Non-Orthogonal Multi-Dimensional Modulation and Nonlinear Distortion Compensation for Beyond 5G

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A thesis submitted in partial fulfillment of the requirements for the Master of Engineering Science degree in Electrical and Computer Engineering
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Abstract

The introduction of new advanced technologies such as higher carrier frequencies, ultra-wide bandwidth, and increased transmission rate in 5G to support ever growing quality-of-service (QoS) demands have brought new challenges such as transmitter-receiver pair specific and domain specific non-orthogonality induced among spatial, time-frequency, and delay-doppler domain radio resource blocks and nonlinear distortions induced among multiple-input multiple-output (MIMO) antennas in spatial domain. In such conditions, current communication systems encounter severe performance degradation and incur higher operational cost. Based on this observation, this thesis aims at creating new multi-dimensional modulation techniques and nonlinear predistortion architectures to achieve higher communication performance with less operational cost.

Firstly, a customized, situation-aware multi-dimensional modulation (MDM) technique is developed with the goal of achieving maximized data rate under joint non-orthogonality conditions in spatial and time-frequency domains. The proposed MDM scheme is designed to take into account the non-orthogonality degrees induced in those domains and jointly optimize the radio resource block attributes to achieve the goal. Secondly, to minimize receiver side operational cost while supporting required data rate under joint non-orthogonality conditions in spatial, time-frequency, and delay-doppler domains, a user-centric multi-dimensional modulation (UC-MDM) technique is developed. The proposed situation-aware and cost-aware UC-MDM is designed to take into account the transmitter-receiver pair specific non-orthogonality degrees and the receiver side operational cost, and intelligently utilize optimum radio resource combination through MDM to achieve the goal. Finally, to reduce complexity of current multi-input digital predistortion (DPD) models for nonlinear distortion compensation, a less complex, decomposed, scalable cross-correlation based DPD (CC-SISO DPD) architecture is proposed for massive MIMO systems.

Through simulation results and analysis, it is demonstrated that the proposed novel, customized, domain specific solutions can achieve higher communication performance with reduced operational cost in future wireless networks.

**Keywords:** 5G, 6G, cross-correlation, crosstalk, multi-dimensional modulation, non-orthogonality, user-centric
Lay Summary

5G and beyond wireless networks are becoming densified and operated with advanced communication technologies to support diverse QoS requirements. However, such advanced technologies have also brought new challenges in multiple domains which needs to be closely investigated to efficiently achieve higher performance communication. Therefore, this thesis mainly focuses on those challenges such as transmitter-receiver pair specific non-orthogonality induced among radio resources in spatial, time-frequency, and delay-doppler domains and the nonlinear distortions induced in spatial domain.

First of all, to achieve maximized communication data rate under joint non-orthogonality in spatial, and time-frequency domains, a multi-dimensional modulation (MDM) technique is proposed. The main premise behind the proposed MDM is to simultaneously utilize spatial, and time-frequency domain radio resources through multi-dimensional modulation and also jointly optimize the spatial-time-frequency domain radio resource block attributes to minimize the overall non-orthogonality degree in those domains and thus achieve maximized data rate.

Then, focusing on transmitter-receiver pair specific non-orthogonality degrees induced among radio resources in spatial, time-frequency and delay-doppler domains and the domain specific receiver side operational costs for orthogonality restoration and demodulation, a user-centric multi-dimensional modulation (UC-MDM) technique is proposed. The situation-aware and cost-aware UC-MDM is designed to intelligently utilize the optimum combination of radio resources with optimum radio resource separation through MDM in either spatial-time-frequency or spatial-delay-doppler domains to minimize the user-device operational cost while achieving the required communication data rate.

Finally, considering the severity of nonlinear distortions induced by nonlinear and reverse crosstalk among MIMO antennas and the exponentially increasing operational cost of current multi-input DPD models for nonlinear distortion compensation in MIMO, a decomposed cross-correlation based single-input DPD (CC-SISO DPD) architecture is proposed. The proposed CC-SISO DPD architecture can estimate the nonlinear and reverse crosstalk with high accuracy and with significantly less operational cost than multi-input DPD models. Furthermore, the proposed CC-SISO DPD eliminates the requirement for signal feedback in transmit path, and thus reduces the overall hardware implementation complexity in massive MIMO arrays.
Acknowledgments

I would like to express my sincere gratitude to Dr. Xianbin Wang for his valuable guidance, support, and care throughout my graduate research. His insightful supervision allowed me to explore interesting valuable research directions and solve new critical challenges encountered in upcoming wireless generations. Dr. Wang has always been very supportive and provided great ideas and strategies to approach and solve real world problems which nurtured my research skills. I’m also grateful for the valuable opportunity to engage in industrial projects during my graduate research. I had a great enjoyable research experience and it is always an honour and a privilege to pursue research under Dr. Wang’s supervision and guidance. The research skills I developed throughout my graduate study and research will certainly help me to tackle and solve critical challenges faced by the industry.

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<tr>
<td>5G</td>
<td>Fifth-generation</td>
</tr>
<tr>
<td>6G</td>
<td>Sixth-generation</td>
</tr>
<tr>
<td>AWGN</td>
<td>Additive white Gaussian noise</td>
</tr>
<tr>
<td>BER</td>
<td>Bit error rate</td>
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<tr>
<td>BS</td>
<td>Base station</td>
</tr>
<tr>
<td>CC-SISO</td>
<td>Cross-correlation based single-input-single-output</td>
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<tr>
<td>CDF</td>
<td>Cumulative distribution function</td>
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<tr>
<td>CFO</td>
<td>Carrier frequency offset</td>
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<td>CP</td>
<td>Cyclic prefix</td>
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<tr>
<td>DPD</td>
<td>Digital predistortion</td>
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<tr>
<td>IAC</td>
<td>Inter antenna correlation</td>
</tr>
<tr>
<td>ICI</td>
<td>Inter carrier interference</td>
</tr>
<tr>
<td>IFFT</td>
<td>Inverse fast fourier transform</td>
</tr>
<tr>
<td>ISFFT</td>
<td>Inverser symplectic fast fourier transform</td>
</tr>
<tr>
<td>ISI</td>
<td>Inter symbol interference</td>
</tr>
<tr>
<td>LTE</td>
<td>Long term evolution</td>
</tr>
<tr>
<td>MDM</td>
<td>Multi-dimensional modulation</td>
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<tr>
<td>MIMO</td>
<td>Multi-input-multiple-output</td>
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<tr>
<td>mmWave</td>
<td>Millimeter wave</td>
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<tr>
<td>OFDM</td>
<td>Orthogonal frequency division multiplexing</td>
</tr>
<tr>
<td>OTFS</td>
<td>Orthogonal time-frequency-space modulation</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<tr>
<td>PA</td>
<td>Power amplifier</td>
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<tr>
<td>PDF</td>
<td>Probability density function</td>
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<tr>
<td>QAM</td>
<td>Quadrature amplitude modulation</td>
</tr>
<tr>
<td>QoS</td>
<td>Quality of service</td>
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<tr>
<td>RF</td>
<td>Radio frequency</td>
</tr>
<tr>
<td>SINR</td>
<td>Signal to interference noise ratio</td>
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<tr>
<td>SM</td>
<td>Spatial modulation</td>
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<td>TO</td>
<td>Time offset</td>
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<tr>
<td>UC-MDM</td>
<td>User-centric multi-dimensional modulation</td>
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<tr>
<td>UE</td>
<td>User equipment</td>
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<tr>
<td>XR</td>
<td>Immersive extended reality</td>
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</table>
List of Symbols

\[ \alpha_e \quad \text{Nonlinear crosstalk coefficient} \]
\[ \beta_e \quad \text{Reverse crosstalk coefficient} \]
\[ \theta \quad \text{Spatial phase separation} \]
\[ \rho_f \quad \text{Frequency domain non-orthogonality} \]
\[ \rho_s \quad \text{Spatial domain non-orthogonality} \]
\[ \rho_t \quad \text{Time domain non-orthogonality} \]
\[ \rho_{de} \quad \text{Delay domain non-orthogonality degree} \]
\[ \rho_{do} \quad \text{Doppler domain non-orthogonality degree} \]
\[ \Upsilon \quad \text{Signal-to-interference-noise ratio} \]
\[ \Delta f \quad \text{Subcarrier spacing} \]
\[ \Delta \tau \quad \text{Delay resolution} \]
\[ \Delta v \quad \text{Doppler resolution} \]
\[ \tau_p \quad \text{Delay period} \]
\[ v_p \quad \text{Doppler period} \]
\[ B \quad \text{Bandwidth} \]
\[ f_c \quad \text{Carrier frequency} \]
\[ g(\cdot) \quad \text{Windowing function} \]
\[ H \quad \text{Channel matrix} \]
\[ H_c \quad \text{Spatially correlated channel matrix} \]
\[ K \quad \text{Power amplifier order} \]
\( N \)  Number of subcarriers  
\( N_s \)  Number of antenna subsets  
\( N_t \)  Number of transmit antennas  
\( Q \)  Power amplifier memory depth  
\( R \)  Data rate  
\( T \)  Total transmission time  
\( T_s \)  Symbol duration  
\( v \)  UE speed
Chapter 1

Introduction

1.1 Overview of 5G and 6G wireless communication technologies

Wireless communication networks have evolved tremendously since 1980s, with a new generation emerging every decade. This remarkable evolution and the promised quality of service (QoS) delivery in each generation has become practicable with outstanding efforts from research, development, experimentation, and standardization fields. Since 2020, the vision and research efforts to frame beyond 5G and 6G wireless communication technologies have begun with several efforts from both academia and leading industrial research partners.

Different from legacy wireless networks, the beyond 5G and 6G are use case driven wireless technologies being developed to support advanced communication scenarios such as machine to machine communication, immersive extended reality (XR), 3D haptic hologram telepresence, unmanned aerial vehicular communication, vehicle-to-everything communication (V2X), digital twin and connected intelligence [1]. These enterprise driven use cases will play a major role towards building a hyper connected world, or in other words “the internet of everything” as illustrated in Fig. 1.1 [2]. The new hyper connectivity wireless communication technologies involving humans and machines will provide ultimate seamless multimedia experience to the end users, and allow the high performance industrial IoT applications to evolve from current industry 4.0 towards industry X.0.
Figure 1.1: Beyond 5G and 6G use cases [2]
In order to support such advanced use cases of future wireless communication systems, the RF front-end is required to deliver stringent yet diverse QoS requirements of different wireless devices such as user experience datarate upto 1 Gbps, increased spectral efficiency upto 100 bps/Hz, aggregated data throughput of 1000 Gbps, ultra connectivity to support 10 billion devices/km², ultra reliability ≥ 99.99999% and user plane latency ≤ 0.1 ms [2].

To achieve such diverse QoS requirements, network densification is being carried out in 5G and beyond wireless networks with advanced wireless technologies such as mmWave carrier frequencies in the range of 30 GHz to 100 GHz, MIMO antennas with hybrid beamforming capabilities, massive MIMO arrays with smaller form factor (20×20 array at 300 GHz of size (1 cm²), multi-dimensional modulation and multi-dimensional multiple-access techniques [4] as illustrated in Fig 1.2.

![Figure 1.2: Network densification and new technologies in beyond 5G](image)

However, with the introduction of such advanced technologies in beyond 5G wireless networks, the non-orthogonality conditions induced among radio resources in multiple domains and the non-linear distortions induced among MIMO antennas in spatial domain have become more severe, and thus consequently affect the overall communication performance. Therefore, to efficiently achieve higher communication performance under such non-orthogonality conditions and nonlinear distortions in different domains, this thesis proposes novel multi-dimensional modulations techniques and nonlinear distortion compensation architectures in the following chapters.
1.2 Research motivation

The operation of beyond 5G wireless networks with mmWave higher carrier frequencies upto 100 GHz, high-speed mobility upto 500 km/h, ultra-wide bandwidth upto 500 MHz, and closely packed MIMO antennas with reduced spatial separation have brought new challenges such as increased non-orthogonality degrees among radio resources in multiple domains, higher operational cost from user-device to restore orthogonality of such non-orthogonal radio resources, and strong nonlinear distortions among transmitters in MIMO arrays, which consequently affect overall communication performance. Such non-orthogonality degrees and non-linear distortions have become device dependent and domain specific due to diverse device attributes and different synchronization capability of devices in different domains. Based on these observations, the most critical challenges due to such transmitter-receiver pair specific non-orthogonality degrees among radio resources in different domains and the severity of spatial domain nonlinear distortions in MIMO arrays are extensively focused in this thesis. The major motivations of this thesis are summarized as follows:

- **Transmitter-receiver pair specific non-orthogonality degrees**: Although orthogonal radio resource blocks comprising radio resources from multiple domains are utilized for multi-domain modulation in 5G, the orthogonality among the radio resource blocks is destroyed due to stochastic nature of wireless channel, imperfect synchronization, complex multi-path signal propagation, and higher spatial channel correlation among antennas. These non-orthogonality conditions have become transmitter-receiver pair specific due to different device dependent attributes such as different carrier frequencies, different transmission rates, different UE speeds, and different synchronization capability of receiver in different domains. Under such transmitter-receiver pair specific non-orthogonality conditions in different domains, current OFDM based communication systems which inherently assume and inevitably rely on preservation of orthogonality among radio resources encounter severe data rate degradation [3]. Therefore, to achieve maximized communication performance under such non-orthogonality conditions in multiple radio resource domains, a more flexible, situation-aware, transmitter-receiver pair specific modulation technique is necessary.
1.2. Research motivation

- **Increased operational cost from user-device:** Considering the fact that multi-domain non-orthogonal radio resources have become a common paradigm for modulation in beyond 5G communication systems, it is imperative to control the overall operational cost from user-device for orthogonality restoration and demodulation in different domains so that the resource constrained user-device can be effectively utilized. The current two-dimensional modulation techniques such as OFDM in time-frequency domains [3] and OTFS in delay-doppler domains [5] do not aim to control the cost required from user device for orthogonality restoration and neither do have the flexibility to utilize the radio resources in different domains for efficient modulation. Therefore, a situation-aware and cost-aware unified multi-dimensional modulation solution which can consider the non-orthogonality degrees among radio resources in different domains and utilize the optimum combination of radio resources from different domains through multi-dimensional modulation is necessary to support the required communication performance with minimal operational cost from user-device.

- **Increased simultaneous nonlinear crosstalk among PAs:** Due to significantly reduced spatial separation in massive MIMO arrays, the crosstalk induced among transmitter paths have become more severe. As a result, severe nonlinear distortion with memory effects are induced from multiple PA branches which significantly affect the output signal quality, and overall spectral efficiency. At such strong crosstalk conditions in MIMO arrays, the current nonlinear distortion compensation techniques such as multi-input digital predistortion (DPD) models encounter sharp increase in operational cost and have poor scalability due to the exponential increase in number of coefficients required for digital predistortion in MIMO arrays larger than \( \geq 2 \times 2 \) [6]. In addition, such multi-input DPD architectures also require signal feedback tap before the PA in each transmitter path for DPD coefficient extraction. Although such signal tap is not a major concern for laboratory based experiments, it is infeasible to be implemented in commercial massive MIMO arrays due to smaller form factor. Therefore, a decomposed, less complex, and scalable DPD architecture which can effectively compensate the nonlinear distortions in spatial domain is paramount for massive MIMO arrays in the next generation wireless networks.
1.3 Research objectives

This thesis is motivated to solve the previously outlined critical challenges due to the transmitter-receiver pair specific and domain specific non-orthogonality and nonlinear distortions encountered in beyond 5G communication systems. Therefore, the research objectives of this thesis is summarized as follows.

- Design a novel situation-aware, customized, transmitter-receiver pair specific multi-dimensional modulation technique that can jointly optimize the modulation attributes of spatial-time-frequency domain radio resource blocks to minimize the non-orthogonal interference in each domain, and thus achieve maximized communication data rate under non-orthogonality conditions in spatial, and time-frequency domains.

- Design a novel, cost-aware, and situation-aware user-centric multi-dimensional modulation scheme that can utilize the optimum combination of radio resources with optimum radio resource separation through multi-dimensional modulation in either spatial-time-frequency or spatial-delay-doppler domains to minimize the overall operational cost (orthogonality restoration and demodulation cost) from user-device while achieving required data rate under transmitter-receiver pair specific non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains.

- Develop a novel, less complex crosstalk estimation method in spatial domain to estimate the nonlinear and reverse crosstalk induced simultaneously from multiple transmit paths in a MIMO array. Using the developed crosstalk estimation method, propose a decomposed, less complex, scalable DPD architecture for massive MIMO arrays.
1.4 Thesis contribution

The main contributions of this thesis to solve the aforementioned critical challenges are summarized as follows:

- Firstly, a transmitter-receiver pair specific, situation-aware multi-dimensional modulation (MDM) scheme that can jointly utilize the spatial and time-frequency domain radio resources is developed. The proposed MDM is designed to take into account the non-orthogonality degrees induced among radio resources in spatial, time-frequency domains and jointly optimize the spatial-time-frequency domain radio resource block modulation attributes to minimize the non-orthogonality degrees in those domains and thus achieve maximized data rate.

- Then, a novel, situation-aware, and cost-aware user-centric multi-dimensional modulation (UC-MDM) scheme is proposed with the goal of minimizing overall operational cost from receiver while supporting required data rate under transmitter-receiver pair specific non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains. The proposed UC-MDM is designed to jointly optimize the separation of radio resources in spatial, time-frequency, and delay-doppler domains and utilize the optimum combination of radio resources from those domains through multi-dimensional modulation in either spatial-time-frequency domains or spatial-delay-doppler domains to achieve the goal under different levels of non-orthogonality degrees in those domains.

- Finally, a less complex cross-correlation based crosstalk estimation method is developed to estimate spatial domain simultaneous nonlinear and reverse crosstalk induced from multiple PA branches in MIMO arrays. Using the developed method, a new scalable single-input (CC-SISO) DPD architecture is designed to overcome the requirement of signal taps in the transmitter paths, and thus significantly reduce the hardware implementation complexity. Furthermore, the decomposed CC-SISO DPD architecture incurs less operational cost compared to existing multi-input DPD models for MIMO systems.
## 1.5 Thesis organization

The following Fig. 1.3 illustrates the organization of remaining chapters of this thesis.

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<th>Objectives</th>
<th>Proposed Solution</th>
<th>Contributions</th>
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<tr>
<td>Achieve maximized data rate in non-orthogonal spatial, time-frequency domains.</td>
<td>Minimize receiver operational cost while achieving the required data rate in non-orthogonal spatial, time-frequency, delay-doppler domains.</td>
<td>Develop a complexity reduced DPD architecture to compensate nonlinear, and reverse crosstalk in spatial domain.</td>
</tr>
<tr>
<td><strong>Chapter 3</strong></td>
<td><strong>Chapter 4</strong></td>
<td><strong>Chapter 5</strong></td>
</tr>
<tr>
<td>Designed situation-aware MDM scheme that optimize the modulation attributes in spatial, time-frequency domains to minimize the non-orthogonal interference, and achieve maximum datarate.</td>
<td>Developed situation-aware and cost-aware UC-MDM scheme that can intelligently utilize optimum radio resource combinations from spatial, time-frequency, and delay-doppler domains through multi-dimensional modulation to achieve required datarate with minimal operational cost from UE.</td>
<td>Developed cross-correlation based technique to estimate linear and reverse crosstalk among multiple transmit paths in MIMO arrays. Designed a scalable, less complex CC-SISO DPD architecture for massive MIMO systems.</td>
</tr>
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</table>

Figure 1.3: Thesis organization

In chapter 3, the transmitter-receiver pair specific non-orthogonality among radio resources in different domains such as inter antenna correlation (IAC) in spatial domain, inter symbol interference (ISI) in time domain, and inter carrier interference (ICI) in frequency domain are modeled. It is shown that although such non-orthogonal interference vary differently in those domains, they combinedly affect the overall achievable data rate. Therefore, to achieve maximized data rate under such joint non-orthogonality conditions in those domains, a novel multi-dimensional modulation (MDM) is proposed. The proposed
MDM scheme is designed to utilize the spatial-time-frequency domain radio resources simultaneously through multi-dimensional modulation and jointly optimize modulation attributes of the resource block to achieve the goal. From simulation results, it is demonstrated that the proposed MDM can achieve maximized datarate under joint non-orthogonality conditions spatial, and time-frequency domains compared to state-of-art MIMO-OFDM systems. Further results show that MDM can gain higher data rate in massive MIMO systems.

In chapter 4, transmitter-receiver pair specific non-orthogonality degrees perceived by a UE in spatial, time-frequency, delay-doppler domains and the overall operational cost incurred by UE receiver to restore orthogonality of such non-orthogonal radio resource blocks, and the subsequent cost for demodulation in those domains are modeled. Considering such different non-orthogonality degrees perceived by UE in those domains, and the associated operational cost from receiver, a new user-centric multi-dimensional modulation (UC-MDM) technique is proposed to minimize the receiver side operational cost while supporting required data rate. The proposed UC-MDM is designed to minimize the non-orthogonality degree in each domain by jointly optimizing the radio resource separation and also utilize optimum combination of radio resources from spatial, time-frequency and delay-doppler domains through multi-dimensional modulation (either spatial-time-frequency or spatial-delay-doppler domains) to achieve the goal. Using simulation results, it is validated that UC-MDM achieves required data rate with minimal operational cost from user-device compared to state-of-art MIMO-OFDM and MIMO-OTFS systems.

In chapter 5, the nonlinear distortions with memory effects induced due to nonlinear and reverse crosstalk from multiple PA branches are modeled. It is shown that under such strong simultaneous crosstalk, the existing multi-input DPD models encounter exponential increase in complexity with the increase in MIMO array size. Therefore, a less complex cross-correlation based method is developed to estimate crosstalk from multiple PA branches. Using the developed method, a decomposed, scalable cross-correlation based single-input (CC-SISO) DPD architecture is designed such that it does not require signal taps before the PA, and therefore reduces the hardware complexity of current multi-input DPD models. Simulation analysis using 5G signals and actual PA configurations demonstrate high accuracy and less complexity of proposed CC-SISO DPD architecture for massive MIMO systems.
Chapter 2

Non-Orthogonality Conditions and Nonlinear Distortions in Beyond 5G

Wireless communication networks have evolved dramatically over the last 40 years with the goal of supporting human-to-human communication, machine-to-machine communication, and now towards connected intelligence. In this tremendous evolution process, modulation techniques to carry data play a key role towards fulfilling the QoS requirements of each wireless generation. To perform modulation in each wireless generation, orthogonal radio resource blocks from different domains such as time domain in 2G, frequency domain in 3G, time-frequency domain in 4G, spatial-time-frequency domains in early 5G systems have been utilized [7]. Although preservation of orthogonality among radio resource block is crucial to achieve efficient modulation and higher performance wireless communication, the orthogonality among radio resources is generally destroyed due to inevitable stochastic nature of wireless channels, imperfect synchronization of the receiver in different domains, and the hardware discrepancies at transmitter-receiver. As a result, non-orthogonal interference are induced among radio resources in multiple domains which combinedly affect the overall communication performance [9, 10, 31, 55]. Owing to the non-orthogonality mitigation techniques, the non-orthogonal interference have been maintained under tolerable level in 4G and predecessor wireless generations [16, 18, 23, 33, 37, 42, 43, 50, 51, 53, 55].

However, the introduction of mmWave higher carrier frequencies, higher transmission rate, smaller radio resource block size and smaller form-factor MIMO antennas in beyond 5G to
support advanced QoS requirements have increased the severity of non-orthogonality degrees induced among radio resources in different domains. As a result, significant non-orthogonal interference are induced in those domains which curtail the communication performance of current systems [11]. Furthermore, such non-orthogonality degrees have become transmitter-receiver pair specific and domain specific due to predominant device dependent attributes such as different carrier frequencies, different inter-antenna correlation, different synchronization performance of devices in different domains and different UE speeds [12, 13]. In addition, such increase in non-orthogonality degrees has also resulted in increase in operational cost from user-device to restore the orthogonality of such non-orthogonal radio resources. Although such operational cost is not a major concern at BS, it affects the power consumption in user-devices [14]. Furthermore, the nonlinear distortions due to nonlinear and reverse crosstalk induced among PA branches in closely packed MIMO arrays have also become significant thus deteriorate the transmit signal quality, and overall spectral efficiency [56].

Based on these observations, the non-orthogonality conditions in spatial, time-frequency, delay-doppler domains, and the nonlinear distortions in spatial domain are investigated in this chapter along with the existing solutions and the drawbacks of the current studies. Finally, this chapter is summarized with highlights of new device dependent, transmitter-receiver pair specific non-orthogonality degrees and nonlinear distortions encountered in beyond 5G wireless systems which will be addressed in the remaining chapters of this thesis.

2.1 Time-frequency domain non-orthogonality and existing solutions

In this section, the principle behind time-frequency domain modulation, the transmitter-receiver specific non-orthogonality conditions induced among time-frequency domain radio resources, the existing solutions to mitigate those non-orthogonality conditions and their drawbacks in mitigating new transmitter-receiver pair specific non-orthogonalities are analyzed.
In the current LTE as well as early 5G wireless communication systems, OFDM based time-frequency domain modulation is utilized to modulate transmission QAM symbols. For this purpose, the time domain radio resource (total transmission time $T$), and the frequency domain radio resource (total bandwidth $B$) are divided into basic time slots and multiple mutually orthogonal narrow-band subcarriers of $\Delta f$ bandwidth placed at equidistant separation in the frequency domain [3] as illustrated in Fig. 2.1.

![Figure 2.1: (a) Time-frequency domain radio resource block, (b) Orthogonal subcarriers](image)

As the name implies, the orthogonality among time-frequency domain radio resources is the main pillar of OFDM based communication systems to achieve high performance communication. For this purpose, mutually orthogonal subcarriers overlapped in frequency domain shown in Fig. 2.1 (b) are utilized during the modulation. The mutually orthogonal subcarrier are individually modulated via parallel $N$-point IFFT operation to carry the QAM information symbols. In order to maintain orthogonality between time-frequency domains, the modulation parameters are chosen such that $T_s \cdot \Delta f = 1$ where, $\Delta f$ is the subcarrier separation, and $T_s$ is the symbol duration. The OFDM systems inherently assume that the orthogonality among subcarriers is preserved or can be completely recovered at the receiver through time/frequency domain synchronization [3].

Although such orthogonal subcarriers are utilized duration modulation, due to imperfect synchronization in time/frequency domain, stochastic nature of wireless channel and inherent hardware discrepancies, the orthogonality among the radio resource blocks is often destroyed. The challenges encountered due to such loss of orthogonality among the time-frequency domain radio resources are analyzed in the following subsection.
2.1.1 Carrier frequency offset

Carrier frequency offset (CFO) occur among the orthogonal subcarriers due to imperfect synchronization at transmitter or receiver, frequency mismatch between the oscillators of receivers and BS, and doppler shifts induced in the wireless channel due to varying relative speed between receiver and BS. Hence, the transmitted orthogonal subcarriers loss their orthogonality, and tend to overlap each other at the receiver [15] as shown in Fig. 2.2.

![Figure 2.2: Carrier frequency offset (CFO) among subcarriers](image)

Such overlap between subcarriers induces inter carrier interference (ICI) in frequency domain which consequently degrades the achievable communication data rate [9]. Therefore, CFO estimation and ICI cancellation techniques are used to mitigate or suppress the ICI, and achieve higher communication performance. The efforts to estimate and suppress the ICI can be mainly classified into two directions, a) Estimating and compensating ICI using modulation attributes b) Blind estimation and suppression.

Following the first direction, authors in [15] proposed a low complex algorithm to process the received subcarriers, and evaluate a cost function at certain time steps in a discrete manner to evaluate CFO degree in each received subcarrier. Authors in [16] proposed an algorithm to estimate the CFO using good auto-correlation characteristic, such as Barker sequence, pseudo-noise (PN) sequence. The CFO was evaluated by measuring correlation peaks, and subcarrier
phase change rate at certain time intervals. Authors in [17] proposed an approach that use BER and MSE of the modulated signals as parameters to evaluate and suppress the CFO.

On the other hand, in fast varying channel conditions, compensation of multiple CFOs at the receiver using received signal attributes is not a feasible solution since the training waveforms lose their correlation properties in such environments. Therefore, to minimize the ICI in such conditions, techniques such as adaptive windowing [10], differential grouping [18], spectrum shaping [23], adaptive subcarrier grouping [24], iterative carrier frequency offset estimation, and compensation [25], and deep learning to compensate unknown carrier frequency offset [26] have been proposed.

Although these proposed methods were proved to perform well, in next generation wireless systems with higher carrier frequencies, high-speed mobility, and large doppler spreads, the CFO becomes more severe and more specific between a transmitter-receiver pair inducing different levels of ICI in different transmitter-receiver pairs. In such conditions, the current ICI compensation techniques that rely on receiver’s capability to mitigate the ICI do not perform well due to their inherent assumption of ICI upto a certain degree, and unawareness of device-dependent CFO levels [27]. Furthermore, the current multi-dimensional modulation techniques [19, 20, 21, 22] do not aim to minimize the transmitter-receiver pair specific ICI in frequency domain.

Therefore, a customized, situation aware, transmitter-receiver pair specific solution is required at BS to efficiently mitigate the CFO, and simultaneously achieve higher communication performance across different transmitter-receiver pairs in future wireless systems.

### 2.1.2 Time offset

In practical wireless channels, the transmitted signals often undergo reflection, refraction, and diffraction due to their inherent electromagnetic wave property when traveling through a wireless channel. Hence, multiple signal copies of the transmitted signal are generated in the communication environment each arriving at the receiver at different time stamps as illustrated in Fig. 2.3.
2.1. Time-frequency domain non-orthogonality and existing solutions

Generally, a cyclic prefix (also known as guard interval) which is a redundant copy of the last symbol of the transmission frame is appended to the beginning of the OFDM symbol in order to avoid loss of information caused by multi-path symbol overlap contamination. The cyclic prefix duration is selected such that it is larger than the expected delay spread of the channel. At the receiver, the redundant CP is dropped before demodulation. The CP duration $T_g$ has to be long enough for an expected delay spread of the channel, also a shorter $T_g$ is often desired since longer $T_g$ reduces the overall transmission efficiency [28]. However, at time-varying channel conditions, the multi-path delay sometimes becomes longer than CP duration leading to destruction of orthogonality among the received symbols in time domain or in other words overlap on CP+OFDM symbol causing time-offset (TO) as shown in Fig. 2.4 [28].
Therefore, to improve communication performance in such time-delayed multi-path propagation conditions, TO estimation and mitigation techniques such as polynomial optimization based blind equalization [29], an iterative data-aided channel estimation [30], CP-based correlation [31], extrinsic neural network equalization [32], raised cosine pulse shaping [33], a neural network based time domain equalizer [34], a pulse shaping filter to suppress the ISI [35], and a semi-blind estimation using expectation-maximization (EM) algorithm in [36] have been proposed in the literature.

Although the proposed methods performed well in their studies, those solutions were designed to suppress the ISI within a certain multi-path time delay, and in specific channel conditions. However, in future wireless environments, the time delay of multi-path signals vary differently in different transmitter-receiver pairs. In such systems, the existing solutions that generally target to tackle the worst delay spread become inefficient. Therefore, a situation aware, transmitter-receiver pair specific ISI mitigation solution is necessary.

2.2 Delay-doppler domain non-orthogonality and existing solutions

In this section, the main premise behind delay-doppler domain modulation, the transmitter-receiver specific non-orthogonality conditions induced among delay-doppler domain radio resource blocks, and the existing solutions to mitigate those non-orthogonality conditions are analyzed.

In the recent years, modulation of QAM symbols in delay-doppler domains instead of time-frequency domains have become popular due to the inherent resiliency of delay-doppler modulation towards time-frequency domain non-orthogonality conditions. The main premise behind OTFS transform based delay-doppler modulation is to convert a time-varying wireless channel into a time-invariant channel with a complex channel gain over all the transmitted symbols [5]. For this purpose, the delay-doppler domain modulation requires inverse symplectic fourier transform (ISFFT) operations in addition to conventional IFFT operations to modulate the QAM symbols into orthogonal delay-doppler domain bins.
2.2. Delay-doppler domain non-orthogonality and existing solutions

2.2.1 Delay-doppler spread

Although orthogonal delay-doppler domain resource blocks are utilized during modulation, due to increased carrier frequencies up to 100 GHz, ultra-wide bandwidth up to 500 MHz, and high-speed mobility scenarios in the range of 500 km/h in future wireless systems, large delay spreads, and large doppler spreads are induced in delay-doppler domain resource blocks thus deteriorate the overall communication performance [37] as shown in Fig. 2.5.

![Figure 2.5: High-speed mobility scenario with large delay, doppler spread](image)

In order to enhance the communication performance under such delay spread, and doppler spread conditions, techniques such as waveform based delay spread cancellation [38], adaptive subcarrier selection and grouping [39], delay-sensitive buffer-aided relay selection [40], delay aware optimization framework [41], pilot aided doppler spread estimation and compensation [42], and angle domain doppler compensation [43] have been proposed recently.

Although these methods perform well under delay spread and doppler spread conditions, they aim to mitigate the overall delay-doppler spread in the wireless channel without considering the transmitter-receiver pair specific different levels of delay-doppler spread due to different UE speeds and different carrier frequencies ($6 \leq f_c \leq 100$GHz). Therefore, a situation-aware, transmitter-receiver pair specific customized modulation technique is necessary to effectively improve the communication performance under such device-dependent delay-doppler spread conditions.
2.3 Spatial domain non-orthogonality and existing solutions

Utilization of higher carrier frequencies in mmWave spectrum has significantly reduced the antenna size, thus now allowing multiple antennas to be packed within a small massive MIMO array as shown in Fig. 2.6.

Following the introduction of such massive MIMO technology, spatial modulation (SM) technique has been acknowledged as a new modulation technique to achieve higher data rate, and higher spectral efficiency. The main premise behind spatial modulation is to utilize those multiple antennas in a MIMO array as an additional radio resource domain to carry additional QAM information symbols. For this purpose, the transmission bits are first divided into two parallel bit streams of length $\log_2(N_t)$ and $\log_2(M)$ as shown in Fig. 2.7. The first bit stream of length $\log_2(N_t)$ is carried through spatial modulation, whereas the latter bit stream of length $\log_2(M)$ is carried through conventional time-frequency or delay-doppler modulation [44].

The spatial modulation techniques proposed initially utilized a unique mapping between the MIMO antenna index and the SM bit pattern such that for each SM input bit pattern, a
corresponding MIMO antenna is activated for transmission [44] as presented in Table 2.1. At the receiver, the index of active transmit antenna, and the information symbol transmitted through the active transmit antenna is demodulated using ML techniques [44].

Table 2.1: Mapping of SM bit pattern, and transmit antenna for 4 × 4 MIMO transmitter

<table>
<thead>
<tr>
<th>SM input bits</th>
<th>Transmit antenna</th>
</tr>
</thead>
<tbody>
<tr>
<td>000</td>
<td>1</td>
</tr>
<tr>
<td>001</td>
<td>2</td>
</tr>
<tr>
<td>010</td>
<td>3</td>
</tr>
<tr>
<td>011</td>
<td>4</td>
</tr>
</tbody>
</table>

From Table 2.1, it is observable that only a single transmitter was activated for transmission in the index based SM schemes. Such utilization of a single antenna at a time for transmission provided robustness against inter channel interference, reduced complexity of transceiver as well as the power consumption. However, since only one antenna was utilized for transmission while the other antennas in the MIMO array were kept in idle state, the initially proposed SM techniques suffered from very low spectral efficiency, and no transmit diversity compared to spatially multiplexed communication systems [44]. Therefore, in order to achieve higher spectral efficiency, and higher throughput, more complex extended versions of SM techniques such as differential-SM (D-SM) [45], generalized-SM (G-SM) [46], quadrature-SM(Q-SM) [47] has been proposed. The table 2.2 presents the comparison on the different spatial modulation schemes on their spectral efficiency, and energy efficiency.

Table 2.2: Comparison on different SM schemes

<table>
<thead>
<tr>
<th>Scheme</th>
<th>No. of RF chains</th>
<th>Spectral efficiency</th>
<th>Energy efficiency</th>
</tr>
</thead>
<tbody>
<tr>
<td>SM</td>
<td>1</td>
<td>low</td>
<td>high</td>
</tr>
<tr>
<td>G-SM</td>
<td>≤ N_t</td>
<td>medium</td>
<td>high</td>
</tr>
<tr>
<td>D-SM</td>
<td>≤ N_t</td>
<td>medium</td>
<td>medium</td>
</tr>
<tr>
<td>Q-SM</td>
<td>1</td>
<td>low</td>
<td>high</td>
</tr>
</tbody>
</table>

Recently, the implementation of SM through phase rotation [48] technique has become popular due to it’s low complexity, higher spectral efficiency, and no reduction in spatial multiplexing gain. However, the dramatic reduction in spatial separation between antennas in
beyond 5G MIMO arrays has increased the severity of inter antenna correlation, and inter antenna crosstalk in spatial domain.

2.3.1 Inter antenna correlation

Simultaneous operation of multiple closely packed antennas in massive MIMO arrays has significantly increased spatial correlation among the antennas. As a result, inter antenna correlation (IAC) occurs at transmitter (BS) as well as receiver (user-device) which affects the phase of modulated symbols [49]. In general, the IAC depends on the device related attributes such as inter-antenna separation distance in arrays, angle of arrival, angle of transmission, and mutual coupling. A typical IAC condition among transmit and receive signals in a $3 \times 1$ linear MIMO array is shown in Fig 2.8.

![Figure 2.8: Inter antenna correlation in a $3 \times 1$ linear MIMO array](image)

Although IAC conditions were not significant in predecessor wireless generations with fewer transmitters, the IAC has become more severe and transmitter-receiver pair specific in massive MIMO antenna arrays. Therefore, it is essential to consider and mitigate the IAC at transmitter, and receiver. Several IAC reduction techniques such as two dimensional phase gradient based partially reflective surfaces [50], frequency selective surface [51], adaptive antenna selection [52], iterative inter-cell coordination to mitigate antenna correlation in multi-cell scenario [53], mutual scattering mode to reduce the IAC by increasing the Q factors of the MIMO antenna elements [54], and design of signature constellations for different transmit antennas [55] have been newly proposed in order to tackle the IAC.

However, the proposed solutions mainly rely on the hardware level design. Since the IAC among antennas in MIMO arrays vary between different transmitter-receiver pairs based on the spatial separation between the antennas, antenna operation model, angle of arrival, and angle of
transmission, a customized transmitter-receiver pair specific solution is essential to effectively mitigate the effect of IAC in different transmitter-receiver pairs.

2.3.2 Inter antenna crosstalk

Power amplifiers (PAs) are crucial components utilized at transmitters to amplify the baseband signal for transmission. To achieve higher power efficiency, and higher spectral efficiency, the PAs are required operate near saturation region. However, due to intrinsic nonlinear behavior exhibited by PAs driven at or close to saturation, nonlinear distortions with memory effects are introduced to the amplified output signal of the PA. Hence, spectral regrowth, and unwanted emissions occur around the carrier frequency which significantly deteriorate the output signal quality, and overall spectral efficiency [56].

Although these nonlinear distortions have been modeled, and suppressed through compensation techniques, the nonlinear distortions have become more severe in massive MIMO arrays due to significantly reduced spatial separation between antenna elements, higher carrier frequencies and ultra-wide bandwidth. As a result, nonlinear and reverse crosstalk from multiple PA branches are induced resulting in severe nonlinear distortions with memory effects in spatial domain [63]. Therefore, the PA’s output signal suffer from unwanted out-of-band emissions which affects the overall output signal quality, and spectral efficiency. The nonlinear crosstalk with coefficients $\alpha_{nl}, \beta_{nl}$ and linear crosstalk with coefficients $\alpha_l, \beta_l$ induced among transmit paths in a $2 \times 2$ MIMO array are illustrated in Fig. 2.9.

Figure 2.9: Spatial domain crosstalk in a $2 \times 2$ MIMO array [63]
To suppress such nonlinear distortions, and obtain a linearized output signal, digital predistorters (DPDs) are widely deployed at BS transmitters. To model, and compensate such nonlinear distortions in MIMO systems, many DPD models have been proposed. These research efforts can be roughly classified into two directions: a) Development of multi-input DPD models considering the crosstalk, and simplifying their complexity with suitable assumptions and b) Estimation, and mitigation of crosstalk followed by DPD for linearization.

Following the first direction, a crossover modified memory polynomial is proposed in [56] replacing higher order terms with canonical piecewise linear functions for a $2 \times 2$ MIMO transmitter while considering only the odd-order nonlinearities, and crosstalk strength upto $-15$ dB from neighbor antenna. The less complex memory polynomial DPD for a $2 \times 2$ MIMO transmitter in [57] also assumes nonlinear crosstalk strength upto $-20$ dB, and that even order nonlinearities have weak effects. In [58], a complexity reduced multi-input DPD model is proposed assuming that the nonlinearity generated due to crosstalk is lesser than by a factor of 2 compared to nonlinearity from the main PA branch. In [59], the impact of backward crosstalk of $-15$ dB, $-10$ dB is modeled and compensated. Although these works with suitable assumptions demonstrated reduction in complexity and better performance, the assumptions inherited in those methods have become no longer valid in scenarios where multiple “nearest neighbors” exist, such as in larger MIMO arrays or non-1D configurations. In addition, the DPD cost of above multi-input models increase exponentially with the increase in transmitters.

An alternate approach to MIMO DPD that has gained some momentum is to estimate and cancel the crosstalk prior to performing DPD. One example of this approach can be found in [60] where a crosstalk estimation block estimated, and mitigated the crosstalk followed by DPD linearization. Further in this direction, the works in [61, 62] aimed at cancelling the cross talk prior to DPD in order to reduce the overall complexity.

However, both approaches require additional signal feedback taps between the input signal and PA for each transmit path. While this is not a major issue in instrument-based laboratory equipment configurations, this approach becomes rather complex to implement when considering massive MIMO systems or non-1D array configurations. Therefore, a more feasible, less hardware complex, scalable DPD architecture is required to estimate and compensate the crosstalk in massive MIMO arrays.
2.4 New transmitter-receiver pair specific and domain specific challenges

Although current studies on non-orthogonality mitigation techniques proposed for the specific radio resource domains, and the nonlinear distortion compensation techniques for MIMO were proved to perform well, the utilization of dramatically smaller radio resource blocks made up of radio resources from different domains to achieve higher transmission rates in 5G encounter transmitter-receiver pair specific, and domain specific non-orthogonality degrees which vary differently in different domains. In such conditions, the existing solutions which mainly rely on receiver side synchronization to mitigate the non-orthogonality among radio resource blocks become less effective due to different synchronization capability of receiver in different domains.

Furthermore, the current multi-dimensional modulation techniques which utilize combinations of radio resources in spatial, time-frequency, and delay-doppler domains [19, 21, 22, 37, 44, 45, 46, 47, 48, 76] neither take into account the transmitter-receiver pair specific non-orthogonality degrees in different domains, nor aim to control the operational cost of user-device for orthogonality restoration and demodulation. Although such operational cost is not a major concern at the BS, it needs to be controlled at the receiver side (user-device) so that the resource constrained user-device can be effectively utilized. Therefore, in order to efficiently achieve higher communication performance under transmitter-receiver pair specific non-orthogonality conditions in spatial, time-frequency, and delay-doppler domains, more customized situation-aware, transmitter-receiver pair specific multi-dimensional modulation techniques are required at the BS.

Furthermore, current multi-input DPD models incur exponential increase in complexity and operational cost due to their proliferating number of DPD coefficients with the increase in MIMO array size. Moreover, they require signal feedback taps before the PA which has become infeasible to be implemented in small form-factor massive MIMO arrays. Therefore, less complex, scalable nonlinear distortion compensation architectures are paramount to enhance communication performance in future massive MIMO system.
2.5 Chapter summary

In this chapter, an overview of principles behind modulation in time-frequency, delay-doppler, and spatial domains, the non-orthogonality conditions that occur in those domains, and the existing solutions to mitigate such severe non-orthogonality conditions as well as nonlinear distortions were analyzed. The drawbacks of those existing solutions to efficiently achieve higher communication performance under transmitter-receiver pair specific non-orthogonality in those domains were outlined. Finally, the requirement for customized, transmitter-receiver pair specific multi-dimensional modulation techniques to efficiently achieve higher communication performance under different levels of non-orthogonality in different domains, and the need for less complex, scalable DPD architecture to compensate spatial domain nonlinear distortions induced among antennas in massive MIMO arrays were highlighted. Such required situation-aware, cost-aware multi-dimensional modulation techniques and nonlinear distortion compensation architectures will be designed in the remaining chapters of this thesis.
Chapter 3

Multi-Dimensional Modulation for Non-Orthogonal Spatial-Time-Frequency Domains

Due to higher carrier frequencies (mmWave bands), closely packed antennas in MIMO arrays, and complex time-delay multi-path signals in beyond 5G wireless networks, the severity of non-orthogonal interference (inter carrier interference (ICI) in frequency domain, inter symbol interference (ISI) in time domain, and inter antenna correlation (IAC) in spatial domain) has significantly increased. As a result, current OFDM systems which inherently rely on preservation of orthogonality between radio resources encounter severe data rate degradation under such non-orthogonal interference from multiple domains. Therefore, a novel, customized, situation-aware multi-dimensional modulation scheme (MDM) is proposed in this chapter with the goal of achieving maximized data rate under joint non-orthogonality conditions in spatial-time-frequency domains. The proposed MDM scheme is situation-aware such that it takes into account the non-orthogonality degrees among spatial, time-frequency domains radio resources, and jointly optimizes the radio resource attributes to minimize the non-orthogonality degrees in those domains, and thus achieve maximized data rate. Simulation results demonstrate that the proposed MDM scheme achieves higher data rate than current state-of-art spatially multiplexed MIMO-OFDM systems. Further result show that the proposed MDM can achieve higher data rate gain in massive MIMO systems.
3.1 Introduction

Although orthogonal radio resource blocks from different domains are utilized for modulation in 5G and beyond, due to stochastic nature of wireless channels, imperfect synchronization of receiver in different domains, and transmitter-receiver dependent hardware discrepancies, the orthogonality among those radio resource blocks is often destroyed inducing non-orthogonal interference such as inter carrier interference (ICI) in frequency domain, inter symbol interference (ISI) in time domain, and inter antenna correlation (IAC) in spatial domain. Current OFDM based LTE and early 5G systems assume that the orthogonality can be completely recovered at the receiver through time-frequency domain synchronization, and the residual non-orthogonal interference is generally tolerated at the receiver.

However, due to increased CFO at higher carrier frequencies (mmWave bands), more complex multi-path signals, and spatial correlation among closely packed antennas in MIMO arrays in next generation networks, the non-orthogonal interference from those domains have become more severe. Although such non-orthogonal interference vary differently in different domains, they combineldy affect the overall achievable data rate [9, 10, 31, 55]. Furthermore, such non-orthogonality conditions have become transmitter-receiver pair specific due to diverse synchronization capability of receiver in different domains [12].

Based on these observations, the contributions of this chapter to achieve maximized communication data rate under transmitter-receiver pair specific non-orthogonality conditions in spatial-time-frequency domains are summarized as follows.

- The transmitter-receiver pair specific non-orthogonality conditions induced among spatial, and time-frequency domain radio resources are modeled. It is shown that such joint non-orthogonal interference from those domains combinedly deteriorate the achievable data rate. Therefore, to maximize the data rate under such joint non-orthogonal interference, a new multi-dimensional modulation (MDM) technique which can utilize spatial domain radio resources simultaneously with time-frequency domain radio resources through multi-dimensional modulation is proposed.

- The proposed MDM is designed to be situation-aware such that it can take into account the transmitter-receiver pair specific non-orthogonality degrees in spatial-time-frequency
domains, and jointly optimize the modulation attributes of the resource block to minimize the non-orthogonality degrees in each domain and thus achieve maximized data rate.

The rest of this chapter is organized as follows: In section 3.2, the system model, the architecture of proposed MDM are presented, and the transmitter-receiver pair specific non-orthogonality in spatial, time-frequency domains are modeled. In section 3.3, the research problem is formulated and the solution is presented in section 3.4. In section 3.5, performance of proposed MDM scheme is analyzed using simulation results and the chapter is summarized in section 3.6.

### 3.2 System model

The specific goal of the proposed MDM technique is to achieve maximized data rate between a BS with $N_t$ MIMO antennas serving a single user under transmitter-receiver pair specific non-orthogonality degrees in spatial, time-frequency ($\rho_s, \rho_t, \rho_f$) domains as shown in Fig. 3.1.

![System model diagram](image-url)

**Figure 3.1: System model**

It is assumed that the channel state information to evaluate non-orthogonality degrees are available at BS using channel reciprocity [65]. Hence, the overall goal of this chapter to achieve
maximized communication data rate under non-orthogonality conditions in spatial, and time-frequency domains is expressed as,

\[
\max(R(\rho_s, \rho_t, \rho_f)),
\]

(3.1)

where \( R \) is the achievable data rate, \( \rho_s, \rho_t, \rho_f \) are the spatial, time, frequency domains non-orthogonality degrees detailed later in this section. The architecture of proposed MDM scheme to achieve the goal is presented in Fig. 3.2. In the proposed architecture, the spatial modulation block is jointly integrated with time-frequency domain modulation blocks to simultaneously utilize spatial domain radio resources with time-frequency domain radio resources for multi-dimensional modulation. The spatial-time-frequency modulation steps in MDM are as follows:

1. At first, transmit antennas \( N_t \) are divided into a \( N_s \) number of antenna subsets.

2. Then, QAM symbols are spatially modulated (SM) by phase rotation technique \cite{48} with phase rotation angle \( \theta_i = i \cdot \pi / N_s \) determined based on the input bit.

3. Finally, SM symbols are modulated into time-frequency domain frame using IFFT.

Hence, spatial-time-frequency modulated transmission signal of MDM through the channel is expressed as,

\[
x(t) = \sum_{n=0}^{N-1} H(k)x(n) \cdot e^{j2\pi nk/N} \cdot g(t - nT),
\]

(3.2)

where \( x(t) \) is the transmission signal, \( N \) is number of subcarriers, \( n \in \{0, 1, 2, \cdots, N - 1\} \) is the subcarrier index, \( x(n) \) is spatially modulated symbols, and \( e^{j2\pi n/} \) represents IFFT operation, \( H(k) \) is the channel matrix and \( g_k(t - nT) \) is time-frequency domain windowing function.
3.2.1 Spatial-time-frequency domain non-orthogonality analysis

In achieving the goal of MDM in (3.1), transmitter-receiver pair specific non-orthogonality conditions in spatial- time-frequency domains are determined in this subsection.

Due to reduced spatial separation between antennas in a MIMO antenna array, the inter antenna correlation (IAC) occurs among the antennas which affects the phase of modulated symbol. The IAC matrix between the BS and UE depending on the mean angle of arrival, angle spread of multiple incoming signals, and the inter antenna distance is expressed as,

$$ R_{T_x} = \begin{bmatrix} 1 & r_{12} & \ldots & r_{1N_t} \\ r_{21} & 1 & \ldots & r_{2N_t} \\ \vdots & \vdots & \ddots & \vdots \\ r_{N_t1} & \ldots & \ldots & 1 \end{bmatrix}, \quad (3.3) $$

where \( r_{i,j} = \alpha e^{j\beta} \) [66], \( 0 \leq \alpha \leq 1, 0 \leq \beta \leq 2\pi \) and \( \alpha, \beta \) are uniformly distributed i.i.d variables [67]. Hence the spatial domain non-orthogonality (IAC matrix) is incorporated into the channel matrix of the BS transmitter and UE receiver using Kronecker model as,

$$ H_c = H R_{T_x}^{1/2}, \quad (3.4) $$

where \( H \) is the Rayleigh channel matrix, of which each entry is independently and identically distributed (i.i.d.) complex Gaussian random variables with \( H(i, j) \sim CN(0, 1) \). Hence using (3.4), the spatial domain interference due to IAC is modeled as,

$$ P_{IAC}(\theta) = P_s \sigma_s^2, \quad (3.5) $$

where \( P_s \) is the signal power, \( \sigma_s^2 = E(||H - H_c||_F^2) \), \( E(.) \) is the statistical expectation, \( ||.||_F \) is the Frobenius norm, \( H \) is the channel matrix, and \( H_c \) is the spatially correlated channel matrix.

On the other hand, the ICI due to CFO among subcarriers can be expressed as a function of modulation attributes of frequency domain as,

$$ P_{ICI}(\Delta f) = \sum_{\substack{k=0 \\ k\neq m}}^{N-1} \left| P_s(f_a + \frac{\hat{\xi}}{\Delta f}) \right|^2, \quad (3.6) $$
where \( \hat{\xi} = f_{a-m} \cdot \Delta f \) is the normalized frequency offset, \( f_{a-m} \) is the carrier frequency offset, \( \Delta f \) is the subcarrier spacing, \( P_s \) is the signal power. Furthermore, the ISI due to multi-path signal propagation is expressed in terms of time domain modulation parameter \( T_s \) as,

\[
P_{\text{ISI}}(T_s) = R_m(0) \cdot \left| \frac{\hat{\delta}}{T_s} \right|^2,
\]

where \( R_m(0) \) is the auto-correlation of multi-path signal, and \( \hat{\delta} = \alpha_m \cdot T_s \) is the time-offset.

### 3.2.2 Analysis of overall SINR and achievable data rate

Although the transmitter-receiver pair specific non-orthogonal interference vary in spatial, time-frequency domains, they combinedly impact the overall SINR of wireless channel. Therefore, incorporating the non-orthogonal interference (3.5), (3.6), and (3.7) from spatial, frequency, and time domains, the overall SINR of the wireless channel between BS and UE is expressed as,

\[
\Upsilon(\Delta f, T_s, \theta) = \frac{P_{\text{desired}}}{P_{\text{IAC}}(\theta) + P_{\text{ISI}}(T_s) + P_{\text{ICI}}(\Delta f) + P_N},
\]

where \( P_{\text{desired}} \) is the desired signal’s power, \( P_{\text{IAC}}, P_{\text{ISI}}, P_{\text{ICI}} \) are the non-orthogonal interference power in spatial, time-frequency domains, and \( P_N \) is the AWGN power. Considering such non-orthogonal interference in the SINR of wireless channel between the transmitter-receiver pair, the achievable data rate of the proposed MDM scheme is expressed in (3.9).

\[
R = \sum_{k=1}^{k=N_f} B_k \left( \log_2 \det \left( I + \frac{\Upsilon_k(\Delta f, T_s, \theta)}{N_t} \cdot H_c \cdot H_c^+ \right) \right) + \log_2(N_s \cdot N_b)
\]

where \( B_k \) is subcarrier bandwidth, \( \Upsilon_k \) is SINR of each subcarrier, \( \Delta f \) is subcarrier spacing, \( T_s \) is symbol duration, \( N_t \) is the number of transmit antennas, \( H_c \) is IAC channel matrix from (3.4), \( N_s \) is number of antenna subsets, \( N_b = \log_2(N_s) \) is number of bits per antenna subset.

Hence from (3.8) and (3.9) it is evident that although different levels of non-orthogonal interference occur in different domains, they combinedly affect the overall SINR of the wireless channel (3.8), and thus consequently affect the achievable data rate (3.9).
3.3 Problem formulation

In this section, the overall research problem of achieving maximized data rate under spatial, and time-frequency domain non-orthogonality condition is formulated. The radio resource attributes from each domain such as subcarrier spacing $\Delta f$ in frequency domain, symbol duration $T_s$ in time domain, and phase separation $\theta$ in spatial domain are utilized as the optimization parameters to achieve the goal under transmitter-receiver pair specific non-orthogonality degrees in those domains. Hence, the overall research problem is formulated as follows:

**Assumption 1**: Number of antennas allocated $N_t$ is fixed.

**Assumption 2**: Allocated bandwidth $B$ is fixed.

**Assumption 3**: Transmit signal power allocated is fixed.

\[
\max_{\{\Delta f, T_s, \theta\}} \left( R(\Delta f, T_s, \theta) \right),
\]

s.t

- **C1**: $\Delta f = 15(n) \times 10^3, n \in \{1, 2, \ldots, 32\}$
- **C2**: $T_s = \frac{1}{\Delta f}$
- **C3**: $1 \leq N_s \leq N_t$
- **C4**: $\log_2(N_s) \in \mathbb{Z}$

where $\Delta f$ is the subcarrier spacing, $T_s$ is the symbol duration, $N_s$ is the number of antenna subsets, $N_t$ is the number of transmit antennas.

**C1** states that the subcarrier spacing $\Delta f$ is a multiple of 15 kHz up to 480 kHz to ensure compatibility with 3GPP standard for 5G and beyond communications [68].

**C2** states that the symbol duration $T_s$ should be the reciprocal of subcarrier spacing $\Delta f$ to maintain orthogonality between time-frequency domains.

**C3** states that number of antenna subsets $N_s$ is between 1 and total number of antennas $N_t$.

**C4** states that $\log_2(N_s)$ should be a positive integer to ensure one-to-one unique mapping between spatial domain symbols and antenna subsets.
### 3.4 Proposed multi-dimensional modulation technique

In achieving the goal of data rate maximization under spatial, and time-frequency domain non-orthogonality conditions, the following **Algorithm 1** is proposed to solve the optimization problem \(3.10\). The **Algorithm 1** is designed to jointly optimize the radio resource block attributes such as phase separation angle \(\theta\) in spatial domain, subcarrier spacing \(\Delta f\) in frequency domain, and symbol duration \(T_s\) in time domain with the objective of minimizing the total non-orthogonal interference, and thus maximize the achievable data rate.

**Algorithm 1:** Data rate maximization under spatial-time-frequency domain non-orthogonality

1. **Input:** estimated normalized non-orthogonality degree \(\hat{\xi}\) in frequency domain, estimated normalized non-orthogonality degree \(\hat{\delta}\) in time domain, estimated IAC matrix \(R_{Tx}\), number of transmit antennas \(N_t\), Bandwidth \(B\), Total transmission time \(T\)
2. initialization \(n, P_{\text{int}} = [], i, N_s\)
3. for \(i = 1\) to \(6\) do
   4. \(n \in \{1,2,4,8,16,32\}\)
   5. Set the subcarrier spacing \(\Delta f = 15n[i] \times 10^3\)
   6. Set \(T_s = 1/\Delta f\)
   7. Evaluate \(P_{\text{ICI}} + P_{\text{ISI}}\) using \(\hat{\xi}, \hat{\delta}, \Delta f,\) and \(T_s\)
   8. Set \(P_{\text{sum}} = P_{\text{ICI}} + P_{\text{ISI}}\)
   9. Update \(P_{\text{int}} = [P_{\text{int}} P_{\text{sum}}]\)
10. end
11. Evaluate \(\Delta f = n(\arg\min(P_{\text{int}})) \cdot 15 \times 10^3\)
12. Evaluate \(T_s = 1/\Delta f\)
13. Set \(N_f = B/\Delta f\)
14. Set \(N_t = T/T_s\)
15. for \(i = \log_2(N_t)\) to \(1\) do
   16. Evaluate \(P_{\text{IAC}}\) using \(R_{Tx}\)
   17. Update \(\theta_i = \arg\min(P_{\text{IAC}})\)
18. end
19. Set \(N_s = 2^i\)
20. **Output:** optimum no. of subcarriers \(N_f\), optimum no. of time slots \(N_t\), and optimum no. of antenna subsets \(N_s\)
In the first loop, the MDM jointly takes into account the ICI and ISI non-orthogonal interference and finds the optimum subcarrier spacing $\Delta f$, symbol duration $T_s$ such that the overall ICI and ISI is minimum. Here, it is important to note that joint optimization of the time-frequency domain modulation attributes $\Delta f, T_s$ is necessary. To elaborate further, in order to maintain the orthogonality between time domain, and frequency domain, the subcarrier spacing, and symbol duration should be reciprocal to each other ($\Delta f = 1/T_s$). Therefore, when $\Delta f$ increases, $P_{ICI}$ will reduce, however on the other hand the $P_{ISI}$ will increase since the $T_s$ decreases when $\Delta f$ increases. Similarly, when $T_s$ decreases, $P_{ISI}$ will decrease, and $P_{ICI}$ will increase. So, it is necessary to jointly take into account the ISI and ICI, and optimize the time-frequency domain modulation attributes $\Delta f, T_s$ simultaneously such that the total non-orthogonal interference ($P_{ICI} + P_{ISI}$) is minimized to achieve maximized data rate in time-frequency domains.

In the second loop, for the evaluated IAC condition, the optimum phase separation $\theta_i$ is found such that the IAC is minimum. Finally, the optimum number of subcarriers $N_f$, optimum number of time slots $N_t$, and optimum number of antenna subsets $N_s$ are output which can guarantee maximized data rate under the non-orthogonality conditions in spatial-time-frequency domains.

Hence, the proposed MDM scheme is designed to simultaneously optimize the modulation attributes of the spatial-time-frequency domain radio resource block such that joint non-orthogonal interference in spatial, and time-frequency domains is minimized. Therefore, by minimizing the total non-orthogonal interference power, the proposed MDM scheme maximizes the overall SINR of the wireless channel between the transmitter-receiver pair, and thus achieves the maximized communication data rate as expressed in (3.11).

\[
\min (P_{ICI} + P_{ISI} + P_{IAC})_{|\Delta f, T_s, \theta_i} \Rightarrow \max (\mathcal{Y}(\Delta f, T_s, \theta_i)) \Rightarrow \max (R) \quad (3.11)
\]

It is important to note that since the non-orthogonality conditions vary between different transmitter-receiver pairs, it is paramount to evaluate the non-orthogonality degrees in each domain for the specific transmitter-receiver pair, and jointly optimize the radio resource block attributes accordingly to achieve the maximized communication data rate.
3.5 Simulation results and analysis

A range of simulation results are presented in this section to demonstrate the performance of the proposed MDM scheme compared with state-of-art MIMO-OFDM systems. It was assumed that the UE is allocated with fixed bandwidth $B$, transmission time $T$, and the number of transmission antennas $N_t = 64$ to resemble a massive MIMO communication scenario. It was also assumed that both BS and UE have the flexibility to operate on a range of subcarrier spacing $\in \{15, 30, 60, 120, 240, 480\}$ kHz. Furthermore, based on the 3GPP recommendations for beyond 5G communications the channel models CDL-A, TDL-A, TDL-D were utilized in the simulation [68]. The simulation was implemented in MATLAB using signal processing and 5G toolbox.

Table 3.1: Simulation parameters

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bandwidth</td>
<td>100 MHz</td>
</tr>
<tr>
<td>Carrier frequency</td>
<td>30 GHz</td>
</tr>
<tr>
<td>Number of antennas</td>
<td>64</td>
</tr>
<tr>
<td>Subcarrier spacing</td>
<td>15, 30, 60, 120, 240, 480 kHz</td>
</tr>
<tr>
<td>Channel model</td>
<td>CDL-A, TDL-A, TDL-D</td>
</tr>
</tbody>
</table>

The CDF plot for data rate achieved by the proposed MDM scheme, the optimized time-frequency modulation scheme, and the state-of-art spatially multiplexed MIMO-OFDM scheme is illustrated in Fig. 3.3. As shown, the optimized time-frequency modulation scheme achieves higher data rate than MIMO-OFDM through joint optimization of time-frequency domain modulation attributes to minimize the non-orthogonal interference in those domains while achieving multiplexed data rate gain in spatial domain. In contrast, MDM scheme utilizes spatial domain as an additional radio resource domain for modulation with time-frequency domain and also jointly optimizes spatial-time-frequency domain modulation attributes ($\theta, T_s, \Delta f$) to minimize the overall non-orthogonal interference in those domains, and thus achieves maximized data rate.

Furthermore, the maximum data rate achieved by MIMO-OFDM, optimized time-frequency modulation, and the proposed MDM for different number of MIMO antennas (8, 16, 32, 64, 128) are plotted in Fig. 3.4. From that simulation result, it can be noticed that
the data rate gain achieved by proposed MDM is significantly higher when \( N_t \geq 32 \). In detail, the MIMO-OFDM, and optimized time-frequency modulation scheme benefits from spatial multiplexing gain, whereas the proposed MDM utilizes the spatial domain as an additional domain for modulation. Therefore, with increase in MIMO transmitters, the MDM utilizes more number of antenna subsets for spatial modulation, and achieve higher data rate gain than MIMO-OFDM and time-frequency modulation schemes. Hence, it is corroborated that the proposed MDM can achieve higher data rate gain in massive MIMO systems.

![CDF of data rate achieved for 64 transmit antennas](image1)

**Figure 3.3: CDF of data rate achieved for 64 transmit antennas**

![Maximum data rate achieved for different no. of transmit antennas](image2)

**Figure 3.4: Maximum data rate achieved for different no. of transmit antennas**
3.6 Chapter summary

In this chapter, the deterioration of achievable data rate due to dynamically varying transmitter-receiver pair specific joint non-orthogonal interference in spatial, time-frequency domains is focused. To achieve maximized communication data rate under such non-orthogonality conditions in those domains, a situation-aware transmitter-receiver pair specific multi-dimensional modulation (MDM) technique was proposed. The proposed MDM scheme is designed to jointly take into account the different levels of non-orthogonal interference induced in spatial-time-frequency domains and jointly optimize the modulation attributes of spatial-time-frequency domain radio resource blocks to minimize the non-orthogonal interference in those domains. Hence, by minimizing the total non-orthogonal interference power, the proposed MDM maximizes the SINR of the wireless channel between the transmitter-receiver pair, and thus achieves maximized communication data rate. Through simulation results, the performance of the proposed MDM scheme was validated and compared with state-of-art spatially multiplexed MIMO-OFDM and optimized time-frequency modulation schemes. Through further results, it was demonstrated that the proposed MDM scheme is highly advantageous for future massive MIMO communication systems with transmit antennas $\geq 32$. 
Chapter 4

User-Centric Multi-Dimensional Modulation for Non-Orthogonal Domains in Beyond 5G and 6G

Following the previous chapter on data rate maximization in non-orthogonal spatial-time-frequency domains, this chapter mainly focuses on user-centric challenges in beyond 5G: a) Non-orthogonality degrees among radio resource blocks have become transmitter-receiver pair specific and domain specific depending on device-dependent attributes in different domains, b) Overall operational cost of receiver for orthogonality restoration and demodulation is also domain specific. Therefore in this chapter, a user-centric multi-dimensional modulation (UC-MDM) technique is proposed to minimize the overall operational cost from receiver while achieving required data rate under transmitter-receiver specific non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains. The proposed situation-aware and cost-aware UC-MDM is designed to jointly optimize the separation of radio resource blocks in those domains and utilize the optimum radio resource combination through MDM in either spatial-time-frequency or spatial-delay-doppler domains to achieve the goal under different non-orthogonality degrees in those domains. Simulation results under simultaneously varied non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains demonstrate that UC-MDM can achieve required data rate with less operational cost from UE compared to state-of-art MIMO-OFDM and MIMO-OTFS systems.
4.1 Introduction

To support the proliferating demand for higher data rates in beyond 5G communication systems, orthogonal radio resource blocks from different domains with dramatically reduced size are utilized for modulation. As a result, the radio resource blocks have become more susceptible towards non-orthogonality conditions in multiple domains leading to non-orthogonal radio resource blocks a common paradigm for modulation.

In addition, such non-orthogonality degrees among radio resources have become specific for a transmitter-receiver pair due to strong influence of device-dependent communication attributes such as different UE speed, different carrier frequency, different reception/synchronization performance of the receiver in different domains, and different levels of antenna correlation at BS transmitter and UE receiver [12, 13]. Hence, the non-orthogonality degrees in the wireless channel perceived by the receiver have become device-dependent and domain specific. Although such non-orthogonality degrees vary differently in different domains, they combinedly affect the achievable communication performance, and the worst domain limits the overall communication performance [11].

Furthermore, the operational cost required from user-device to restore orthogonality of such non-orthogonal radio resources and for demodulation in corresponding domains has also become domain specific [14]. This operational cost can be in the form of power consumption when receiver is implemented in hardware, or computational complexity when the receiver is implemented in software. Hence, it has also become imperative to consider and control such operational cost so that the resource-constrained user-device can be effectively utilized. As highlighted in Chapter 2, current MIMO-OFDM and MIMO-OTFS systems are not situation-aware and cost-aware to effectively achieve required communication performance under these non-orthogonality conditions in multiple domains.

Motivated by these new, transmitter-receiver pair specific, and user-centric challenges in beyond 5G communication systems, the contributions of this chapter to minimize the overall operational cost from user-device while achieving required communication datarate under transmitter-receiver pair specific non-orthogonality in spatial, time-frequency and delay-doppler domains are summarized as follows.
4.1. Introduction

- A new user-centric multi-dimensional modulation (UC-MDM) is proposed. Unlike existing OFDM systems which inevitably depend on radio resource orthogonality for higher communication performance, the proposed UC-MDM is situation-aware such that it is designed to take into account the transmitter-receiver pair specific non-orthogonality degrees among radio resource blocks in spatial, time-frequency, and delay-doppler domains and effectively utilize the optimum combination of radio resources through multi-dimensional modulation.

- The operational cost at receiver side to restore the orthogonality of non-orthogonal radio resources vary in different domains depending on the non-orthogonality degree perceived by receiver in those domains. The current OFDM and OTFS systems do not aim to control the user-device specific operational cost. In contrast, proposed UC-MDM is designed to take into account such user-device specific and domain specific orthogonality restoration cost and jointly optimize the radio resource block separation attributes to minimize the non-orthogonality degree in each domain, and thus reduce the receiver side orthogonality restoration operational cost.

- Furthermore, the proposed UC-MDM is also designed to take into account the domain specific operational cost at receiver for demodulation in spatial, time-frequency, and delay-doppler domains. Therefore, the cost-aware UC-MDM is designed to utilize the optimum combination of radio resources with optimum separation through multi-dimensional modulation in either spatial-time-frequency domains or spatial-delay-doppler domains to achieve the required communication data rate with minimal overall operational cost (orthogonality restoration and demodulation cost) from the user-device.

The rest of this chapter is organized as follows: In section 4.2, architecture of proposed UC-MDM technique is presented, and then the domain specific non-orthogonality degrees and receiver side overall operational cost are evaluated. In section 4.3, the overall research problem is formulated. In section 4.4, the proposed UC-MDM technique is presented. Simulation results are presented in section 4.5, and this chapter is concluded with future research directions in section 4.6.
4.2 System model

The overall goal of the proposed UC-MDM is to achieve required communication data rate with minimal operational cost from user-device under transmitter-receiver pair specific non-orthogonality conditions in spatial, time, frequency, delay, and doppler domains.

The system model with a BS serving a single user-device over a wireless channel with transmitter-receiver pair specific non-orthogonality degrees $\rho_s, \rho_t, \rho_f, \rho_{de}, \rho_{do}$ in spatial, time-frequency, and delay-doppler domains is considered as shown in Fig. 4.1. It is assumed that the channel state information to evaluate transmitter-receiver pair specific non-orthogonality degrees are available at BS using channel reciprocity [65]. The radio resource blocks in either spatial-time-frequency or spatial-delay-doppler domains will be effectively utilized with optimum separation to achieve the goal.

Hence, the overall goal with primary objective of achieving required data rate and secondary objective of minimizing the receiver operational cost is expressed as,

$$R \geq R', \quad s.t \quad C_1: \min (C)$$

(4.1)
where $R$ is the achievable data rate, $R'$ is required the data rate, and $C$ is the overall receiver side operational cost for orthogonality restoration and demodulation in the respective domains. The achievable data rate $R$, and the operational cost $C$ will be detailed later in this section.

In achieving the goal (4.1), it is aimed to flexibly utilize spatial, time-frequency, and delay-doppler domain radio resources jointly through multi-dimensional modulation. Therefore, the architecture of proposed UC-MDM is presented in Fig. 4.2. As shown, the spatial modulation block is jointly integrated with switched delay-doppler domain modulation block, and time-frequency domain modulation block so that the UC-MDM can switch and perform multi-dimensional modulation in either spatial-time-frequency or spatial-delay-doppler domains to flexibly utilize the optimum radio resource blocks based on the evaluated non-orthogonality degrees and receiver operational cost in those domains.

![Figure 4.2: Proposed UC-MDM architecture.](image)

To perform spatial-time-frequency modulation in the proposed UC-MDM, the available transmit antennas $N_t$ are divided into a number of antenna subsets $N_s$ and then the QAM symbols are spatially modulated using phase rotation technique [48]. Hence, the unique spatially modulated symbols are obtained for transmission as detailed in the previous chapter. Furthermore, the total available bandwidth $B$ is divided into $N$ narrow-band orthogonal subcarriers with equidistant frequency separation $\Delta f$ for modulation in frequency domain using IFFT technique [70]. For spatial-delay-doppler domain modulation in UC-MDM, additional inverse symplectic Fourier transform (ISFFT) operations are implemented to modulate SM symbols into delay-doppler domain transmission frame.
Since the proposed UC-MDM is designed to operate in either spatial-time-frequency domain modulation mode or spatial-delay-doppler domain modulation mode, the transmitted signal of the proposed UC-MDM is jointly expressed as,

\[ x[k, l] = \frac{1}{\sqrt{\alpha N + (1 - \alpha)NM}} \sum_{n=0}^{N-1} \sum_{m=0}^{M-1} x[n, m] e^{j2\pi \left( \frac{nk}{N} - \frac{(1 - \alpha)ml}{M} \right)}, \]

where \( x[k, l] \) is the UC-MDM transmission signal, \( \alpha = 1 \) in spatial-time-frequency domain modulation mode, \( \alpha = 0 \) in spatial-delay-doppler domain modulation mode, \( x[n, m] \) is the spatially modulated symbol, \( e^{j2\pi(x-y)} \) represents ISFFT operation, \( e^{j2\pi(x)} \) represents IFFT operation, \( N \) is number of subcarriers when \( \alpha = 1 \), and \( n, m, N, M \) are the delay index, doppler index, number of delay bins and number of doppler bins when \( \alpha = 0 \).

The proposed UC-MDM is designed to take into account the transmitter-receiver pair specific non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains and the receiver operational cost for orthogonality restoration as well as for subsequent demodulation in those domains. Based on such non-orthogonality degrees and the evaluated overall operational cost, the UC-MDM is designed to intelligently utilize the optimum combination of radio resource blocks for modulation in either spatial-time-frequency or spatial-delay-doppler domains with optimum radio resource separation \((\theta, T_s, \Delta f)\) or \((\theta, \Delta \tau, \Delta v)\) from each domain to achieve the goal.

Therefore, considering such non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains, the overall goal in (4.1) is extended as,

\[ R(\rho, \rho, \rho, \rho, \rho) \geq R', \]

\[ s.t \]

\[ C1: \min \ (C) \]

where \( \rho, \rho, \rho, \rho, \rho \) are the non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains respectively.
4.2. System model

4.2.1 Domain specific non-orthogonality and operational cost analysis

To achieve the system specific goal in (4.3), transmitter-receiver pair specific and domain specific non-orthogonality degrees and the corresponding operational cost of receiver are evaluated in this subsection.

Due to varying device dependent communication attributes such as different UE speed, different carrier frequency, multi-path signal propagation and synchronization mismatch, the receiver experiences loss of orthogonality among time-frequency domain radio resources. Such non-orthogonality degree in time-frequency domain is expressed in terms of time-frequency domain radio resource separation parameters as,

\[ \rho_f = \frac{v_i}{\Delta f} = \frac{v \cdot f_c}{c}, \]

\[ \rho_t = \frac{\tau_i}{T_s}, \]  

(4.4)

(4.5)

where \( \rho_f, \rho_t \) are the frequency, time domain non-orthogonality degrees, \( v_i \) is frequency offset indice, \( v \) is the UE speed, \( f_c \) is carrier frequency, \( c \) is speed of light, \( \tau_i \) is time delay indice, and \( T_s \) is the symbol duration. From (4.4), and (4.5) it is apparent that the non-orthogonality degree perceived by UE in time-frequency domain depends on corresponding radio resource parameters such as subcarrier spacing \( \Delta f \), and symbol duration \( T_s \) in those domains.

Therefore, considering the no. of time-frequency domain window interval sampling shift operations required to restore the orthogonality radio resources in time-frequency domains [72, 73], and the following demodulation computational operations [74] in time-frequency domain, the overall operational cost (no. of operations) required at receiver for orthogonality restoration and demodulation of a non-orthogonal time-frequency domain resource frame can be expressed as,

\[ C_{\text{time-frequency}} = \sum_{i=1}^{N} \rho_f, N_i + \sum_{j=1}^{M} \rho_t, M_j + N \log_2 N, \]  

(4.6)

where \( C_{\text{time-frequency}} \) is the receiver operational cost in time-frequency domain, \( \rho_f \) is the frequency domain non-orthogonality degree of \( N_i \)-th non-orthogonal subcarrier in (4.4), and \( \rho_t \) is the time domain non-orthogonality degree of \( M_j \)-th time domain non-orthogonal symbol in (4.5).
On the other hand, spatial correlation occurs between antennas at transmitter as well as receiver due to reduced spatial separation in MIMO arrays. As a result, phase noise is induced which affects the phase of spatially modulated symbols [49]. Such spatial correlation have become more specific towards the transmitter-receiver pair depending on the antenna reception capability, angle of receive signal angle, and angle of transmit signal at transmitter and receiver. Hence, the inter antenna correlation matrix at the BS transmitter as well as the user-device receiver is expressed as,

\[
R_{T/R_x} = \begin{bmatrix}
1 & r_{12} & \ldots & r_{1N_t}
\vline &
&
\hline
r_{21} & 1 & \ldots & r_{2N_t}
\vline &
&
\hline
\vdots & \vdots & \vdots & \vdots \\
\hline
r_{N_t1} & \ldots & \ldots & 1
\end{bmatrix},
\]

(4.7)

with IAC coefficient \(r_{i,j} = \alpha e^{j\beta}\) [71] where, \(0 \leq \alpha \leq 1, 0 \leq \beta \leq 2\pi\) and \(\alpha, \beta\) are uniformly distributed i.i.d variables. It is important to note that the elements of IAC matrix vary between different transmitter-receiver pairs depending on the device related attributes such as inter-antenna separation distance in arrays, angle of arrival/transmission, and mutual coupling.

Using Kronecker model, the IAC matrix of BS transmitter \(R_{T_x}\) and UE receiver \(R_{R_x}\) is incorporated in to the channel matrix of the transmitter-receiver pair as,

\[
H_c = R_{T_x}^{1/2} H R_{R_x}^{1/2},
\]

(4.8)

where \(H_c\) is the spatially correlated matrix, \(H\) is the Rayleigh channel matrix of which each entry is independently and identically distributed (i.i.d.) complex Gaussian random variables with \(H(i, j) \sim CN(0, 1)\), \(R_{T_x}\), \(R_{R_x}\) are the IAC matrix at transmitter and receiver. Hence using (4.8), the spatial domain non-orthogonality degree perceived by a UE due to the transmitter-receiver pair specific IAC is expressed as,

\[
\rho_s = \frac{\Delta\theta_i}{\theta_i},
\]

(4.9)

where \(\Delta\theta_i = \mu(H - H_c)\) is phase offset of \(i\)-th antenna subset and \(\theta_i\) is ideal phase separation.
Hence, considering IAC at transmitter and receiver, the received symbol vector $Y$ between the transmitter-receiver pair is expressed as,

$$Y = \sqrt{\frac{\Upsilon}{N_t}} H_c X + W,$$

where $Y$ is the received symbol vector, $\Upsilon$ is the signal to interference noise ratio (SINR), $N_t$ is the number of transmit antennas, $H_c$ is the spatially correlated channel matrix, $X$ is the transmitted symbol vector, and $W$ is the additive white Gaussian noise (AWGN) vector. From (4.10) it is evident that the received symbol vector is affected by the phase offset induced by spatially correlated matrix $H_c$. Therefore, such phase offset needs to be compensated before demodulation in spatial domain. Considering the number of phase shift operations required to restore the orthogonality of such non-orthogonal SM symbols, and the number of operations required for demodulation [75], the overall operational cost incurred by the receiver for orthogonality restoration and demodulation in spatial modulation is evaluated as,

$$C_{\text{spatial}} = \sum_{i=1}^{N_s} \rho_s N_s \log_2(N_s) + \log_2(N_s),$$

where $N_s$ is the number of antenna subsets, $\rho_s$ is the spatial domain non-orthogonality degree of $N_s$-th non-orthogonal antenna subset in (4.9).

Furthermore, the large delay spread and doppler spread that occurs between a transmitter-receiver pair considering varying UE speed upto 500 km/h, higher carrier frequencies upto 100 GHz [68], and imperfect synchronization in delay-doppler domains is expressed in terms of delay-doppler domain separation parameters as,

$$\rho_{do} = \frac{v_i B}{N} = \frac{v_p}{\Delta v} = \frac{v \cdot f_c}{c},$$

$$\rho_{de} = \frac{\tau_i T}{M} = \frac{\tau_p}{\Delta \tau},$$

where $M, N$ are number of delay-doppler domain modulation bins, $v_i$ is doppler spread indices, $\Delta v$ is doppler domain resolution, $v_p$ is doppler period, $v$ is UE speed, $f_c$ is carrier frequency, $c$ is speed of light, $\tau_i$ is delay spread indice, $\Delta \tau$ is delay domain resolution, $\tau_p$ is delay period.
Therefore, to restore the orthogonality of received symbols in delay-doppler domain bins, additional delay domain and doppler domain equalization with windowing operations are required at the receiver [76, 77]. Also, in addition to the conventional FFT operations subsequent SFFT operations are required for demodulation in delay-doppler domains. Hence, the overall operational cost incurred by UE for orthogonality restoration and demodulation of delay-doppler domain transmission frame can be evaluated as,

\[ C_{\text{delay-doppler}} = \sum_{i=1}^{M} \rho_{dei} M_i + \sum_{j=1}^{N} \rho_{doj} N_j + M \log_2 M + 2N \log_2 N, \]  

(4.14)

where \( M, N \) are the number of delay-doppler bins, \( \rho_{dei} \) is the non-orthogonality degree of \( M_i \)-th non-orthogonal delay bin in (4.13), and \( \rho_{doj} \) is the non-orthogonality degree of \( N_j \)-th non-orthogonal doppler bin (4.12).

Deducing from the cost analysis for orthogonality restoration and demodulation in time-frequency domain (4.6), spatial domain (4.11) and delay-doppler domains (4.14), the overall operational cost (no. of computational operations) required from the receiver side in UC-MDM is expressed in (4.15).

\[ C_{\text{UC-MDM}} = \sum_{i=1}^{N_s} \rho_{si} N_{si} \log_2 (N_{si}) + \log_2 (N_{si}) + \alpha \left( \sum_{i=1}^{N} \rho_{ji} N_i + \sum_{j=1}^{M} \rho_{fi} M_j + N \log_2 N \right) \]

\[ + (1 - \alpha) \left( \sum_{i=1}^{M} \rho_{dei} M_i + \sum_{j=1}^{N} \rho_{doj} N_j + M \log_2 M + 2N \log_2 N \right), \]  

(4.15)

where \( \rho_{si}, \rho_{ti}, \rho_{fi}, \rho_{dei}, \rho_{doj} \) are the transmitter-receiver pair specific non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains.

Hence from (4.15) it is evident that increase in non-orthogonality degrees increases the overall operational cost from user device for orthogonality restoration. Therefore, the proposed UC-MDM is designed to control such operational cost so that the user-device can be effectively utilized. For this purpose, the overall SINR of the wireless channel and achievable data rate of the proposed UC-MDM is analyzed in the following subsection.
4.2. System model

4.2.2 Analysis on impact of non-orthogonality on overall SINR and achievable data rate

The transmitter-receiver pair specific non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains induce non-orthogonal interference in those domains which impact the overall SINR of wireless channel, and thus consequently deteriorate the achievable data rate. Considering such non-orthogonal interference in spatial, time-frequency, and delay-doppler domains, the overall SINR of wireless channel between the transmitter-receiver pair is expressed as,

\[ \Upsilon = \frac{P}{P_s(\sigma^2_s) + \alpha P_{tf}(\rho_t, \rho_f) + (1 - \alpha)P_{dd}(\rho_{de}, \rho_{do}) + \sigma_n^2}, \]  

(4.16)

where \( \alpha = 1 \) in spatial-time-frequency domain modulation, and \( \alpha = 0 \) in spatial-delay-doppler domain modulation mode, \( P_s \) is the spatial domain interference, \( P_{tf} \) is the time-frequency domain interference, \( P_{dd} \) is delay-doppler domain interference, and \( \sigma_n^2 \) is the AWGN power.

Considering such non-orthogonal interference in SINR of wireless channel between the transmitter-receiver pair, the achievable data rate of the proposed UC-MDM scheme is expressed in 4.17.

\[ R_{UC-MDM} = \sum_{k=1}^{k=N_f} B_k \log_2 \left( \det \left( I + \frac{\Upsilon_k(\rho_s(\theta), \rho_{de}(\Delta \tau), \rho_{do}(\Delta \nu), \rho_t(T_s), \rho_f(\Delta f))}{N_f} \cdot H_c \cdot H_c^+ \right) \right) + \log_2(N_s \cdot N_b), \]

(4.17)

where \( B_k \) is subcarrier bandwidth, \( \Upsilon_k \) is the SINR of subcarrier, \( \rho_s, \rho_{de}, \rho_{do}, \rho_t, \rho_f \) are the non-orthogonality degrees in spatial, delay-doppler or time-frequency domains, \( H_c \) is spatially correlated channel matrix, \( N_s \) is number of antenna subsets, \( N_b \) is number of SM bits per antenna subset.

Hence from (4.16) and (4.17), it is evident that although different levels of non-orthogonal interference occur in different domains, they combinedly affect the overall SINR of the wireless channel, and consequently impact the achievable data rate.
4.3 Problem formulation

Considering the system specific goal of achieving required data rate with minimum operational cost from UE under transmitter-receiver pair specific non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains domains, the research problem in (4.3) is extended as a dual optimization problem in (4.18) with primary objective of achieving required data rate and minimization of receiver operational cost as secondary objective. Radio resources separation parameters $\theta, \Delta \tau, \Delta v, \Delta f, T_s$ are used as optimization parameters.

Assumptions 1: Allocated time frame $T$, bandwidth $B$, transmission power $P$ are fixed.

\[
R_{\text{UC-MDM}}(\rho_s, \rho_{de}, \rho_{do}, \rho_t, \rho_f) \geq R',
\]

s.t

\begin{align*}
C1: & \min (C_{\text{UC-MDM}}), \quad \alpha \in \{0, 1\} \\
C2: & v_p = 1/\tau_p \\
C3: & \Delta f = 1/T_s \\
C4: & 6 \leq f_c \leq 100 \text{ GHz}
\end{align*}

where $R'$ is required data rate, $(\rho_s, \rho_{de}, \rho_{do}, \rho_t, \rho_f)$ are non-orthogonality degrees in spatial, time-frequency, delay-doppler domains, $\theta$ is phase separation of SM symbols, $\Delta \tau, \Delta v$ is resolution of delay-doppler domain modulation bins, $T_s$ is symbol duration, $\Delta f$ is subcarrier spacing and $\alpha \in \{0, 1\}$ is modulation mode switching parameter of value $\alpha = 0$ in spatial-delay-doppler modulation and $\alpha = 1$ in spatial-time-frequency modulation mode.

C1 states that intial solution set of optimal number of radio resources from spatial, time-frequency, and delay-doppler domains that achieve required data rate should achieve minimal operational cost from receiver as secondary objective.

C2 states that doppler period $v_p$ is reciprocal of delay period $\tau_p$ to maintain orthogonality between delay-doppler domains.

C3 states that subcarrier spacing $\Delta f$ in frequency domain should be the reciprocal of symbol duration $T_s$ in time domain to maintain orthogonality between time-frequency domains.

C4 states that operating carrier frequency should be between 6 GHz, and 100 GHz to comply with 3GPP specifications for beyond 5G mmWave communication systems [68].
4.4 Proposed user-centric multi-dimensional modulation technique

In achieving the goal (4.3), the overall research problem (4.18) is decomposed into two sub-problems with closed form solutions in order to reduce the overall computational complexity. Considering the primary objective, Algorithm 2 is proposed to find the optimum radio resource combinations with optimum radio resource separation to achieve the required data rate.

Algorithm 2: Achieving required data rate with optimum radio resource separation

1. **Input:** UE speed ($v$), carrier frequency ($f_c$), Bandwidth $B$, IAC matrix at transmitter-receiver $(H_{Tx}, H_{Rx})$, estimated doppler spread ($v_i$), and estimated delay spread ($\tau_i$)

2. for $i = 1$ to $6$
   
   3. $n \in \{1,2,4,8,16,32\}$
   
   4. Set the subcarrier spacing $\Delta f = 15 \cdot n[i] \times 10^3$
   
   5. Set $T_s = 1/\Delta f$
   
   6. Set delay grids $N = 15 \cdot n[i] \Delta v \times 10^3$
   
   7. Set doppler grids $M = T_s/\Delta \tau$ where $\tau_p \cdot \Delta \tau = v_p \cdot \Delta v$
   
   8. Evaluate $[\rho_f, \rho_t, \rho_s, \rho_{do}, \rho_{de}]$ using (4.4),(4.5),(4.9),(4.12),(4.13)
   
   9. Find $\Delta f_i, T_s, M_i, N_i, \theta_i = \arg \min [\rho_{de}, \rho_{do}, \rho_f, \rho_f]$
   
   10. Calculate $(N_f, N_t, M, N_s) \forall (\Delta f_i, T_s, M_i, N_i, \theta_i)$
   
   11. Evaluate $(R)$ from (4.17) $\forall \{N_f, N_t, M, N_s, N_{si}, \alpha\}$
   
   12. while $(R) \geq R'$ do
       
       13. Append $N_f, N_t, M, N_s, \alpha \rightarrow \{S\}$
   
   14. end

16. **Output:** $S \equiv \{\text{number of delay bins } M, \text{doppler bins } N, \text{number of subcarriers } N_f, \text{number of time slots } N_t, \text{and number of antenna subsets } N_s, \text{modulation mode switching parameter } \alpha\}$

In the proposed Algorithm 2, at first the non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains ($\rho_s, \rho_t, \rho_f, \rho_{do}, \rho_{de}$) are evaluated. Then, the radio resource separation parameters $\theta, \Delta f, T_s, \Delta \tau, \Delta v$ are jointly optimized such that the non-orthogonality degrees are minimum in those domains.
In detail, re-arranging (4.12), and (4.13) yield,

\[ v_i = \rho_{do} \cdot \Delta f = \rho_{do} \cdot \frac{\Delta v}{N} \]  
(4.19)

\[ \tau_i = \rho_{de} \cdot T_s = \rho_{de} \cdot \frac{\Delta \tau}{M} \]  
(4.20)

Since IAC in spatial domain affects the phase of modulated symbol, the spatial domain non-orthogonality parameter is reformulated as,

\[ \sigma_s^2(\theta) = E\{|(H - H_c) \cdot H|^2\} \]  
(4.21)

Hence from (4.19), (4.20), and (4.21) it can be inferred that non-orthogonality degree in each domain depends on the corresponding radio resource separation values. Therefore, the proposed Algorithm 2 is designed to jointly find the optimum radio resource separation values \( \Delta f_i, T_s, \Delta v_i, \Delta \tau_i, \theta_i \) for which the non-orthogonality degree is minimal in each domain. Then, considering the obtained optimum radio resource separation values, and the total available radio resources such as bandwidth \( B \), transmission time \( T \), and transmit antennas \( N_t \), the number of radio resources in each domain such as number of subcarriers \( N_f \) in frequency domain, time slots \( N_t \) in time domain, delay bins \( M \) in delay domain, doppler bins \( N \) in doppler domain, and antenna subsets \( N_s \) in spatial domain are determined. Subsequently, the achievable data rate (4.17) is evaluated based on the non-orthogonality degree in those domains and the different combinations of radio resources in those domains. Finally, the combinations of radio resources which achieve equal or higher data rate are appended to the initial solution set \( S \).

To achieve the secondary objective of minimizing the receiver side operational cost, the following Algorithm 3 is proposed. The overall operational cost from receiver is evaluated for all combination of radio resources from the initial solution set \( S \). Then, optimum combination of radio resources that achieve minimal overall operational cost is output as the optimum solution in \( S_{opt} \). Hence, through the proposed two-staged sub-optimal optimization, UC-MDM can effectively achieve required communication data rate with minimum operational cost from receiver.
4.4. **Proposed User-Centric Multi-Dimensional Modulation Technique**

**Algorithm 3:** Achieving minimal operational cost with optimum radio resources

```plaintext
1 Input: Feasible radio resource sets S from Algorithm 1
2 for i = 1 to length(S) do
   3 evaluate UE cost C in (4.15) ∀S;
   4 update \([C_{UC-MDM}] \rightarrow [C_{UC-MDM} C]\)
5 end
6 find \(S_{opt} \subseteq \{S[\arg \min(C_{UC-MDM})]\}\)
7 Output: \(S_{opt} \equiv \{\text{optimum number of delay bins } M, \text{ doppler bins } N, \text{ optimum number of subcarriers } N_f, \text{ optimum number of time slots } N_t, \text{ optimum number of antenna subsets } N_s, \text{ modulation mode switching parameter } \alpha\}
```

The overall complexity of proposed two stage optimization algorithms are analyzed with respect to the total number of search and combination operations associated. In Algorithm 2, finding the feasible solution takes \(\mathcal{K}N_{iter1}\) search iterations which requires the computational complexity \(O(\mathcal{K}N_{iter1})\), where \(\mathcal{K}\) is the number of domains. The subsequent combination operation to find the set of radio resources that achieve required data rate incurs \(O(C_N^K)\) complexity. Similarly, in Algorithm 3, finding the optimum set of radio resources \(S_{opt} \in S\) that achieves minimum operational cost requires \(O(N_{iter2})\) complexity where \(N_{iter2} = \text{length}(S)\). Hence, overall complexity of proposed sub-optimal solutions to achieve the goal is \(O(\mathcal{K}N_{iter1} + C_N^K + N_{iter2})\), whereas the complexity of conventional grid search methods with \(\mathcal{K}\) dimensions to find the optimum combination of radio resources requires the computational complexity \(O(\mathcal{K}N_{iter1}N_{iter2}C_N^K)\).

Therefore, through the proposed decomposed implemented two stage optimization with sub-optimal solutions, the overall computational complexity is reduced from \(O(\mathcal{K}N_{iter1}N_{iter2}C_N^K) \rightarrow O(\mathcal{K}N_{iter1} + C_N^K + N_{iter2})\). Although, the search space is small \((N_{iter1}, N_{iter2} = 6, \text{ and } \mathcal{K} = 3)\) in the system model, with the increase in number of radio resources \(N\) and modulation domains \(\mathcal{K}\), the proposed two-stage optimization approach can achieve lower computational complexity compared to conventional grid search methods with \(\mathcal{K}\) dimensions [69].
4.5 Simulation results and analysis

In this section, a range of simulation results are presented to demonstrate the performance of the proposed UC-MDM scheme compared with state-of-art MIMO-OFDM, and MIMO-OTFS systems. It is assumed that the BS and UE can flexibly operate on a range of subcarrier spacing $\Delta f \in \{15, 30, 60, 120, 240, 480\}$ kHz. As presented in Table 4.1, CDL-A, TDL-A, and TDL-D channel models and communication parameters recommended for mmWave 5G communications [68] were incorporated in the simulation analysis. The simulation was implemented using MATLAB signal processing and 5G toolbox.

Table 4.1: Simulation parameters

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bandwidth</td>
<td>100 MHz</td>
</tr>
<tr>
<td>Number of antennas</td>
<td>64</td>
</tr>
<tr>
<td>Carrier Frequency</td>
<td>6 – 100 GHz</td>
</tr>
<tr>
<td>UE speed</td>
<td>0 – 500 km/h</td>
</tr>
<tr>
<td>Subcarrier spacing</td>
<td>15, 30, 60, 120, 240, 480 kHz</td>
</tr>
<tr>
<td>Channel model</td>
<td>CDL-A, TDL-A, TDL-D</td>
</tr>
</tbody>
</table>

In Fig. 4.3, the overall operational cost required at receiver to achieve 128 Mbps data rate through the proposed UC-MDM, and state-of-art MIMO-OFDM, MIMO-OTFS systems under the spatial, time-frequency, and delay-doppler domain normalized non-orthogonality degrees are illustrated. As shown, the MIMO-OFDM encounters higher operational cost (orthogonality restoration cost) at higher non-orthogonality degree range due to it’s high sensitivity towards higher non-orthogonality degrees in time-frequency domains. On the other hand, MIMO-OTFS incurs comparatively less operational cost than the MIMO-OFDM due to the intrinsic robustness of OTFS towards time-frequency domain non-orthogonality degrees. Compared to MIMO-OFDM and MIMO-OTFS systems, the proposed situation-aware and cost-aware UC-MDM scheme requires comparatively less operational cost since it utilizes optimum combination of radio resources with optimum radio resource separation such that the overall operational cost for orthogonality restoration and demodulation is minimized while achieving the required communication data rate.
4.5. Simulation results and analysis

Figure 4.3: Number of computational operations required to achieve 128 Mbps data rate for varying normalized non-orthogonality degree.

The CDF plot in Fig. 4.4 shows the overall operational cost incurred by the proposed UC-MDM, MIMO-OFDM, and MIMO-OTFS to support random data rate requirements under simultaneously varied transmitter-receiver specific non-orthogonality degrees in spatial, time-frequency, and delay-doppler domains. Hence it is validated that UC-MDM incurs comparatively less operational cost than MIMO-OFDM and MIMO-OTFS systems to achieve the required data rate through cost-aware utilization of optimum radio resource combination with optimum radio resource separation through multi-dimensional modulation in either spatial-time-frequency or spatial-delay-doppler domains.

Figure 4.4: CDF of operational cost from user-device to achieve required data rates
Furthermore, considering the use cases in 5G and beyond where UEs have higher data rate requirements and can tradeoff the operational cost, the proposed UC-MDM can jointly optimize the radio resource separation parameters in each domain according to Algorithm 2 to minimize the overall non-orthogonality degree in each domain, and thus achieve maximized data rate compared to MIMO-OFDM and MIMO-OTFS systems as shown in Fig. 4.5 PDF plot.

![Figure 4.5: PDF of data rate achieved by proposed UC-MDM, MIMO-OFDM, MIMO-OTFS](image)

The plot in Fig. 4.6 shows the data rate achieved for different number of transmit antennas (8, 16, 32, 64, 128). It can be noted that the data rate gain achieved by UC-MDM compared to MIMO-OFDM, and MIMO-OTFS systems is significantly higher when $N_t \geq 32$. Although the data rate of MIMO-OFDM and MIMO-OTFS increases through higher spatial multiplexing gain, the MIMO-OFDM and MIMO-OTFS still utilize only the available radio resources in either time-frequency or delay-doppler domains for modulation. Different from such two-dimensional MIMO-OFDM and MIMO-OTFS systems, the proposed UC-MDM utilizes the spatial domain transmit antenna subsets as additional radio resources through multi-dimensional modulation jointly with time-frequency or delay-doppler modulation radio resources. Hence, the proposed UC-MDM can achieve comparatively higher data rate gain by utilizing higher number of transmit antennas subsets through spatial modulation. Therefore, it can be corroborated that the proposed UC-MDM scheme is advantageous for massive MIMO arrays in future wireless networks.
4.5. Simulation Results and Analysis

Figure 4.6: Data rate achieved for different number of transmit antennas \( N_t \).

As further illustrated in Fig. 4.7, the proposed UC-MDM scheme achieves higher spectral throughput than MIMO-OFDM, and MIMO-OTFS systems through simultaneous utilization spatial domain radio resources with the time-frequency or delay-doppler domain radio resources in multi-dimensional modulation, whereas the current MIMO-OFDM and MIMO-OTFS systems only achieve spatial multiplexing gain in the spatial domain.

Figure 4.7: Spectral throughput achieved for different number of transmit antennas \( N_t \).
4.6 Chapter summary

This chapter extensively focused on the user-device specific new challenges in beyond 5G communications: a) The non-orthogonality degrees perceived by a receiver in spatial, time-frequency, and delay-doppler domains are transmitter-receiver pair specific and domain specific due to predominant device related attributes such as different carrier frequency, different mobility speed, and different synchronization capability of devices in different domains, b) the operational cost required from user-device are also device specific and domain specific due to different synchronization capabilities of receiver in different domains. To minimize the operational cost from receiver while achieving required data rate, a user-centric multi-dimensional modulation (UC-MDM) technique was proposed. The proposed UC-MDM was designed to effectively utilize the optimum radio resource combinations through multi-dimensional modulation in either spatial-time-frequency or spatial-delay-doppler domains with optimum radio resource separation in each domain to achieve the goal. From simulation results, the usefulness of the proposed UC-MDM scheme was demonstrated. As future works, additional radio resource domains such as power domain, and code domain can be incorporated into multi-dimensional modulation to achieve higher communication performance.
Chapter 5

Scalable Digital Predistortion Architecture to Compensate Spatial Domain Nonlinear Distortions

The inherent nonlinearity exhibited by power amplifiers (PAs) driven at or close to saturation has become more severe due to strong nonlinear crosstalks and reverse crosstalks with memory effects induced from neighboring PAs in closely packed MIMO arrays. As a result, the output signal encounter severe deterioration in the signal quality and overall spectral efficiency. The current nonlinear distortion compensation techniques (eg: multi-input digital predistortion (DPD)) models have become increasingly complex and suffer from sharp increase in DPD cost as the number of transmitter path increases. To circumvent this impediment, a decomposed, less complex, scalable cross-correlation based single input DPD architecture is proposed in this chapter for MIMO systems. The proposed CC-SISO DPD architecture estimates and cancels the nonlinear and reverse crosstalk prior to the DPD process. Furthermore, the proposed architecture also eliminates the requirement for feedback signal taps before the PA, and thus reduces the overall hardware implementation complexity. Through simulations using 5G signals and PAs with memory effects, it is demonstrated that the proposed CC-SISO DPD architecture is capable of efficiently linearizing MIMO systems while achieving higher accuracy and less complex than state-of-art multi-input DPD models for MIMO systems.
5.1 Introduction

Power amplifier (PA) is a key transmit antenna component which intrinsically exhibits nonlinear distortions to amplified output signal of a PA when driven in saturation mode for higher power efficiency. In current communication systems PAs are operated close to saturation and almost linearized output signal is obtained with digital predistortion (DPD) techniques. The DPD, and PA are implemented in a cascaded model so that inverse nonlinearity upstream generated by the DPD, and the nonlinear components in PA’s output cancel each other, thus DPD combined with PA act as a single linear amplification system.

However, with significantly reduced spatial separation between antennas in MIMO arrays, the nonlinear and reverse crosstalk induced simultaneously from multiple transmit paths have become more severe. As highlighted in chapter 2 literature analysis, under such simultaneous crosstalks induced from multiple PA branches, current multi-input DPD models encounter significant DPD operational cost due to sharp increase in DPD coefficients. Furthermore, they also need feedback paths to be added between input signal and PA for each transmit path. While this is not a major issue in instrument-based laboratory configurations, this approach becomes rather infeasible to implement in massive MIMO or non-1D array configurations.

Based on these observations, the contributions of this chapter to overcome this impediment in massive MIMO systems are summarized as,

- Firstly, a less complex cross-correlation based method is developed to estimate nonlinear as well as reverse crosstalk induced simultaneously from multiple PA branches. The cross-correlation based nonlinear crosstalk estimation is developed such that it uses only the output signals of the PA. Hence, it eliminates the need for input signal feedback tap which has become impracticable to be implemented in advanced MIMO RF boards.
- Using the developed cross-correlation method, a decomposed, multiple single input single output (CC-SISO) DPD architecture is proposed that requires lower operational cost compared to current multi-input DPD models for MIMO systems. Furthermore, an optimized CC-SISO DPD architecture is proposed for equispaced MIMO arrays. It is designed to reuse the estimated nonlinear crosstalk coefficient among similar PAs, and also reduce the overall DPD cost based on varying levels of crosstalk.
The rest of this chapter is organized as follows: In section 5.2, the system model is presented. Then, the nonlinear and reverse crosstalk is modeled. In section 5.3, the proposed CC-SISO DPD architectures are presented. In section 5.4, simulation results are presented and this chapter is summarized with recommendations for future directions in section 5.5.

5.2 System model

In traditional SISO systems, the PA’s output is a function of its input, and the inherent nonlinear distortive characteristics of the amplifier. However in practical MIMO systems, the output signal of a PA is also affected by the nonlinear and reverse crosstalk from other signal paths [63] as shown for a $3 \times 3$ MIMO array in Fig. 5.1

\[ x_1 + \alpha_1 x_c + \gamma x_1 + \alpha_2 x_2 + \beta_1 y_1 + \beta_2 y_2 \]

\[ x_2 + \rho x_2 \]

\[ y_1 \]

\[ y_2 \]

\[ y_c \]

\[ \alpha_1, \alpha_2, \beta_1, \beta_2, \gamma, \rho \]

Figure 5.1: Nonlinear and reverse crosstalk at $3 \times 3$ MIMO array

5.2.1 Spatial domain nonlinear and reverse crosstalk analysis

As shown in system model, PA$_c$ is worst affected by simultaneous crosstalk (nonlinear and reverse crosstalk) from both neighbor PAs (PA$_1$, and PA$_2$). Therefore, considering the worst case scenario, PA$_c$ is focused in the following analysis. In general, the nonlinear crosstalk is proportional to the spatial separation distance between the PAs, and the weak crosstalk below $-30$ dB is not significant [63]. Therefore, nonlinear and reverse crosstalk values of range $\{-5, -10, -15, -20, -25, -30\}$ dB is taken into account in the system model.
Using power series model [64], the output signal of a PA of order $K$ and memory depth $Q$ is expressed as,

$$y_{MP[n]} = \sum_{k=1}^{K} \sum_{q=0}^{Q} a_{kq} (x_{[n-q]}) |x_{[n-q]}|^{K-1}, \quad (5.1)$$

where $K$ is the PA order, $Q$ is the memory depth, and $n$ is the time index. Hence, using (5.1), the output signal $y_c$ of PA, incorporating the memory effect components from input signal, nonlinear crosstalk signal, and the reverse crosstalk from neighbor PAs is expressed as,

$$y_{c[n]} = \sum_{k=1}^{K} \sum_{q=0}^{Q} a_{kq} (x_{c[n-q]} + \alpha_1 x_{1[n-q]} + \alpha_2 x_{2[n-q]} + \beta_1 y_{1[n-q]} + \beta_2 y_{2[n-q]})$$

$$\times |x_{c[n-q]} + \alpha_1 x_{1[n-q]} + \alpha_2 x_{2[n-q]} + \beta_1 y_{1[n-q]} + \beta_2 y_{2[n-q]}|^{k-1} \quad (5.2)$$

On the other hand, using (5.1), the multi-input DPD model including the crosstalk from neighbor PAs with memory effects for a $n \times n$ MIMO transmitter is expressed as,

$$y_{MN[n]} = \sum_{m=1}^{M} \sum_{n=1}^{N} \sum_{k_1=0}^{k-R} \cdots \sum_{r_{N-1}=0}^{r_N-1} a_{m,n,r_1,r_2,\ldots,r_{N-1}}^{1} x_{1[n-m]} |x_{1[n-m]}|^{k-R} \cdots |x_{m[n-m]}|^{R_i}$$

$$+ \sum_{m=1}^{M} \sum_{n=1}^{N} \sum_{k_1=0}^{k-R} \cdots \sum_{r_{N-1}=0}^{r_N-1} a_{m,n,r_1,r_2,\ldots,r_{N-1}}^{2} x_{2[n-m]} |x_{2[n-m]}|^{k-R} \cdots |x_{m[n-m]}|^{R_i}$$

$$\vdots$$

$$+ \sum_{m=1}^{M} \sum_{n=1}^{N} \sum_{k_1=0}^{k-R} \cdots \sum_{r_{N-1}=0}^{r_N-1} a_{m,n,r_1,r_2,\ldots,r_{N-1}}^{R} x_{m[n-m]} |x_{m[n-m]}|^{k-R} \cdots |x_{m[n-m]}|^{R_i},$$

where $R = (r_1 + r_2 + \cdots + r_{N-1}), R_i \in r_1, r_2, \ldots r_{N-1})$. From (5.3), the operational cost of multi-input DPD model is evaluated as,

$$C_{DPD-PH} = K(M + 1) [(M + 2)(M + 3) \cdots (M + n_t)], \quad (5.3)$$

where $n_t$ is the number of transmit paths. Hence from 5.3, it is evident that the operational cost of multi-input DPD models increase exponentially with the increase number of MIMO transmitters $n_t$. 
5.2. System model

5.2.2 Proposed cross-correlation based crosstalk estimation technique

To achieve our previously-stated objectives, a less complex, novel, cross-correlation based technique is developed in this subsection to estimate crosstalk in spatial domain as follows:

1. At first, PA$_1$ is excited with an input signal while keeping the remaining PAs in off state to obtain the individual output signal $y_{10[n]}$ from neighboring PA$_1$ with order $K = 5$, and memory depth $M = 3$ as expressed in (5.4). Since the other PAs were in idle state, the individual output signal from PA$_1$ does not contain any nonlinear crosstalk signals.

$$y_{10[n]} = \sum_{k=1}^{5} \sum_{m=0}^{3} a_{km} (x_{1[n-m]}) |x_{1[n-m]}|^{K-1}$$

$$= a_{10} (x_{1[n]}) + a_{20} (x_{1[n]}) + \cdots + a_{50} (x_{1[n]}) |x_{1[n]}|^4$$

$$+ a_{11} (x_{1[n-1]}) + a_{21} (x_{1[n-1]}) + \cdots + a_{51} (x_{1[n-1]}) |x_{1[n-1]}|^4$$

$$+ a_{12} (x_{1[n-2]}) + a_{22} (x_{1[n-2]}) + \cdots + a_{52} (x_{1[n-2]}) |x_{1[n-2]}|^4$$

$$+ a_{13} (x_{1[n-3]}) + a_{23} (x_{1[n-3]}) + \cdots + a_{53} (x_{1[n-3]}) |x_{1[n-3]}|^4$$

(5.4)

2. Then, all PAs are excited simultaneously in normal operation condition, and signal outputs $y_c$ from PA$_c$ and $y_1$ from PA$_1$ which also consists nonlinear crosstalk and reverse crosstalk components from neighbor PA branches is obtained using (5.2).

3. Next, the individual signal output $y_{10[n]}$ in (5.4) is cross-correlated with the output signal $y_{c[n]}$ obtained in step (2) and the nonlinear crosstalk signal components are extracted from 1$^{st}$ branch (signal components with $\alpha x_1$ terms) as formulated in (5.5).

$$y_{1e} = y_{10} \otimes y_c$$

$$\approx a_{10} (\alpha x_{1[n]}) + a_{20} (\alpha x_{1[n]}) + \cdots + a_{50} (\alpha x_{1[n-1]}) |\alpha x_{1[n]}|^4$$

$$+ a_{11} (\alpha x_{1[n-1]}) + a_{21} (\alpha x_{1[n-1]}) + \cdots + a_{51} (\alpha x_{1[n-1]}) |\alpha x_{1[n-1]}|^4$$

$$+ a_{12} (\alpha x_{1[n-2]}) + a_{22} (\alpha x_{1[n-2]}) + \cdots + a_{52} (\alpha x_{1[n-2]}) |\alpha x_{1[n-2]}|^4$$

$$+ a_{13} (\alpha x_{1[n-3]}) + a_{23} (\alpha x_{1[n-3]}) + \cdots + a_{53} (\alpha x_{1[n-3]}) |\alpha x_{1[n-3]}|^4$$

(5.5)

4. Then, the normal output signal $y_{1[n]}$ is cross-correlated with the output signal $y_{c[n]}$ and the reverse crosstalk signal components from 1$^{st}$ branch (signal components with $\beta_1 y_1$
terms) is extracted as formulated in (5.6).

\[ y_{1a} = y_1 \otimes y_c \]
\[ \approx a_{10} (\beta_1 y_{1[n]}^1) + a_{20} (\beta_1 y_{1[n]}^1) + \cdots + a_{50} (\beta_1 y_{1[n]}^1) \beta_1 y_{1[n]}^1 \]
\[ + a_{11} (\beta_1 y_{1[n-1]}^1) + a_{21} (\beta_1 y_{1[n-1]}^1) + \cdots + a_{51} (\beta_1 y_{1[n-1]}^1) \beta_1 y_{1[n-1]}^1 \]
\[ + a_{12} (\beta_1 y_{1[n-2]}^1) + a_{22} (\beta_1 y_{1[n-2]}^1) + \cdots + a_{52} (\beta_1 y_{1[n-2]}^1) \beta_1 y_{1[n-2]}^1 \]
\[ + a_{13} (\beta_1 y_{1[n-3]}^1) + a_{23} (\beta_1 y_{1[n-3]}^1) + \cdots + a_{53} (\beta_1 y_{1[n-3]}^1) \beta_1 y_{1[n-3]}^1 \]

(5.6)

5. Finally, the nonlinear crosstalk coefficient \( \alpha_1 \), and reverse crosstalk coefficient \( \beta_1 \) are estimated using nonlinear polynomial regression fit method as illustrated in Fig. 5.2. Similarly, crosstalk coefficients for other neighbor PA branch can be estimated as presented in Algorithm 1.

![Figure 5.2: Nonlinear and reverse crosstalk estimation](image-url)

**Algorithm 4:** Nonlinear and reverse crosstalk estimation

1. **Repeat**
2. Excite neighbor PA\( _i \) individually
3. Obtain individual PA\( _i \) output (5.4)
4. Excite neighbor PA\( _i \) in normal operation
5. Obtain PA\( _i \) output (5.2)
6. Extract reverse crosstalk components \( y_{1a} \) by cross-correlating \( y_c \) with \( y_1 \)
7. Extract nonlinear crosstalk components \( y_{1b} \) by cross-correlating \( y_c \) with \( y_{10} \)
8. Estimate \( \alpha_i \) by nonlinear polynomial regression fit (\( y_{1a}, y_1 \))
9. Estimate \( \beta_i \) by nonlinear polynomial regression fit (\( y_{1b}, y_{10} \))
10. **until** all crosstalk coefficients are estimated
11. **Output** \( \alpha_i, \beta_i \in \{1, 2, \ldots, n_i\} \)
5.2.3 Proposed cross-correlation based single-input (CC-SISO) DPD architecture for MIMO

Using the developed cross-correlation based nonlinear crosstalk estimation method in the previous subsection, a novel decomposed multiple SISO DPD architecture for MIMO systems is proposed as shown in Fig. 5.3.

Figure 5.3: (a) Existing Multi-input DPD [56], (b) proposed CC-SISO DPD architecture in $3 \times 3$ MIMO array.

In the proposed architecture, the crosstalk estimation block outputs the estimated nonlinear and reverse crosstalk coefficient corresponding to each neighbor PA branch. To elaborate more, in a scenario in which all PAs suffers from and reverse crosstalk from both neighbor PA branches, the crosstalk coefficient estimation block estimates the crosstalk from the corresponding neighbor PAs, and subsequently outputs the estimated crosstalk coefficients in a switched manner to the corresponding DPDs as required. Hence, the overall cost of proposed CC-SISO DPD architecture incurs cost for crosstalk estimation as well as for DPD is evaluated as,

$$C_{CC-SISO} = \frac{K(M+1)n_t + 5K(M+1)n_{ct}}{n_t + n_{ct}},$$ \hspace{1cm} (5.7)
where $n_t$ is the number of transmit antennas, $n_{ct}$ is the number of crosstalk neighbor antennas. From (5.7), it can be inferred that the cost for DPD and crosstalk estimation increases linearly with the increase in number of transmit antennas ($n_t$). Furthermore, the crosstalk estimation component incurs the cost to estimate nonlinear and reverse crosstalk from neighbor PAs only. In other words, it is sufficient to estimate the nonlinear crosstalk from neighbor PAs which induce crosstalk up to $-30$ dB since lower values are not significant.

Since the operational cost of current multi-input DPD model increases exponentially with the increase in number of transmitters ($n_t$) as expressed in (5.3), such multi-input models scale poorly in MIMO systems larger than $2 \times 2$. On the other hand, the operational cost of proposed CC-SISO DPD model increases linearly with the increase in number of MIMO antennas as expressed in (5.7). Furthermore, the proposed CC-SISO DPD does not require signal feedback paths $z_1, z_2, z_3$ (Fig. 5.3), thus reduces the overall hardware implementation complexity.

In addition, the proposed CC-SISO DPD architecture is further optimized to reduce the overall cost based on varying nonlinear crosstalk levels at equispaced MIMO arrays. The optimization steps are performed as follows:

1. In the proposed CC-SISO DPD architecture, the cross-correlation operation to estimate the nonlinear crosstalk requires all $(K(M + 1))$ signal components in (5.4). As presented in Table 5.5, and illustrated in Fig. 5.10, all the cross-correlation terms are required to estimate the nonlinear crosstalk with high accuracy at strong nonlinear crosstalk level such as $-5$ dB. However, at weak crosstalk levels such as $-30$ dB, a fewer crosstalk terms is sufficient to estimate the nonlinear crosstalk with high accuracy. Therefore, in the optimized CC-SISO DPD architecture, based on the nonlinear crosstalk strength, the number of cross-correlation terms for cross-correlation operation is adaptively reduced to reduce the overall operational cost.

2. Furthermore, nonlinear crosstalk is a RF board dependent variable which is almost constant for PAs with equidistant separation in equispaced MIMO arrays. Therefore, it is efficient to estimate only the nonlinear crosstalk at PAs with distinct spatial separation and reuse estimated nonlinear crosstalk coefficient value across other PAs with similar spatial separation to reduce overall nonlinear crosstalk coefficient estimation cost.
5.3 Simulation results and analysis

In this section, a range of simulation results are presented using 5G signals and PA with actual configurations to evaluate their performance of our CC-SISO DPD architectures with state-of-art multi-input DPD models for MIMO systems.

5.3.1 PA coefficient estimation

At first the PA coefficients were estimated for actual 5G input, output signal pairs. The estimated PA coefficients for PA of order 5, and memory depth 3 for the input signals of 10 MHz, and 20 MHz bandwidth are presented in Table 5.1 and Table 5.2.

Table 5.1: Estimated PA coefficients for 10 MHz bandwidth input signal

<table>
<thead>
<tr>
<th>M, K</th>
<th>k = 1</th>
<th>k = 2</th>
<th>k = 3</th>
<th>k = 4</th>
<th>k = 5</th>
</tr>
</thead>
<tbody>
<tr>
<td>m = 0</td>
<td>-5.5490+0.3417i</td>
<td>0.3485-0.2759i</td>
<td>-0.0263+0.0123i</td>
<td>0.0013+0.0030i</td>
<td>0.0000-0.0002i</td>
</tr>
<tr>
<td>m = 1</td>
<td>-6.088-0.0464i</td>
<td>0.0273-0.0015i</td>
<td>-0.0051+0.0017i</td>
<td>0.0014-0.0018i</td>
<td>-0.0001+0.0001i</td>
</tr>
<tr>
<td>m = 2</td>
<td>-0.0635+0.0626i</td>
<td>0.0715-0.0578i</td>
<td>-0.0167+0.0112i</td>
<td>0.0007+0.0007i</td>
<td>0.0000-0.0001i</td>
</tr>
<tr>
<td>m = 3</td>
<td>-0.0093-0.0178i</td>
<td>-0.0412+0.0179i</td>
<td>0.0124-0.0024i</td>
<td>-0.0012-0.0005i</td>
<td>0.0000+0.0000i</td>
</tr>
</tbody>
</table>

Table 5.2: Estimated PA coefficients for 20 MHz bandwidth input signal

<table>
<thead>
<tr>
<th>M, K</th>
<th>k = 1</th>
<th>k = 2</th>
<th>k = 3</th>
<th>k = 4</th>
<th>k = 5</th>
</tr>
</thead>
<tbody>
<tr>
<td>m = 0</td>
<td>-5.1611+0.2882i</td>
<td>0.3221-0.2625i</td>
<td>-0.0231+0.0138i</td>
<td>0.0013+0.0018i</td>
<td>-0.0000-0.0001i</td>
</tr>
<tr>
<td>m = 1</td>
<td>-0.8759+0.1545i</td>
<td>0.0510-0.0224i</td>
<td>-0.0130+0.0058i</td>
<td>0.0024-0.0010i</td>
<td>-0.0001+0.0001i</td>
</tr>
<tr>
<td>m = 2</td>
<td>-0.1136-0.1679i</td>
<td>0.0935-0.0627i</td>
<td>-0.0187+0.0118i</td>
<td>0.0009-0.0004i</td>
<td>0.0000-0.0000i</td>
</tr>
<tr>
<td>m = 3</td>
<td>0.0313+0.0984i</td>
<td>-0.0383-0.0066i</td>
<td>0.0078+0.0051i</td>
<td>-0.0004-0.0008i</td>
<td>-0.0000+0.0000i</td>
</tr>
</tbody>
</table>
The estimated output signal using the obtained PA coefficients, and the actual output signal from the PA is shown in time domain (Fig. 5.4, 5.5), and in frequency domain (Fig. 5.6, 5.7). Hence, it can be observed that the actual, and estimated signals fit well in time domain as well as in frequency domain.

Figure 5.4: Actual, and estimated output signal for 10 MHz bandwidth input signal.

Figure 5.5: Actual, and estimated output signal for 20 MHz bandwidth input signal.
5.3. Simulation results and analysis

![Figure 5.6](image1.png)

Figure 5.6: Spectrum of actual, estimated 10 MHz bandwidth output signal.

![Figure 5.7](image2.png)

Figure 5.7: Spectrum of actual, estimated 20 MHz bandwidth output signal.

Furthermore, the goodness of fit between the actual output signal, and the estimated output signal is evaluated using the normalized mean square error (NMSE) metric as,

$$\text{NMSE} = 10 \log_{10} \left( \frac{\sum_{n=1}^{N} \|y - \hat{y}\|^2}{\|y - \bar{y}\|^2} \right),$$

(5.8)

where $y$ is actual output signal, and $\hat{y}$ is the estimated output signal. Hence, the evaluated NMSE values for both 10 MHz, and 20 MHz signals are presented in Table 5.3.
Table 5.3: NMSE value between actual, estimated signals

<table>
<thead>
<tr>
<th>Signal bandwidth (MHz)</th>
<th>NMSE (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>-38.9140</td>
</tr>
<tr>
<td>20</td>
<td>-39.5435</td>
</tr>
</tbody>
</table>

Since the NMSE values obtained are approximately equal to −40 dB, it can be confirmed that the estimated output signal, and actual output signal fit well. Hence, the accuracy of the obtained PA coefficients are validated, and the estimated output signal, and the estimated PA coefficients are utilized for further simulation analysis.

5.3.2 Nonlinear and reverse crosstalk coefficients estimation

Using the obtained PA coefficients, simultaneous nonlinear and reverse crosstalk were induced from both 1st, and 2nd PA branches to the input signal of PA. Then, using the proposed cross-correlation based crosstalk estimation method, the nonlinear crosstalk coefficient α₁, and reverse crosstalk coefficient β₁ from 1st PA branch were evaluated. The simultaneous nonlinear and reverse crosstalk levels of range {−30, −25, −20, −15, −10, −5} dB [63] from both neighboring PA branches (PA₁, PA₂) were considered for different configurations such as memoryless PA, PA of memory depth 3, and PA orders 3, 5. Estimated nonlinear crosstalk coefficient αₑ, and reverse crosstalk coefficient βₑ for each PA configuration is given in Table 5.4 and illustrated in Fig 5.8 and Fig. 5.9.

Table 5.4: Actual and estimated nonlinear crosstalk coefficient α

<table>
<thead>
<tr>
<th>Nonlinear crosstalk strength (dB)</th>
<th>-5</th>
<th>-10</th>
<th>-15</th>
<th>-20</th>
<th>-25</th>
<th>-30</th>
</tr>
</thead>
<tbody>
<tr>
<td>α,β (actual)</td>
<td>0.6325</td>
<td>0.2000</td>
<td>0.0632</td>
<td>0.0200</td>
<td>0.0063</td>
<td>0.0020</td>
</tr>
<tr>
<td>αₑ (K = 3, M = 0)</td>
<td>0.6307</td>
<td>0.2001</td>
<td>0.0633</td>
<td>0.0200</td>
<td>0.0063</td>
<td>0.0020</td>
</tr>
<tr>
<td>αₑ (K = 5, M = 3)</td>
<td>0.6523</td>
<td>0.2054</td>
<td>0.0638</td>
<td>0.0201</td>
<td>0.0063</td>
<td>0.0020</td>
</tr>
<tr>
<td>βₑ (K = 5, M = 3)</td>
<td>0.6293</td>
<td>0.1992</td>
<td>0.0629</td>
<td>0.0200</td>
<td>0.0063</td>
<td>0.0020</td>
</tr>
</tbody>
</table>
5.3. Simulation results and analysis

Figure 5.8: Estimated $\beta_e$ for varying reverse crosstalk levels

Figure 5.9: Estimated $\alpha_e$ for varying nonlinear crosstalk levels
Chapter 5. Scalable DPD to Compensate Spatial Domain Nonlinear Distortion

From Fig. 5.8 and Fig. 5.9, it can be verified that the nonlinear and reverse crosstalk from 2nd branch did not affect the accuracy of estimated nonlinear crosstalk coefficient $\alpha_e$, and reverse crosstalk coefficient $\beta_e$ for branch 1 thus validating the robustness of the proposed cross-correlation based method to estimate the simultaneous nonlinear and reverse crosstalk from multiple PA branches in spatial domain. Even at high complex PA configuration such as PA order of 5 with memory depth of 3, and under simultaneous strong nonlinear crosstalk condition of $-5$ dB from both neighbor PA branches, the proposed cross-correlation based method achieved estimation accuracy of 95.16 %.

5.3.3 Analysis on operational cost and scalability of CC-SISO DPD for MIMO

From (5.5), it can be inferred that all cross-correlation signal terms are used to estimate the nonlinear crosstalk coefficient. To reduce the nonlinear crosstalk estimation cost in proposed optimized CC-SISO method for equispaced MIMO arrays, further analysis is carried out between the operational cost, and estimation accuracy for reduced number of cross-correlation signal terms. The results of evaluation analysis on number of cross-correlation signal terms, crosstalk strength from neighbor PA branch, and the achieved nonlinear crosstalk coefficient estimation accuracy is presented in Table 5.5, and illustrated in Fig 5.10.

<table>
<thead>
<tr>
<th>Terms</th>
<th>$\alpha$ (dB)</th>
<th>-5</th>
<th>-10</th>
<th>-15</th>
<th>-20</th>
<th>-25</th>
<th>-30</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$\alpha$</td>
<td>0.6325</td>
<td>0.2000</td>
<td>0.0632</td>
<td>0.0200</td>
<td>0.0063</td>
<td>0.0020</td>
</tr>
<tr>
<td>20</td>
<td>$\alpha_e$</td>
<td>0.6324</td>
<td>0.1999</td>
<td>0.0632</td>
<td>0.0200</td>
<td>0.0063</td>
<td>0.0020</td>
</tr>
<tr>
<td></td>
<td>Accuracy %</td>
<td>99.98</td>
<td>99.95</td>
<td>99.97</td>
<td>100.0</td>
<td>100.0</td>
<td>100.0</td>
</tr>
<tr>
<td>16</td>
<td>$\alpha_e$</td>
<td>0.3665</td>
<td>0.1441</td>
<td>0.0632</td>
<td>0.0198</td>
<td>0.0063</td>
<td>0.0020</td>
</tr>
<tr>
<td></td>
<td>Accuracy %</td>
<td>57.94</td>
<td>72.05</td>
<td>87.97</td>
<td>99.0</td>
<td>100.0</td>
<td>100.0</td>
</tr>
<tr>
<td>12</td>
<td>$\alpha_e$</td>
<td>0.2110</td>
<td>0.1027</td>
<td>0.048</td>
<td>0.0192</td>
<td>0.0063</td>
<td>0.0020</td>
</tr>
<tr>
<td></td>
<td>Accuracy %</td>
<td>33.36</td>
<td>51.35</td>
<td>75.95</td>
<td>96.0</td>
<td>100.0</td>
<td>100.0</td>
</tr>
<tr>
<td>8</td>
<td>$\alpha_e$</td>
<td>0.1195</td>
<td>0.0712</td>
<td>0.0396</td>
<td>0.0179</td>
<td>0.0062</td>
<td>0.0020</td>
</tr>
<tr>
<td></td>
<td>Accuracy %</td>
<td>18.89</td>
<td>35.60</td>
<td>62.66</td>
<td>89.50</td>
<td>98.41</td>
<td>100.0</td>
</tr>
<tr>
<td>4</td>
<td>$\alpha_e$</td>
<td>0.0636</td>
<td>0.0448</td>
<td>0.0144</td>
<td>0.0116</td>
<td>0.0057</td>
<td>0.0019</td>
</tr>
<tr>
<td></td>
<td>Accuracy %</td>
<td>10.06</td>
<td>22.40</td>
<td>22.78</td>
<td>58.0</td>
<td>90.48</td>
<td>95.0</td>
</tr>
</tbody>
</table>
5.3. Simulation results and analysis

Figure 5.10: Number of cross-correlation terms used vs estimation accuracy.

In Fig. 5.10, it is shown that to estimate nonlinear crosstalk coefficients at weak crosstalk level (−30) dB with high accuracy ≥ 90 %, a fewer number of signal terms are sufficient. Therefore, by reducing the number of cross-correlation terms based on nonlinear crosstalk strength, the proposed optimized CC-SISO DPD achieves significantly lesser cost than multi-input DPD models. Furthermore, the evaluated scalability of proposed CC-SISO DPD architecture compared with state-of-art multi-input MIMO DPD models is shown in Fig. 5.11.

Figure 5.11: Number of operations for different no. of MIMO transmitters.
From Fig. 5.11, it can be observed that the DPD operational cost for current MIMO-MDM models increase exponentially with the increase in MIMO transmitters, whereas the cost of proposed CC-SISO DPD increases almost linearly. Hence, CC-SISO DPD, and optimized CC-SISO DPD achieve significantly lower DPD cost, and scale well with MIMO array size.

5.4 Chapter summary

In this chapter, the nonlinear distortions with memory effects induced due to simultaneous nonlinear and reverse crosstalk from multiple PA branches in spatial domain was extensively focused. To estimate reverse and nonlinear crosstalk from multiple PA branches, a cross-correlation based approach to estimate the nonlinear and reverse crosstalk in MIMO systems was developed. Using the developed the cross-correlation based crosstalk estimation method, a cost efficient, highly scalable CC-SISO DPD architecture was proposed for MIMO systems. Furthermore, by adaptively reducing the number of cross-correlation terms used for crosstalk estimation, and by re-utilizing the nonlinear crosstalk estimation effort, an optimized CC-SISO architecture was proposed for equispaced MIMO arrays. In addition, the proposed CC-SISO DPD architectures were designed to eliminate the requirement for signal feedback taps in the transmit paths which are indispensable in current multi-input MIMO DPD models. Therefore, the proposed CC-SISO DPD architectures reduce the overall DPD implementation complexity and the operational cost. Through simulation results using 5G signals, and actual PA configuration with memory effects, the high accuracy, low complexity and scalability of the proposed CC-SISO DPD architectures was demonstrated with state-of-art multi-input MIMO DPD models. Hence, it was corroborated that the proposed CC-SISO DPD architectures are highly advantageous, and scalable for massive MIMO applications. As future research, crosstalk estimation, and cancellation could be explored in higher order massive MIMO systems.
Chapter 6

Conclusion and Future Works

In this thesis, the non-orthogonality conditions and nonlinear distortions induced in multiple domains in 5G and beyond wireless networks are extensively focused. To achieve higher communication performance, multi-dimensional modulation techniques and nonlinear predistortion architectures are proposed.

In Chapter 3, the joint non-orthogonality conditions in spatial-time-frequency domains and the data rate deterioration due to such non-orthogonality conditions was focused. To achieve maximized communication data rate under joint non-orthogonality conditions in spatial, and time-frequency domains, a situation-aware multi-dimensional modulation (MDM) technique was proposed. The proposed MDM scheme was designed to take into account the transmitter-receiver pair specific non-orthogonality conditions that occur in those domains, and jointly optimize the spatial-time-frequency domain radio resource block attributes to minimize the non-orthogonality degree in each domain and thus achieve maximized data rate. Through simulation results, it was demonstrated that the proposed MDM achieves higher data rate than spatially multiplexed MIMO-OFDM systems under spatial, and time-frequency domain non-orthogonality conditions. As future work, additional radio resource domains such as power domain and code domain radio resources can be incorporated and optimized to enhance wireless communication performance.

In Chapter 4, the transmitter-receiver pair specific and domain specific non-orthogonality degrees induced in spatial, time-frequency and delay-doppler domains and the increased operational cost from receiver side to process such non-orthogonal radio resources was
focused. To minimize the overall operational cost from user-device while achieving required data rate under joint non-orthogonality degrees among radio resources in spatial, time-frequency and delay-doppler domains, a user-centric multi-dimensional modulation (UC-MDM) technique was proposed. The proposed cost-aware and situation-aware UC-MDM scheme is designed to jointly take into account the transmitter-receiver pair specific spatial, time-frequency, and delay-doppler domain non-orthogonality degrees perceived in the wireless channel by UE. Hence, the UC-MDM jointly optimizes the radio resource separation and utilizes the optimum radio resource combination with optimum separation in each domain through multi-dimensional modulation in spatial-time-frequency or spatial-delay-doppler domains to achieve the goal. Using simulation results, it was demonstrated that the proposed UC-MDM can achieve the required data rate with minimal operational cost from user-device compared to state-of-art MIMO-OFDM and MIMO-OTFS systems. As future work, the proposed UC-MDM can be expanded to multiple transmitter-receiver pairs scenario which encounter additional non-orthogonality due to overlap between the radiated beams and imperfect successive interference cancellation in power domain NOMA.

In Chapter 5, the increased severity of nonlinear distortions induced due to nonlinear and reverse crosstalk among MIMO antennas was focused. To estimate and cancel such crosstalk, a single-input cross-correlation based DPD architecture was proposed. The proposed CC-SISO DPD architecture achieves less complexity, and demonstrates high scalability than current multi-input DPD models for MIMO systems. Furthermore, the proposed CC-SISO DPD architecture also eliminates the requirement for signal feedback taps before the PA thus reducing the overall hardware implementation complexity. In addition, an optimized CC-SISO DPD which can re-utilize the estimated crosstalk coefficients for crosstalk cancellation was proposed for equispaced MIMO arrays. Through simulation results, the high accuracy, less complexity, and high scalability of the proposed architectures were validated and compared with state-of-art multi-input DPD models. As future work, the proposed CC-SISO DPD architectures can be extended to 2-D MIMO arrays and fully connected hybrid beamforming arrays for efficient crosstalk estimation and nonlinear distortion compensation.
Bibliography


[68] 3GPP TR 38.901, “5G Study on channel model for frequencies from 0.5 to 100 GHz”, pp. 70-76.


# Curriculum Vitae

<table>
<thead>
<tr>
<th><strong>Name:</strong></th>
<th>Thakshan Uthayakumar</th>
</tr>
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<tbody>
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<td><strong>Post-Secondary Education and Degrees:</strong></td>
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<td>London, Ontario, Canada</td>
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<td>2020-2022 M.E.Sc</td>
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<td></td>
<td>2014 - 2018 B.Sc</td>
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<tr>
<td><strong>Honours and Awards:</strong></td>
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<td></td>
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<tr>
<td><strong>Related Work Experience:</strong></td>
<td>Research Assistant</td>
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<tr>
<td></td>
<td>Engineer</td>
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<td>Dialog Axiata PLC, Sri Lanka</td>
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**Publications:**

