}}/8)}{2f_c [\sin \hat{\theta}^{(k)} - \sin(\hat{\theta}^{(k)} - \Omega_{3\text{dB}}/8)]}. \quad (3.81)$$

- Within beam k , the groups that are turned inactive to mitigate the effects of IBI in the system are determined by $N_I^{(k)} = \log_2(N_g^{(k)})$. This strategic selection of inactive groups plays a crucial role in minimizing interference and optimizing overall system performance.

By strategically deactivating specific groups, the proposed approach effectively mitigates the impact of IBI, contributing to improved signal quality and enhanced system efficiency. The location of inactive groups is meticulously chosen to minimize IBI, thereby boosting the overall system capacity, as discussed herein. Assuming $s^k = 1, 2, \dots, N_g^{(k)}$ and $s' = s'^1, s'^2, \dots, s'^{M'}$, where s'^k represents the active group index, a subset of set s^k , the optimal

Algorithm 3.3 Sub-grouped beam-based scheduling algorithm.

Require: f_c, W, M, M' .

▷ Input

Ensure: s'_{opt} .

▷ Output

- 1: Define s'_{opt} as an empty set, $s'_{\text{opt}} = \{\}$.
 - 2: **for** $k = 1, 2, \dots, M'$ **do**
 - 3: Calculate $N_g^{(k)}$ as given in (3.81).
 - 4: Calculate $N_I^{(k)}$ as $N_I^{(k)} = \log_2 N_g^{(k)}$.
 - 5: Calculate $N_A^{(k)}$ as $N_A^{(k)} = N_g^{(k)} - N_I^{(k)}$.
 - 6: Find the optimum set s_{opt}^k using (3.83).
 - 7: Fill the set $s'_{\text{opt}} = \{s'_{\text{opt}}, s_{\text{opt}}^k\}$.
 - 8: **end for**
-

locations of the active groups are computed as follows

$$s'_{\text{opt}} = \arg \max_{s'^q} \sum_{k=1}^{M'} \sum_{n=1}^N b_n C_k(f_n), \quad (3.82)$$

where $b_n \in \{0, 1\}$ is an indicator associated with the subcarrier index that belongs to a specific group, equal to 1 if the index n belongs to the active group, otherwise $b_n = 0$.

In order to solve (3.82), an optimum way is to exhaustively search for the optimum location of the active groups for all beams together, where this joint maximization problem is very complex. Therefore, for a more practical approach, the optimization problem in (3.82) is simplified to

$$s'_{\text{opt}} = \arg \max_{s'^k} \prod_{n=1}^N b_n \text{SINR}_k(f_n). \quad (3.83)$$

The optimization process commences with the first beam ($k = 1$), considering that the other beams remain ungrouped at this stage. Subsequently, the algorithm determines the optimal location for the active groups within the current beam. Once this optimization is complete, the algorithm shifts its focus to the next beam, initiating the optimization process anew. This sequential progression persists until all beams successfully identify their optimal locations for the active groups. This approach is summarized in **Algorithm 3.3**.

Broadly speaking, in the far-field propagation, the distance-dependent channel steering vector (3.58) can be approximated to the distance-independent one as in (3.52). Hence, the proposed sub-grouped beam-based scheduling approach can be used in both near-field and far-field propagation since the users are assumed to be distributed uniformly within the cell coverage where they can be inside or outside the Fraunhofer zone (i.e., $\leq d_F$).

As discussed, the iterative optimization algorithm provides adaptability by dynamically adjusting to each beam's unique characteristics, optimizing active group locations based on specific requirements. Notably, this iterative optimization algorithm effectively addresses the beam squinting problem, ensuring accurate and optimized beam configurations. Focusing on one beam at a time ensures precise and tailored optimization, enhancing computational efficiency and avoiding unnecessary complexity. This approach facilitates convergence towards optimal solutions for each beam, gradually refining active group locations.

3.3.4.2. Complexity evaluation

An optimal solution to problem (3.83) is to exhaustively search all possible locations of the inactive groups for all beams jointly. This requires $\prod_{k=1}^{M'} C_{N_I^{(k)}}^{N_g^{(k)}}$ possible sets to search for the optimum set that maximizes the overall system capacity, where $C_{N_I}^{N_g} = \frac{N_g!}{N_I!(N_g - N_I)!}$. For mMIMO systems with a very large antenna array, the number of generated beams M' is high enough, which makes the computational cost prohibitive in the practical implementations. For example, for a UWB mMIMO system with $f_c = 28$ GHz, $W = 3$ GHz, $M = 256$, and with a scanning range between $-\pi/3$ and $\pi/3$ the number of possible beams $M' = 86$, by calculating the N_g of each beam using (3.81) and their corresponding N_I values, the total possible sets are immeasurable. On the other hand, when the proposed simplified search mechanism is applied, the required possible sets become $\sum_{k=1}^{M'} C_{N_I^{(k)}}^{N_g^{(k)}}$. For the same example mentioned above, the total possible sets in the proposed scheme can be measured and is around 5.7×10^{11} .

While the proposed iterative optimization algorithm may introduce some overhead due to its sequential approach, the potential benefits in terms of reduced computational load and targeted optimization make it a favorable alternative, particularly in scenarios where computational efficiency is a crucial consideration.

3.4. Study IV: PLA In BeamSpace MIMO

3.4.1. System model and assumptions

A mmWave MIMO system is considered a system model that consists of a legitimate node, Bob, that wants to authenticate the identity of transmitting node Alice, where an

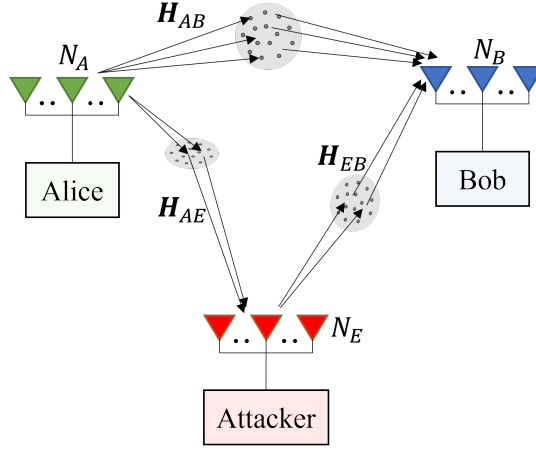


Figure 3.16: The mmWave MIMO system model.

attacker wants to impersonate the identity of Alice as shown in **Figure 3.16**. All the nodes are assumed to be equipped with a multi-antenna array. Alice, Bob, and the attacker are equipped with ULA of N_A , N_B , and N_E antenna elements, respectively. The channel is assumed to be a sparse multipath channel represented by a geometric channel model between different nodes and can be given as [81]

$$\mathbf{H}_{XY} = \sum_{l=1}^{L_p^{(XY)}} \beta_l \mathbf{a}_R^{(Y)}(\theta_{R,l}) \mathbf{a}_T^{(X)*}(\theta_{T,l}), \quad (3.84)$$

where $X \in \{A, E\}$ and $Y \in \{E, B\}$, such that $\mathbf{H}_{AB}$, $\mathbf{H}_{AE}$, and $\mathbf{H}_{EB}$ are the channels between Alice-Bob, Alice-attacker, and attacker-Bob, respectively. $L_p^{(XY)}$ is the number of channel path between node X and node Y , β_l is a complex channel gain which has complex normal distribution $\mathbb{CN}(0, 1)$, $\mathbf{a}_T^{(X)} \in \mathbb{C}^{N_X \times 1}$ is the steering vector of the transmitter, $\mathbf{a}_R^{(Y)} \in \mathbb{C}^{N_Y \times 1}$ is the steering vector of the receiver, θ_R is the AoA, and θ_T is the AoD, where both angles follow the uniform distribution $\sim \mathcal{U}(-\pi/2, \pi/2)$. For a ULA, the steering vector is given as

$$\mathbf{a}(\theta) = \frac{1}{\sqrt{N}} [1, e^{-j2\pi \frac{d_c}{\lambda_c} \sin(\theta)}, \dots, e^{-j2\pi(N-1) \frac{d_c}{\lambda_c} \sin(\theta)}]^T, \quad (3.85)$$

where λ_c is the carrier wavelength and d_c is the antenna element spacing, $N \in \{N_A, N_B, N_E\}$, and $\theta \in \{\theta_T, \theta_R\}$.

3.4.2. Proposed PLA scheme

This work exploits the joint effect of the hardware features and beamspace channel characteristics to build a robust PLA scheme for mmWave MIMO networks to tackle spoofing attacks.

3.4.2.1. Proposed distance signature

Multi-antenna technology enables the exploitation of the beamspace domain for enhancing communication and security. The beamspace channel representation is given as

$$\mathbf{H}_b^{(XY)} = \mathbf{U}_Y \mathbf{H}_{XY} \mathbf{U}_X^*, \quad (3.86)$$

where $\mathbf{U}_Y \in \mathbb{C}^{N_Y \times N_Y}$ and $\mathbf{U}_X \in \mathbb{C}^{N_X \times N_X}$ are unitary DFT matrices [106]. Enabling beamspace domain representation for the channel allows for more channel characteristics to be extracted. Using the beamspace channel matrix as a channel signature causes storage overhead when it is stored in the database, especially in a massive antenna array system. Therefore, this work proposes to extract a distance signature from the beamspace channel matrix based on a number of channel paths, paths' location, and the relative gain and phase of these paths, as explained herein. First, L_p ⁴ maximum principal components along with their location pairs (V_{AoD}, V_{AoA}) are extracted, where V_{AoD} is the virtual AoD and V_{AoA} is the virtual AoA. Then, the distance from reference point $(V_{AoD} = 0, V_{AoA} = 0)$ to the peak point $(V_{AoD}^{(l)}, V_{AoA}^{(l)})$ is calculated for each chosen principal component, $l = 1, 2, \dots, L_p$, as

$$d_l = \sqrt{\left(V_{AoD}^{(l)}\right)^2 + \left(V_{AoA}^{(l)}\right)^2}. \quad (3.87)$$

Afterward, the L_p distances are sorted based on the phase value of their corresponding principal component in descending order, and the result is stored in vector $\mathbf{d}$. **Figure 3.17** shows an example of a beamspace channel representation with $L_p = 6$, and $N = 64$ at both transmitter and receiver, where d_1 corresponds to the distance of the pairs (44,13) calculating as in (3.87). Accordingly, the other distances are calculated as d_2 all the way to d_6 . Then the distances are sorted in $\mathbf{d}$ vector based on the corresponding angles as

⁴To illustrate the superiority of the proposed scheme, we assume the worst case where the number of channel paths between nodes equals $L_p = L_p^{(AB)} = L_p^{(AE)} = L_p^{(EB)}$. Thus, the index (XY) is dropped here and in the following equations. If the number of channel paths at each node is different, this would be better for detecting spoofing attacks since the size of the proposed signature for the non-legitimate node differs from the one stored in the database.

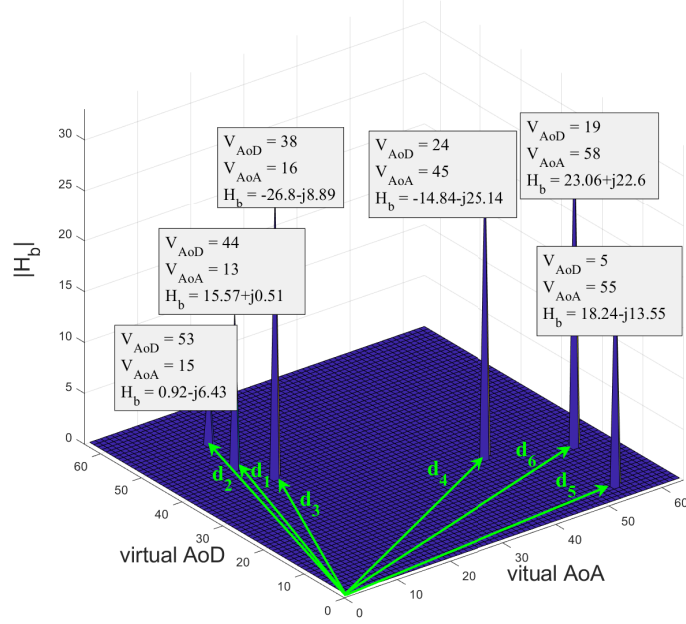


Figure 3.17: An example of extracting the proposed distance signature for the beamspace channel representation with $L_p = 6$ and $N = 64$.

$$\mathbf{d} = [d_6 \ d_1 \ d_5 \ d_2 \ d_4 \ d_3]^T.$$

3.4.2.2. Proposed hardware signature

Due to the use of multiple antenna elements in the MIMO system design, while placing the elements in a specific geometry, the electromagnetic waves transmitted/received from different antenna elements interact with each other, allowing some energy leakages between the elements, causing an effect that is known as the mutual coupling effect. Mutual coupling happens when the electromagnetic field generated from one antenna element permutes the circuit current distribution and distorts the radiation pattern of the other elements compared to their isolated radiation patterns. This effect degrades the overall system performance where different techniques are presented to solve this issue [107], [108].

In this work, the mutual coupling effect is considered a hardware feature for multiple antenna array design. It is investigated and used as a hardware signature to build a PL authentication system. The mutual coupling matrix $\mathbf{C} \in \mathbb{C}^{N \times N}$ can be defined as [109]

$$\mathbf{C} = (\mathbf{Z}_s + \mathbf{Z}_L)(\mathbf{Z} + \mathbf{Z}_L \mathbf{I}_N)^{-1}, \quad (3.88)$$

where Z_s is the antenna self impedance, Z_L is the load impedance connected to each antenna element, $\mathbf{Z}$ is the mutual impedance matrix for a specific antenna geometry design with diagonal elements equal to Z_s , and $\mathbf{I}_N$ is $N \times N$ identity matrix. In an ideal antenna array design, Z_L at all antenna elements is chosen to match Z_s for a maximum energy transfer; however, in order to strengthen the hardware signature-based authentication, a controlled or a random mismatching can be produced $Z_L = Z_L^{\text{ideal}} + e$ where e is controlled, or random error added to the system⁵. Similarly, the error can be generated at Z_s side instead, $Z_s = Z_s^{\text{ideal}} + e$. For different antenna polarization, antenna geometry design, and specific relative antenna separation, the self-impedance and corresponding mutual impedance matrix generate a unique mutual coupling signature for the hardware design. Hence, the mutual coupling effect for a specific MIMO device can be taken as a unique hardware signature to identify the given device.

3.4.2.3. Proposed hybrid signature

In this work, to have a more robust system against spoofing attacks, mutual coupling effect-based hardware signature and distance-based beamspace channel signature are jointly employed in the proposed PL authentication process under the name of hybrid signature. It is noticed that the mutual coupling of an antenna array interacts with the steering vector of the channel, causing changes in the gain and phase of the original channel [108]. Hence, an effective channel must be considered on Bob's side in order to build an authentication system. Assume that the channel between the Alice node and Bob node is $\mathbf{H}_{AB}$ and follows the channel model given in (3.84), with the mutual coupling effect, the effective channel that Bob needs to observe is

$$\mathbf{H}'_{AB} = \mathbf{H}_{AB}\mathbf{C}_A, \quad (3.89)$$

where $\mathbf{C}_A$ is the mutual coupling matrix of Alice's antenna array. Note that the signature now is termed as hybrid signature $\mathbf{d}'$ since it is extracted from a channel that is under the effect of mutual coupling (the effective channel $\mathbf{H}'_{AB}$).

⁵In this work, only ideal case is considered since any mismatching in the antenna array can lead to performance degradation in the system. However, a mismatching within a certain limit can be done as a future work considering the system error rate while enhancing the PLA performance.

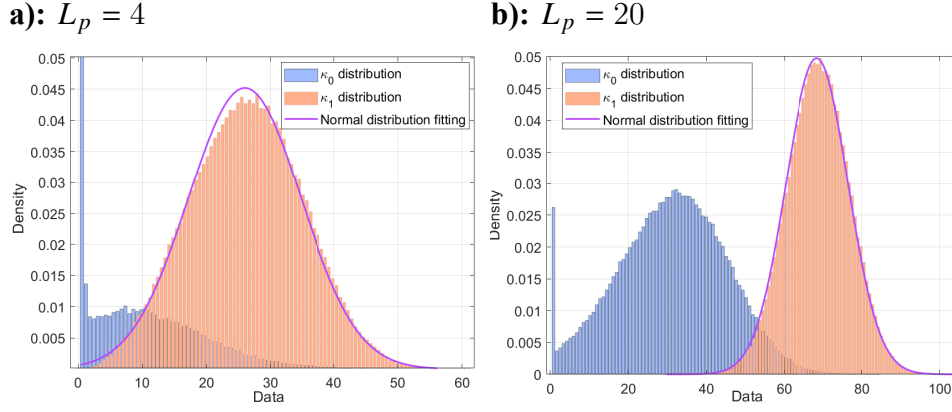


Figure 3.18: The distribution of η_L for the normal case K_0 and spoofing case K_1 for different channel sparsity levels. The distribution is obtained by simulating 500000 realizations generated from a geometric-based channel model with channel gain of complex normal distribution $\mathbb{CN}(0, 1)$ and the angles follow uniform distribution $\sim \mathcal{U}(-\pi/2, \pi/2)$. Then, the proposed distance signature is extracted from the channel. Lastly, fitting tools in [1] are utilized to find the best matching distribution for the fading amplitude of the actual hypothesis.

3.4.2.4. Detection model

The authentication scheme comprises two stages: training and verification. In the training stage, at time t , features of the legitimate node are extracted to make its fingerprints' database, where the identity of the legitimate node is verified through standard higher-layer protocols [110] before making the database. This can be done at Bob's side from the received noisy signal, where the received signal can be given as

$$y = \mathbf{C}_B \mathbf{H}'_{AB} \mathbf{s} + \mathbf{n}, \quad (3.90)$$

where $\mathbf{n}$ is additive white Gaussian noise (AWGN) with zero mean and variance σ^2 , $\mathbf{n} \sim \mathcal{CN}(0, \sigma^2)$. Note that Bob needs to compensate for his mutual coupling effect $\mathbf{C}_B$ before detecting the effective channel and starting the authentication process.

In the verification stage, occurring at time $t + 1$, Bob verifies the identity of received signal's source by detecting the effective channel $\mathbf{H}'^6$ and then extracting the signature $\mathbf{d}'(t + 1)$ from a noisy version of unknown transmitter's signal and comparing it with the

⁶Note that the channels at training stage and verification stage are estimated at Bob side using pilot signals.

database signature $\mathbf{d}'_{st}(t)$ using a binary hypothesis test given as

$$\begin{aligned} K_0 : \eta_L &= \|\mathbf{d}'(t+1) - \mathbf{d}'_{st}(t)\| \leq \eta_{th}, \\ K_1 : \eta_L &= \|\mathbf{d}'(t+1) - \mathbf{d}'_{st}(t)\| > \eta_{th}, \end{aligned} \quad (3.91)$$

where η_{th} is a threshold used to separate between K_0 and K_1 , where K_0 represents that the legitimate node Alice sent the received signal while K_1 represents that the legitimate node Alice did not send the received signal (an attacker may have sent the received signal). η_{th} is obtained using Neyman-Pearson testing-based spoofing attack detection. Using a fitting tool from [1] to observe the distribution of η_L for K_1 , a normal distribution curve can fit under it with specific mean and variance $\sim \mathcal{CN}(\mu_1, \sigma_1^2)$ as shown in **Figure 3.18**, where the mean and variance can be calculated from the used fitting tool. However, it is noticed from **Figure 3.18** that the distribution of η_L for K_0 can't have a fixed normal distribution as the channel sparsity level changes. Therefore, η_{th} can be easily calculated from the detection rate γ as

$$\begin{aligned} \gamma &= Pr(\eta_L > \eta_{th} | K_1), \\ &= \int_{\eta_{th}}^{\infty} Pr_{K_1}(\eta_L) dK_1, \\ &= Q\left(\frac{\eta_{th} - \mu_1}{\sigma_1}\right), \end{aligned} \quad (3.92)$$

where $Q(\cdot)$ is the tail distribution function of the standard normal distribution. Consequently, the threshold is given as

$$\eta_{th} = Q^{-1}(\gamma)\sigma_1 + \mu_1. \quad (3.93)$$

Accordingly, a threshold for deciding between the hypothesis K_1 and the hypothesis K_0 is obtained for a given value of γ .

It is assumed that the channel $\mathbf{H}_{AB}$ is updated as in [72] for time $t+1$ giving that the transmitter is in a mobility state, where the channel parameters follow first-order Gauss-Markov model [111] for updating.

The proposed PLA scheme is summarized in **Algorithm 3.4**.

Algorithm 3.4 The proposed PLA scheme in beamspace MIMO systems.

Require: Estimated effective channel $\check{\mathbf{H}}'_{AB}$. ▷ Input At time t

- 1: Convert $\check{\mathbf{H}}'_{AB}$ from spatial domain to beamspace domain using (3.86).
- 2: Extract the L_p maximum principal components with their location pairs.
- 3: Calculate the signature d'_i for each principal component as in (3.87) and sort them in descending order based on their principal component phase.
- 4: Store the proposed $\mathbf{d}'$ signature in the database of Bob under the name $\mathbf{d}'_{st}(t)$.

Require: Estimated effective channel $\check{\mathbf{H}}'$, η_{th} . ▷ Input At time $t + 1$

- 5: Convert $\check{\mathbf{H}}'$ from spatial domain to beamspace domain using (3.86).
- 6: Repeat the steps 2 and 3 to get the signature vector $\mathbf{d}'(t + 1)$.
- 7: Apply the binary hypothesis test given in (3.91).
- 8: **if** $\eta_L \leq \eta_{th}$ **then**
- 9: Bob authenticates Alice successfully
- 10: **else**
- 11: Bob detects an attacker
- 12: **end if**

3.5. Study V: Tracking-Based PLA In Beamspace Massive MIMO

3.5.1. System model and assumptions

A dynamic mmWave MIMO system model is considered with three entities: Alice, Bob, and an attacker, as shown in **Figure 3.19**. Bob is the legitimate base station that needs to authenticate Alice's identity, while the attacker seeks to impersonate Alice. All nodes in the system are assumed to have a multi-antenna array, specifically a one-dimensional ULA with N_A , N_B , and N_E antenna elements, respectively⁷. The sparse channel between different nodes is modeled as a time-varying geometric model [81] and expressed by the following equation

$$\mathbf{H}_{XY} = \sum_{l=1}^{L_p^{(XY)}} \beta_l \mathbf{a}_R^{(Y)}(\theta_{R,l}, v_{R,l}) \mathbf{a}_T^{(X)*}(\theta_{T,l}, v_{T,l}), \quad (3.94)$$

where $X \in \{A, E\}$ and $Y \in \{E, B\}$ represent the nodes in the system, and $\mathbf{H}_{AB}$, $\mathbf{H}_{AE}$, and $\mathbf{H}_{EB}$ denote the channels between Alice-Bob, Alice-attacker, and attacker-Bob, respectively. $L_p^{(XY)}$ is the number of channel path between node X and node Y . β_l is a complex channel gain with a complex normal distribution $\mathbb{CN}(0, 1)$. The steering vectors of the transmitter and receiver are denoted by $\mathbf{a}_T^{(X)} \in \mathbb{C}^{N_X \times 1}$ and $\mathbf{a}_R^{(Y)} \in \mathbb{C}^{N_Y \times 1}$. The relative Doppler shift of the transmitter and receiver movement, $v_{T,l}$ and $v_{R,l}$, are chosen from

⁷Note that the system can also be applied to a two-dimensional antenna array such as UPA and UCA.

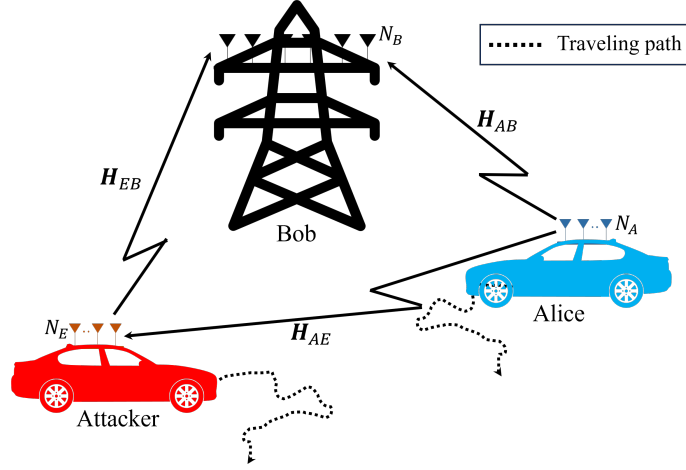


Figure 3.19: System model for a mmWave MIMO scenario with an attacker.

the set $v_l \in [-\frac{v_{\max}}{2}, \frac{v_{\max}}{2}]$, θ_R is the AoA, and θ_T is the AoD, where both angles follow the uniform distribution $\sim \mathcal{U}(-\pi/2, \pi/2)$. For a ULA structure, the steering vector is given as

$$\mathbf{a}(\theta, v) = \frac{1}{\sqrt{N}} \left[1, e^{-j2\pi \frac{d_c}{\lambda_c} \sin(\theta)(1+\frac{v}{f_c})}, \dots, e^{-j2\pi(N-1) \frac{d_c}{\lambda_c} \sin(\theta)(1+\frac{v}{f_c})} \right]^T, \quad (3.95)$$

where λ_c is the carrier wavelength and d_c is the antenna element spacing, $N \in \{N_A, N_B, N_E\}$, and $\theta \in \{\theta_T, \theta_R\}$.

The sparsity of mmWave channels in the beamspace domain makes the beamspace representation a better choice since it allows efficient representation and estimation of channel parameters, low overhead processing, and noise robustness. The spatial and beamspace domains are related by a spatial DFT represented by the unitary matrix $\mathbf{U}$ [106]. Hence, the beamspace representation of the spatial channel in (3.94) can be given by

$$\begin{aligned} \mathbf{H}_b^{(XY)} &= \mathbf{U}_Y \mathbf{H}_{XY} \mathbf{U}_X^*, \\ &= \sum_{l=1}^{L_p^{(XY)}} \beta_l \tilde{\mathbf{c}}_{Y,l} \tilde{\mathbf{c}}_{X,l}^*, \end{aligned} \quad (3.96)$$

where $\mathbf{U}_Y \in \mathbb{C}^{N_Y \times N_Y}$ and $\mathbf{U}_X \in \mathbb{C}^{N_X \times N_X}$ represent the unitary DFT matrices at receiver and transmitter side, respectively. $\tilde{\mathbf{c}}_l$ denotes the l -th path component in the beamspace, and is determined as

$$\begin{aligned} \tilde{\mathbf{c}}_l &= \mathbf{U} \mathbf{a}(\theta_l, v_l), \\ &= [\Upsilon(\bar{\psi}_l - \psi_1), \Upsilon(\bar{\psi}_l - \psi_2), \dots, \Upsilon(\bar{\psi}_l - \psi_N)], \end{aligned} \quad (3.97)$$

where $\tilde{\mathbf{c}}_l \in \{\tilde{\mathbf{c}}_{X,l}, \tilde{\mathbf{c}}_{Y,l}\}$, $\mathbf{U} \in \{\mathbf{U}_X, \mathbf{U}_Y\}$, $\theta_l \in \{\theta_{T,l}, \theta_{R,l}\}$, $v_l \in \{v_{T,l}, v_{R,l}\}$, and $\Upsilon(\Lambda) = \frac{\sin N\pi\Lambda}{\sin \pi\Lambda}$ is the Dirichlet sinc function. $\bar{\psi}_l = \frac{d_c}{\lambda_c}(1 + \frac{v_l}{f_c}) \sin \theta_l$, $\bar{\psi}_l \in \{\bar{\psi}_{T,l}, \bar{\psi}_{R,l}\}$, and $\psi_n = \frac{1}{N}(n - \frac{N+1}{2})$ is the spatial direction in the beamspace domain for $n \in \{n_X = 1, 2, \dots, N_X, n_Y = 1, 2, \dots, N_Y\}$.

3.5.2. Proposed narrowband tracking-based PLA approach

The PLA process proposed in this work consists of two stages: training and verification. In the training stage, Bob verifies Alice's identity using higher-layer protocols and extracts a unique physical layer signature from the received signal, which is then stored in a database [110]. This stage assumes that only Alice and Bob nodes exist. In the verification stage, Bob employs pilots to detect and extract the signature from a received signal, where a hypothesis test is built to compare the stored signature in the database with the extracted one and verify the identity of the transmitter node as legitimate or not. However, due to the high channel variation in dynamic environments where Alice, Bob, or both are moving at a specific speed, database signature mismatching can occur, leading to authentication failures. To address the signature mismatching problem, an additional stage, called the tracking stage, is introduced in this work. Through iterative updates over time, the tracking stage minimizes the mismatch between the signature observed in the training stage and the one in the verification stage, thereby improving the authentication system's performance. It should be noted that both the training and verification stages estimate the channels at Bob's end using pilot signals $\mathbf{s}$.

In this section, the procedure for extracting the proposed Enhanced Distance (ED) signature is discussed, and then the detection model is described. After that, the tracking process is presented.

3.5.2.1. ED signature extraction in narrowband transmission

MmWave channel exhibits a limited scattering nature, resulting in a sparse channel property while changing more rapidly than the sub-6GHz channels. Hence, a virtual sparse channel representation, also called a beamspace representation, is the natural choice for the mmWave MIMO channels. In this beamspace representation, channel parameters including AoDs and AoAs can be easily estimated and extracted using sparse estimation methods with low overhead processing. It is proved that the angle estimation is robust

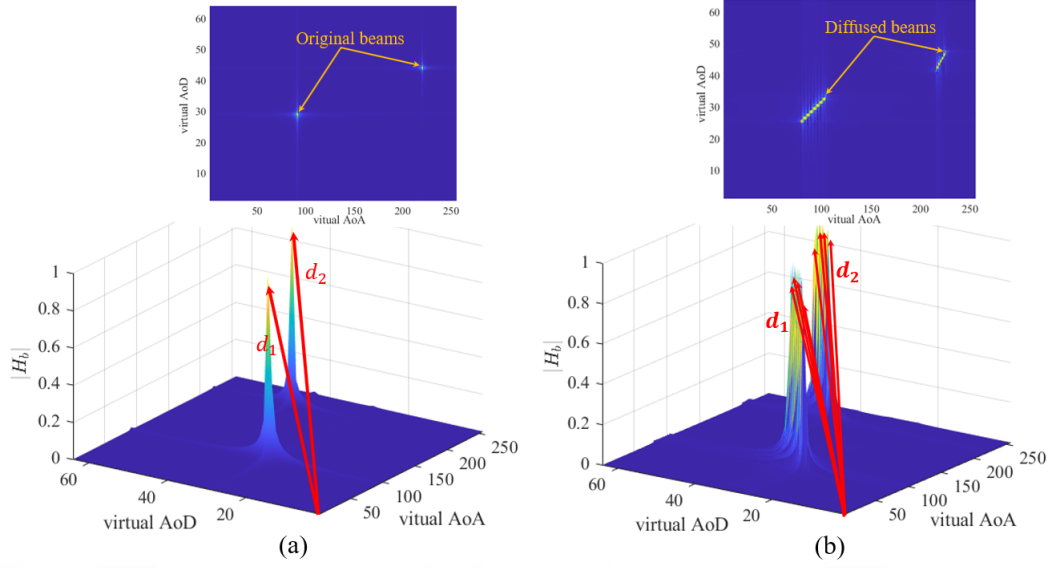


Figure 3.20: The beamspace channel representation **a)** without considering beam squint effect, and **b)** with beam squint effect.

to noise while presenting an accurate angle reciprocity even at low SNR levels. In this section, we propose a ED signature extraction method based on the channel parameters, AoDs and AoAs, in the beamspace representation of the MIMO channel.

The process of extracting the ED signature in the narrowband transmission is done during the training stage and can be described as follows:

1. At the first step, the system processes a channel estimation method to estimate the parameters AoAs, AoDs, and Doppler shift ν such as methods in [112]⁸.
2. Determine the index of the strongest element in the beamspace representation corresponding to the estimated AoAs and AoDs. This is done by finding the n value that makes Λ inside the Dirichlet sinc function in (3.97) equal to zero, i.e., optimum n for path l can be determined as

$$n^*(l) = \arg \min_{1 \leq n \leq N} \left| \frac{d_c}{\lambda_c} \left(1 + \frac{\nu_l}{f_c} \right) \sin(\theta_l) - \psi_n \right|. \quad (3.98)$$

Note that $n \in \{n_X, n_Y\}$, $N \in \{N_X, N_Y\}$, $\nu_l \in \{\nu_{T,l}, \nu_{R,l}\}$, and $\theta_l \in \{\theta_{T,l}, \theta_{R,l}\}$.

Since we are in the training stage, there are only Alice and Bob nodes, hence, X

⁸Due to the multi-dimensionality and sparsity of mmWave massive MIMO channels, tensor low-rank decomposition techniques can be fully exploited to obtain higher channel estimation accuracy with less pilot overhead. Therefore, a two-stage parallel factor tensor decomposition presented in [112] is employed to estimate the channel parameters in our work.

here refers to Alice (A) and Y refers to Bob (B). Therefore, $n_X = n_A$, $N_X = N_A$, $n_Y = n_B$, and $N_Y = N_B$.

3. Assuming that $(\psi_{n_X}^*(l), \psi_{n_Y}^*(l))$ is the optimum spatial direction pair at path l for AoA and AoD, respectively, refers to the optimum index selecting in (3.98), $n_X^*(l)$ and $n_Y^*(l)$. Assuming the reference point (0,0) as shown in **Figure 3.20a**, the distance for the pair from the reference point is calculated as

$$d_l = \sqrt{(\psi_{n_X}^*(l))^2 + (\psi_{n_Y}^*(l))^2}. \quad (3.99)$$

4. The distances in the ED vector signature $\mathbf{d}_e = [d_1, d_2, d_3, \dots, d_{L_p}]^T$ are sorted in descend order, i.e.,

$$d_1 > d_2 > d_3 > \dots > d_{L_p}. \quad (3.100)$$

5. At the end, the resulting ED signature $\mathbf{d}_e^{\text{st}}(t_0)$ is stored in the database at Bob's side.

3.5.2.2. Detection model

In the verification process that is based on the proposed ED signature, Bob first estimates the beamspace channel $\hat{\mathbf{H}}_b$ from the received signal $\mathbf{y}_b$ to detect its parameters. The received signal in the beamspace domain is given by

$$\mathbf{y}_b = \underbrace{\mathbf{U}\mathbf{H}\mathbf{U}^*}_{g(\mathbf{x}_{t_i})}\mathbf{s} + \mathbf{U}\mathbf{n}, \quad (3.101)$$

where $\mathbf{n}$ is a Additive White Gaussian Noise (AWGN) with zero mean and variance σ^2 , $\mathbf{n} \sim \mathcal{CN}(0, \sigma^2)$. The novel ED signature, denoted as $\mathbf{d}_e(t_1)$, is then extracted from the estimated beamspace channel.

Next, a binary hypothesis test is performed to verify the identity of the transmitted signal:

$$\begin{aligned} K_0 : \eta_L &= \|\mathbf{d}_e(t_1) - \mathbf{d}_e^{\text{st}}(t_0)\| \leq \eta_{th}, \\ K_1 : \eta_L &= \|\mathbf{d}_e(t_1) - \mathbf{d}_e^{\text{st}}(t_0)\| > \eta_{th}, \end{aligned} \quad (3.102)$$

where $\mathbf{d}_e^{\text{st}}(t_0)$ is the ED signature extracted and stored in the database in the training stage, and η_{th} is a threshold used to separate between K_0 and K_1 . Hypothesis K_0 corresponds to the case where the legitimate node, Alice, sent the received signal, while K_1 corresponds to the scenario where the legitimate node, Alice, did not send it (an attacker may have

done).

By observing the distribution of η_L as analyzed in [6], the value of η_L can be approximated to a normal distribution under a spoofing attack case $\sim \mathcal{CN}(\mu_1, \sigma_1^2)$. Hence, for a given detection rate value γ , the threshold η_{th} can be obtained following $\eta_{th} = Q^{-1}(\gamma)\sigma_1 + \mu_1$, where $Q(\cdot)$ is the tail distribution function of the standard normal distribution. Here, the detection probability γ is defined as the probability of correctly detecting a forged signal when Alice did not transmit, i.e., $\gamma = \Pr(\eta_L > \eta_{th} | K_1)$.

Under the assumption that the legitimate node channel and the attacker channel are independent, based on the first-order Gauss-Markov model [111], the channel correlation of the legitimate node can be represented

$$\mathbf{H}_{AB}^{t_i} = \sum_{l=1}^{L_p} \beta_l(t_i) \mathbf{a}_R^{(B)}(\theta_{R,l}(t_i), v_{R,l}) \mathbf{a}_T^{(A)*}(\theta_{T,l}(t_i), v_{T,l}), \quad (3.103)$$

where the channel parameters, $\theta_{R,l}$, $\theta_{T,l}$, and β_l , at successive time slots t_{i-1} and t_i can be mathematically expressed by

$$\begin{aligned} \beta_l(t_i) &= \rho \beta_l(t_{i-1}) + \varepsilon_1(t_{i-1}), \\ \theta_{T,l}(t_i) &= \theta_{T,l}(t_{i-1}) + \varepsilon_2(t_{i-1}), \\ \theta_{R,l}(t_i) &= \theta_{R,l}(t_{i-1}) + \varepsilon_3(t_{i-1}), \end{aligned} \quad (3.104)$$

with ρ is channel correlation coefficient between channel gain samples, $\varepsilon_1(t_{i-1}) \sim \mathcal{CN}(0, 1 - \rho^2)$, $\varepsilon_2(t_{i-1})$ and $\varepsilon_3(t_{i-1}) \sim \mathcal{U}(-\varrho, \varrho)$ [72].

3.5.2.3. Tracking process for the ED signature in narrowband transmission

In mmWave transmission, the channel information obtained may last for a short period due to the short channel coherence time at these bands. Then, the channel has to be estimated again in the later transmission. Hence, a frequent estimation and authentication process is needed. This increases the complexity overhead in the system, especially in an environment where the transmission nodes or the scatters are in a movement, i.e., a dynamic environment. Therefore, a tracking approach can be applied to reduce the estimation overhead.

A tracking stage is integrated into the PLA process to enhance the authentication performance in dynamic environments such as high-speed railways. This mechanism tracks and

updates the physical layer database signature in response to channel variations, resulting in a longer authorized connection period. This is possible because the beamspace channel parameters corresponding to Alice's node and signature change slowly, while an unauthorized node likely causes abrupt changes.

In our system model, the authentication process begins with initial channel acquisition to estimate the channel and authorize transmission nodes using upper-layer authentication techniques. Once channel estimation is completed and physical-layer channel features are extracted, the system enters a tracking mode. Various channel tracking techniques are employed to monitor temporal channel variations using a limited amount of radio resources. This tracking procedure is repeated in each transmission cycle and can be sustained for a defined period, denoted as T_o [113], aligning with processes outlined in the 5G NR standard [114]. The tracking mode concludes either when the transmission cycle exceeds T_o or when the connection between the nodes is lost due to sudden blockage or shadowing.

In the tracking stage, the first step is to define the state evolution and measurement models, as shown below

$$\begin{aligned} \mathbf{x}_{t_i} &= f(\mathbf{x}_{t_{i-1}}) + \mathbf{w}_{t_i}, \\ \mathbf{z}_{t_i} &= g(\mathbf{x}_{t_i}) + \mathbf{v}_{t_i}. \end{aligned} \tag{3.105}$$

Here, f denotes the state transition function, g represents the measurement model given in (3.101), $\mathbf{w}$ and $\mathbf{v}$ denote the process and observation noises, respectively. The noise vectors follow a Gaussian distribution with $\mathbf{w} \sim \mathcal{N}(0, Q)$ and $\mathbf{v} \sim \mathcal{N}(0, R)$. It is important to note that the state vector $\mathbf{x}$ captures the tracking parameters related to the physical layer characteristics of the developed signature, i.e., $\mathbf{x} = [\boldsymbol{\theta}_T, \boldsymbol{\theta}_R]^T$.

The model of the state transition function f for the tracked parameters $\mathbf{x}$ follows the first-order Gauss-Markov model, i.e., $\boldsymbol{\theta}_{t_i} = \boldsymbol{\theta}_{t_{i-1}} + \boldsymbol{\zeta}_{t_i}$ where $\boldsymbol{\zeta} \sim \mathcal{N}(0, \Sigma^2)$ [111], $\boldsymbol{\theta} \in \{\boldsymbol{\theta}_T, \boldsymbol{\theta}_R\}$ and $\Sigma \in \{\Sigma_T, \Sigma_R\}$. Note that the vectors $\boldsymbol{\theta}_T$ and $\boldsymbol{\theta}_R$ represent the angle values at all channel paths.

For a time-varying channel, channel gain β changes rapidly, while the angles θ_T and θ_R vary slowly. The angular variation can be measured according to the variation in Σ . Even with high Σ variation, since the correlation between the current angle and the previous one is high, the variation of the angle can be easily tracked over time.

To deal with the nonlinearity of the system, an Extended Kalman Filter (EKF) [115] is employed in this work due to its excellent performance and low computational complexity. The EKF effectively addresses error accumulation in track filtering by combining predictions based on a dynamic model with corrections from noisy measurements, thereby adapting to nonlinearity and noise in the system. It uses feedback to estimate the process state and divides the main iteration model into two stages: time propagation and observation update [111].

In time propagation, the filter updates the current state vector and error covariance according to the previous time information, i.e.,

$$\begin{aligned} \mathbf{x}_{t_i}^- &= f(\mathbf{x}_{t_{i-1}}), \\ \mathbf{P}_{t_i}^- &= \mathbf{A}_{t_{i-1}} \mathbf{P}_{t_{i-1}} \mathbf{A}_{t_{i-1}}^T + \mathbf{Q}, \end{aligned} \quad (3.106)$$

where $\mathbf{x}_{t_i}^-$ represents the priori estimate of the state vector at time t_i , $\mathbf{P}_{t_i}^-$ represents the covariance matrix of priori estimate, and $\mathbf{A}_{t_{i-1}} = \frac{df}{dx}|_{\mathbf{x}_{t_{i-1}}^-}$ is the Jacobian matrix of $f(\mathbf{x})$ at $\mathbf{x}_{t_{i-1}}^-$.

In the observation update, the filter gain is combined with the prior estimate and observation values to form the posterior estimates, i.e.,

$$\begin{aligned} \mathbf{K}_{t_i} &= \mathbf{P}_{t_i}^- \mathbf{H}_{t_i}^T \frac{1}{\mathbf{H}_{t_i} \mathbf{P}_{t_i}^- \mathbf{H}_{t_i}^T + \mathbf{R}} \\ \mathbf{x}_{t_i} &= \mathbf{x}_{t_i}^- + \mathbf{K}_{t_i} (\mathbf{y}_b - g(\mathbf{x}_{t_i}^-)) \\ \mathbf{P}_{t_i} &= (\mathbf{I} - \mathbf{K}_{t_i} \mathbf{H}_{t_i}) \mathbf{P}_{t_i}^- \end{aligned} \quad (3.107)$$

where $\mathbf{K}_{t_i}$ is the filter gain and $\mathbf{H}_{t_i} = \frac{dg}{dx}|_{\mathbf{x}_{t_i}^-}$ is the Jacobian matrix of $g(\mathbf{x})$ at $\mathbf{x}_{t_i}^-$.

Note that the initial stage of the filter is taken from the physical layer characteristics extracted from the estimated channel at time t_0 . As long as the transmitter node is authorized (i.e., the case K_0 occurs showing that the authentication is successful), within tracking duration T_o , Bob tracks the legitimate node's physical layer characteristics and updates its database. The process returns to the first step if authenticating the transmitted node fails (i.e., the case K_1 occurs pointing out that the authentication has failed) or the system exceeds the tracking period. **Figure 3.21** summarizes the proposed tracking-based PLA mechanism.

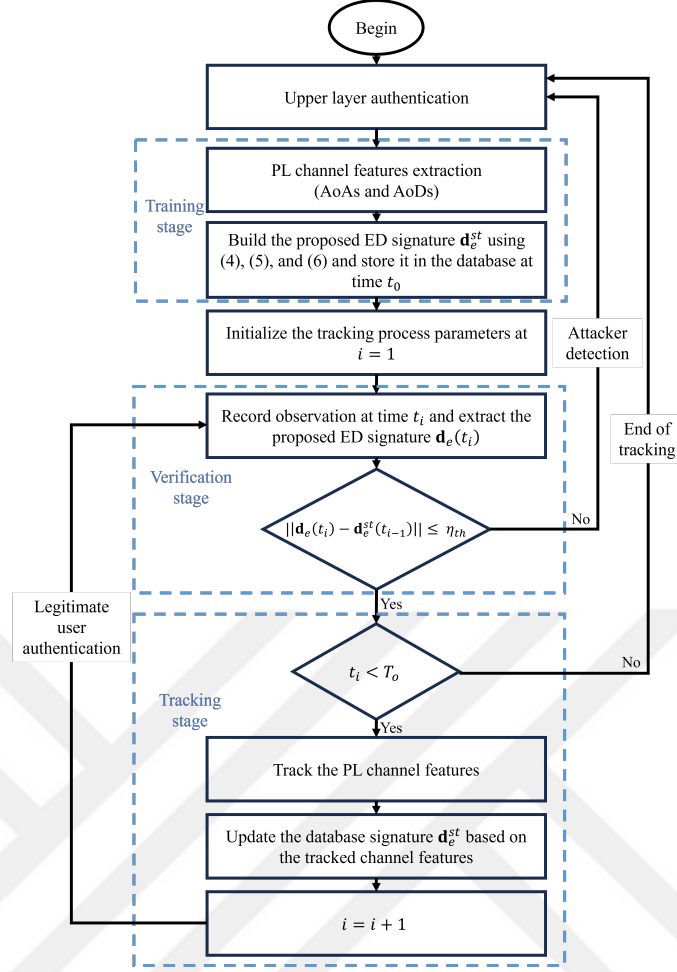


Figure 3.21: The proposed tracking-based PLA flowchart.

3.5.3. Proposed wideband tracking-based PLA approach

In practical wideband mmWave MIMO systems, frequency selectivity leads to wideband characteristics, resulting in the occurrence of the beam squint effect [18]. This effect causes the channel steering vector to become frequency-dependent, resulting in a modified expression for (3.95) as follows

$$\mathbf{a}(\theta, f) = \frac{1}{\sqrt{N}} \left[1, e^{-j2\pi \frac{d_c}{\lambda_c} (1 + \frac{v+f}{f_c}) \sin(\theta)}, \dots, e^{-j2\pi (N-1) \frac{d_c}{\lambda_c} (1 + \frac{v+f}{f_c}) \sin(\theta)} \right]^T. \quad (3.108)$$

This leads to having different beam pointing directions for a different frequency at both transmission and reception, where the angle at a specific frequency is expressed as [50]

$$\theta(f) = \sin^{-1} \left(\frac{f}{f_c} \sin \hat{\theta} \right). \quad (3.109)$$

With a beam squint effect, the phase adjustment of the analog beamforming varies as a function of frequency as in (3.109) which limits the system performance. Therefore,

numerous approaches have been proposed in the literature to deal with this problem. For example, some approaches utilize TTD lines [47], [55] to mitigate the beam squint effect in terahertz massive MIMO systems. On the other side, analog and hybrid beamforming designs [45], [46], [49], [54] are also proposed to compensate for this effect. Although all these works tend to eliminate the beam squint effect, the squinting effect can be exploited to facilitate some level of security in communication networks. In this work, we propose the use of the beam squint effect in enhancing the proposed PLA scheme in wideband transmission.

Figure 3.20 provides an example of beamspace channel representation for a two-path MIMO channel. As shown in **Figure 3.20a**, a single path beam points to a particular grid point. However, due to the beam squint effect, the same path beam diffuses to multiple grid points, as demonstrated in **Figure 3.20b**. This phenomenon introduces limitations to the communication performance in wideband massive MIMO systems and poses challenges to traditional narrowband techniques. However, in this work, the beam squint effect is utilized to improve the authentication performance by increasing the richness of the enhanced distance signature. The details of this approach will be explained in the following sections.

In the following subsections, the process to detect and extract the ED in the wideband transmission is explained. Then, the modification of the tracking model for the wideband transmission is discussed.

3.5.3.1. Beam squint-based ED signature in wideband transmission

The beam diffusion phenomenon in wideband MIMO systems allows for extracting multiple distance values, as shown in **Figure 3.20**. In OFDM system with N_{FFT} subcarriers, ideally, $2N_{\text{FFT}}$ different distances can be extracted within a single path, as the beam is diffused in both the transmitter and receiver nodes, covering multiple grid points. However, due to the limited resolution of grid points in beamspace representation, resulting from a limited number of antenna elements at both ends, the number of the extracted distances is lower than in the ideal case. To address this, M distance signatures are extracted from a single beam (within one path), where $M \ll N_{\text{FFT}}$, is chosen based on the squinting amount in each beam, to obtain distinct distance values.

The process of extracting the enhanced distance signature in the wideband transmission is done during the training stage and can be described as follows:

1. First, the system processes a channel estimation method to estimate the parameters AoAs, AoDs, and Doppler shift ν such as methods in [116], [117] under wideband massive MIMO system design⁹.
2. Assume a multicarrier system as in OFDM with N_{FFT} subcarriers, the directions (AoAs and AoDs) of the beam within one path are calculated for all subcarriers using (3.109). The step is repeated for all paths to get $2N_{\text{FFT}}$ angles for each path (N_{FFT} values for AoDs and N_{FFT} values for AoAs).
3. Determine the index of the strongest element in the beamspace representation corresponding to the estimated channel parameters. This is done by finding the n value that makes Λ inside the Dirichlet sinc function in (3.97) equal to zero, i.e., optimum n for path l at specific subcarrier can be determined as

$$n^* = \arg \min_{1 \leq n \leq N} \left| \frac{d_c}{\lambda_c} \left(1 + \frac{f + \nu}{f_c}\right) \sin(\theta(f)) - \psi_n \right|, \quad (3.110)$$

for $n \in \{n_X, n_Y\}$ and $\theta \in \{\theta_R, \theta_T\}$. The optimum supports (n_X^*, n_Y^*) give the optimum virtual pair $(\psi_{n_X}^*, \psi_{n_Y}^*)$ for AoA and AoD.

4. Since the beamspace representation has a resolution limitation issue based on the antenna array size at both transmitter and receiver nodes, some of the optimum virtual pairs may equal each other. Hence, we drop the repeated pairs and remain with the unique ones which are assumed to be M pairs¹⁰.
5. Assume that $(\psi_{m_x}^*, \psi_{m_y}^*)$, with m_x and $m_y \in m$ and $m = 1, 2, \dots, M$, is the optimum spatial direction pair in a single path for AoA and AoD, respectively. Given a reference point of (0,0), the distance of each pair from the reference point is calculated as

$$d_m = \sqrt{(\psi_{m_x}^*)^2 + (\psi_{m_y}^*)^2}, \quad (3.111)$$

⁹In the work here, we utilize the off-grid sparse Bayesian learning-based channel estimation algorithm, as presented in [116], to estimate the wideband mmWave MIMO channel with the beam squint effect. The algorithm works on the continuous angle-delay parameter domain to avoid grid mismatch problems and estimate the channel parameters with good accuracy.

¹⁰Note that, for the virtual pair corresponding to the angles AoA and AoD at carrier frequency f_c , if there are similar virtual pairs to this pair, the other pairs are dropped and the pair corresponds to f_c remains for the next step.

for $m = 1, 2, \dots, M$.

6. Steps 3 to 5 are repeated on each channel path to get all acceptable pairs to build the distance signature. Then, the distances are sorted in descending order, that is,

$$d_1 > d_2 > d_3 > \dots > d_{M'}, \quad (3.112)$$

where $M' = \sum_{l=1}^{L_p} M_l$ with M_l is the number of pairs M that are extracted from the path l .

7. At the end, the resulting enhanced distance signature $\mathbf{d}_{st}(t_0)$ in the wideband system is stored in the database on Bob's side.

Detection process: In the verification stage, the discussion in the detection model section is followed here where the channel is estimated from the received signal, then the characteristics of the physical layer are detected and extracted. After that, a hypothesis test is applied to the extracted features using (3.102) to detect the legitimate node and authenticate it.

3.5.3.2. Tracking process for the ED signature in wideband transmission

The proposed approach for tracking the beam squint-based enhanced distance signature is extended from the narrowband case discussed in the previous section where the state vector $\mathbf{x}$ must include the angles of all paths and the angles of different frequencies used for constructing the signature for a given path.

As a result, the size of the $\boldsymbol{\theta}$ vector needs to be expanded to include the angles from different frequencies on each channel path for both $\boldsymbol{\theta}_T$ and $\boldsymbol{\theta}_R$. This expansion increases the complexity and time overhead of the tracking approach. To mitigate this complexity, we exploit the correlation between the angles of a single path, which are determined based on a fixed formula (i.e., a function of frequency and angle at the carrier frequency f_c , as given in (3.109)). Therefore, only the angles corresponding to f_{ref} within are tracked, and the other angles are calculated as $\theta(f_m) = \sin^{-1}(f_m/f_{\text{ref}} \sin \theta_{\text{ref}})$, where θ_{ref} is the reference angle used in the tracking process with its corresponding frequency f_{ref} .

In detail, the state vector $\mathbf{x}$ is divided into two parts: $\mathbf{x}^A$ and $\mathbf{x}^B$. Vector $\mathbf{x}^A = [\boldsymbol{\theta}_T^A, \boldsymbol{\theta}_R^A]^T$ is $2L_p \times 1$ a state vector that captures the tracking parameters related to the physical layer

characteristics of the developed signature in the reference subcarrier f_{ref} . Vector $\mathbf{x}^B = [\boldsymbol{\theta}_T^B, \boldsymbol{\theta}_R^B]^T$ is $2(M' - L_P) \times 1$ state vector that captures the parameters related to the physical layer characteristics of the developed signature, which can be extracted and calculated from the reference angles in the state vector $\mathbf{x}^A$ using (3.109). This approach reduces the tracking complexity to be similar to that of narrowband systems.

The process of tracking the enhanced distance signature in the wideband transmission is done as follows:

1. For the time propagation stage, the tracking filter updates the current state vector $\mathbf{x}^A$ and its covariance matrix as in (3.106), i.e.,

$$\begin{aligned}\mathbf{x}_{t_i}^{A-} &= f(\mathbf{x}_{t_{i-1}}^A), \\ \mathbf{P}_{t_i}^{A-} &= \mathbf{A}_{t_{i-1}}' \mathbf{P}_{t_{i-1}}^A \mathbf{A}_{t_{i-1}}'^T + \mathbf{Q},\end{aligned}\tag{3.113}$$

where $\mathbf{A}_{t_{i-1}}' = \frac{df}{dx}|_{\mathbf{x}_{t_{i-1}}^{A-}}$ is the Jacobian matrix of $f(\mathbf{x})$ at $\mathbf{x}_{t_{i-1}}^{A-}$. On the other hand, the current state vector $\mathbf{x}^B$ is updated as

$$\mathbf{x}_{t_i}^{B-} = Q(\mathbf{x}_{t_i}^{A-}),\tag{3.114}$$

where Q is the transition function for the parameters $\boldsymbol{\theta}_T^B$ and $\boldsymbol{\theta}_R^B$ given as $\theta(f_m) = \sin^{-1}(f_m / f_{\text{ref}} \sin \theta_{\text{ref}})$ where $\theta(f_m) \in \{\boldsymbol{\theta}_T^B, \boldsymbol{\theta}_R^B\}$ and $\theta_{\text{ref}} \in \{\boldsymbol{\theta}_T^A, \boldsymbol{\theta}_R^A\}$.

2. For the observation updating stage, the tracking filter gain is given as

$$\mathbf{K}_{t_i}^A = \mathbf{P}_{t_i}^{A-} (\mathbf{H}_{t_i}^A)^T \frac{1}{\mathbf{H}_{t_i}^A \mathbf{P}_{t_i}^{A-} (\mathbf{H}_{t_i}^A)^T + \mathbf{R}},\tag{3.115}$$

where $\mathbf{H}_{t_i}^A = \frac{dg}{dx}|_{\mathbf{x}_{t_i}^{A-}}$ is the gradient of the measurement function $g(\mathbf{x})$ at $\mathbf{x}_{t_i}^{A-}$ and $\mathbf{x}_{t_i}^{B-}$. Furthermore, the posterior estimates are given as

$$\begin{aligned}\mathbf{x}_{t_i}^A &= \mathbf{x}_{t_i}^{A-} + \mathbf{K}_{t_i}^A (\mathbf{y}_b - g(\mathbf{x}_{t_i}^{A-}, \mathbf{x}_{t_i}^{B-})) \\ \mathbf{P}_{t_i}^A &= (\mathbf{I} - \mathbf{K}_{t_i}^A \mathbf{H}_{t_i}^A) \mathbf{P}_{t_i}^{A-} \\ \mathbf{x}_{t_i}^B &= Q(\mathbf{x}_{t_i}^A)\end{aligned}\tag{3.116}$$

After tracking the developed signature reserved in the database, the database is updated and the verification process is started again to authenticate the legitimate user.

CHAPTER 4

4. RESULTS AND DISCUSSION

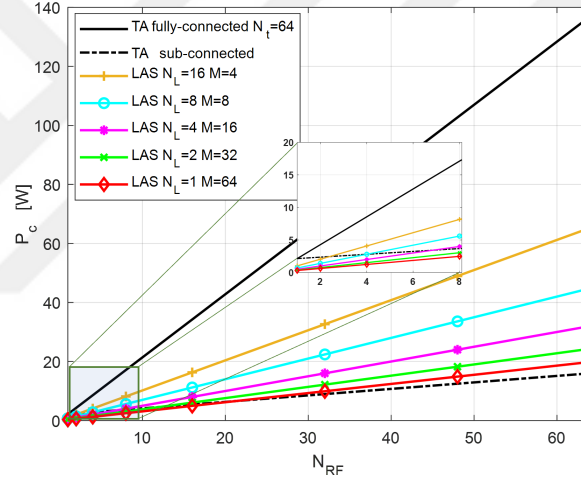
4.1. Results Of Study I

In this section, the spectral and energy efficiency performances of the proposed sub-grouped LAS-MIMO system are evaluated when the system uses the proposed hybrid precoder. The performances are compared with those of the TA-MIMO systems when the TA systems use three well-established precoders, which are defined as 1) TA fully-connected: optimal unconstrained precoding based on the Single Value Decomposition (SVD) of the channel matrix for fully-connected hybrid beamforming MIMO structure, 2) TA sparse: spatially sparse precoding algorithm introduced in [24] and applied on a fully-connected hybrid beamforming MIMO structure, and 3) TA sub-connected: the SIC-based precoding introduced in [90] and applied on a sub-connected hybrid beamforming MIMO structure. Furthermore, the proposed precoding algorithm is applied on a fully-connected LAS design with $N_L = 4$ by assuming one group in the proposed sub-grouped LAS structure. TA performance evaluations assume that full-resolution PSs are used in all three system designs. In addition, as in [118], a sub-connected MIMO system model with one lens in each subgroup is considered, where each RF chain is connected to only one lens.

Simulation parameters set for all results are shown in **Table 4.1**. The transmit power is assumed to be $P_x = \text{EIRP} - G_{dB}$, where $\text{EIRP} = 45$ dBm is the effective isotropic radiated power and G represents the antenna array gain calculated with aperture efficiencies of 90% and 80% for the two architectures TA and LAS, respectively [37]. Each lens in the LAS system has an area of $M(\frac{\lambda}{2})^2$ with M antenna elements uniformly connected under this area. The AoD is assumed to follow a uniform distribution within $[-\pi/2, \pi/2]$ with 5° spreading. Finally, the two related spectral and energy efficiency results are averaged over 500 random channel realizations and reported.

Table 4.1: Simulation parameters for the proposed LAS-MIMO design

Parameters	Value
Bandwidth Bw	500 MHz
Carrier frequency	28 GHz
Phase shifter bit resolution B	4
BS antenna elements N_t	64
N_{RF} at BS side	1, 4, 8
N_{RF} at User Equipment (UE) side	1
Channel paths P_k for each UE	4
P_{PS}	30 mW [23], [119]
P_{SW}	10 mW [37]
P_{RF}	220 mW [37]
η_{PA}	0.2 [37]
IL_{SW}	1 dB [37]
Number of series switches ζ	$\log_2(M)$

**Figure 4.1:** Power consumption vs N_{RF} for different N_L in the fully-connected LAS-MIMO system architecture.

In **Figure 4.1**, a comparison for the power consumption at different numbers of RF chains is provided for the different fully-connected LAS-MIMO architectures, fully-connected TA-MIMO architecture, and sub-connected TA-MIMO architecture when the total number of antennas is $N_t = 64$. It is noticed that the fully-connected TA-MIMO architecture has the worst performance in terms of power consumption even with a high number of lenses in the LAS architectures. The power consumption of the LAS architectures increases with increasing the number of lenses due to the need for more PSs in the system to connect and control the added lenses. For example, an LAS architecture with 16 lenses consumes around 16 W at $N_{RF} = 16$ while the fully-connected TA-MIMO architecture consumes

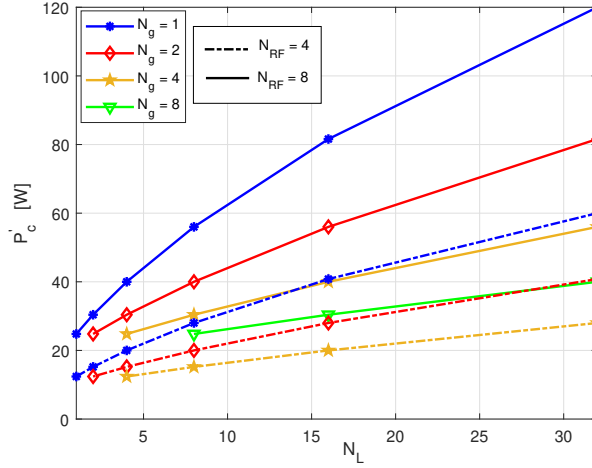


Figure 4.2: Power consumption vs N_L for different sub grouping at $N_{RF} = 4$ (dash lines) and $N_{RF} = 8$ (solid lines) in the proposed sub-grouped LAS-MIMO architecture.

around 34 W. At a low number of RF chains, the LAS-MIMO architectures have lower power consumption than the sub-connected TA-MIMO architecture which makes it more applicable in a system with a high sparsity level since small numbers of RF chains can be only activated in the system if the channel is sparse as in mmWave channels [120].

The power consumption for the sub-grouped LAS-MIMO architecture versus N_L is illustrated in **Figure 4.2** at different numbers of grouping. Given that the number of sub-group is between 1 (the design becomes equal to a fully-connected one) and N_{RF} (the design becomes similar to sub-connected design), it is noticed that as the number of sub-grouping increases, the power consumption reduces even at high N_L lenses for both 4 and 8 RF chains. Note that the best performance is obtained when the number of sub-grouping is equal to the number of RF chains where each sub-group has only one RF chain since decreasing the number of sub-grouping converts the hybrid beamforming structure from sub-grouped model to a fully-connected one.

Figure 4.3 shows the spectral efficiency comparison in mmWave MIMO system, where the number of RF chains is set to 4, $N_t = 64$, and $N_g = N_{RF}$. Note that N_g is set to be equal to the number of RF chains in the system to have a fair comparison with the sub-connected TA-MIMO system. The TA and TA sparse methods are referred to as fully-connected models where each RF path is connected to all antenna elements. In contrast, the LAS fully-connected method refers to our proposed precoding method when $N_g = 1$. This means that our proposed sub-grouped LAS design becomes equal to conventional

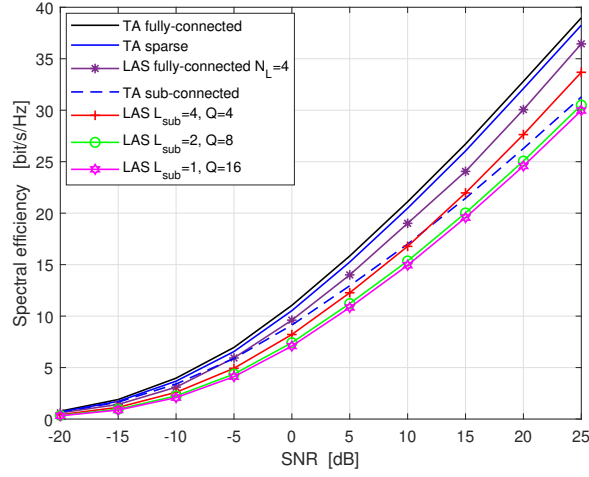


Figure 4.3: Spectral efficiency vs SNR for $N_{\text{RF}} = 4$ in the proposed sub-grouped LAS-MIMO architecture.

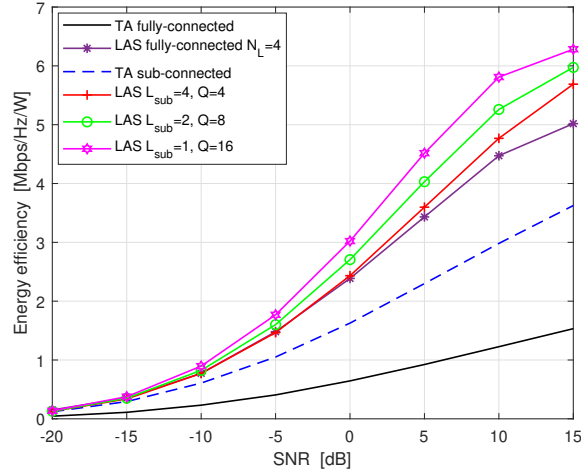


Figure 4.4: Energy efficiency vs SNR for $N_{\text{RF}} = 4$ in the proposed sub-grouped LAS-MIMO architecture.

fully-connected LAS design as in [37]. In TA sub-connected model, each RF path is connected to N_t/N_{RF} antenna elements only independently from other RF path connections. **Figure 4.3** shows that the proposed sub-grouped precoder achieves a reasonable spectral efficiency performance with less than 30% difference from the fully-connected TA-MIMO model at 0 dB SNR with much less power consumption and exceeds the sub-connected TA precoder's performance as the number of lenses in each sub-grouped and the SNR level increase while achieving better energy efficiency as shown in **Figure 4.4**. Note that increasing the number of lenses within the sub-group enhances the performance due to better beam flexibility in which the multiple lenses can provide to steer the beam to the desired direction compared to a single lens structure.

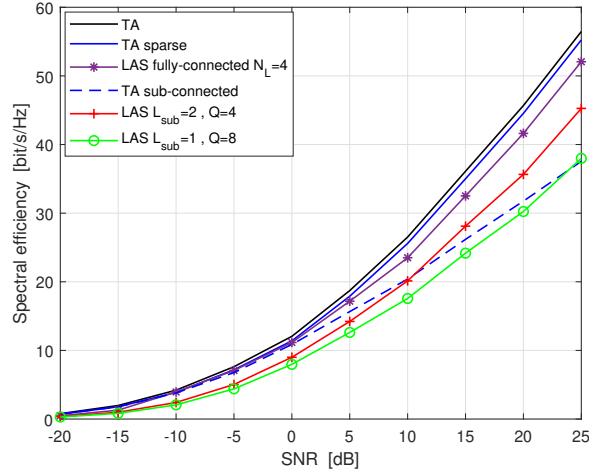


Figure 4.5: Spectral efficiency vs SNR for $N_{RF} = 8$ in the proposed sub-grouped LAS-MIMO architecture.

Figure 4.4 shows the energy efficiency comparison against the SNR levels for the same system design given in **Figure 4.3**. It is observed that the proposed sub-grouped LAS-MIMO system for all the cases can achieve higher energy efficiency than the fully-connected and sub-connected TA MIMO systems. Higher energy efficiency for a given SNR is achieved as the number of lenses decreases due to the decrease in the number of PSs. This increase in energy efficiency can be observed at high SNR levels. From **Figure 4.3**, it is noticed that a system of $L_{sub} = 1$ (one lens in each subgroup) has the worst performance in terms of the spectral efficiency for a given SNR; however, it has the lowest consumed power value compared to other system designs which make it preferable for the energy-efficient systems. It can be stated that the energy efficiency and spectral efficiency trade-off depends on the model design. If the priority is the spectral efficiency, the number of lenses in the design is preferred to be high since each lens has a smaller size and is compact and provides more control on the beam compared to the $L_{sub} = 1$ case. On the other hand, if the priority is energy efficiency, the number of lenses is required to be less to get an energy-efficient system design. Furthermore, if a scenario requires a spectral and energy-efficient system, a joint optimization problem should be formulated and solved. In this problem, the system's spectral efficiency and energy efficiency metrics should be maximized to extract the optimum sub-grouping structure in the proposed sub-grouped LAS-MIMO design.

Figure 4.5 compares the spectral efficiency vs SNR for 8 RF chains in the LAS-MIMO systems. Again, it is shown that the performance of the proposed sub-grouped precoding

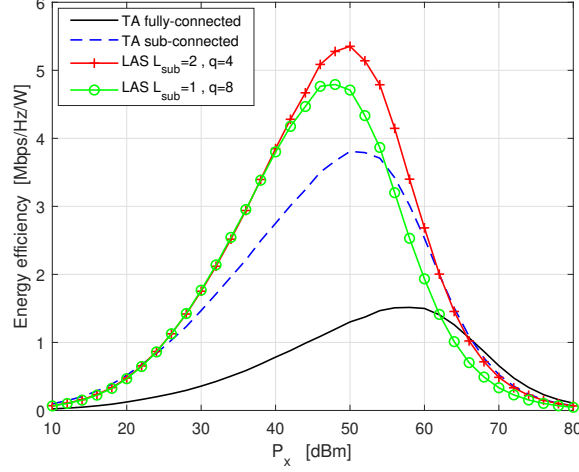


Figure 4.6: Energy efficiency vs transmitted power P_x for $N_{\text{RF}} = 8$ in the proposed sub-grouped LAS-MIMO architecture.

algorithm is close to the traditional sub-connected and fully-connected MIMO systems and can outperform the sub-connected traditional MIMO system for a larger number of lenses in the system at high SNR levels.

The energy efficiency performance vs transmitted power P_x is compared in **Figure 4.6** considering the same assumptions given in **Figure 4.5**. The figure shows that the proposed sub-grouped LAS-MIMO system outperforms the fully-connected and sub-connected traditional MIMO systems when the number of RF chains is set to 8. This result confirms that the proposed system can be exploited even with the higher number of RF chains. From **Figure 4.6**, it is clear that the cases of subgroups consisting of $L_{\text{sub}} = 1$ or $L_{\text{sub}} = 2$ lenses perform near-identical energy efficiency. Comparing these $N_{\text{RF}} = 8$ cases with the $N_{\text{RF}} = 4$ cases of **Figure 4.4** shows that increasing number of RF chains causes the energy efficiency of systems to become more independent of the subgroup configuration. This is due to the fact that the gain in spectral efficiency with increasing N_{RF} is much higher than the increase in power consumption. In the presented $N_{\text{RF}} = 8$ cases, $L_{\text{sub}} = 2$ clearly becomes a better choice over $L_{\text{sub}} = 1$ by providing higher spectral efficiency with near-identical energy efficiency. Although energy efficiency can be increased by increasing the transmitted power, after a certain power level, the energy efficiency will decrease again to a level worse than that of traditional MIMO systems in unimodal function behavior. Therefore, an optimum transmitted power level needs to be defined in each system design. It is noticed from the figure that the optimum transmitted power points of the proposed sub-grouped LAS designs are less than those of the conventional TA-MIMO design,

which makes our design more appealing to green cellular networks.

4.2. Results Of Study II

In this section, we analyze the performance of the proposed precoder design in terms of beam gain, complexity, and overall system capacity. Afterward, considering the inclusion of some extra hardware components in the proposed architecture which may increase the system's power consumption, we delve into the detailed examination of the system's energy efficiency. Subsequently, we analyze the impact of parameter selection on the overall system performance, aiming at understanding the nuances and trade-offs involved in choosing specific design parameters. Finally, we extend our analysis to the multiuser case, exploring how the proposed design adapts and performs in a scenario involving multiple users. This comprehensive analysis aims to provide a holistic understanding of the proposed precoder design across various performance metrics and operational scenarios.

Table 4.2: Simulation parameters of the proposed LAS design.

Parameters	Value
Operating frequency f_c	60 GHz
System bandwidth W	5 GHz
Channel paths L_p	1
Antenna elements M	128
Distance between antenna elements d	$\lambda_c/2$
Number of RF chains N_{RF}	1
Number of subcarrier frequencies N	2048
Number of lenses N_L	4
Number of analog subband filters N_F	16
Number of antenna per lens P	M/N_L
Desired direction $\hat{\theta}$	$\pi/4$
Power efficiency of amplifier η_{PA}	0.2 [37]
Insertion loss of filter $\text{IL}_{\text{filter}}$	1 dB
Power consumption of PS P_{PS}	30 mW [37]
Power consumption of one RF chain P_{RF}	220 mW [37]

Unless stated otherwise, we use the numerical values of the system parameters from **Table 4.2** in this analysis. We benchmark the performance of the proposed LAS-based design against two configurations of conventional mMIMO systems: one considering the full beam squint effect, denoted as “Full-squint mMIMO,” and the other disregarding beam squint effect (i.e., the ideal case), referred to as “No squint mMIMO.” Additionally, we

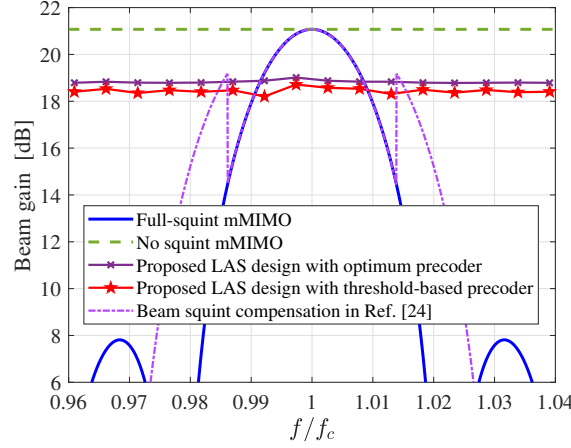


Figure 4.7: Beam gain vs normalized frequency.

compare the performance of the proposed design with existing state-of-the-art solutions, including the designs presented in [45], [46], and [54]. To ensure a fair and meaningful comparison, we align the parameters used in [45], [46], and [54] with those detailed in **Table 4.2**. It’s important to note that the power consumption model for [54] follows (3.47), excluding the terms related to switching. For a single-user scenario, we adopt the array vector of a single subcarrier approach from [45] for the analog precoding design, while Mean Channel Matrix (MCM) and Mean Channel Covariance Matrix (MCCM)-based digital precoding approaches are adopted for both single-user and multi-user scenarios. This thorough comparative analysis aims to provide a rigorous understanding of the proposed LAS-based design’s performance with conventional mMIMO systems and other contemporary beam squint solutions. It ensures a fair and insightful assessment across various precoding strategies.

4.2.1. Proposed precoder design performance

Here, we evaluate the performance of our proposed precoder design in terms of beam gain, complexity, and capacity.

In beam gain evaluation, the achievable beam gain with the proposed LAS-based design is given in **Figure 4.7**, benchmarked with “Full-squint mMIMO” and “No squint mMIMO” as well as with [54]. The beam gain provided by the proposed design is analyzed with both the full exhaustive search and the proposed threshold-based search approach. While both cases significantly improve the gain (to less than ≈ 3 dB of ideal performance), the exhaustive search, as expected, performs relatively better, though at the expense of high

complexity, as analyzed in **Figure 4.8**. However, due to the loss introduced by lens structure as discussed in [37], the beam gain of the proposed design can not reach the performance of an ideal mMIMO system (with no squinting). Note that the proposed design achieves significant gain enhancement compared to the design presented in [54] as the deviation of the subcarrier frequency from f_c increases. The rationale behind this behavior is due to the frequency-dependent property of the analog precoder introduced in the proposed analog LAS design. Unlike the work of [54] where each subband filter and its associated PS perform the beam squint correction with respect to the center frequency of that filter only which limits the achievable beam gain performance, especially for the edge subbands, the analog precoder in our proposed design corrects the beam squint pertained to each subband with respect to the system's carrier frequency f_c .

The exhaustive search-based precoder complexity is calculated as the total number of possible beam sets in the system for a given subband. The precoder needs to search over all possible sets and choose the set that has the minimum beam deviation that fulfills the condition (3.41). On the other hand, the proposed threshold-based method determines the precoder complexity by counting the number of beam sets that are examined to achieve the optimal set that reduces the beam deviation to a specific threshold value that satisfies the condition (3.42). **Figure 4.8** shows a huge difference in complexity between the exhaustive search (which is in the order of millions) and the proposed threshold-based precoder (only hundreds to few thousand) at each subband. This is due to the simplified searching mechanism adopted by the proposed precoder.

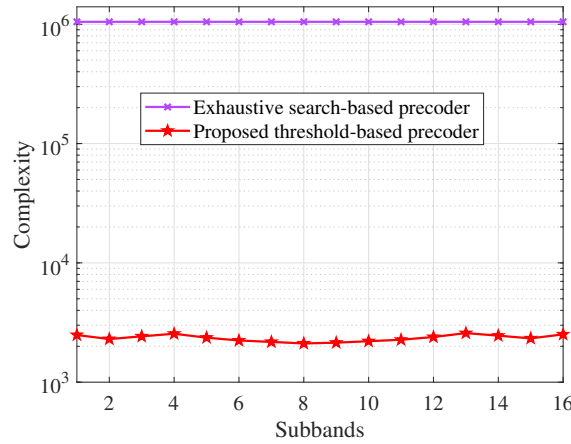


Figure 4.8: Complexity analysis.

Figure 4.9 illustrates the capacity versus bandwidth performance of the proposed LAS design in comparison to the contemporary designs presented in [45] and [54] under 10 dB SNR. Notably, our proposed design maintains a stable and superior capacity performance for different bandwidths, whereas the design in [45] and [54] experience high capacity degradation as bandwidth increases. Furthermore, the capacity performance of [54] oscillates as the bandwidth increases due to the inherent number of filters in the design.

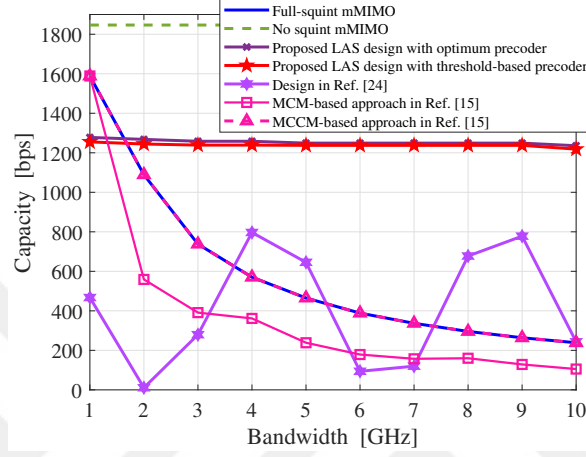


Figure 4.9: Capacity vs bandwidth analysis.

4.2.2. Energy efficiency performance

In this subsection, we analyze the power consumption and energy efficiency of the proposed LAS design compared to other related designs in the literature (i.e., the design in [54]).

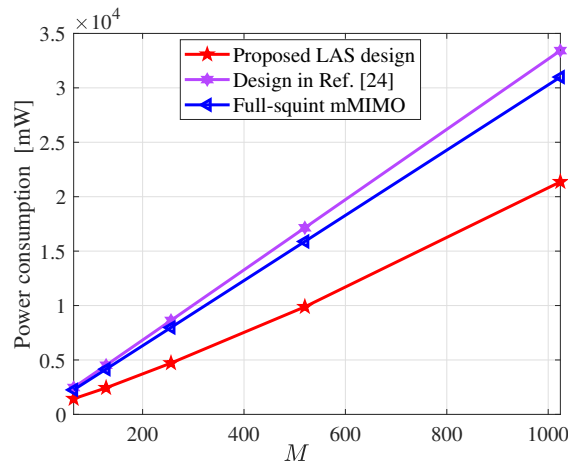


Figure 4.10: Power consumption vs M analysis.

In practice, since the values of P_{SW} are only available for the few ports MPMT switches, we model the N_F -poles $\times$ P -throws MPMT switch considered in our design as a series of SP2T switches for each pole for power consumption analysis. Therefore, the term P_{SW} in (3.47) becomes $\log_2(P)N_F P_{\text{SW}}^{(s)}$, where $P_{\text{SW}}^{(s)} = 10$ mW is the power consumption of an SP2T switch, while $\text{IL}_{\text{SW}} = \log_2(P) \times 1$ dB [37]. The calculated power consumption for different array sizes is summarized in **Figure 4.10**. The performance is compared with that of the conventional mMIMO system (i.e., (3.46)), and with reference [54] which also presents subband-based beam squinting compensation design. It is clear from **Figure 4.10** that the proposed LAS design is the most power-efficient design while profitably compensating for the beam squinting, which makes it suitable for commercial applications of mMIMO networks.

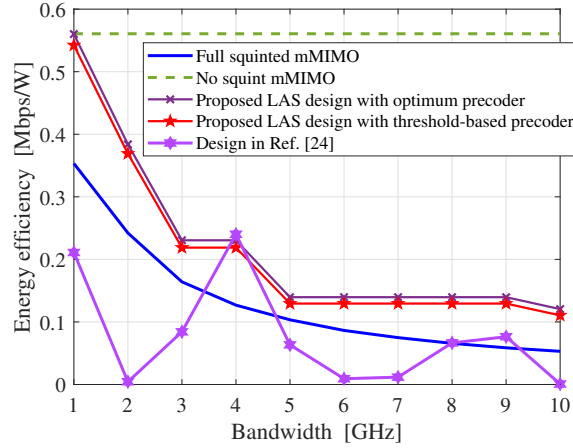


Figure 4.11: Energy efficiency vs bandwidth analysis.

In **Figure 4.11**, the energy efficiency performance of the proposed LAS design across various bandwidths is illustrated. It is compared with the design presented in [54] under a SNR of 10 dB. In particular, it is observed that while the performance of the proposed design experiences a decline with an increase in system bandwidth, it consistently outperforms existing approaches. The proposed LAS design demonstrates superior energy efficiency compared to UWB mMIMO systems without squint compensation, even in scenarios with elevated bandwidths. This observation underscores the effectiveness of the proposed LAS design in achieving commendable energy efficiency across a spectrum of bandwidths, positioning it as a competitive solution compared to existing approaches.

a): $N_F = 16, N_L = 2$, different angles
b): $N_F = 16, \hat{\theta} = \pi/4$, different N_L
c): $N_L = 2, \hat{\theta} = \pi/4$, different N_F

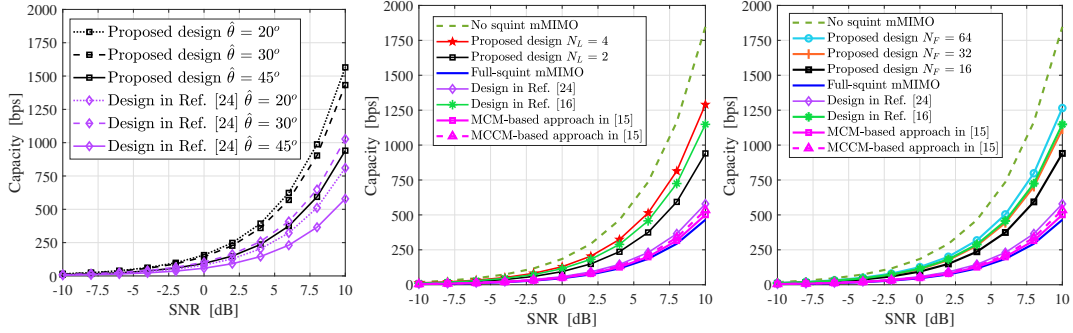


Figure 4.12: Capacity vs SNR analysis in a single user scenario.

4.2.3. Impact of the design parameters selection

In this subsection, we turn our attention to examining the influence of varying hardware parameters on the proposed design's performance, specifically assessing their impact on system capacity and energy efficiency. The parameters under scrutiny include the transmit direction $\hat{\theta}$, the number of subband filters N_F , and the number of lenses N_L incorporated in the design. This systematic analysis aims to elucidate the sensitivity of the proposed design to alterations in these crucial hardware parameters, providing valuable insights into the trade-offs and considerations associated with different parameter choices.

Figure 4.12 gives some insights regarding the effects of the design parameters $\hat{\theta}$, N_F , and N_L on the system capacity of the proposed design for a point-to-point link scenario with a single-antenna receiver. The system performance here is compared with the performances of the designs presented in [54] and [46]. Since the beam squinting amount suffered by the system is a function of the desired direction $\hat{\theta}$ and increases as $\hat{\theta}$ deviates further from the array's foresight, see **Figure 3.3a** to **Figure 3.3c**, the proposed design in section III-A considers the worst case which is $\hat{\theta}_{\max} = \pi/4$ for LAS systems [37]. **Figure 4.12a** analyzes the capacity performance of the proposed design for different $\hat{\theta}$ ($\leq \hat{\theta}_{\max}$). Clearly, the system capacity performance improves for all $\hat{\theta} \leq \hat{\theta}_{\max}$. This implies that, with the proposed approach, the system can be designed once, based on the worst-case scenario, and work for all other scenarios without the need for changing the design parameters. It is also clear from the figure that the proposed design outperforms that of [54] for different $\hat{\theta}$. In **Figure 4.12b**, we only analyze the proposed design when both steering mechanisms, i.e., PS and switching networks, have a notable effect in controlling the beam squint amount. The

figure shows that there is capacity gain as the number of lenses (i.e., N_L) is increased. We reckon that this improvement comes from increasing the number of individual beams that are responsible for providing the resultant beam while keeping the switching network dominant over the PS network, which in return enhances the controllability of beam squint amount.

Likewise, increasing the number of filters N_F improves system capacity as shown in **Figure 4.12c**. With the same W , larger N_F means reducing the subband sizes, which in turn reduces the residual squint effect within each subband. However, since increasing N_F also elevates the system's energy consumption, it is imperative to weigh the achieved capacity gain against the energy efficiency as a function of N_F . This analysis is presented in **Figure 4.13**. While it is clear that the capacity enhancement through increasing N_F comes at the expense of degraded energy efficiency, the proposed design still outperforms the conventional system in terms of both capacity and energy efficiency for a wide range of N_F values. It is also clear from the figure that, although (3.35) gives wide flexibility on the choice of N_F , the best choice remains to be the minimum N_F that satisfies (3.35) when system's energy efficiency is of great concern.

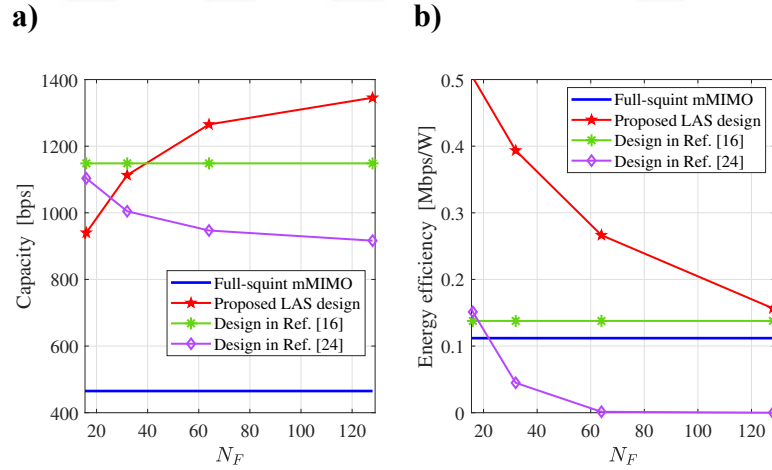


Figure 4.13: a) Capacity and b) energy efficiency vs number of subband filters N_F analysis.

4.2.4. Performance in multiuser case

Expanding the applicability of our proposed approach to multiuser networks, this section undertakes an evaluation of system capacity within scenarios involving multiple users. In this context, the proposed LAS design is assumed to be assessed utilizing a threshold-based

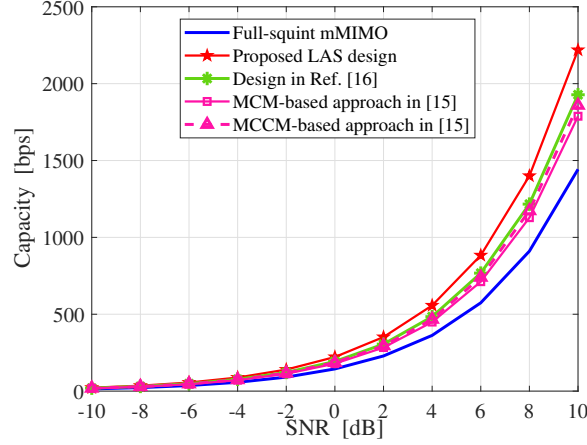


Figure 4.14: Capacity vs SNR analysis in multi-user scenario.

precoder approach, incorporating the use of ZF precoder in its digital baseband domain.

The system capacity vs SNR is analyzed in **Figure 4.14** for 10 users (i.e., $U = N_{\text{RF}} = 10$) and compared with the precoding design presented in [45] and [46]. The performance is also compared with the full-squint mMIMO system capacity with an SVD-based approach. The system capacity for the proposed LAS design is given as

$$C_{\text{LAS}} = \sum_{k'=1}^{N_F} \frac{W_b}{N'} \sum_{k''=0}^{N'-1} \log_2 \left(\left\| \mathbf{I}_{N_{\text{RF}}} + \frac{\gamma N'}{W_b} (\mathbf{H}(k'') \mathbf{F}_{\text{RF}}(k') \mathbf{F}_{\text{BB}}(k'')) \right. \right. \quad (4.1)$$

$$\left. \left. \mathbf{F}_{\text{BB}}^*(k) \mathbf{F}_{\text{RF}}^*(k') \mathbf{H}^*(k'') \right\| \right).$$

Our proposed analog LAS transceiver design has demonstrated superior performance compared to alternative methods and designs discussed in the existing literature. This outcome substantiates the effectiveness of our proposed design in successfully mitigating the beam squint effect in a multi-user environment.

4.3. Results Of Study III

In this section, simulations are provided to verify the effect of beam squinting on the overall IBI in the near-field and far-field propagation and demonstrate the performance of the proposed sub-grouped beam-based scheduling scheme. We consider the UWB mMIMO system, where the multi-beam transmitter is transmitting to K single-antenna receivers distributed uniformly within the cell coverage with only LoS connection; unless stated otherwise, the key system parameters are given in **Table 4.3**.

Table 4.3: Simulation parameters for the IBI-based scenarios

Parameters	Value
Operating frequency f_c	28 GHz
System bandwidth W	5 GHz
Number of antenna elements M	256
Channel paths L_p for all beams	1
Standard deviation σ	0.04 [105]
The intercept value γ	45.3 [105]
The slop β	2.9 [105]
the channel power β_o	10^{-4} [101]
Received antenna gain G_r for all K receivers	0 dBi
Number of receivers/users K	20
Transmitted power P_t for all beams	27 dBm [58]
Number of subcarriers N	1024
Noise Temperature	298 K [58]
Cell radius	1500 m

4.3.1. IBI analysis

Scenario I: Assume a scenario where a user is located at $\rho_o = d_F$ m away from the massive antenna array while the other users are distributed uniformly in angle, $\hat{\theta}^{(k)} \sim \mathcal{U}(-\pi/3, \pi/3)$, and distance, $\rho_k \sim \mathcal{U}(d_F, \text{Cell}_{\text{edge}})$, in the far-field region (for **Figure 4.15a** and **Figure 4.15c**), and for the near-field propagation, the interested user is located at $\rho = d_F/16$ m while the other users are distributed uniformly in angle, $\hat{\theta}^{(k)} \sim \mathcal{U}(-\pi/3, \pi/3)$, and distance, $\rho_k \sim \mathcal{U}(1, d_F)$, over the near-field region (for **Figure 4.15b** and **Figure 4.15d**). **Figure 4.15** analyzes the IBI model in near-field and far-field propagation with and without beam squint effect. In a narrowband transmission where the beam squinting is not considered in the system, the IBI varies along different beam angles where the minimum interference levels are considered near to $\pm 20^\circ$ in far-field propagation as shown in **Figure 4.15a**, while in near-field propagation, only the edge beams face lower interference level compared to other beams in the design as presented in **Figure 4.15b**, where the non-stationarity of the near-field propagation creates uncontrollable interference over all beams in the design. In wideband signaling under the beam squint effect, the squinting effect dominates in the system in both near-field and far-field propagation, as illustrated in **Figure 4.15c** and **Figure 4.15d**. The figures show that the interference at its minimum is in the range of 0° beam at the center frequency f_c where the amount of squinting of the beam is around zero. Note that at a given pair of subcarrier and

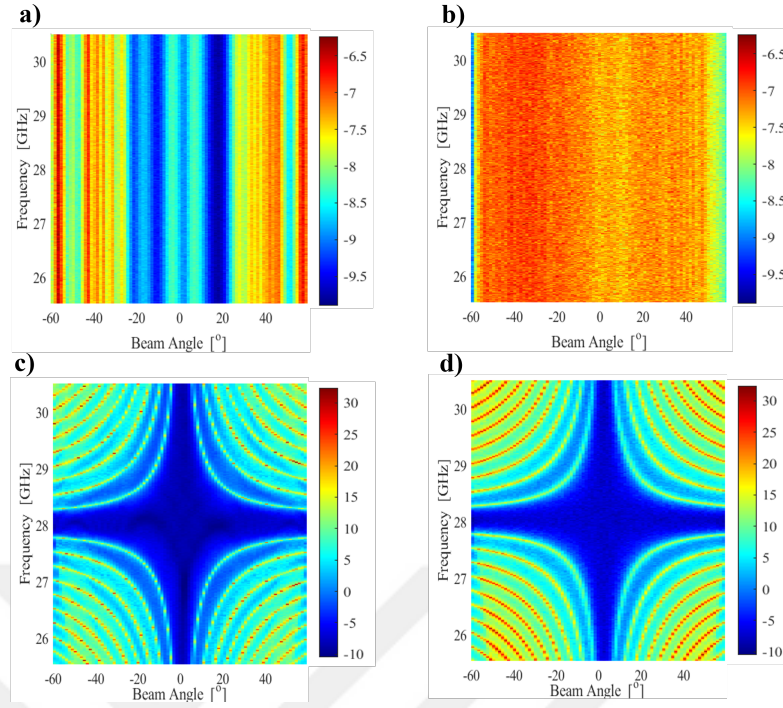


Figure 4.15: The normalized IBI pattern in the space (in dB) for Scenario I for: **a)** far-field propagation without beam squinting, **b)** near-field propagation without beam squinting, **c)** far-field propagation under beam squint effect, and **d)** near-field propagation under beam squint effect.

beam angle, the interference can be at its minimum or maximum in a periodic way leading to an opportunity to multiplex another user over a specific beam and subcarrier in case the user has a low requirement as the case of Internet-of-Things (IoT) devices. However, due to the IBI behavior, it is impossible to find an optimum beam that experiences the lowest IBI level at all frequencies at a specific distance or a specific frequency at all distances.

Scenario II: Assume a scenario where a user is located at $\rho_o = d_F$ m away from the massive antenna array with an angle $\hat{\theta} = 45.5^\circ$ while there is only interfering user in the system located at the same distance as the interested user ($\rho_I = \rho_o = d_F$) with different angles in the far-field propagation assumption (for **Figure 4.16a** and **Figure 4.16c**), and for the near-field propagation, the interested user is located at $\rho_o = d_F/32 \simeq 22$ m similar to the location of the interference user, while the beam angle of the interested user is $\hat{\theta} = 45.5^\circ$ in the near-field region (for **Figure 4.16b** and **Figure 4.16d**). **Figure 4.16** analyzes the IBI model for a single interfering user located over different beams in the system in near-field and far-field propagation with and without beam squint effect. In

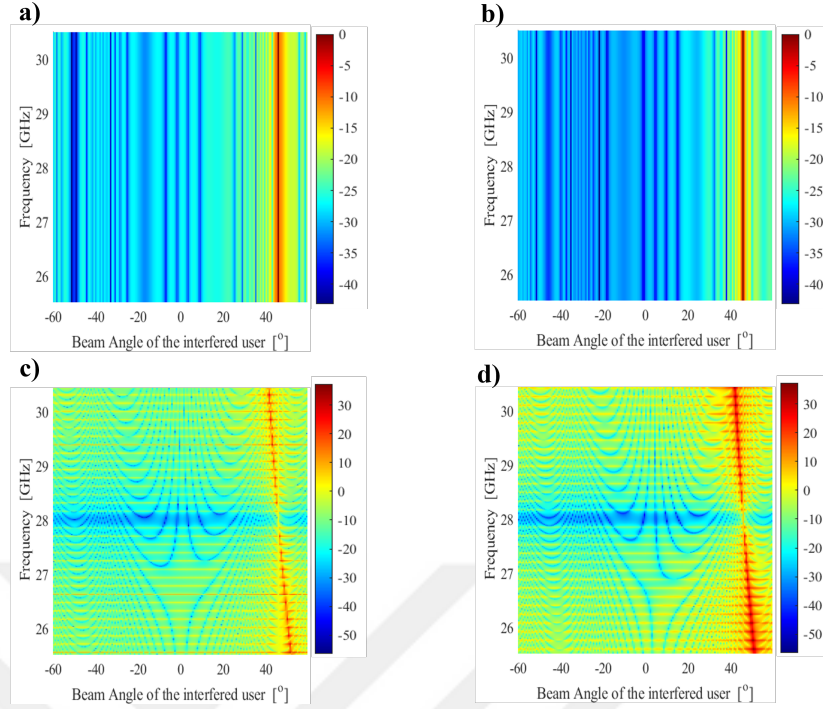


Figure 4.16: The normalized IBI pattern in the space (in dB) for Scenario II for: **a)** far-field propagation without beam squinting, **b)** near-field propagation without beam squinting, **c)** far-field propagation under beam squint effect, and **d)** near-field propagation under beam squint effect.

narrowband transmission, due to beam focusing phenomena in near-field propagation, the desired user faces less interference from the interfered one over all beams, specifically, around the desired beam angle ($\hat{\theta} = 45.5^\circ$) as shown in **Figure 4.16b** compared to the interference in the far-field propagation shown in **Figure 4.16a**. However, in the wideband transmission where the beam squinting is considered in the design, the overall interference level in the system increases for both near-field and far-field propagation as illustrated in **Figure 4.16c** and **Figure 4.16d**. Unlike the case of narrowband near-field propagation, wideband transmission spreads the desired beam and the interfering one, which increases the interference level around the desired beam as shown in **Figure 4.16d**.

4.3.2. Capacity analysis

In this section, the system capacity is first analyzed for the proposed IBI model under beam squint effect and compared with the model without squinting (referring to the IBI model presented in [58]). Then, since there is no method previously proposed to mitigate the

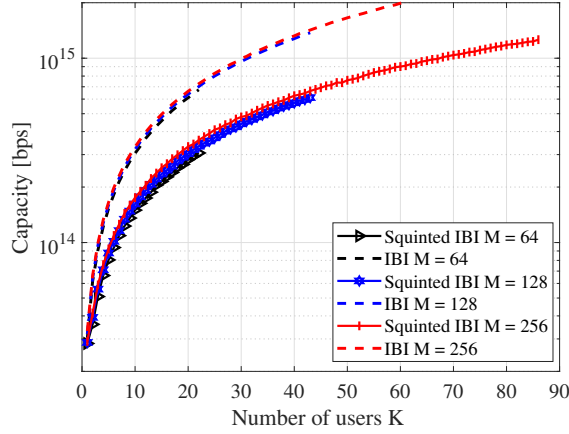


Figure 4.17: System capacity vs number of users analysis.

IBI under beam squint effect, the system capacity when the proposed sub-grouped beam-based scheduling algorithm is applied is evaluated and compared with two benchmarks: system with a beam squint effect in the IBI model (Squinted IBI) and a system where the IBI is removed (No IBI).

In **Figure 4.17**, the system capacity vs number of users is analyzed with different antenna array sizes. The solid lines represent the system capacity under the IBI effect modeled under the consideration of the distance-dependence property in near-field propagation and the beam squint effect as presented in this work, while the dashed lines represent the system capacity under the IBI effect in far-field propagation without considering the beam squint effect as in [58]. It is noticed from the figure that increasing the number of transmitted beams significantly increases the system capacity; however, in a wideband system with a beam squint effect, the capacity drops significantly as the number of users increases compared to the narrowband non-squinted design. For example, with $K = 50$ users, the capacity drops around 54% from its maximum when the beam squinting is considered in the design. This leads to the need for a mechanism to reduce the IBI in wideband systems.

Figure 4.18 represents the system capacity vs SNR under different antenna array sizes. The dashed lines indicate the system capacity without considering the impact of IBI, while the solid lines reflect the capacity of the system under the influence of IBI, and the dot lines point to the system capacity after using the proposed sub-grouped beam-based scheduling algorithm. It is assumed that users are uniformly distributed throughout the cell area, with

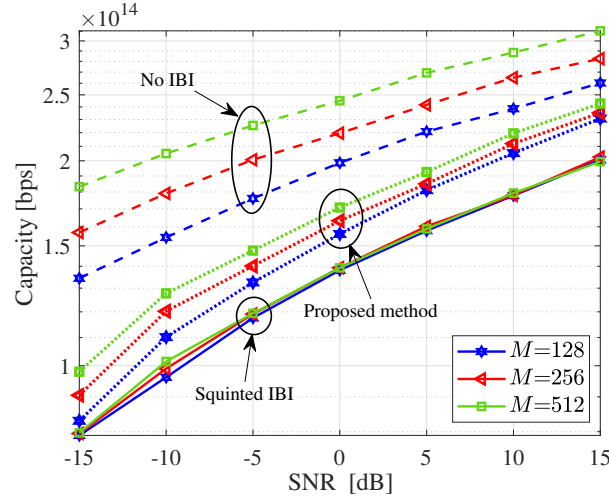


Figure 4.18: System capacity vs SNR analysis.

consideration given to the non-stationarity of near-field propagation for users nearby. The figure shows that as the number of antenna elements increases, the performance of the system increases linearly for a perfect system assumption. However, this assumption is not practical as issues such as beam squinting and the non-stationarity property of near-field propagation severely degrade the system capacity performance, as shown in **Figure 4.18**. Therefore, the insertion of extra antenna elements does not necessarily enhance the system performance, as new challenges are introduced in next-generation wireless networks. In **Figure 4.18**, the effectiveness of the proposed sub-grouped beam-based scheduling approach is also evaluated. It is noticed that the proposed method enhances the capacity of the UWB mMIMO system under the beam squint effect overcoming the aforementioned degradation.

4.4. Results Of Study IV

This section evaluates the proposed PLA scheme's performance through link-level simulation analysis. The probability of detection γ and the probability of false alarm α are considered for the performance analysis. During the simulation, it is considered that all nodes have the same number of antenna elements $N = N_A = N_B = N_E$, the carrier frequency $f_c = 28$ GHz, $Z_s = (73 + j42.5) \Omega$ for half-wavelength dipole antenna and $Z_L = 50 \Omega$ at all nodes. The performance of the proposed $\mathbf{d}$ signature with and without mutual coupling effect is compared with the beamspace channel $\mathbf{H}_b$ signature introduced

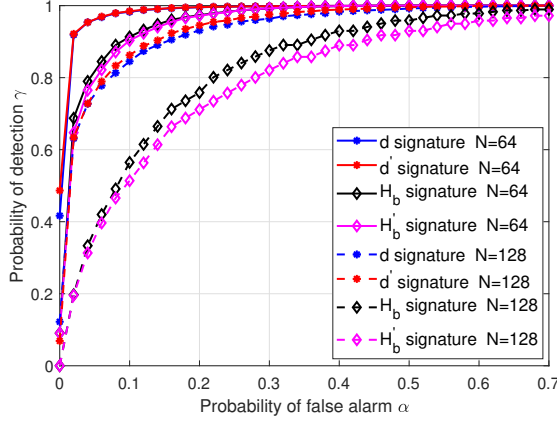


Figure 4.19: Probability of detection vs probability of false alarm at $L_p = 4$.

in [73]. The mutual coupling effect is added to the signatures of MC in the figures.

Figure 4.19 presents the probability of detection γ versus the probability of false alarm α , also known as Receiver Operating Characteristics (ROC) curve. Generally speaking, the closer the ROC curve is to the top left corner, the better the performance. It is observed that the proposed **d** outperforms the **H_b** signature at different antenna array sizes ($N = 64$ and $N = 128$), noting that the mutual coupling effect slightly improves the performance of both **H_b** and **d** based authentication approaches.

Figure 4.20 shows the detection probability γ versus SNR. It is observed that the proposed **d** signature outperforms the **H_b** signature for both $L_p = 4$ (sparser) and $L_p = 20$ (less sparse) without and with mutual coupling effect. The proposed signatures (**d** and **d'**) clearly show the advantages of virtual AoA and virtual AoD since the proposed signature takes the virtual (AoD, AoA) pairs of the specific principal component and converts them into a vector of distances. Moreover, the figure shows that the beamspace channel signature **H_b** experiences performance degradation under imperfect channel estimation. In contrast, the proposed signatures exhibit negligible performance degradation, highlighting the robustness of the proposed signatures to noise and channel imperfection.

Figure 4.21 presents the probability of false alarm α versus SNR for the proposed **d** signature with and without mutual coupling effect when attacker's channel is correlated to Bob's channel, where the correlation model is defined as in [121] with a correlation factor ρ of $\rho = 0, 0.8, 0.9, 0.99$. From **Figure 4.21**, it can be seen that the mutual coupling effect of the ideal array design slightly amplifies the proposed **d** signature as the degree

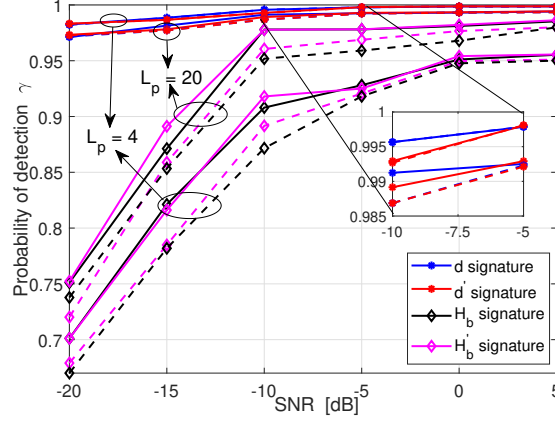


Figure 4.20: Probability of detection vs SNR analysis at $\alpha = 0.5$ and $N = 64$ for $L_p = 4$ and $L_p = 20$. Solid lines represent the signatures with perfect channel estimation, while the dash lines represent the signatures with imperfect channel estimation.

of channel correlation between Bob and the attacker increases, suggesting that the mutual coupling effect is more effective for the proposed signature in a correlated environment where stronger amplification is expected when controlled mismatch of the array design is added. For example, at SNR = 0 dB, in the uncorrelated case, $\rho = 0$, the proposed **d** signature with mutual coupling effect has a performance improvement of about 2.5% in α compared to the signature without coupling effect. In contrast, in the highly correlated channel case, $\rho = 0.99$, the performance with coupling effect exceeds up to 3.7%.

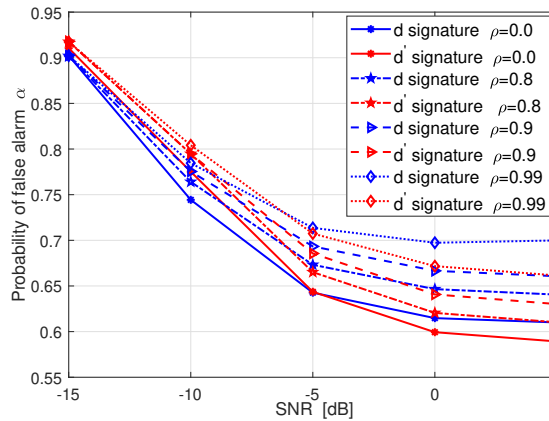


Figure 4.21: Probability of false alarm vs SNR analysis for correlated and uncorrelated channels at $\gamma = 0.5$, $N = 64$, and $L_p = 4$.

4.5. Results Of Study V

In this section, we evaluate the performance of the proposed tracking-based PLA scheme through simulation analysis in terms of the probability of detection γ and probability of false alarm α . We assume that the attacker has the same capability as Alice, with $N_E = N_A = 128$ antenna elements, while Bob has $N_B = 256$ antenna elements, and uses $M = 3$ for the wideband systems. The EKF parameters utilized in this study are adopted from [111] where Σ_T and Σ_R are assumed to be 0.25° in a static environment, and 0.5° and 0.75° in a dynamic environment. The initial filter state is $\mathbf{x}_{t_o} = \hat{\mathbf{x}}_o$, where $\hat{\mathbf{x}}_o$ is the estimated values of AoDs and AoAs, and its covariance is $\mathbf{P}_{t_o} = \text{diag}([\Sigma_T, \Sigma_R])$.

Table 4.4: Simulation parameters for the tracking-based PLA scheme.

Parameters		Value
Operating frequency f_c		60 GHz
Antenna elements N_E and N_A		128
Antenna elements N_B		256
SNR		20 dB
Number of subcarrier frequencies N_{FFT}		1024
Number of channel paths L_p		1, 4, 8, 20, 30
$\Sigma_T = \Sigma_R$		$0.25^\circ, 0.5^\circ, 0.75^\circ$
v_{max}		0, $0.01f_c$, $0.1f_c$
Tracking period T_o		6
Channel updates	ρ	0.99, 0.8, 0.7
	ϱ	0.01, 0.1, 0.5
Narrowband	System bandwidth W	100 MHz
Wideband	System bandwidth W	5 GHz
	$M = M_1 = M_2 = \dots = M_{L_p}$	3

Unless stated otherwise, the numerical values of the used system parameters are given in **Table 4.4**. Note that $L_p = 4$ corresponds to a high sparse channel environment while $L_p = 20$ corresponds to a low sparse one. When the system is static, the parameters are set to $v_{\text{max}} = 0$, $\rho = 0.99$, and $\varrho = 0.01$. In a low-dynamic environment, the parameters are adjusted to $v_{\text{max}} = 0.01f_c$, $\rho = 0.8$, and $\varrho = 0.1$, while in a high-dynamic environment, $v_{\text{max}} = 0.1f_c$, $\rho = 0.7$ and $\varrho = 0.5$.

The results of the ED signature-based PLA scheme $\mathbf{d}_e$ are benchmarked against the distance signature proposed in [6] referred to $\mathbf{d}$ and the sparse beamspace channel signature presented in [72] referred to $\mathbf{H}_b$.

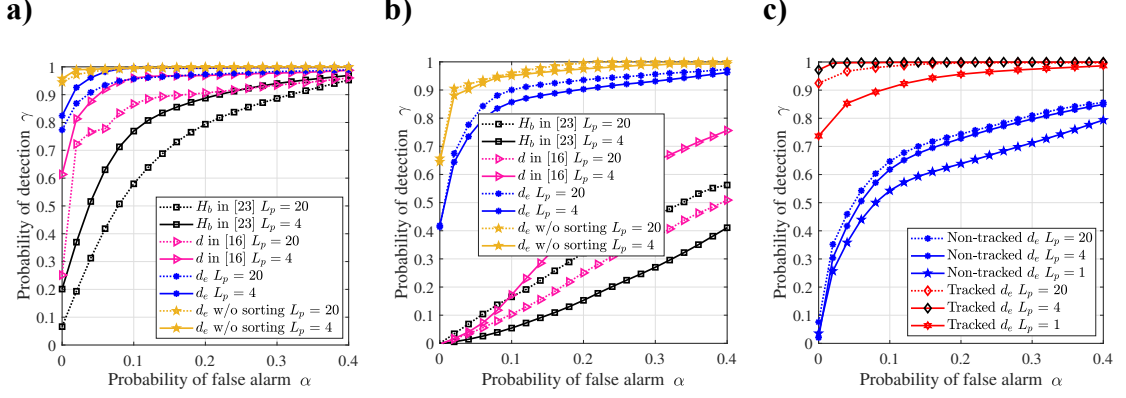


Figure 4.22: Probability of detection vs probability of false alarm analysis for **a)** static, **b)** low-dynamic, and **c)** high-dynamic environment in narrowband transmission.

To provide a more comprehensive evaluation of our proposed approach, we evaluate the performance of the proposed ED signature with and without the sorting step in (6) and (18). The performance of the proposed ED signature with the sorting step refers to “ $\mathbf{d}_e$ ” and without the sorting step refers to “ $\mathbf{d}_e$ w/o sorting” in the figures below.

4.5.1. Narrowband transmission

In this section, we evaluate the authentication performance in narrowband transmission where the gain remains the same for each frequency in the range, given that the system bandwidth does not exceed 100 MHz. Narrowband communication is the perfect option for any application that needs reliable, low-power, and long-range communications.

In **Figure 4.22**, we present the analysis of the probability of detection (γ) versus the probability of false alarm (α) across different sparsity levels in narrowband transmission. In **Figure 4.22a**, we observe that the proposed ED signature $\mathbf{d}_e$ consistently outperforms signatures $\mathbf{d}$ and $\mathbf{H}_b$ across varying levels of channel sparsity in static environments. This superior performance underscores the effectiveness of our proposed approach, particularly in scenarios with sparse channels.

Moving to **Figure 4.22b**, we delve into the performance of the ED signature in dynamic environments. Here, our approach demonstrates significant improvement over benchmark signatures. This enhancement is primarily attributed to the robustness of the ED signature in adapting to arbitrary channel variations, facilitated by the refinement of the sorting mechanism within our algorithm.

Comparing the performance of our proposed ED signature with and without the sorting step in **Figure 4.22a** and **Figure 4.22b**, it is observed that while the sorting mechanism enhances the consistency between Alice and Bob, it introduces a trade-off by slightly reducing the signature's randomness. This reduction in randomness can, in some scenarios, impact the robustness of the signature, particularly in environments with high interference or noise levels thus slightly degrading the PLA system performance.

Figure 4.22c presents the performance of our tracking-based PLA approach in high-dynamic environments. It is noticed that the tracked ED signature exhibits enhanced performance and resilience across different sparsity levels. Notably, as the number of paths increases, the degradation in performance is mitigated due to the application of the tracking approach. Specifically, the tracking mechanism plays a pivotal role in continuously updating the stored signature for the intended user based on channel variation statistics. This proactive adjustment reduces the mismatch between the intended user's stored signature and their current signature. Consequently, the need for frequent authentication processes for that user is diminished, contributing to a more efficient and seamless authentication experience. Remarkably, from **Figure 4.22c**, our approach remains applicable even in single-path systems, $L_p = 1$, unlike previous methods such as those based on signatures $\mathbf{d}$ and $\mathbf{H}_b$, which fail to operate effectively in such scenarios.

4.5.2. Wideband transmission

The authentication performance in the wideband transmission is evaluated in this section, where the system bandwidth goes beyond the channel coherence bandwidth and reaches the GHz bands. It is preferable in applications with high data rates such as IoT devices, criminal surveillance applications, and public safety services. Here, we assume that the system bandwidth is $W = 5$ GHz. Therefore, multicarrier signaling is used, such as OFDM with 1024 subcarriers ($N_{\text{FFT}} = 1024$). For simplicity, it is assumed that the number of extracted distances from all paths is similar (i.e., $M = M_1 = M_2 = \dots = M_{L_p}$) and equal to 3 ($M = 3$) where we utilize the first, last, and middle subcarriers (f_1 , $f_{N_{\text{FFT}}/2}$ and $f_{N_{\text{FFT}}}$) in the system band to extract the distance signature within one path.

Figure 4.23 represents the probability of detection γ versus probability of false alarm α analysis at different sparsity levels in wideband transmission under the beam squint effect. **Figure 4.23a** shows the ED signature performance in a static environment, while

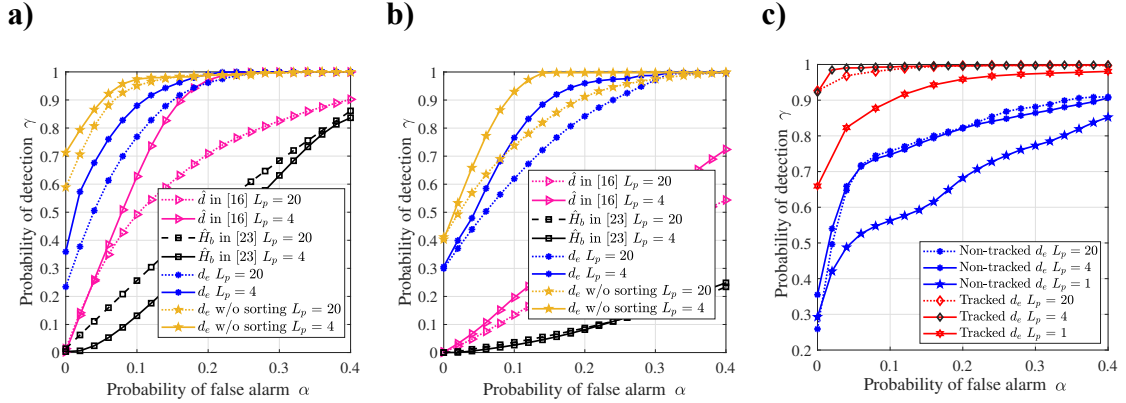


Figure 4.23: Probability of detection vs probability of false alarm analysis for **a)** static, **b)** low-dynamic, and **c)** high-dynamic environment in wideband transmission.

Figure 4.23b represents the performance in a low-dynamic environment.

For a fair comparison, both the distance signature in [6] and the beamspace channel signature in [72] are modified to be applicable to wideband transmission. For the distance signature, the channel gain-based sorting mechanism is applied to the extracted distance vector at each utilized subcarrier frequency (first, middle, and last subcarriers). Then, the sorted distance vectors are arranged in one vector called $\hat{\mathbf{d}}$. Similarly, the beamspace channel at the utilized subcarriers (first, middle, and last subcarriers) is extracted and stored in one matrix signature called $\hat{\mathbf{H}}_b$. It is noticeable from both figures, **Figure 4.23a** and **Figure 4.23b**, that the proposed ED signature-based wideband transmission outperforms the existing ones $\hat{\mathbf{d}}$ and $\hat{\mathbf{H}}_b$ in both static and dynamic environments. This superiority arises from the fact that the proposed ED signature is constructed based only on virtual angle indices in the beamspace representation corresponding to estimated angles and the distance between the indices and the original point. It is proven that the virtual angle estimation is robust to noise and exhibits high reciprocity even in low SNR regimes. Consequently, sudden channel variations do not significantly affect the estimated virtual angles, ensuring that the proposed ED signature remains stable. In contrast, $\hat{\mathbf{d}}$ and $\hat{\mathbf{H}}_b$ signatures depend on the path gain at each virtual angle, which reduces their stability and makes them more susceptible to noise and channel variations. In wideband transmission with a frequency-selective fading channel, these variations become more pronounced, further diminishing the performance of these signatures. Additionally, the figure shows that the proposed approach makes the authentication performance less dependent on the channel's sparsity level.

Figure 4.23c showcases the performance of our tracking-based PLA approach in a highly dynamic environment. Mirroring the trends observed in narrowband transmission, the tracked ED signature-based wideband transmission exhibits a substantial improvement in authentication performance compared to non-tracked signatures across different sparsity levels. Moreover, our proposed tracking-based PLA approach demonstrates adaptability even in single-path channels, further extending its applicability. In summary, the tracking-based approach presents a promising solution for ensuring secure and reliable authentication in wideband mmWave MIMO systems, offering robust performance from the lowest to the highest sparsity levels.

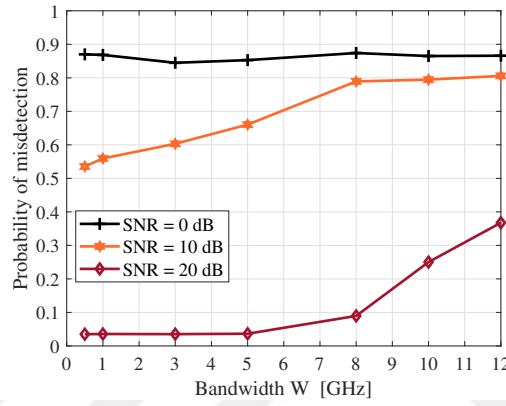


Figure 4.24: Probability of misdetection vs system bandwidth analysis.

In order to evaluate the performance of our proposed approach within wideband transmission, **Figure 4.24** illustrates the analysis of the probability of misdetection versus system bandwidth over different SNR levels. The misdetection probability is given as $P_{\text{md}} = 1 - \gamma = 1 - \Pr(\eta_L > \eta_{th} | K_1)$. Here, we assume that the system has only 4 paths ($L_p = 4$), and the false alarm rate is 0.1 ($\alpha = 0.1$) operating in a low-dynamic scenario. In our proposed beam squint-based ED signature extraction, we assume M_l is different for different paths and is calculated based on the different d values that are extracted using (3.110) and (3.111) from each path. The figure reveals that with an increase in bandwidth, the misdetection probability of our proposed signature increases, holding true for both low and high SNR levels. This phenomenon is due to the escalating squinting effect with larger bandwidths, causing the squinted angles to exhibit greater variations, especially in dynamic environments. Consequently, the tracking process becomes necessary to address these variations effectively.

To evaluate the robustness of our proposed approach against imperfect channel estimation

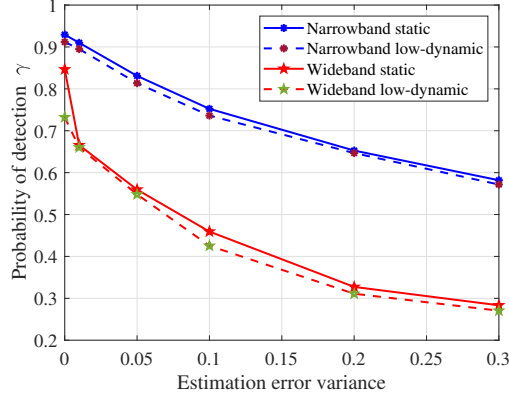


Figure 4.25: Probability of detection vs channel estimation error analysis.

parameters, **Figure 4.25** shows the analysis of the probability of detection γ versus channel estimation error in both narrowband and wideband transmissions at a false alarm rate of 0.1. We observe that the detection rate of our proposed signature decreases by approximately 33% in narrowband systems and around 64% in wideband systems for both static and low-dynamic scenarios. Despite high error variance, our proposed signature demonstrates robustness to imperfect channel parameter detection in narrowband transmissions. However, the impact of imperfect estimation is more significant on the proposed PLA approach in wideband transmissions. This difference in robustness between narrowband and wideband systems can be attributed to the inherent characteristics of wideband channels. Wideband systems experience more pronounced frequency selectivity and beam squint effects, which exacerbate the sensitivity to estimation errors. These factors lead to greater variation in the channel response across different frequencies, making accurate channel estimation more challenging.

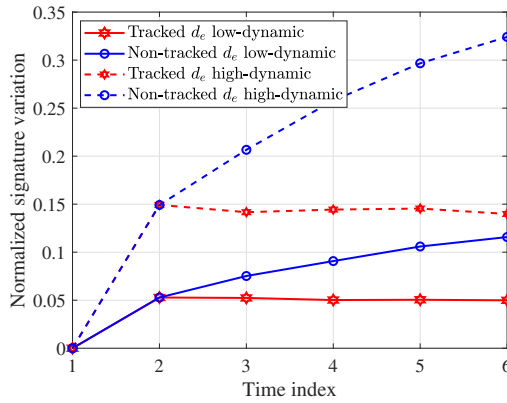


Figure 4.26: The normalized signature variation vs time analysis.

Figure 4.26 displays the behavior of our enhanced distance signature $\mathbf{d}_e$ during the tracking period in low- and high-dynamic environments. The variation of the normalized enhanced signature over time is given as

$$\delta(t_i) = \left\| \frac{\mathbf{d}_e(t_i) - \mathbf{d}_e(t_{i-1})}{\mathbf{d}_e(t_{i-1})} \right\|. \quad (4.2)$$

The figure illustrates a noteworthy trend: the enhanced signature variation converges towards a fixed value over time with the adoption of the tracking process in the authentication scheme. Conversely, in higher dynamic systems without tracking, signature variation experiences a significant increase. This divergence in behavior can be attributed to the fundamental design of the proposed signature, which leverages low-variation physical layer features to build a unique signature to legitimate nodes. By implementing the tracking process, we exert control over this variation, effectively stabilizing the signature over time. This stabilization not only extends the duration of authorized connections but also ensures sustained authentication performance. Essentially, the tracking mechanism acts as a stabilizing force, mitigating the impact of dynamic channel conditions and preserving the integrity of the authentication process.

CHAPTER 5

5. CONCLUSION AND FUTURE WORK

In this section, we present the key conclusions of our study on beamforming design and efficient PLA schemes in current and future wireless networks. We discuss the improvements in beamforming techniques to address challenges such as beam squint and ANS in XL-MIMO systems. Furthermore, we explore the development of robust PLA schemes that enhance security and authentication in wireless communications, paving the way for more reliable and efficient networks in the 6G era and beyond.

The summary of contributions and findings of each contribution part of this thesis is given as follows:

A sub-grouped LAS-MIMO architecture with low-complexity beam selection and associated precoding based on the partial exhaustive search and SIC-based algorithm are proposed. The proposed sub-grouped system was introduced to improve the energy efficiency of a recently presented fully-connected LAS-MIMO system. In the proposed design, the LAS grouping can be selected to maximize energy efficiency while achieving a functional sum-rate performance with the proposed precoding method. In addition, the power consumption models for the fully-connected and the proposed sub-grouped LAS-MIMO system are presented. Comparisons with fully- and sub-connected hybrid TA-MIMO systems show that the proposed sub-grouped LAS-MIMO architecture provides significant improvement in energy efficiency while achieving spectral efficiency approaching that of the traditional sub-connected MIMO system as the number of lenses in each sub-group increases. A trade-off between energy and spectral efficiency should be considered for an optimal design of the sub-grouped LAS-MIMO system. This is because when the number of lenses in each subgroup increases, the spectral efficiency increases to a sub-optimal solution, but a degradation in the energy efficiency of the system is observed.

To mitigate the beam squint effect in UWB mMIMO systems, a novel analog transceiver design leveraging the LAS architecture and subband filters is proposed. The design is practical with traditional exhaustive search-based precoding and a more straightforward threshold-based precoding technique. Simulation results confirm the robust system performance of the proposed LAS-based transceiver design under both precoding strategies.

The challenge of IBI is addressed in UWB mMIMO systems, considering the influence of the beam squint effect. We thoroughly analyzed IBI, exploring its implications in near-field and far-field propagation and considering both narrowband and wideband transmission scenarios. Additionally, we introduced a novel sub-grouped beam-based scheduling algorithm to alleviate IBI's impact on system performance. Our simulations offer insights into IBI analysis across various system scenarios and their consequences for system capacity. The proposed algorithm proves effective in enhancing the capacity of UWB mMIMO systems.

A novel distance and hybrid signatures-based PLA scheme is proposed for mmWave MIMO systems against spoofing attacks. The proposed distance signature is extracted from the beamspace representation of the MIMO channel. In contrast, this representation and the mutual coupling effect are jointly used to generate the hybrid signature. The robustness of the proposed signatures in terms of detection rate and false alarm is evaluated. It was found that the proposed signatures outperformed the beamspace channel signature in different sparse environments and antenna array sizes. This is due to the robustness of the beamspace channel characteristics to noise effects. Moreover, the hybrid signature outperforms the distance signature in a scenario where the channels of the attacker and the legitimate nodes are correlated, with an improvement of about 3.7% in an environment with high channel correlation.

In mobility scenarios, a novel tracking-based PLA scheme is presented for narrowband and wideband mmWave MIMO networks. The proposed scheme introduces an ED signature that can be tracked and updated, providing more extended authentication periods in both TDD and FDD transmissions within static and dynamic environments. Simulation results demonstrate the superior authentication performance of the proposed scheme compared to existing approaches in narrowband and wideband scenarios. This proposed system

provides an efficient authentication solution for future mobile networks prevalent in next-generation wireless networks.

Possible extensions of the current works of this thesis are as follows:

- **Hybrid Precoding Design for Beam Squint Compensation:** Hybrid precoding combines analog and digital processing to optimize beamforming under the effect of beam squint, which should be considered with the involvement of the analog LAS design. In multi-user and multi-cell scenarios, beam squint effects become more severe due to several factors: varying user distances and angles cause beam misalignment; overlapping coverage areas increase interference; user mobility introduces dynamic changes in beamforming; frequency-dependent distortions lead to inconsistent performance across bands; and increased system complexity makes effective beamforming more challenging. Future research should focus on designing hybrid precoding techniques that adapt to varying distances and frequencies, ensuring consistent beam performance.
- **Exploiting Grating Lobe Characteristics in Spatial Aliasing:** Exploiting GLs for advanced sensing applications could enable more precise environmental monitoring and localization. Future studies should also investigate techniques for managing GLs to minimize interference while harnessing their potential benefits, ensuring their characteristics are used effectively in communication and sensing applications. Integrating GL management with spatial reuse techniques will be crucial for optimizing the use of the available spectrum in high-density environments. Researchers should explore how GL characteristics can be leveraged for more efficient spatial reuse, allowing multiple users and cells to operate concurrently without significant performance degradation. This approach can improve spectral efficiency and better support for future communication networks' high data rate demands. Future research should also consider the mobility of users in both multi-user and multi-cell scenarios. As users move, the relative positions of GLs and the main beams will change, necessitating real-time adjustments to beamforming and interference management strategies. Researchers can ensure consistent performance and reliability in dynamic environments by developing adaptive techniques that account for user mobility.

- **Utilizing Non-Stationarity in XL-MIMO Channels:** Integrating non-stationarity-aware designs with advanced interference management and spatial reuse strategies will be crucial for maximizing the benefits of XL-MIMO systems in high-density environments. In multi-user scenarios, non-stationary channels can lead to uneven signal quality across users due to the varying propagation conditions at different array elements. Algorithms must dynamically adjust beamforming weights and allocate resources in real time to cater to these variations, ensuring that each user receives optimal signal quality. This also involves dealing with the increased interference potential when multiple users are served simultaneously, requiring sophisticated interference mitigation techniques considering the channels' non-stationary nature. The challenge is even more significant for multi-cell scenarios, where multiple base stations serve overlapping areas. ANS can cause substantial differences in channel characteristics across the cells, making coordinated beamforming and resource allocation essential. Techniques like coordinated multi-point transmission and reception can be adapted to handle ANS, ensuring a seamless user experience as users move across cell boundaries. Moreover, implementing ANS-aware designs can improve spectral efficiency by enabling more effective spatial reuse and reducing inter-cell interference. Incorporating ANS into the design of XL-MIMO systems for multi-user and multi-cell scenarios also means considering users' mobility patterns. High mobility users may experience rapid changes in channel conditions, necessitating faster adaptation mechanisms in the beamforming and resource allocation processes. By leveraging the ANS properties, researchers can develop more robust systems that maintain high performance even in dynamic environments. Future research should also explore integrating machine learning techniques to predict and adapt to ANS channel conditions. Using real-time data and historical patterns, machine learning algorithms can anticipate changes in the channel and proactively adjust the system parameters to maintain optimal performance. This approach can significantly enhance XL-MIMO systems' efficiency and reliability in multi-user and multi-cell scenarios.

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PUBLICATIONS, PRESENTATIONS, AND PATENTS ON THE THESIS:

- L. Afeef, H. M. Furqan, and H. Arslan, "Robust tracking-based physical layer authentication for narrowband and wideband mmWave MIMO systems," in IEEE Transactions on Information Forensics and Security, October 2024.
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