

UNIVERSITY OF BAHRAIN

College of Engineering



**Department of Electrical
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Millimeter Wave Beamforming for Wireless Communications

A Thesis Submitted in Partial Fulfillment of the Requirements for the degree of Master of Science in Electrical and Electronics Engineering

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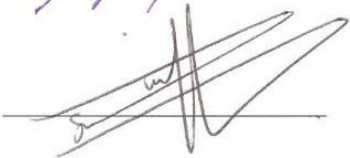
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ABSTRACT

Millimeter Wave (mmWave) wireless communications is considered an enabling technology to allow fifth generation (5G) to achieve data rates in the order of giga-bits-per-seconds for future mobile applications. The main objective of this thesis is to design a low complexity hybrid beamforming system taking into consideration the corresponding mmWave hardware constraints.

In this thesis, hybrid precoders and combiners are reviewed and newly modified implantations are proposed. Also, spatial characteristics of the channel model in multiple input multiple output (MIMO) fully-connected structure are studied to be used in system model. The design problem is presented as a sparse reconstruction and solved using Orthogonally Based Matching Pursuit (OBMP) algorithm. Maximum Likelihood (ML) fixed weight beamforming is used along with OBMP for direction of departure estimation. The hybrid precoder design is simplified by combining the design of ML beamforming, orthogonal RF codebook and the beamspace full digital precoder.

Performance of modified hybrid precoding-combining design is presented by evaluating spectral and energy efficiency. System performance is investigated by varying the following system parameters, number of data streams, radio frequency (RF) chains, RF codebook layers, signal to noise ratio (SNR) and power consumption values. Simulation results shows near optimal efficiency (above 90%) compared to unconstrained full digital optimal beamforming for higher SNR and RF chains. The results are compared with variable hardware complexity and it is shown that with the modified designs an efficient decrease of iterative process is achieved. Also, less matrix inversion calculation is needed compared to the state-of-art approach.

Moreover, the effect of using uniform planner array (UPA) antenna at the receiver is investigated. The thesis shows that beamspace acquisition with ML fixed weight beamforming is an optimum implementation to achieve similar to optimum mmWave full digital precoder.

Dedication

To my wife and our children. Their love is a gift from Allah that I am thankful every day.

Acknowledgments

First and Foremost praise, I thank my God “Allah” for providing me the power and conciliation to do this research. Special thanks to my family, my supervisor Dr. Mohab Mangoud for his continuous support and guidance. I also appreciate the encouragement of all academic and faculty stuff of the Electrical and Electronics Engineering Department at the University of Bahrain. I would also like to express my gratitude to my friend Hamad Al-Doseri for discussion.

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Notations

Table of Abbreviations

4G	4th Generation
5G	5th Generation
ADC	Analog to Digital Converter
AoA	Angle of Arrival
AoD	Angle of Departure
BS	Base Station
BSVD	Beamspace Singular Value Decomposition
CSI	Channel State Information
DFT	Discrete Fourier Transform
LOS	Line of Site
MIMO	Multiple Input Multiple Output
MMIMO	Massive Multiple Input Multiple Output
MmWave	Millimeter Wave
ML	Maximum Likelihood
MS	Mobile Station
NLOS	Non Line of Site
OMP	Orthogonal Matching Pursuit
OBMP	Orthogonally based Matching Pursuit
Rx	Receiver
RF	Radio Frequency
SNR	Signal to Noise Ratio
SVD	Singular Value Decomposition
Tx	Transmitter
ULA	Uniform Linear Array
UPA	Uniform Planner Array

Table of Mathematical Operators

$\mathbf{A}$	Matrix
$\mathbf{a}$	Vector
a	Scalar
$\mathcal{A}$	Set
$\ \mathbf{A}\ _F$	Frobenius norm of $\mathbf{A}$
$ \mathbf{A} $	Determinant of $\mathbf{A}$
$\text{trace}(\mathbf{A})$	Trace of $\mathbf{A}$
$\text{rank}(\mathbf{A})$	Rank of $\mathbf{A}$
$\mathbf{A}(:,l)$	l th column of $\mathbf{A}$
$\mathbb{E}[\mathbf{A}]$	Expectation of $\mathbf{A}$
$\text{diag}(\mathbf{A})$	Vector composed of diagonal elements of $\mathbf{A}$
$\mathbf{A}(u, j)$	Sub matrix of matrix $\mathbf{A}$ with index sets u and j
$\mathbf{A}^T$	Transpose
$\mathbf{A}^H$	Conjugate transpose
$\mathbf{A}^*$	Conjugate operation
$\mathbf{A}^{-1}$	Inverse of a matrix $\mathbf{A}$
$\mathcal{CN}(\mathbf{a}, \mathbf{A})$	Complex Gaussian vector with mean $\mathbf{a}$ and covariance matrix $\mathbf{A}$
$[\mathbf{A} \mathbf{B}]$	Horizontal concatenation
$\otimes$	Kronecker product
$\mathbf{a}(i)$	i th element of a vector $\mathbf{a}$
$ \mathcal{A} $	Cardinality of set $\mathcal{A}$
$\mathcal{O}(N)$	Order of N
$\mathbf{I}_N$	$N \times N$ identity matrix

Chapter 1

Introduction

- 1.1 MmWave Wireless Communications
- 1.2 Overview of Research
- 1.3 Contributions of This Work
- 1.4 Outline of Thesis

Chapter 1

Introduction

Wireless communications are facing challenges with new applications for future cellular systems, hence the need for higher data rates becomes a vital issue for the next fifth generation (5G) designs. Furthermore, the faster increase of mobile data usage motivates the need for updated wireless technology standards that overcome the bandwidth limitations. As introduced by Rappaport, et al. (2013), the next generation for communications systems will move the spectrum from copper to millimeter wave (mmWave) connectivity, allowing more cooperation between base stations (BS).

Massive multiple input multiple output (MMIMO) systems that contain hundreds of steerable directional antenna arrays at BS will overcome the rain and propagation losses experienced by common wireless technologies by allocating the transmitted energy to the mobile users. Another approach for the mmWave technology is the use of small cells that extend the distance between BS and mobile users which enhance the energy efficiency and data rate, as well as limits the propagation loss, these two challenging trends are the most important issues in mmWave systems. The small wavelength antenna arrays, which can be packed in small physical dimensions, and small cell technologies evolution will overcome the rain attenuation drawback. Extensive measurements for mmWave channel propagation were conducted by New York University research team at New York University campus and University of Texas at Austin campus with different frequency bands and different urban environments. Based on this work on channel modeling, low complexity transceiver

techniques are needed in mmWave system to achieve the desired design performance.

1.1 MmWave Wireless Communications

The use of huge amount of unlicensed spectrum especially at 28, 38 and 72 GHz bands is the main target for the communications industry to provide more common resources sharing for multiple users. The first mmWave system was conducted in 1895 by Jagadish Chandra Bose, who developed a project of electromagnetic waves transceiver system at 60 GHz and a distance of 23 meters. In the same year, a second mmWave pioneer, Pyotr N. Lebedew, came with his investigation of the propagation for 4 to 6 millimeter wavelength radio transmission. Nowadays, mmWave spectrum is currently occupied by radar, backhaul and military applications. As stated by Rappaport, Heath, Daniels, & Murdock (2014), mmWave wireless communications evolution will begin to be regulated all over the world. The governments of the USA, Europe, Korea, Japan and Australia have allocate a total of approximately 7 GHz spectrum for 60 GHz applications.

Although the mmWave cellular systems have large potential, here are some key limitations facing these band of frequencies:

- Penetration losses of the materials and surfaces, mainly diffraction and reflection, which as a result enables non line of site (NLOS) communications.
- Doppler spread for multi path channels which depends on mobility and carrier frequency, especially for the case of rich scattering environment with the use of omnidirectional steerable antennas.
- Coordination of multi transmission with adjusted cell interference nulling.

- Power consumption in the Analog to Digital converter (ADC) devices.
- Hardware complexity of the transceiver design unit in MIMO systems.

Applications of massive mmWave spectrum ranging from home to office including information showers, where the data can be transferred to people whenever they go, drive or walk. Moreover, vehicle to infrastructure and vehicle to vehicle applications which enable the collision avoidance between cars and traffic information exchange. Federal communications commission (FCC) regulates the use of 60 GHz spectrum for inter-satellite communications, which can be used for aerospace multimedia applications for aircrafts.

1.2 Overview of Research

Beamforming was presented by Pi & Khan (2011) as a key enabling technology for hybrid mmWave mobile broadband and fourth generation (4G) standard using spatial division multiple access and MIMO operation. Roh, et al. (2014) discussed the conducted channel measurements results and system prototype in both USA and Korea to prove the reliability of mmWave band for cellular usage. Rangan, Rappaport, & Erkip (2014) considered the capacity and channel measurements for small cell urban environments and proved that there was street level coverage for large transceiver distance, further, more capacity was achieved using statistical channel models based on these measurements compared to current state-of-art 4G cellular system. The practical hybrid beamforming algorithm design presented by El Ayach, Rajagopal, Abu-Surra, Pi, & Heath (2014) which showed that the design performance can approach the unconstrained beamforming case with the given hardware constraints. Alkhateeb, El

Ayach, Leus, & Heath (2014a) formulated channel estimation problem with hierarchical codebook hybrid precoding for single user mmWave system with comparison of efficiency and precoding gain for exhaustive search. Yu, Shen, Zhang, & Lataief (2016) compared the performance results of proposed alternating minimization algorithm for partially connected and fully connected hybrid precoding structures in terms of spectral and energy efficiency. Liu, Chen, Tsai, & Wu (2016) proposed a low complexity hybrid precoding algorithm based on multilevel discrete Fourier transform (DFT) codebook which eliminates iterative process and matrix inversion for calculating hybrid precoders. Chen, Tsai, Liu, & Wu (2017) presented a low complexity compressed sensing channel estimator algorithm based on DFT-based radio frequency (RF) beamforming codebook introduced by Hung, Chen, Liao, Tsai, & Wu (2015) and reduced-dimension beamspace singular value decomposition (BSVD) full digital precoder that saves the power for the BS and the mobile station (MS).

According to the above previous recent research topics, the extended work on beamforming techniques that consider the hardware complexity issue of mmWave MIMO system under certain conditions need to be discussed and investigated for the approximated maximum near-optimal performance will be presented in this thesis.

1.3 Contributions of This Work

The major contributions of this thesis are summarized as follows:

- Present a modified layered radio frequency (layered-RF) orthogonal beamforming codebook as a predefined candidate matrix for the hybrid precoding optimization problem. This candidate matrix consists of a conventional RF codebook matrix and a series of layered

RF matrices. Then with this layered-RF codebook matrix, it can be shown that the BS steers the signals during downlink transmissions in more quantized resolution beamforming directions with the considered hardware constraints.

- Present a maximum likelihood (ML) fixed weight beamforming for the hybrid design. This method is used to approximate the desired quantized angle of the antenna array response vector with the existing noise. The calculated optimum weights used to maximize the desired signal power and minimizing the power of noise signal.
- Modification of the BSVD algorithm for acquiring the full digital precoder. This algorithm is based on perfect channel state information (CSI) and corresponded to the transceiver array response vectors with lower matrix dimension which reduced the hardware design complexity and is achieved the desired acceptable performance.
- Modifications of the existing orthogonally based matching pursuit (OBMP) hybrid precoding algorithm based on the proposed layered-RF codebook, ML fixed weight beamforming and BSVD full digital precoder. The modification of this design targets the lower hardware complexity in terms of multiplication and addition processes while achieving near-optimal system performance compared to the state of art hybrid precoding algorithm. The performance will be numerically evaluated according to different system settings, including number of streams, number of RF chains, signal to noise ratio (SNR) ranges and total power consumption.

1.4 Outline of Thesis

This thesis is organized as follows:

In **Chapter 1**, introduction followed by mmWave wireless communications background, then overview of research, and contributions of the thesis are demonstrated.

In **Chapter 2**, a survey of mmWave MIMO system is presented. First, propagation and channel models including path loss, random blockage, spatial characteristics and Beamspace representation are outlined. Then, mmWave MIMO systems architectures with the comparison of analog, digital and hybrid architectures are discussed. After that, different hybrid precoding-combining techniques are discussed for single user and multi user MIMO systems with analog processing of hybrid beamforming based on phase shifters, switches and 1-bit ADC.

In **Chapter 3**, the system model and the corresponding channel model is presented followed by hybrid precoding/combining algorithm design. At first, the formulation of an optimization problem for hybrid precoding and combining algorithm is introduced and discussed. Then, the layered-RF codebook for uniform linear array (ULA) antenna response vectors for single user mmWave MIMO system is implemented. After that, the ML fixed weight beamforming is presented as the solution to estimating angles in the presence of noise. At the end of this chapter, the proposed lower complexity hybrid precoding-combining algorithm is designed for different system settings.

In **Chapter 4**, the system model with channel model for hybrid precoder and full digital minimum mean square error (MMSE) combiner will be presented for a single user mmWave MIMO system. First, the BSVD design is outlined. Then, with layered-RF codebook design for

ULA-UPA antenna response vectors, the process of BSVD ML-based hybrid precoding algorithm will be explained and derived. Finally, simulation results are presented with different system and beamforming algorithm parameters. The results of spectral and energy efficiency for the hybrid beamforming algorithm in terms of SNR, RF chains, RF codebook layers and number of streams, are investigated, studied and discussed for the verification of 5G cellular system.

In **Chapter 5**, the work is concluded with a summary of the obtained results and the recommended extensions of the future work.

Chapter 2

Millimeter wave MIMO systems

- 2.1 MmWave Propagation and Channel Models
- 2.2 MmWave MIMO Systems Architectures
- 2.3 Hybrid Precoding/Combining Techniques
- 2.4 Conclusions

Chapter 2

Millimeter Wave MIMO Systems

MmWave frequencies are considered as the frontier for wireless communications due to their higher bandwidth channels compared with current commercial microwave wireless systems. In this chapter, a study on mmWave propagation and channel models is presented in section 2.1 including free space path loss, random blockage, spatial channel characteristics and beamspace-based system representation. In section 2.2, an explanation of MIMO system architectures for mmWave channels is carried out. Analog, digital and hybrid beamforming with emphasis on analog processing of hybrid analog-digital beamforming is presented. Then, hybrid beamforming algorithms are discussed in section 2.3. The cases of single user and multi user with intra cell and inter-cell interference are analyzed. Finally, the conclusion of this chapter is presented in section 2.4.

2.1 Millimeter Wave Propagation and Channel Models

Propagation in mmWave frequencies is an important issue due to the small wavelength. This wavelength, which is smaller than one centimeter or even fingernail of human, is very sensitive to the obstacles over the distance between the transmitter (Tx) and receiver (Rx). Furthermore, atmosphere absorption in the air, and temperature naturally attenuate electromagnetic wave signal in high GHz bands.

2.1.1 Free Space Path Loss

In free space environments, when there is line of site (LOS) links, i.e. there are no obstacles between Tx and Rx, according to Rappaport (2002), the large scale wireless channel characteristics are modeled by the Friis' Law

$$P_r = \frac{P_t G_t G_r}{L} \left(\frac{\lambda}{4\pi d} \right)^2, \quad (2.1)$$

where P_r is received power as a function of the distance d between Tx and Rx, P_t is the transmitted power, G_t is the Tx antenna gain, G_r is the Rx antenna gain, L is the system propagation loss factor and λ is the wavelength given by c/f , where c is the speed of light in space (3×10^8 m/s) and f is the carrier frequency. From the powers ratio (P_t/P_r), the path loss increase is proportional to the Tx-Rx separation and inversely proportional to wavelength squared (λ^{-2}). Correspondingly, antenna gains scale as $G_t, G_r \propto \lambda^{-2}$, implies that more antennas need to be fitted in same physical area structure and this can be considered as an attractive factor for path loss reduction in mmWave communications. Path loss measurements for outdoor links were studied by Azar, et al. (2013) in New York City for mmWave pico-cellular networks.

2.1.2 Random Blockage and Obstacles

In addition to the path loss challenge, mmWave frequencies are very sensitive to blockage, for instance, brick material, foliage loss, humidity and rain fades. These kinds of obstacles cause reflection, diffraction and scattering. Figure 2.1 illustrates the beamforming technique from the sector antenna. The BS pointing will affects the coverage as the direction of antenna and placement are changed. Signal to interference and noise ratio

for rate analysis is important factor as it was studied by Bai & Heath (2015).

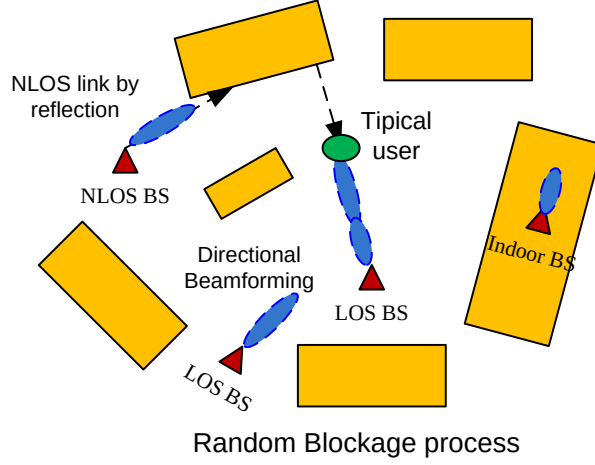


Figure 2.1 Illustration of LOS and NLOS

The buildings are the main source of reflection in NLOS links. The directional beamforming is the main established link in the LOS communication. Motivated by Andrews, et al. (2016) for analytically modeling the blockage in suburban areas, the LOS probability function (not a traditional probability density function) P_{LOS} given by

$$P_{LOS}(d) = e^{-\frac{d}{C}}, \quad (2.2)$$

where d is the positive distance and C can be computed using statistical modeling of the building that are distributed uniformly in the space using random shape theory as

$$C = \frac{\pi}{\lambda_{bldg} \mathbb{E}[L_{bldg}]}, \quad (2.3)$$

where λ_{bldg} stands for number of buildings in unit area and L_{bldg} is perimeter of one building. Using this model, coverage and rate analysis can be derived using stochastic geometry.

2.1.3 Spatial Channel Characteristics

For describing system capacity of mmWave channels, geometric channels with clusters are commonly used for system design. As described by Heath, Gonzalez-Prelcic, Rangan, Roh, & Sayeed (2016), the mmWave nature is non selective in time and frequency domains, however, the current research is in the area of frequency selective channels. The narrowband downlink spatial channel model has a form of

$$\mathbf{H} = \mathbf{A}_r \mathbf{H}_\alpha \mathbf{A}_t^*, \quad (2.4)$$

where $\mathbf{H}_\alpha$ is the channel complex gain matrix. $\mathbf{A}_r$ and $\mathbf{A}_t$ are antenna array response vectors for MS and BS, respectively. For 2D channel models, the antenna array response vectors as a function of departure or arrival plane wave angular direction (θ) can be represented as

$$\mathbf{A}_\zeta = [\mathbf{a}_\zeta(\theta_1^\zeta), \mathbf{a}_\zeta(\theta_2^\zeta), \dots, \mathbf{a}_\zeta(\theta_L^\zeta)], \text{ for } \zeta \in \{t, r\}, \quad (2.5)$$

where $\mathbf{a}_\zeta(\theta_L^\zeta)$ is the uniform linear array (ULA) steering vector of BS($\zeta = t$) and MS($\zeta = r$) of path L . In a 3D channel model, the uniform planar array (UPA) steering vector to the corresponding BS/MS is given by

$$\mathbf{a}(\theta, \phi) = \mathbf{a}_{azimuth}(\theta) \otimes \mathbf{a}_{elevation}(\phi). \quad (2.6)$$

The study of large scale path loss measurements for different orientations of angles of arrival (AoA) and angle of departure (AoD) (θ, ϕ) between Tx and Rx was investigated using power angular spectrum by Rappaport, MacCartney, Samimi, & Sun (2015). For multi path MIMO channels, the delay, power, AoA and AoD are the channel testing conditions for cluster distribution.

2.1.4 Beam-space-based channel Representation

In mmWave frequencies, the impact of high directional propagation along with massive MIMO antenna motivates the choice of beam-space-based spatial channel representation. The spatial Fourier transform is the common point of beam-space-based and antenna space. The beam-space-based system representation for 1D N -element ULA is given by

$$\mathbf{U} = \frac{1}{\sqrt{N}} [\mathbf{a}(\theta_0), \dots, \mathbf{a}(\theta_1), \dots, \mathbf{a}(\theta_{N-1})]^T, \quad (2.7)$$

where θ_i is defined as

$$\theta_i = \sin^{-1} \left(\frac{\lambda \delta_i}{d} \right). \quad (2.8)$$

The DFT matrix which results in a set of orthonormal set of matrices with uniformly distributed quantization spatial angles $\delta_i = \{0, 1/N, \dots, (N-1)/N\}$ has the following unitary property

$$\mathbf{U}^* \mathbf{U} = \mathbf{U} \mathbf{U}^* = \mathbf{I}. \quad (2.9)$$

Consider the beam-space-based channel representation

$$\mathbf{H} = \mathbf{U}_R \mathbf{H}_b \mathbf{U}_T^*, \quad (2.10)$$

where $\mathbf{U}_R$ and $\mathbf{U}_T$ are the orthonormal matrices at the transceiver and $\mathbf{H}_b$ is the equivalent beam-space channel. The beam-space-based channel representation derived by Heath, Gonzalez-Prelcic, Rangan, Roh, & Sayeed (2016) and Bajwa, Haupt, Sayeed, & Nowak, (2010) as

$$\mathbf{H} = \mathbf{A}_R(\theta_{R,i}) [\mathbf{H}_b]_{i,k} \mathbf{A}_T^*(\theta_{T,k}), \quad (2.11)$$

where $1 \leq i \leq N_r$ and $1 \leq k \leq N_t$. Virtual AoD and AoA corresponding to uniformly spaced normalized angles $\{\delta_{T,k}\}$ and $\{\delta_{R,i}\}$ are $\{\mathbf{A}_T(\theta_{T,k})\}$

and $\{\mathbf{A}_R(\theta_{R,i})\}$, respectively. Further, Brady & Sayeed (2014) and Sayeed & Behdad (2010) derived a 2D UPA beamspace-based system representation.

From the channel spatial structure point of view, where there is no selectivity in time or frequency domain ($\mathbf{H}(t, f) = \mathbf{H}$), the channel power can be denoted as

$$\sigma^2 = \text{trace}(\mathbf{H}^* \mathbf{H}) = \text{trace}(\mathbf{H}_b^* \mathbf{H}_b) = \sum_{l,m} |[\mathbf{H}_b]_{l,m}|^2, \quad (2.12)$$

then, the low dimensional communication subspace that corresponds to channel realization can be captured by the singular value decomposition (SVD) of the channel, for a given value of threshold η , i.e 0.8 to 0.9, the channel power captured by singular vectors of the channel to achieve optimal communication is given by

$$\sum_{i=1}^{p_{eff}} \sigma_i^2(\mathbf{H}) \geq \eta \sigma_c^2, \quad (2.13)$$

where p_{eff} is the effective channel rank. Brady & Sayeed (2014), Sayeed & Behdad (2010), Song, Brady, & Sayeed (2013) and Raghavan & Sayeed (2011) formulated an approximated low dimensional communication subspace singular vectors for the spatial signal space.

Consider the extended beamspace-based representation for narrowband channel as an alternative representation to the array geometry, by recalling equation (2.4), Heath, Gonzalez-Prelcic, Rangan, Roh, & Sayeed (2016) defines new array response matrices in the uniform grid δ , as a result, the derived approximated channel matrix has a form of

$$\mathbf{H} = \bar{\mathbf{A}}_R \tilde{\mathbf{H}}_b \bar{\mathbf{A}}_T^*. \quad (2.14)$$

This equation implies that the channel has non-zero elements in the corresponding AoAs and AoDs that are taken from a uniform grid.

However, there will be some grid error between the uniform grid that can be neglected by increasing the size of the grid.

2.2 MmWave MIMO Systems Architectures

As introduced by Kim & Lee (2015), MIMO technology has been standardized for commercial use for current WLAN, i.e. IEEE 802.11n-ac, and cellular (LTE and LTE advanced) at below 6GHz frequency bands. The maximum antenna elements supported for these standards are up to 8. However, due to smaller wavelength in mmWave systems, the larger antenna elements (up to 256 is common) are occupied in small physical size (Heath, Gonzalez-Prelcic, Rangan, Roh, & Sayeed, 2016). In terms of system architecture, all signal processing of MIMO communications at lower frequencies are processed in the baseband, as shown in Figure 2.2, where N_s is the number of data streams, N_{RF} is the number of RF chains and N_t/N_r denotes the BS/MS antennas.

At mmWave frequencies, however, due to hardware constraints, it is unpractical to have dedicated RF chain for each antenna.

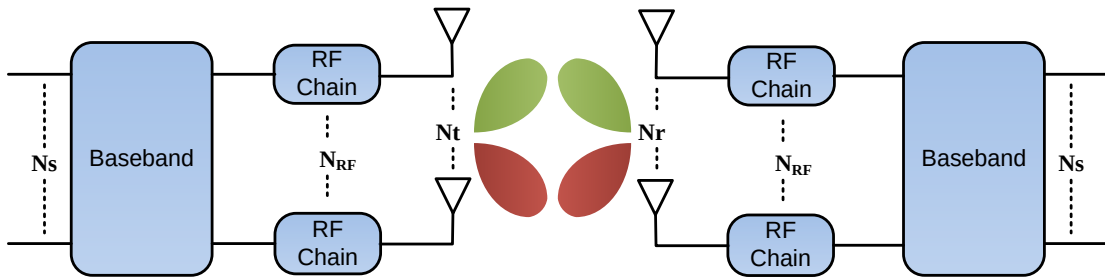


Figure 2.2 Traditional MIMO architecture at sub-6 GHz frequencies

2.2.1 Analog Beamforming

In mmWave channels, analog beamforming is the easiest approach for transceivers operations. It is considered recently as the adopted solution

supported in IEEE 802.11ad WiGig (sub 60-GHz band) standard. As illustrated in Figure 2.3, a single RF chain is connected to a dedicated antenna through phase shifter. This phased network is controlled electronically to form a beam steering of the processed signal, which is maximizing the signal power using signal processing techniques. The adjustment of signal amplitude and quantization of phase shifters degrade and limit the system performance.

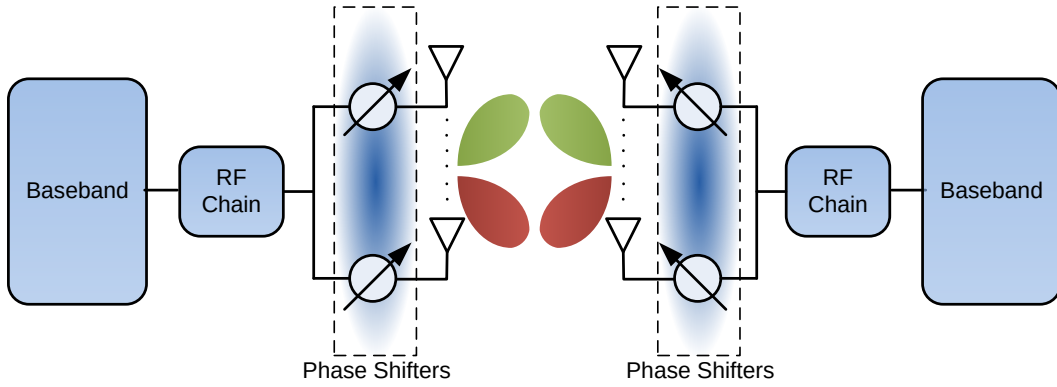


Figure 2.3 MmWave MIMO system using analog beamforming architecture

Analog beamforming supports only single user and single stream transmission. Moreover, multi training algorithms are used for steering the beams of the signal in the direction of arrival. El Ayach, Heath, Abu-Surra, Rajagopal, & Pi (2012b) presented the results for optimality of beamsteering for both single user and multi stream transmissions in realistic spatial channel model.

Hur, et al. (2013) Proposed a beam alignment technique using hierarchical codebook for point to point outdoor mmWave communications using single analog to digital converter as it practically requires low hardware cost. The concept of hierarchical codebook will be included in the next section.

2.2.2 Digital Beamforming

Digital beamforming for conventional single antenna system compared to multiple antenna system with transmit and receive antennas is done by applying several techniques to coordinating the signal detection on the receiver side and cancelling interference at BS side through different transmission method, i.e. channel inversion, block diagonalization and zero forcing based on SVD of the channel matrix as discussed by Cho, Kim, Yang, & Kang (2010). However, mmWave communication is different from lower frequency solutions as precoding/combining cannot be performed at digital baseband due to higher cost and power consumption of components of ADCs and RF chains.

2.2.3 Hybrid Beamforming

Hybrid beamforming supports multi user and multi stream transmission and it is widely used in mmWave communications. In Hybrid precoding systems, users experience enhanced performance compared to analog designs by maximizing desired signal power and eliminating interference at the transmitter.

The system architecture of hybrid analog/digital precoding and combining is shown in Figure 2.4. For the hybrid precoding and combining case these conditions hold: $N_s < N_{RF}^t < N_t$ and $N_r > N_{RF}^r > N_s$ for $N_s > 1$. Analog beamforming case can be predicted by considering the special case of $N_s = N_{RF}^t = N_{RF}^r = 1$, where N_{RF}^t and N_{RF}^r are the number of RF chains at BS and MS, respectively.

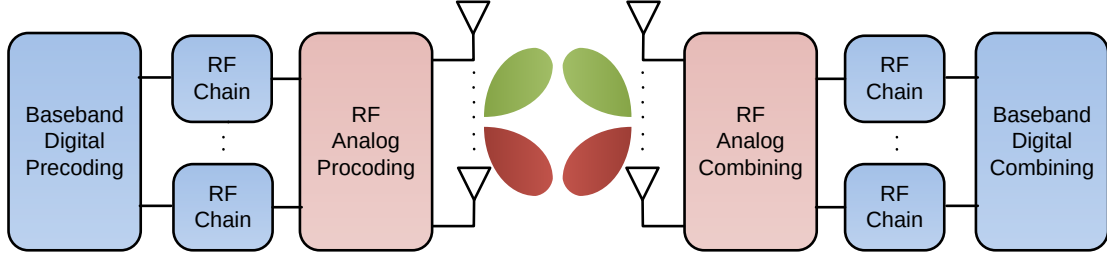


Figure 2.4 Hybrid analog/digital precoding and combining mmWave MIMO architecture

The conventional full digital precoding is not preferred in mmWave systems since it requires a dedicated RF chain for every antenna which requires large installation hardware cost and energy consumption. To overcome such hardware limitations, El Ayach, Rajagopal, Abu-Surra, Pi, & Heath (2014) formulated the hybrid precoder/combiner problem as a sparse reconstruction problem for large antenna arrays using the basis pursuit algorithm. Hybrid precoding is a combination between digital precoding and analog RF beamforming.

The alternative design of traditional digital precoding is to use a large number of analog beamformers which are composed of phase shifters to maximize the antenna gain. Alternatively, another approach of full digital precoder design is to use limited number of digital precoders with a small number of RF chains to achieve minimum interference rate.

The received signal during downlink transmission before combining is given by

$$\mathbf{y} = \mathbf{H}\mathbf{P}_{RF}\mathbf{P}_{BB}\mathbf{s} + \mathbf{n}, \quad (2.15)$$

where $\mathbf{P}_{RF}$ is RF precoder, $\mathbf{P}_{BB}$ is baseband digital precoder, $\mathbf{s}$ is transmitted signal, and $\mathbf{n}$ denotes the noise. The challenge in designing the hybrid precoders is to balance the tradeoff between hardware cost/energy

consumption and system performance as discussed by Gao, Dai, & Sayeed (2016).

One of the benefits from hybrid precoding is the reduced energy consumption thanks to the limited RF chains. The mmWave channel is a low rank-based, hence for $N_{RF} \gg \text{rank}(\mathbf{H})$, the small-sized digital-based hybrid precoder can achieve near-optimal system performance with limited RF chains compared to full digital precoder with a large number of RF chains. This motivates the design of hybrid precoders to meet the targeted maximum achievable sum rate. The hybrid precoder matrix can be reformulated as

$$\mathbf{P} = \mathbf{P}_{RF} \mathbf{P}_{BB}. \quad (2.16)$$

The hybrid combiner matrix can be written in same manner.

Alkhateeb, Leus, & Heath (2015b) extended the hybrid precoder design to two stage multi user scenarios. In the first stage the desired user power is maximized through searching of optimal beamformers from predefined codebook. In second stage, the zero forcing digital precoder is implemented to cancel multi user interference. Gao, et al. (2016) propose a successive interference cancellation for hybrid precoding design. This algorithm decomposes sum rate problem into simple and convex sub-problems.

2.2.4 Analog Processing for Hybrid Beamforming

This stage in hybrid precoding-combining can be implemented using phase shifters, switches or lenses. Two different possible categories of structures can be considered, fully-connected and partially-connected. For the fully connected structure, each RF chain signal output is connected to

all antennas through phase shifters, while for partially connected structure, each RF chain is connected to N_t/N_{RF}^t antennas.

In terms of hardware complexity, the fully-connected structure consumes full beamforming gain with high installation connection cost larger than partially-connected structure where the array contains sub-arrays which reduce the connection complexity but with limited overall array flexibility.

The energy and spectrum efficiency performance for the two structures was analyzed by Han, I, & Rowell (2015) in large scale antenna systems. Yu, Shen, Zhang, & Lataief (2016) developed semi-definite relaxation alternating minimization algorithm and showed that it outperforms the analog beamforming for higher SNR. Accordingly, analog RF precoders P_{RF} can be represented as

$$P_{RF} = \begin{bmatrix} P_1 & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & P_{N_{RF}^t} \end{bmatrix}, \quad (2.17)$$

with $i = 1, 2, \dots, N_{RF}^t$, P_i formulated as

$$P_i = \begin{bmatrix} e^{j\theta \left[(i-1)\frac{N_t}{N_{RF}^t} + 1 \right]}, \dots, e^{j\theta \left[i\frac{N_t}{N_{RF}^t} \right]} \end{bmatrix}, \quad (2.18)$$

where θ_i denotes the phase of the i th phase shifter.

As a comparison between the two discussed structures, the corresponding number of RF phase shifters N_{PS} can be expressed as

$$N_{PS} = \begin{cases} N_t N_{RF}^t & \text{for fully conneceted} \\ N_t & \text{for partially conneceted} \end{cases} \cdot \quad (2.19)$$

Zhang, Huang, Dtadyuk, & Guo (2015) exploit the structure of partially-connected in hybrid precoding for massive MIMO system.

As shown in Figure 2.5, each single RF chain is connected to N_t set of phase shifters in the fully connected structure, this phase shifters are quantized digitally and can be used to cancel multi stream interference in the hybrid design.

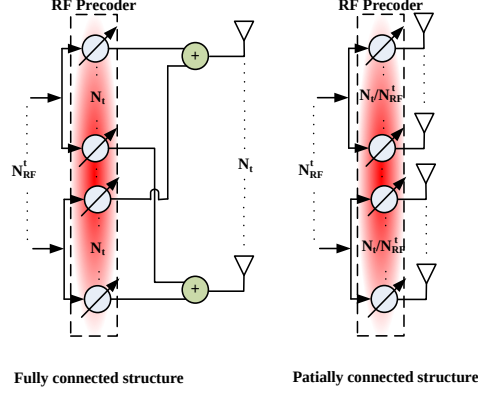


Figure 2.5 Two mapping structures for analog processing of hybrid precoding using phase shifters

Another approach to mmWave hybrid architecture is based on switches to reduce the hardware complexity of the phase shifters-based architecture in terms of power consumption as shown in section 2.3.3.2. Figure 2.6 shows the structure of switching-based hybrid precoding which exploits the sparse nature of mmWave channel.

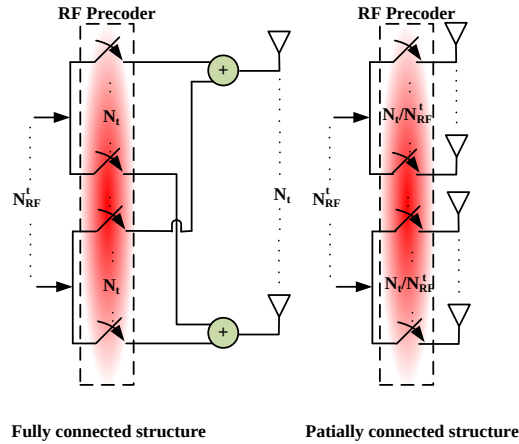


Figure 2.6 MmWave precoding based on switches

Analog processing can be implemented by installing lens antenna. This lens allows direct channel access using continuous aperture phased MIMO transceiver, as illustrated in Figure 2.7. This architecture supposed to further reduces hardware complexity compared to the previously discussed structures. The mmWave beam sector and continuous aperture phased shown in Figure 2.7 will take place of antennas and precoder/combiner unit in Figure 2.4.

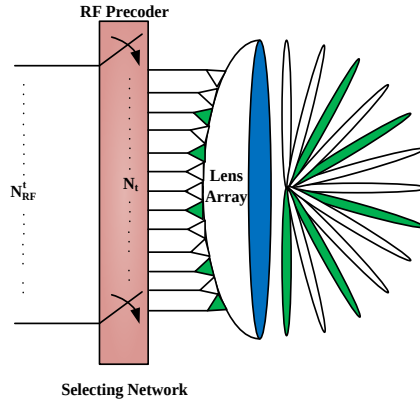


Figure 2.7 Lens-based front-end beamspace analog processing for hybrid beamforming

Brady, Behdad, & Sayeed (2013) modeled and prototyped LOS continuous aperture phased MIMO for exploiting the nature of propagation at mmWave frequencies through beamspace representation discussed in section 2.1.4. Two scenarios were considered: point to point and point to multi point using discrete lens array analog beamforming with achievable near-optimal performance and lower transceiver complexity.

2.2.5 Limited Bits Receivers

The third alternative method instead of using analog or hybrid architecture at the receiver is to reduce the resolution and power consumption of the ADC to a few bits. Figure 2.8 illustrates the one-bit

receiver architecture at mmWave frequencies. The sampling of in-phase and quadrature components of the received signal at the output of RF chain is done through ADCs.

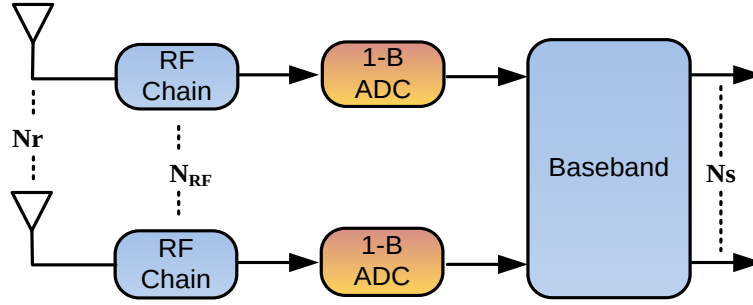


Figure 2.8 Illustration of one-bit ADC receiver at mmWave communication system

Although having the fewer bits of ADC will reduce the power consumed by the front end in the MIMO combiner design, it also reduces the power consumption of baseband circuit. The capacity gap between 1-bit ADC and high resolution ADC is 1.96dB at lower SNR (Heath, Gonzalez-Prelcic, Rangan, Roh, & Sayeed, 2016). Fettweis (2014) conducted the experiment for mmWave system prototype (60 GHz) at Dresden 5G lab studies the power consumption of ADC circuits in terms of time resolution case, he figures out the estimated power of 10mW at a sampling rate of 240 Giga resolution per second and he came with a conclusion of using oversampling to improve design performance.

2.3 Hybrid Precoding/Combining Techniques

Hybrid analog/digital precoding and combining at mmWave is essential and has received recently a huge attention from both academia and industry as it is considered an optimal solution to full dimensional array in mmWave massive MIMO systems as a result of cost reduction compared

to the analog (digital) beamforming. However, based on the different architectures discussed in the last section, many parameters need to be analyzed and tested for mmWave system design using such architecture-based algorithm(s). The channel at the receiver can be investigated through analog beamforming which leads to channel estimation problem as it is topic of current researches. Moreover, the sparsity channel structure can be modeled by signal processing-based techniques algorithms. In the next subsections, a quick review of beam training algorithms which do not require a full knowledge of the mmWave channel, followed by a descriptive survey of different hybrid precoding/combining techniques and scenarios for mmWave narrowband channel.

2.3.1 Beam Training Protocols

Analog beamformers vectors in mmWave are mostly designed using codebook which consists of beam patterns with different resolution levels. This requires iterative processes between Tx and Rx to estimate the angular directions for the best strongest signals of AoAs and AoDs without considering the channel estimation. The codebook beam training strategies which direct the beam in the desired direction have been implemented in IEEE 802.15.3c, IEEE 802.11ad and WirelessHD for analog only beamforming with single stream transmission.

2.3.1.1 Hierarchal Codebook Structure

Hierarchal codebook is a generalization of beam training for hybrid precoding based on RF phase shifters. This algorithm is used to estimate the channel path coefficients, i.e. AoA, AoD and path gain using adaptive measurements from multi resolution dictionary. Multi resolution beamforming codebook was developed by Alkhateeb, El Ayach, Leus, & Heath (2014a), Alkhateeb, El Ayach, Leus, & Heath (2014b), Eltayeb,

Alkhateeb, Heath , & Al-Naffouri (2015) , Xiao, Xia, & Xia (2016) and Cheng, Liao, & Wu (2016). The main contribution was to approximate the ideal sectorized beam pattern based on hybrid precoders. Figure 2.9 shows the structure of multi resolution codebook consists of S levels and N resolution with each level contains a set of beamforming vectors used to jointly design the hybrid precoder and combiner.

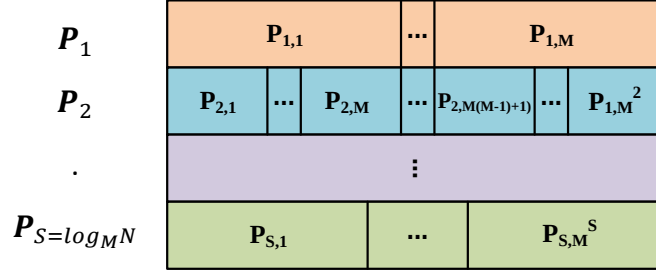


Figure 2.9 Hierarchical codebook structure

2.3.2 Hybrid Precoding

As discussed before, the main target of hybrid precoding design is to have a balance between hardware complexity and system performance in terms of power consumption and energy efficiency. The precoding/combining processes happen in analog and digital domains with the number of RF chains required to be not more than the total number of antennas. The received signal with discrete time model for a single symbol is given by

$$y = \mathbf{C}^* \mathbf{H} \mathbf{P} s + \mathbf{C}^* \mathbf{n}, \quad (2.20)$$

where $\mathbf{P}$ is the hybrid precoder composed of RF and digital precoder and $\mathbf{C}$ is the hybrid combiner.

2.3.3 Single User Hybrid Beamforming

2.3.3.1 Phase Shifters

The case scenario of single user mmWave MIMO system is shown in Figure 2.10 where a MS is connected with BS in single cell, assuming no interference from the desired user cell and from other users on the other cells in the same network.

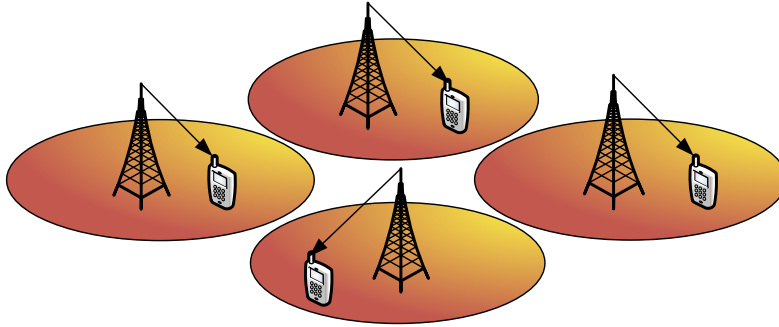


Figure 2.10 A mmWave MIMO system where each BS serves only one MS in each independent cell

In this system setting, a network of phase shifters can be used to approximate analog beamformers whereas the digital hardware is done at baseband. Alkhateeb, El Ayach, Leus, & Heath (2013) developed an iterative downlink channel training algorithm that accounts for hardware limitations assuming partial channel knowledge at BS and MS. This algorithm exploits the sparse nature of the mmWave channel and provides approximated approaches to digital beamforming rates assuming Tx (Rx) knows only its AoDs (AoAs). The scaled complex path gain was approximated by Alkhateeb, El Ayach, Leus, & Heath (2013) using a linear least squares estimator implemented at MS's hybrid combiner as

$$\tilde{\alpha} = (\mathbf{A}_r^* \mathbf{A}_r)^{-1} \mathbf{A}_r^* \mathbf{y}. \quad (2.21)$$

Then the hybrid combiners were represented as

$$\mathbf{C}_{RF} = \mathbf{A}_r \quad (2.22)$$

and

$$\mathbf{C}_{BB} = (\mathbf{A}_r^* \mathbf{A}_r)^{-1}. \quad (2.23)$$

The most common approach for designing hybrid precoding is maximizing the rate expression over all possible solutions of hybrid beamformers, that is

$$I(\mathbf{P}_{RF}, \mathbf{P}_{BB}, \mathbf{C}_{RF}, \mathbf{C}_{BB}) = \log|\mathbf{I} + \mathbf{R}_n^{-1} \mathbf{C}^* \mathbf{H} \mathbf{P} \mathbf{P}^* \mathbf{H}^* \mathbf{C}|, \quad (2.24)$$

where $\mathbf{R}_n$ is the noise covariance matrix. Another optimization approach of hybrid precoders is to approximate the optimal precoders given by SVD definition, i.e. minimizing the Frobenius norm between the optimal and approximated solutions, that is

$$(\mathbf{P}_{RF}^{opt}, \mathbf{P}_{BB}^{opt}) = \arg \min \left[\left\| (\mathbf{P}_{opt} - \mathbf{P}_{BB} \mathbf{P}_{RF}) \right\|_F \right]. \quad (2.25)$$

This problem statement is subject to RF hardware constraints that belongs to a set of feasible RF precoders that corresponds to the phase shifters hardware architecture. The orthogonal matching pursuit (OMP) algorithm was proposed by El Ayach, Rajagopal, Abu-Surra, Pi, & Heath (2014) for configuring RF beamforming vectors from a dictionary of RF steering vectors based on AoDs. The solution was very close to optimal precoders with preferable gain over analog only beamsteering. Figure 2.11 shows the spectral efficiency comparison between analog, digital and hybrid beamforming using UPA at BS and MS assuming 4 data streams, 64 antennas at BS and 16 antennas at MS.

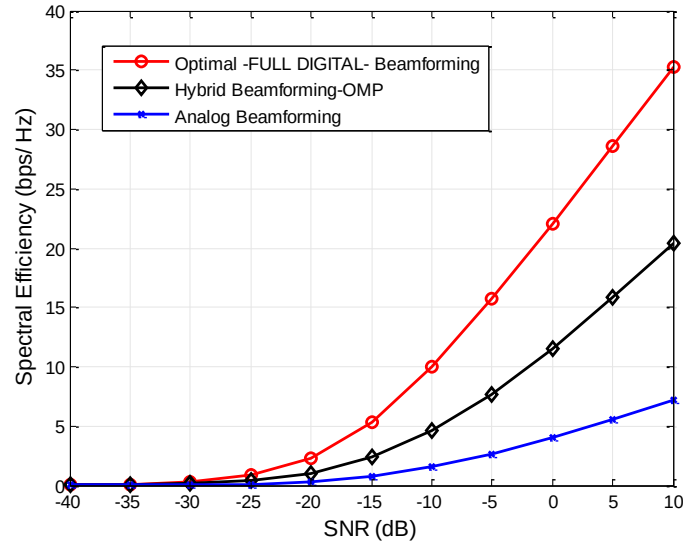


Figure 2.11 Spectral efficiency achieved by analog, digital and hybrid beamforming using OMP algorithm

An Extended work presented by Mendez-Rial, Rusu, Gonzalez-Prelcic, & Heath (2015) with greedy-based OMP which avoids the assumptions on the array geometry. Singh, & Ramakrishna (2015) applies the dictionary-based approach to an architecture based on partially-connected structure. A novel dynamic dictionary learning OMP developed by Donno, Beltran, & Widmer (2016) to synthesize the beam pattern using 2 bits RF phase shifters. Rahman, & Josiam (2014) discussed results of two classes of linear algorithm and codebook based approach for better selection of hybrid beamformers. Rusu, Mendez-Rial, Gonzalea-Prelcic, & Heath (2015) presented a novel low complexity hybrid sparse algorithm based on orthogonal matching and local search that exploits sparsity in the received signal, with emphasis on the goal of reducing the hardware limitations (reduce number of RF chains), i.e. reducing the search complexity of the algorithm. Hung, Chen, Liao, Tsai, & Wu (2015) proposed orthogonal beamforming candidate matrix for the orthogonally based matching pursuit to omit the pseudo-inverse operations. Figure 2.12

shows four beam patterns for 16-elements ULA antenna phase shifters steered by four orthogonal beamforming codes of DFT codebook matrix.

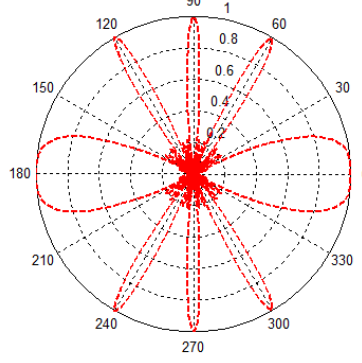


Figure 2.12 ULA antenna beam pattern steered by DFT orthogonal beamforming codebook

Sohrabi & Yu (2015) considered two cases where N_{RF}^t and $N_{RF}^r \geq N_s$, the system performance was close to optimal scheme and the second case when N_{RF}^t and $N_{RF}^r = N_s$. A simple search for every column in analog precoders was predicted by Chen (2015), however, more design computational complexity was reported.

2.3.3.2 Switches

Switch-based structure for hybrid combiners at receiver was investigated and reported by Mendez-Rial, Alkhateeb, Gonzalez-Prelcic, & Heath (2015). The columns of analog beamforming matrices $\mathbf{P}_{RF}$ and $\mathbf{C}_{RF}$ are single ones or zeros. This switching-based architecture comes with the following benefits compared with phase shifters:

- It is a low complexity solution with reduction of 45-70% of power consumption compared to phase shifters.

- Lower array gain than phase shifters, but with lower spectral efficiency.
- The reconstruction of channel estimation approach at the receiver performing well and better than the case of phase shifters.

The power consumption for the switches P^{sw} and phase shifters P^{ps} given as

$$P^{sw} = (13N_{sw}^r + 10)P_{ref} \quad (2.26)$$

and

$$P^{ps} = (N_{ps}^r(N_r + 12) + N_r + 10)P_{ref}, \quad (2.27)$$

where N_{sw}^r is number of switches at the receiver, N_{ps}^r is the number of phase shifters at the receiver and P_{ref} denotes the single power value. Chiang, Kadur, Rave, & Fettweis (2016) extended the work done by Mendez-Rial, Alkhateeb, Gonzalez-Prelcic, & Heath (2015) for estimating the channel by OMP at the receiver, for which use it to calculate the approximated SVD of the estimated channel at the receiver.

2.3.3.3 1-Bit ADCs

Considering the previous work done by Mo & Heath (2015), where the channel state information is assumed to be available at the transmitter, the new simple channel inversion results in nearly optimal solution for the full rank-based channel. It was investigated by this work that the MIMO precoding can eliminate the difference between the quantized and unquantized achievable rate at low or medium SNR. In addition, it was stated that the design of 1-bit ADC resolution is simple as it requires only a comparator and can be implemented with lower power consumption circuitry. In this related work, it was shown that the capacity of mmWave

channel is limited by number of propagation paths. The infinite SNR capacity satisfies the following bound

$$\log_2(K(N_r + 1)) \leq R_{mmW}^{1-bit} \leq \log_2(K(N_r + 1) + 1), \quad (2.28)$$

where K is any arbitrary value.

2.3.4 Multiuser Hybrid Beamforming

The system scenario for multi user mmWave system is illustrated in Figure 2.13. A single BS is assumed to be connected to single user MS_k taking into account the interference from other users in the same cell.

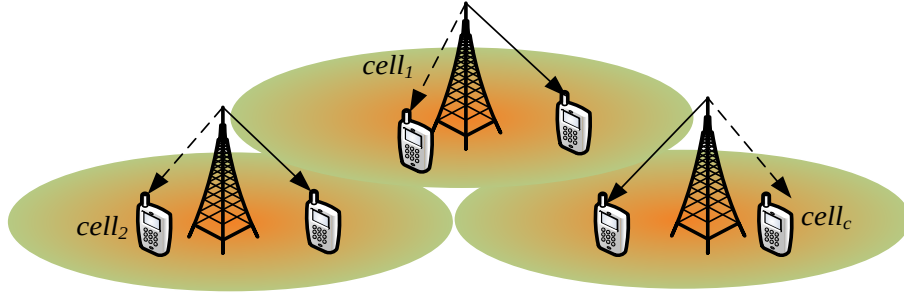


Figure 2.13 Illustration of mmWave multi user MIMO system downlink transmission with intra cell interference

The algorithm that configures hybrid precoders at Tx and analog-only combiners at multiple receivers is considered by Alkhateeb, Leus, & Heath (2015b). This algorithm was considered for full dimensional array and single path channels and designed to reduce training and feedback overhead. Two assumptions were presented in that work:

- BS communicates with MS via one stream transmission, thus the total number of K streams.
- Maximum number of served users simultaneously by BS not more than BS-RF chains ($K \leq N_{RF}^t$).

The processed received signal at k^{th} MS after combining is given by

$$y_k = \mathbf{c}_k^* \mathbf{H}_k \sum_{n=1}^K \mathbf{P}_{RF} \mathbf{p}_n^{BB} s_n + \mathbf{c}_k^* \mathbf{n}_k, \quad (2.29)$$

where $\mathbf{c}_k$ is the k^{th} MS RF combiner, $\mathbf{p}_k^{BB}$ is the k^{th} MS baseband precoder and $\mathbf{H}_k$ denotes the mmWave channel between BS and k^{th} MS.

The achievable rate for user k is

$$R_k = \log_2 \left(1 + \frac{\frac{P}{K} |\mathbf{c}_k^* \mathbf{H}_k \mathbf{P}_{RF} \mathbf{p}_k^{BB}|^2}{\frac{P}{K} \sum_{n \neq k} |\mathbf{c}_k^* \mathbf{H}_k \mathbf{P}_{RF} \mathbf{p}_n^{BB}|^2 + \sigma^2} \right), \quad (2.30)$$

where P is the average total transmitted power.

Then the system sum rate is given as

$$R_s = \sum_{n=1}^K R_k. \quad (2.31)$$

The main design objective is then to design hybrid precoders at BS and analog combiners at MSs to maximize the sum rate system capacity. The precoding design problem can be formulated as

$$\{\mathbf{P}_{RF}^*, \{\mathbf{p}_k^*\}_{k=1}^K, \{\mathbf{c}_k^{*BB}\}_{k=1}^K\} = \arg \max(R_s), \quad (2.32)$$

which could be considered as mixed integer programming problem.

The two-stage hybrid precoding algorithm was designed by Alkhateeb, Leus, & Heath (2015b) to achieve approximated optimal performance compared to fully digital zero forcing or block diagonalization approaches. In the first stage, a single user beam training is used to maximize the desired user power, while in the second stage inter-user interference was reduced by designing the baseband digital precoder from the estimated channel at the user's side. By implementing the limited feedback strategy, it was shown that the performance of full digital precoder is better than analog only beamsteering strategies which could be

improved by quantization. Figure 2.14 shows the spectral efficiency of various precoding algorithms for single user and multi user MIMO in mmWave channels using 64×16 UPA antennas.

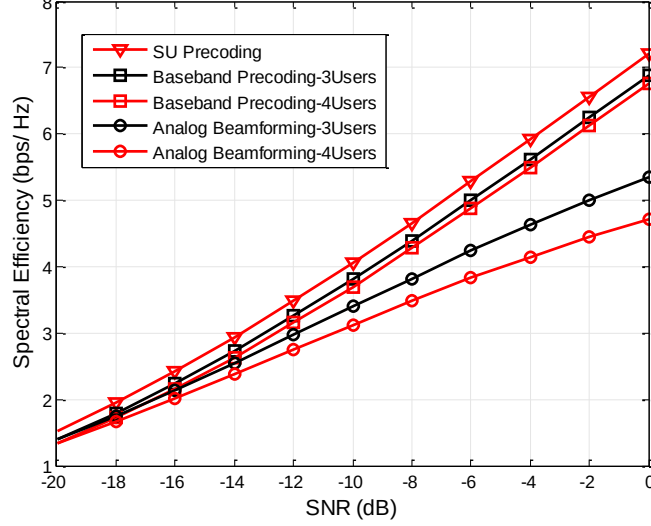


Figure 2.14 Achievable rates for different number of users with perfect channel knowledge

Multi-layer precoding was considered a potential solution to the full-dimensional massive MIMO systems by Alkhateeb, Leus, & Heath (2014) where inter-cell interference management was done by leveraging the directional characteristics of large-scale MIMO channels. The system assumes there was L cells and one BS communicates with K MS in each L cell as shown in Figure 2.15.

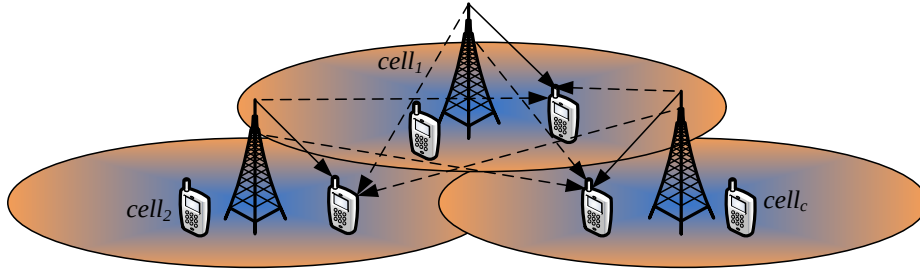


Figure 2.15 Illustration of mmWave multi user MIMO system downlink transmission with intra and inter cell interference

In that work, precoding matrix is further divided into three sub matrices (layers) where each layer is designed for one precoding objective, i.e. maximizing desired user signal power, or eliminating the inter-cell interference or eliminating multi-user interference. The 3-layer precoding matrix in cell c constructed as

$$\mathbf{P}_c = \mathbf{P}_c^{(1)} \mathbf{P}_c^{(2)} \mathbf{P}_c^{(3)}. \quad (2.33)$$

The achievable rate of user k in cell c is given by

$$R_{ck} = \log_2 \left(1 + \frac{SNR}{\left(\mathbf{C}_c^* \mathbf{P}_c^{(2)*} \mathbf{P}_c^{(2)} \mathbf{P}_c^{(2)*} \mathbf{P}_c^{(2)} \mathbf{C}_c \right)^{-1}} \right), \quad (2.34)$$

where $\mathbf{C}_c$ contains the covariance matrix of user k in cell c at BS b in azimuth and elevation direction.

Further, work was presented by Kim & Lee (2015) of hybrid MIMO processor for mmWave communications over multi user interference channels via OMP and MMSE-based hybrid beamforming. The RF beamformers are selected from six sets of candidate matrices in the least square operation using various beamforming approaches. The RF beamformers were designed at the baseband precoder in the proposed algorithm in which it outperforms the designs of baseband precoder after RF beamsteering.

2.4 Conclusion

This chapter starts with a general background of the propagation channel models of mmWave communications starting with free space path loss model and random blockage model. After that, Analog and digital beamforming architectures were presented followed by hybrid

analog/digital beamforming with its analog processing architectures including phase shifters, switches, lenses and 1-bit ADC.

A brief survey on hybrid precoding/combining algorithm was discussed for single user MIMO systems based on phase shifters, switches and 1-bit ADC in section 2.3. A quick explanation of beam training protocol in current wireless standards was outlined and the design of hierarchical codebook structure was presented and investigated. The presentation of optimal optimization design problem was presented for optimal precoders and combiners, based on maximizing system rates or approximating optimal full digital beamforming. Then, a review of past work on OMP algorithm was discussed for comparing the achievable rates for analog, optimal full-digital and hybrid beamforming.

Finally, the system scenario of multi user MIMO system in mmWave channels was illustrated with its achievable sum rate and precoder design optimization problem. Furthermore, a detailed review was conducted for multilayer precoding that accounts for the inter cell interference in mmWave network.

Chapter 3

Hybrid Precoding/Combining with Maximum Likelihood Fixed Weight Beamforming

3.1 System Model

3.2 Channel Model

3.3 Hybrid Precoding/Combining Design

3.4 Simulation Results and Performance Evaluation

3.5 Conclusion

Chapter 3

Hybrid Precoding/Combining with ML Fixed Weight Beamforming

MmWave communication systems can support high data rate at large unlicensed spectrum for the next wireless generation. The small wavelength in mmWave enables tens of hundreds of antennas to be packed at the transceivers for massive MIMO system construction. To deal with RF hardware complexity issues, El Ayach, Heath, Abu-Sara, Rajagopal, & Pi (2012a) developed a precoding algorithm that uses the principle of basis pursuits to approximate the unconstrained optimal precoders at low dimensional basis representation. This method exploits the sparsity nature of the mmWave propagation environment and limits the number of RF chains to be less than the number of scattering paths of the channel. However, the challenge in hybrid beamforming systems is to jointly design RF/baseband precoding/combining with reduced computational complexity in terms of matrix inversion and iterative selection of beamforming basis in traditional OMP. Lee, Wang, & Huang (2015) proposed an advanced method to reduce this effort by updating the least squares solution and parallel selection of basis vectors with acceptable system performance.

In this chapter, a modified hybrid precoding/combining algorithm is developed via ML fixed weight beamforming based on the predefined orthogonal beamforming codebook and the mapping process proposed by Liu, Chen, Tsai, & Wu (2016). Section 3.1 covers the hybrid precoding/combining system model and spectral efficiency formulation. Then, in section 3.2, mmWave channel model is expressed based on the

well-known Saleh-Valenzuela model. Section 3.3 presents the design of hybrid precoding/combining, including hybrid precoder problem formulation and the proposed hybrid precoding method based on layered-RF codebook design. Finally, the simulation results and performance analysis are discussed in section 3.4.

3.1 System Model

Consider the single user mmWave system shown in Figure 3.1, where N_s data streams are sent from the BS with N_t transmit antennas to a single MS with N_r receive antennas. The number of RF chains at Tx and Rx are denoted as N_{RF}^t and N_{RF}^r , respectively. The RF chains are bounded by $N_s \leq N_{RF}^t \leq N_t$ at BS and $N_s \leq N_{RF}^r \leq N_r$ at MS.

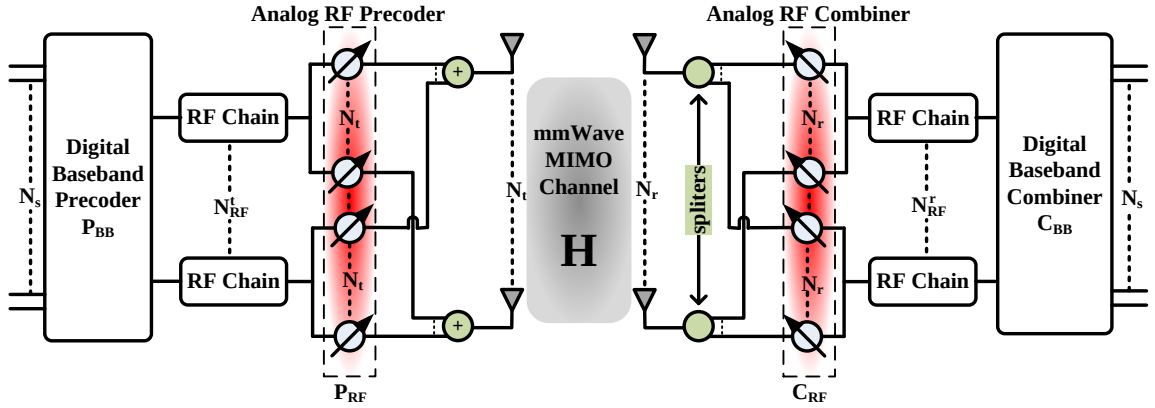


Figure 3.1 Fully-connected structure of single user mmWave system hybrid precoding-combining implemented using phase shifters

The discrete time transmitted signal can be represented as

$$\mathbf{x} = \mathbf{P}_{RF} \mathbf{P}_{BB} \mathbf{s}, \quad (3.1)$$

where $\mathbf{s}$ is $N_s \times 1$ symbol vector with the condition of $\mathbb{E}[\mathbf{s}\mathbf{s}^H] = \frac{1}{N_s} \mathbf{I}_{N_s}$.

The hybrid precoder at BS which is composed of a $N_{RF}^t \times N_s$ digital baseband precoder $\mathbf{P}_{BB}$ and an $N_t \times N_{RF}^t$ analog RF precoder $\mathbf{P}_{RF}$ is

subject to the transmit power constraints $\|\mathbf{P}_{RF}\mathbf{P}_{BB}\|_F^2 = N_s$. The $N_r \times 1$ narrowband post processing received signal at MS is given by

$$\mathbf{y} = \sqrt{\rho}\mathbf{C}_{BB}^H\mathbf{C}_{RF}^H\mathbf{H}\mathbf{P}_{RF}\mathbf{P}_{BB}\mathbf{s} + \mathbf{C}_{BB}^H\mathbf{C}_{RF}^H\mathbf{n}, \quad (3.2)$$

where $\mathbf{H}$ is $N_r \times N_t$ channel matrix with $\|\mathbf{H}\|_F^2 = N_t N_r$, $\mathbf{C}_{BB}$ is $N_{RF}^r \times N_s$ digital baseband combiner and $\mathbf{C}_{RF}$ is $N_r \times N_{RF}^r$ analog RF combiner. The independent and identically distributed (i.i.d) $\mathcal{CN}(0, \sigma_n^2)$ noise denoted as $\mathbf{n}$. The average received power ρ is given by

$$\rho = P \left(\frac{\lambda}{4\pi d} \right)^2 \left(\frac{d_0}{d} \right)^{n_{PL}}, \quad (3.3)$$

where P is the average total transmitted power, d_0 is defined by Rappaport (2002) as the close-in distance or received power reference point related to the BS-MS separation d and n_{PL} is the path loss exponent. The work in this thesis is based on the assumption of perfect knowledge of CSI at Tx and Rx, however, practically, CSI can be analyzed accurately via channel estimation method at the Rx and then shared at the Tx using effective limited feedback. For Gaussian transmission-based symbols, the achievable spectral efficiency can be written as follows

$$R = \log_2 \left| \mathbf{I}_{N_s} + \frac{\rho}{N_s} \mathbf{R}_n^{-1} \mathbf{C}_{BB}^H \mathbf{C}_{RF}^H \mathbf{H} \mathbf{P}_{RF} \mathbf{P}_{BB} \mathbf{P}_{BB}^H \mathbf{P}_{RF}^H \mathbf{H}^H \mathbf{C}_{RF} \mathbf{C}_{BB} \right|, \quad (3.4)$$

where $\mathbf{R}_n = \sigma_n^2 \mathbf{C}_{BB}^H \mathbf{C}_{RF}^H \mathbf{C}_{RF} \mathbf{C}_{BB}$ is the noise covariance matrix after combining and ρ/σ_n^2 is the received SNR. According to the phase control of the signal via RF precoder, the all non-zero entries of $\mathbf{P}_{RF}$ and $\mathbf{C}_{RF}$ satisfies the condition of $\mathbf{P}_{RF} = \mathbf{C}_{RF} = \mathbf{1}$.

3.2 Channel Model

The characteristics of mmWave propagation, i.e. the high free-space path loss, resulted in scattering environment. Furthermore, there is high

antenna correlation due to large antenna arrays packaging which describes the characteristics of mmWave transceivers. As a result of this phenomena, the statistical fading distributions that are applicable for MIMO systems are not suitable in mmWave channel modeling. Accordingly, for adopting a narrowband clustered channel representation, based on previous work of Saleh & Valenzuela (1987), which is well known as Saleh-Valenzuela model, an extended research was done by Rauhavan & Sayeed (2011) to obtain accurate mmWave channels mathematical structure capturing.

As explained by Rappaport, Heath, Daniels, & Murdock (2014), for the clustered channel model, the channel matrix $\mathbf{H}$ represents the sum of N_{cl} scattering clusters with the contribution of N_{ray} propagation path rays in each cluster. Thus, the mmWave channel matrix is depicted as

$$\mathbf{H} = \sqrt{\frac{N_t N_r}{N_c N_{ray}}} \sum_{i=1}^{N_{cl}} \sum_{l=1}^{N_{ray}} \alpha_{il} \Lambda_r(\phi_{il}^r) \Lambda_t(\phi_{il}^t) \mathbf{a}_r(\phi_{il}^r) \mathbf{a}_t(\phi_{il}^t)^H, \quad (3.5)$$

where α_{il} denotes the complex gain of the l^{th} ray in the i^{th} cluster. α_{il} are assumed to be i.i.d. $\mathcal{CN}(0, \sigma_{\alpha,i}^2)$. The average cluster powers are $\sum_{i=1}^{N_{cl}} \sigma_{\alpha,i}^2 = \gamma$ where γ is the normalization factor that satisfy the condition $\|\mathbf{H}\|_F^2 = N_t N_r$. The receive and transmit array response vectors are $\mathbf{a}_r(\phi_{il}^r)$ and $\mathbf{a}_t(\phi_{il}^t)$, respectively, where ϕ_{il}^r is azimuth AoA and ϕ_{il}^t is azimuth AoD. The N_{ray} of ϕ_{il}^ζ are assumed to be uniformly-random distributed with their mean cluster angles (ϕ_i^ζ) and angular spread (σ_{ϕ^ζ}) , $\forall \zeta \in \{t, r\}$. Finally, the vectors $\Lambda_r(\phi_{il}^r)$ and $\Lambda_t(\phi_{il}^t)$ denotes receive and transmit antenna element gain at the corresponding AoA and AoD for the l^{th} ray. The Tx's and Rx's antenna elements can be modeled like ideal sectorized elements as by Singh, Mudumbai, & Madhow (2011) where each of them has the form

$$\Lambda_{\zeta}(\phi_{il}^{\zeta}) = \begin{cases} 1 & \forall \phi_{il}^{\zeta} \in [\phi_{min}^{\zeta}, \phi_{max}^{\zeta}], \\ 0 & elsewhere, \end{cases} \quad (3.6)$$

without loss of generality, the unity gain is assumed over the sector $(\phi_l^{\zeta}, \phi_l^{\zeta})$.

According to Balanis (2005), the one dimensional N_{ζ} -element ULA at BS and MS can be expressed as

$$\mathbf{a}_{\zeta}^{ULA}(\phi_l^{\zeta}) = \frac{1}{\sqrt{N_{\zeta}}} \left[1, e^{j\frac{2\pi}{\lambda}d \sin(\phi_l^{\zeta})}, \dots, e^{j\frac{2\pi}{\lambda}d(N_{\zeta}-1) \sin(\phi_l^{\zeta})} \right]^T, \quad (3.7)$$

where d represents the antenna spacing and the signal wavelength denoted as λ . As shown in Figure 3.2, the ULA response vector is invariant in the elevation domain, hence θ not included in the argument.

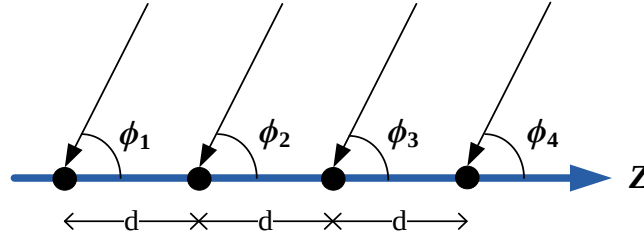


Figure 3.2 Antenna array geometry of ULA

3.3 Hybrid Precoding/Combining Design

The design target of hybrid precoder/combiner is to maximize the spectral efficiency over all possible solutions of four matrix variables $(\mathbf{P}_{RF}, \mathbf{P}_{BB}, \mathbf{C}_{RF}, \mathbf{C}_{BB})$. However, finding perfect solution for similar constraints for optimization problems is not straight forward. Alternatively, the optimization problem can be solved by maximizing the mutual information expression

$$I(\mathbf{P}_{RF}, \mathbf{P}_{BB}) = \log_2 \left| \mathbf{I}_{N_s} + \frac{\rho}{\sigma_n^2 N_s} \mathbf{H} \mathbf{P}_{RF} \mathbf{P}_{BB} \mathbf{P}_{BB}^H \mathbf{P}_{RF}^H \mathbf{H}^H \right| \quad (3.8)$$

3.3.1 Problem Formulation

In mmWave beamforming, the precoders and combiners can be separated into two sub-designs, i.e., the precoding and combining design. Based on mathematical derivations, the precoder/combiner designs are decoupled in Tx-Rx optimization problem to avoid intractable solutions (El Ayach, Rajagopal, Abu-Surra, Pi, & Heath, 2014). Therefore, it is possible to focus on the design of the hybrid precoder only. The corresponding problem formulation of the hybrid precoder design which can be written according to equation (3.9) in terms of distance between analog-digital precoders $\mathbf{P}_{RF}\mathbf{P}_{BB}$ and the $N_t \times N_s$ optimal unconstrained precoder $\mathbf{P}_{opt}$ given as

$$\begin{aligned} (\mathbf{P}_{RF}^{opt}, \mathbf{P}_{BB}^{opt}) &= \arg \min_{\mathbf{P}_{RF}, \mathbf{P}_{BB}} \|\mathbf{P}_{opt} - \mathbf{P}_{RF}\mathbf{P}_{BB}\|_F \\ \text{s.t. } &\begin{cases} \mathbf{P}_{RF}(:, i) \in \{\mathbf{A}_{can}(:, k), 1 \leq k \leq N_{cl} N_{ray}\}, \forall 1 \leq i \leq N_{RF}^t, \\ \|\mathbf{P}_{RF}\mathbf{P}_{BB}\|_F^2 = N_s. \end{cases} \end{aligned} \quad (3.9)$$

With perfect channel knowledge, $\mathbf{P}_{opt} = [\mathbf{V}_1, \mathbf{V}_2, \dots, \mathbf{V}_{N_s}]$, which is composed of the first N_s columns of the unitary matrix $\mathbf{V}$. The unitary matrix derived from the channel's SVD by Chen, Tsai, Liu, & Wu (2017) as

$$\mathbf{H} = \mathbf{U}\mathbf{\Sigma}\mathbf{V}^H = \sum_{m=1}^{\min(N_r, N_t)} \sigma_m \mathbf{u}_m \mathbf{v}_m^H, \quad (3.10)$$

where $\mathbf{U} = [\mathbf{u}_1, \mathbf{u}_2, \dots, \mathbf{u}_{N_r}]$ and $\mathbf{V} = [\mathbf{v}_1, \mathbf{v}_2, \dots, \mathbf{v}_{N_t}]$ are the $N_r \times N_r$ left and the $N_t \times N_t$ right singular matrices of the channel $\mathbf{H}$, respectively. The $N_r \times N_t$ diagonal matrix $\mathbf{\Sigma}$ is with all eigenvalues of $\mathbf{H}$ in non-increasing order, i.e., $\sigma_1 \geq \sigma_2 \geq \dots \geq \sigma_{\min(N_r, N_t)}$. $\mathbf{A}_{can}$ is a $N_t \times N_{cl}N_{ray}$ candidate matrix with N_{RF}^t transmit response vectors chosen from

$$\mathbf{A}_t = [\mathbf{a}_t(\phi_1^t), \mathbf{a}_t(\phi_2^t), \dots, \mathbf{a}_t(\phi_n^t)], \forall 1 \leq n \leq N_{cl}N_{ray}, \quad (3.11)$$

which results in finding the best low dimensional representation of unconstrained optimal digital precoder using the basis vectors $\mathbf{a}_t(\phi_{il}^t)$ from matrix $\mathbf{A}_t$. Minimizing the objective function $\|\mathbf{P}_{opt} - \mathbf{P}_{RF}\mathbf{P}_{BB}\|_F$ leads to maximization of $I(\mathbf{P}_{RF}, \mathbf{P}_{BB})$. Further, the resulting hybrid precoders should be close to the unconstrained optimal full digital precoder, in other words, the Euclidean distance between them should be minimized as much as possible. The design process is to find the projection of the optimal precoder $\mathbf{P}_{opt}$ into a set of hybrid analog/digital precoders $\mathbf{P}_{RF}\mathbf{P}_{BB}$ with $\mathbf{P}_{RF} \in \mathcal{A}$, where $\mathcal{A}$ is the set that corresponds to columns of candidate matrix $\mathbf{A}_{can}$.

3.3.2 Modified Hybrid Precoding Method

In order to minimize the Frobenius norm between the optimal and hybrid precoders, an approximated solution known as simultaneous orthogonal matching pursuit algorithm was considered to reconstruct the hybrid precoder. Although this algorithm achieves performance approximated to unconstrained optimal precoder with limited number of bits, the need for iterative process for selecting N_{RF}^t beamforming vectors of RF precoder and matrix inversion for baseband precoder makes the design impractical since the matrix inversion of $N \times N$ matrix requires a complexity of $\mathcal{O}(N^3)$. Sliding window-index-selection matrix-inversion-bypass simultaneous orthogonal matching pursuit was proposed by Hsu, Wang, Lee, & Huang (2015) to reduce the matrix inversion overhead, however, still it is not suitable for practical hardware implementation due to smaller degree of parallelism.

3.3.2.1 Layered-RF Codebook Design

The layered-RF codebook is preferred as this candidate beamforming matrix has the following properties:

- The columns of the RF matrix are highly uncorrelated compared to traditional candidate matrix for simultaneous orthogonal matching pursuit.
- Ability to synthesizing the space of array response vectors in any direction due to linear combinations of the columns in the RF matrix.

Consider the $N_t \times N_t$ ULA candidate array response vector of RF codebook, also known as columns of RF matrix, defined as

$$\mathbf{A}_{t,RF}(:, k) = \frac{1}{\sqrt{N_t}} \left[1, e^{j\frac{2\pi}{N_t}(k-1)}, \dots, e^{j\frac{2\pi}{N_t}(k-1)(N_t-1)} \right]^T, \quad (3.12)$$

where $1 \leq k \leq N_t$. The weightings of the phase shifters can be selected from the columns of $\mathbf{A}_{t,RF}$ at the BS to steer signals at N_t independent beamforming directions. Figure 3.3 shows the beam pattern of $\mathbf{A}_{t,RF}$ assuming $N_t = 8$, also it can be predicted that there is no correlation between $\mathbf{A}_{t,RF}$ columns vectors, as a result, RF matrix is suitable for practical implementation due to its orthogonal property.

Now, transformation to a new orthonormal matrix is predicted by implementing $(2\pi z / \delta N_t)$ phase shift to the previous $\mathbf{A}_{t,RF}$ matrix

$$\mathbf{A}_{t,RF,z}(:, k) = \frac{1}{\sqrt{N_t}} \left[1, e^{j\frac{2\pi}{N_t}\{(k-1)+\frac{z}{\delta}\}}, \dots, e^{j\frac{2\pi}{N_t}\{(k-1)+\frac{z}{\delta}\}(N_t-1)} \right]^T \quad (3.13)$$

where $1 \leq k \leq N_t$ and $1 \leq z \leq \delta - 1$ is the rotation unit of the phase. The beam patterns of the rotated RF matrix of 8 elements ULA are shown in Figures 3.4, 3.5 and 3.6, assuming $(\delta = 4)$ and $(z = 1, 2, 3)$.

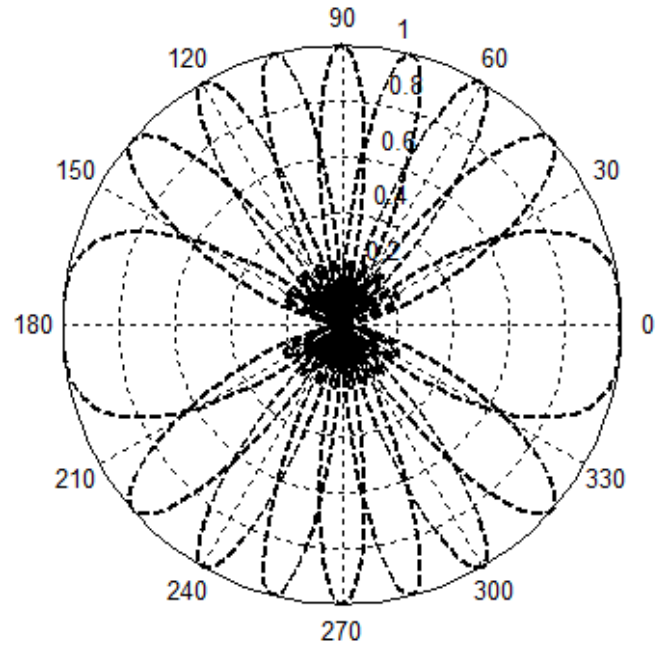


Figure 3.3 Beam pattern of RF matrix $\mathbf{A}_{t,RF}$

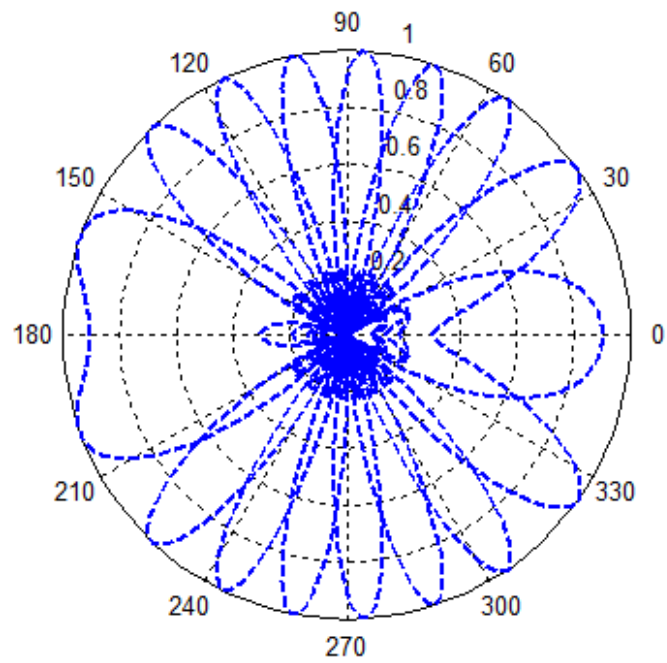


Figure 3.4 Beam pattern of RF matrix $\mathbf{A}_{t,RF,1}$

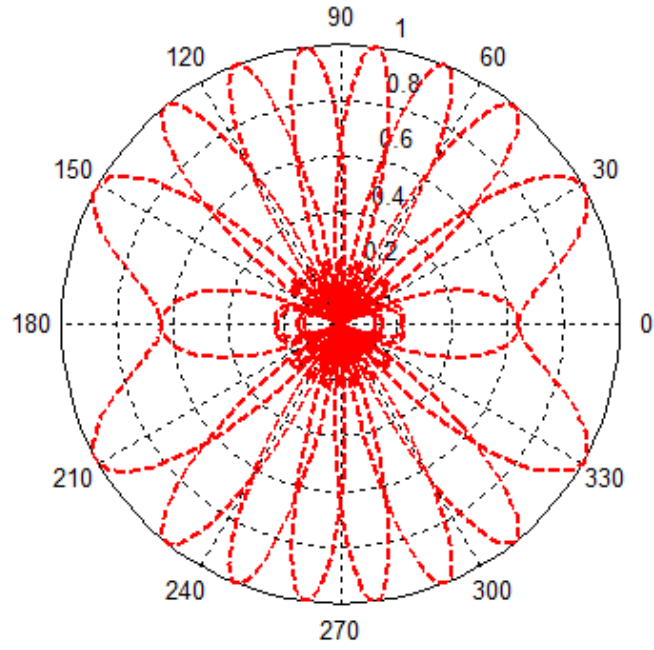


Figure 3.5 Beam pattern of RF matrix $\mathbf{A}_{t,RF,2}$

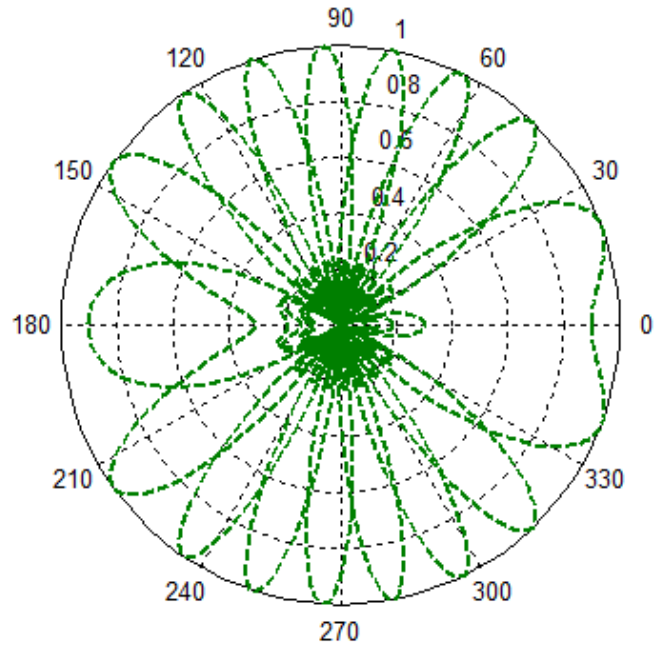


Figure 3.6 Beam pattern of RF matrix $\mathbf{A}_{t,RF,3}$

Finally, by combining all the δ orthonormal matrices $\mathbf{A}_{t,RF}$ and $\mathbf{A}_{t,RF,z}$ for $1 \leq z \leq \delta - 1$, the resulting $N_t \times \delta N_t$ predefined layered-RF codebook $\mathbf{L}_{t,RF,z}$ has the form

$$\mathbf{L}(:, k) = \frac{1}{\sqrt{N_t}} \left[1, e^{j\frac{2\pi}{\delta N_t}(k-1)}, \dots, e^{j\frac{2\pi}{\delta N_t}(k-1)(N_t-1)} \right]^T, \quad (3.14)$$

where, $1 \leq k \leq \delta N_t$. The polar plot of $\mathbf{L}$ beam pattern is shown in Figure 3.7.

Notably, for getting a corresponding orthonormal matrix as $\mathbf{A}_{t,RF,z}$ in equation (3.14), clearly it can be predicted by performing sampled of δ -spacing in $\mathbf{L}$ with initial index $1 \leq z \leq \delta - 1$, as follows

$$\begin{aligned} \mathbf{G}_z &= [\mathbf{L}(:, z), \mathbf{L}(:, z + \delta), \dots, \mathbf{L}(:, z + \delta \times (N_t - 1))] \\ &= \mathbf{A}_{t,RF,z}. \end{aligned} \quad (3.15)$$

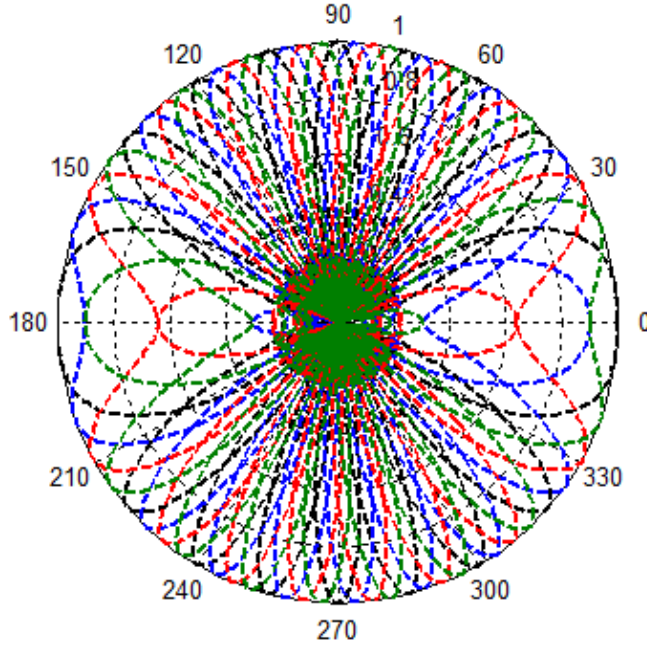


Figure 3.7 Beam pattern of RF matrix $\mathbf{L}$ that combines the four orthonormal matrices

Now with this RF-layered codebook $\mathbf{L}$, the BS will have more accurate steering beamforming directions for the downlink transmitted signals, while also considering the RF hardware constraints as well.

3.3.2.2 ML Fixed Weight Beamforming

The scanning of the main beam pattern of an array of antennas is known as beam steering. The phase shifters, where the beam steered via this values, are implemented at RF frequencies and used to electronically scan the phase of interest of the electric current at each antenna element. Alternatively, as for modern wireless communications, smart antennas beamforming are controlled via algorithms under certain conditions. One of these conditions, is steering toward a signal of interest. ML is one type of fixed weight beamforming which can be applied to maximize the power of the received signal and minimizing the noise signal.

Assuming the desired signal $\mathbf{x}_s$ and the interferer undesired signal $\mathbf{n}$ $\mathcal{CN}(0, \sigma_n^2)$ are considered, the main objective of the likelihood function is to give an approximate value for the desired signal with no feedback for the antenna elements. Referring to Figure 3.8, the input signal is represented as

$$\mathbf{x} = \mathbf{a}_0 s + \mathbf{n} = \mathbf{x}_s + \mathbf{n}. \quad (3.16)$$

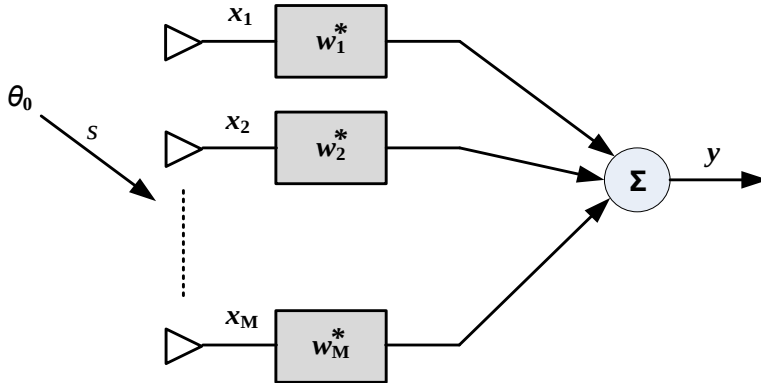


Figure 3.8 Block diagram of M-elements array

The likelihood function defined by Haykin, Justice, Owsley, Yen, & Kak (1985) can be used to describe the joint probability density function between $\mathbf{x}$ and $\mathbf{x}_s$ as

$$p(\mathbf{x}|\mathbf{x}_s) = \frac{1}{\sqrt{2\pi\sigma_n^2}} e^{-((\mathbf{x}-\mathbf{a}_0s)^H \mathbf{R}_{nn}^{-1}(\mathbf{x}-\mathbf{a}_0s))}, \quad (3.17)$$

where $\mathbf{R}_{nn} = \sigma_n^2 \mathbf{I}$ is the noise correlation matrix. Gross (2005) defined the log likelihood function which is the negative of logarithmic of density function as

$$L[\mathbf{x}] = -\ln[p(\mathbf{x}|\mathbf{x}_s)] = C(\mathbf{x} - \mathbf{a}_0s)^H \mathbf{R}_{nn}^{-1}(\mathbf{x} - \mathbf{a}_0s), \quad (3.18)$$

where C is constant and $\mathbf{R}_{nn}^{-1} = \mathbb{E}[\mathbf{nn}^H]$. The maximum value of the likelihood function defined above can be found by solving the partial derivative of likelihood function with respect to s as

$$\frac{\partial L[\mathbf{x}]}{\partial s} = -2\mathbf{a}_0^H \mathbf{R}_{nn}^{-1} \mathbf{x} + 2\hat{s} \mathbf{a}_0^H \mathbf{R}_{nn}^{-1} \mathbf{a}_0 = 0. \quad (3.19)$$

Then, solving for desired estimated signal $\hat{s}$ implies

$$\hat{s} = \frac{\mathbf{a}_0^H \mathbf{R}_{nn}^{-1} \mathbf{x}}{\mathbf{a}_0^H \mathbf{R}_{nn}^{-1} \mathbf{a}_0} \quad \mathbf{x} = \mathbf{w}_{ML}^H \mathbf{x} \quad (3.20)$$

Thus, the optimum array weights can be formulated as

$$\mathbf{w}_{ML} = \frac{\mathbf{R}_{nn}^{-1} \mathbf{a}_0}{\mathbf{a}_0^H \mathbf{R}_{nn}^{-1} \mathbf{a}_0}. \quad (3.21)$$

Once the array vector is calculated from the above equation, the targeted angle can be estimated from a set of quantized points. Figure 3.9 shows the array factor plot for 16 elements ULA with half wavelength spacing, which receives a signal at $\theta_0 = 160^\circ$ arriving angle with noise variance of $\sigma_n^2 = 0.01$. The angles assumed to be distributed between $(\pi/2, 3\pi/2)$.

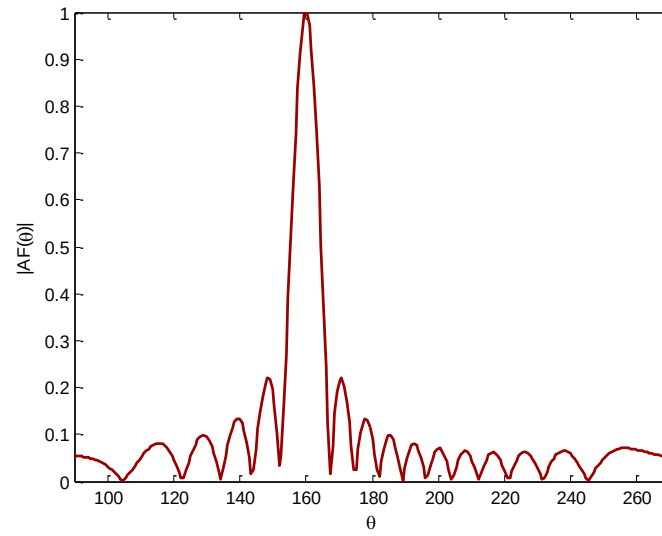


Figure 3.9 ML pattern plot for 16 elements ULA

For mmWave system configuration validity, another setting is considered in Figure 3.10, where 32 ULA antenna elements assumed with an arriving angle of $\theta_0 = 220^\circ$.

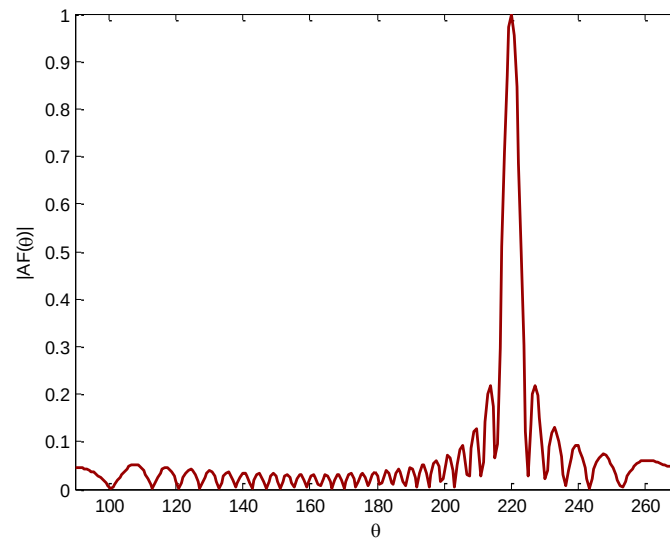


Figure 3.10 ML pattern plot for 32 elements ULA

3.3.2.3 OBMP via ML Fixed Weight Beamforming (MLOBMP)

With layered-RF codebook $\mathbf{L}$ and ML fixed weight beamforming, hybrid precoder-combiner design will be defined based on orthogonal mapping method. Consider the quantized information of the candidate matrix $\mathbf{A}_{can,q}$ which is given as

$$\mathbf{A}_{can,q} = [\mathbf{a}_t(\phi_{q1}^t), \mathbf{a}_t(\phi_{q2}^t), \dots, \mathbf{a}_t(\phi_{qn}^t)], \forall 1 \leq n \leq N_{cl}N_{ray}. \quad (3.22)$$

Given each phase of $\mathbf{a}_t(\phi_{qn}^t)$ that is quantized by finite number of bits, optimum weight for each antenna array response vector is calculated to estimate the best AoD at BS denoted as $\hat{\phi}_{qn}^t$. Then, the smallest Euclidean distance between $\mathbf{a}_t(\hat{\phi}_{qn}^t)$ and $\mathbf{L}(:, k_n)$ is calculated by equating terms in equation (3.7) and equation (3.14) as described by Liu, Chen, Tsai, & Wu (2016), that is

$$k_n = \text{round} \left\{ \left(\left(\frac{d\delta N_t}{\lambda} \right) \sin(\hat{\phi}_{qn}^t) \right) + 1 \right\} \quad (3.23)$$

$$k_n = \begin{cases} k_n + \delta N_t, & \text{for } k_n < 0 \\ k_n, & \text{elsewhere} \end{cases}, \forall 1 \leq n \leq N_{cl}N_{ray},$$

where k_n is the corresponding column index of array response vectors in equation (3.22) for layered-RF codebook. The resulting $\mathbf{G}_n$ matrix now has the largest correlation with $\mathbf{A}_{can,q}$. The operation is shown in Figure 3.11 assuming $N_t = 8$, $N_{cl}N_{ray} = 4$ and $\hat{\phi}_{qn}^t$ are uniformly distributed over $(0, 2\pi)$.

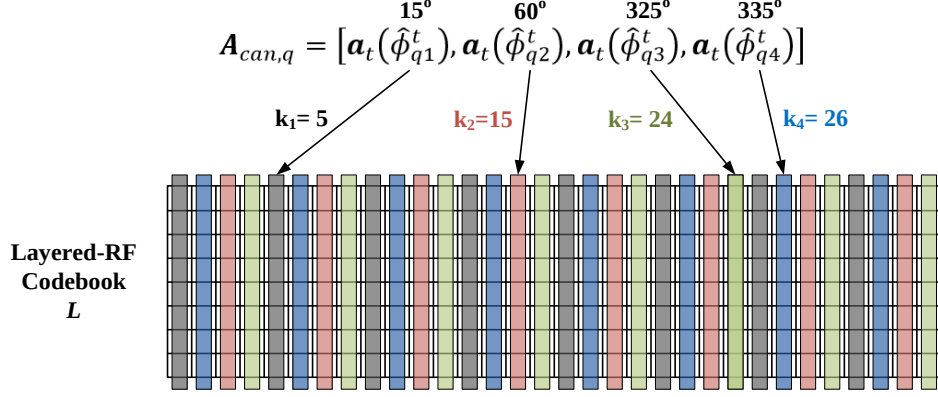


Figure 3.11 Orthogonal mapping process of array response vectors to the vectors in L

The new orthonormal matrix $\mathbf{G}$ can be used in hybrid precoding design in equation (3.9). Motivated by Liu, Chen, Tsai, & Wu (2016), the pseudo-code of the algorithm is shown in Table 3.1 and the illustration of the corresponding process shown in Figure 3.12 assuming $N_s = 1$ and $N_{RF}^t = 2$. First, the ML fixed weight beamforming implemented to derive the corresponding index set $\mathcal{K}$ of orthonormal matrix $\mathbf{G}$, the index set defined by

$$\mathcal{K} = \{k_1, k_2, \dots, k_n\}, \forall 1 \leq n \leq N_{cl}N_{ray}. \quad (3.24)$$

Then, the correlation between optimal precoder $\mathbf{P}_{opt}$ and orthonormal matrix $\mathbf{G}$ is calculated at step (2), which is given by

$$\mathbf{\Psi} = \mathbf{G}^H \mathbf{P}_{opt}, \quad (3.25)$$

where $\mathbf{\Psi}$ is a $N_{cl}N_{ray} \times N_s$ correlation matrix. Then, at step (3), the power spread on each beamforming direction is calculated as

$$\boldsymbol{\mu} = \text{diag}(\mathbf{\Psi}\mathbf{\Psi}^H), \quad (3.26)$$

where $\mathbf{\mu}$ is a $N_{cl}N_{ray} \times 1$ vector which the i^{th} entry represents the total optimal precoder $\mathbf{P}_{opt}$ spread at the i^{th} beamforming direction of $\mathbf{G}$. Due to un-correlation property of beamforming vectors columns, the N_{RF}^t beamforming vectors of $\mathbf{P}_{RF}$ can be approximated and selected in parallel with N_{RF}^t largest values of $\mathbf{\mu}$ from step (4) to (9)

$$\mathbf{P}_{RF} = \mathbf{G}(:, \mathcal{S}), \quad (3.27)$$

where $\mathcal{S}$ is the corresponding set of indexes of the largest N_{RF}^t values of $\mathbf{\mu}$ and $|\mathcal{S}| = N_{RF}^t$. This completely avoids the iterative process for finding optimal analog precoder $\mathbf{P}_{RF}$. After forming $\mathbf{P}_{RF}$, the corresponding baseband precoder $\mathbf{P}_{BB}$ is found as in El Ayach, Rajagopal, Abu-Surra, Pi, & Heath (2014) by least square solution

$$\mathbf{P}_{BB} = (\mathbf{P}_{RF}^H \mathbf{P}_{RF})^{-1} \mathbf{P}_{RF}^H \mathbf{P}_{opt}. \quad (3.28)$$

Based on orthogonal property of $\mathbf{P}_{RF}$ precoder, it is considered as unitary matrix, which is represented as $N_{RF}^t \times N_{RF}^t$ identity matrix, thus the least square solution in equation (3.28) is further simplified according to step (10) as

$$\begin{aligned} \mathbf{P}_{BB} &= (\mathbf{P}_{RF}^H \mathbf{P}_{RF})^{-1} \mathbf{P}_{RF}^H \mathbf{P}_{opt} = \mathbf{I}_{N_{RF}^t} \mathbf{P}_{RF}^H \mathbf{P}_{opt} \\ \mathbf{P}_{RF}^H \mathbf{P}_{opt} &= \mathbf{G}(:, \xi)^H \mathbf{P}_{opt} = \mathbf{\Psi}(\mathcal{S}, :). \end{aligned} \quad (3.29)$$

As a result, this totally omits the need of matrix inversion and replaced by simple correlation for calculating baseband precoder $\mathbf{P}_{BB}$. Finally, at step (11), the baseband precoder is normalized to unity norm column as

$$\mathbf{P}_{BB} = \sqrt{N_s} \frac{\mathbf{P}_{BB}}{\|\mathbf{P}_{RF} \mathbf{P}_{BB}\|_F}. \quad (3.30)$$

The same algorithm can be implemented for calculating the hybrid combiners $\mathbf{C}_{RF}$ and $\mathbf{C}_{BB}$, except the last normalization step.

Table 3.1 Hybrid precoder reconstruction via MLOBMP

Input: the optimal precoder $\mathbf{P}_{opt}$, the layered-RF codebook matrix $\mathbf{G}$, the quantized AoDs $[\mathbf{a}_t(\phi_{q1}^t), \mathbf{a}_t(\phi_{q2}^t), \dots, \mathbf{a}_t(\phi_{qn}^t)]$ and the number of BS RF chains N_{RF}^t. Output: $\mathbf{P}_{RF}$ and $\mathbf{P}_{BB}$ that minimize the objective function in equation (3.9) under the given hardware constraints.	
Step	Process
1:	Implement ML to estimate AoDs then generate the set $\mathcal{K}$ and select the corresponding columns of layered-RF codebook: $\mathbf{G} = \mathbf{L}(:, \mathcal{K})$
2:	Compute all correlations: $\mathbf{\Psi} = \mathbf{G}^H \mathbf{P}_{opt}$
3:	Perform the power spread on certain direction by calculating $diag(\mathbf{\Psi} \mathbf{\Psi}^H)$
4:	Initiate an empty set $\mathcal{S}$
5:	for $i = 1$ to N_{RF}^t do:
6:	Find the largest value of diagonal: $k = \arg \max_{n=1,2,\dots,N_{cl}N_{ray}} \{\mu(n)\}$
7:	Concatenate the set: $\mathcal{S} = [\mathcal{S} k]$ and delete the corresponding row: $\mu(k) = 0$
8:	end for.
9:	Build $\mathbf{P}_{RF}$ as $\mathcal{S}$ columns of $\mathbf{G}$
10:	Select $\mathcal{S}$ rows of $\mathbf{G}$ to compute $\mathbf{P}_{BB}$
11:	Perform normalization to $\mathbf{P}_{BB}$

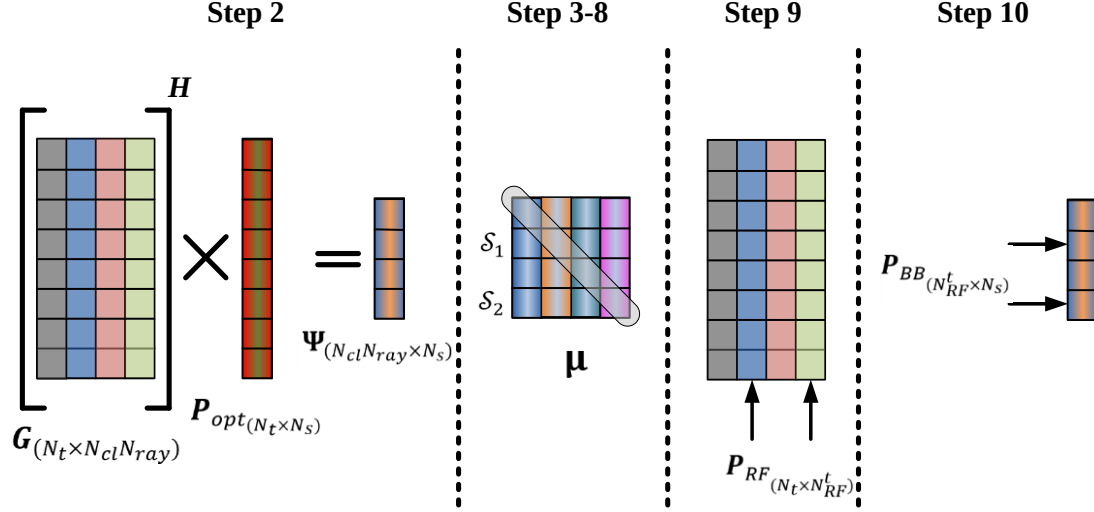


Figure 3.12 Illustration of the correlation, power spread and hybrid precoder selection steps for MLOBMP algorithm

3.4 Simulation Results and Performance Evaluation

In this section, numerical simulation results for the performance analysis of the proposed MLOBMP based on layered-RF codebook and ML fixed weight beamforming will be presented.

3.4.1 System Configuration

In the simulations, a single user MIMO system in mmWave channel is considered. Assume the BS is equipped with $N_t = 64$ antennas to support single and multiple data streams, and the MS with $N_r = 16$ antennas. Both BS and MS adopt ULA with d antenna spacing of 0.5λ . The mmWave channel is modeled by two clusters ($N_{cl} = 2$) with four propagation path rays in each cluster ($N_{ray} = 4$). Further, AoDs and AoAs are assumed to be uniformly distributed between $(\pi/2, 3\pi/2)$. The common power of the transmitter P_{com} is set to 10W, whereas the power of each RF chain P_{RF} , phase shifter P_{PS} and power amplifier P_{PA} are 100mW, 10mW and 100mW(300mW), respectively. The calculation of spectrum efficiency based on equation (3.4) with hybrid precoders

(combiners) adopted at BS (MS). The simulations are performed through 1000 channel realizations. Table 3.2 summarizes the parameters of system configuration.

Table 3.2 System parameters used in the simulation

Simulation Parameter	Value
BS antenna: N_t	64
MS antenna: N_r	16
antenna element's spacing: d	0.5λ (mm)
path clusters: N_{cl}	2
path rays in each cluster: N_{ray}	4
AoDs (AoAs) distribution	$(\pi/2, 3\pi/2)$
common power of the transmitter: P_{com}	10 (W)
power of each RF chain: P_{RF}	100 (mW)
Power of phase shifter: P_{PS}	10 (mW)
Power of power amplifier: P_{PA}	100, 300 (mW)
channel realizations (iterations)	1000

3.4.2 System Performance

Assume that the hybrid precoder/combiner algorithm have the optimal digital precoder (combiner); $\mathbf{P}_{opt}$ ($\mathbf{C}_{opt}$), and the quantized candidate matrix with $q = 8$ bits resolution for AoDs (AoAs) that are distributed uniformly in the set $\{2\pi k/2^q \mid 1 \leq k \leq 2^q\}$. Also, the ML AoDs estimator has a resolution of $\delta N_t(\delta N_r)$ for the BS (MS).

3.4.2.1 Spectral Efficiency versus SNR

First, the spectral efficiency of the proposed MLOBMP versus SNR is plotted in Figure 3.13 and Figure 3.14. For fair comparison, the fully

connected structure was considered for 64×16 mmWave MIMO system with $N_s = 2$ data streams. The same RF candidate matrix discussed in the last section was used with the proposed algorithm. It can be observed that the hybrid beamforming efficiency can achieve $\approx 90\%$ of unconstrained optimal precoder for $N_{RF}^t = N_{RF}^r = 4$ and $\delta = 4$ which is lower than the case for $N_{RF}^t = N_{RF}^r = 6$, i.e., $\approx 95\%$, from this observation it can be noted that the number of RF chains affect the system performance. Moreover, as can be expected, the more layers of RF beamforming candidate matrix, the more accurate approximation of unconstrained optimal digital precoder. Furthermore, based on this results, the proposed algorithm can efficiently approximates the full digital precoder with acceptable rate.

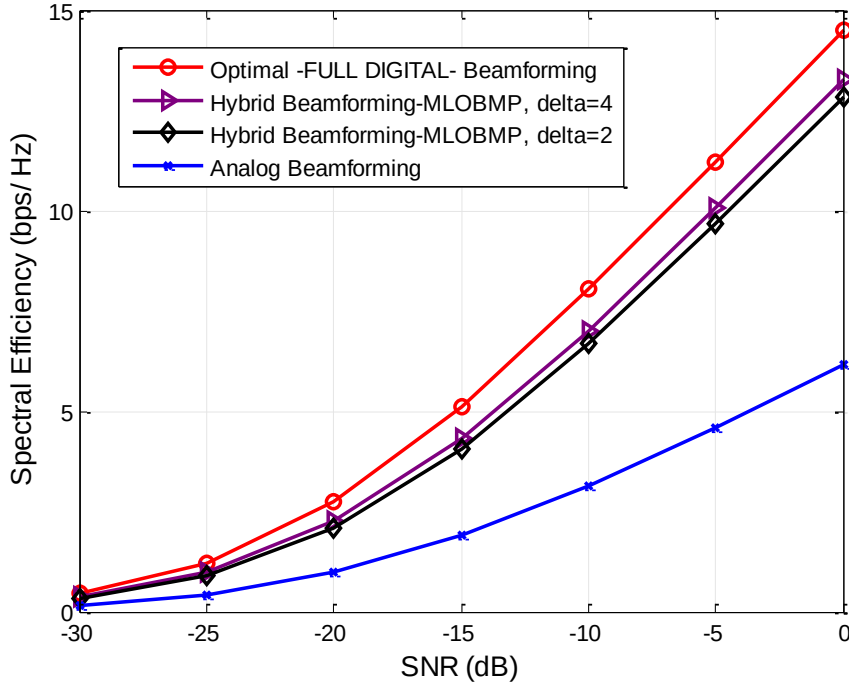


Figure 3.13 Achievable rate of MLOBMP algorithm given $N_s = 2$ and

$$N_{RF}^t = N_{RF}^r = N_{RF} = 4$$

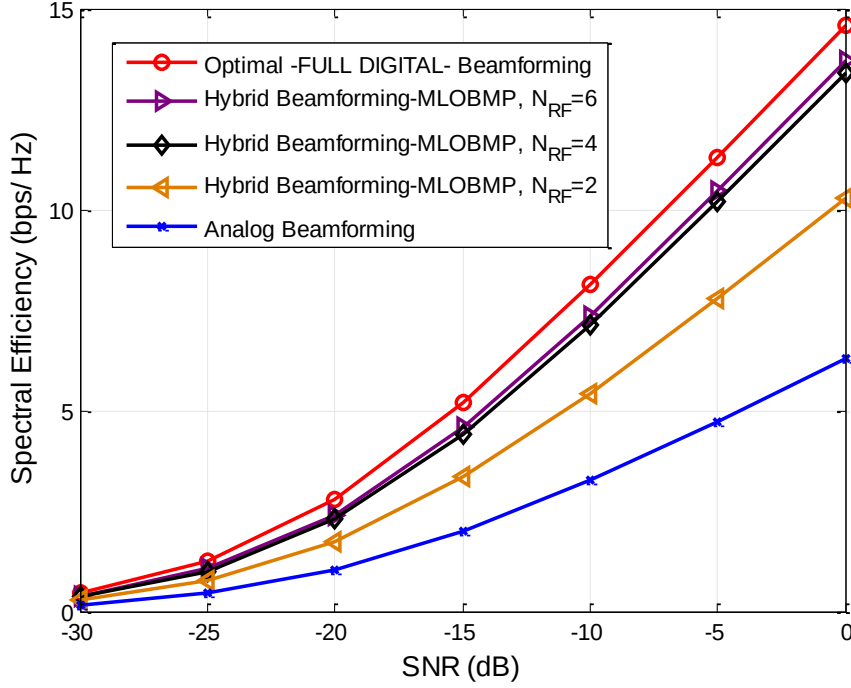


Figure 3.14 Achievable rate of MLOBMP algorithm given $N_s = 2$

3.4.2.2 Spectral Efficiency versus RF Chains

Based on the parameters in this section, Figure 3.15 and Figure 3.16 compare the performance of the proposed algorithm with respect to N_{RF} RF chains. Both figures are showing that the proposed algorithm can achieve near optimal performance between 91% and 94% when $N_{RF}^t = N_{RF}^r \geq 6$, this is clear when $N_{RF}^t \geq 6$ and $N_{RF}^r \geq 6$, a closed-form solution to the design problem exists which results in approximated spectral efficiency to the optimal digital precoding. Moreover, the capacity at 3 data streams is higher compared to the case of $N_s = 2$, this is expected as the columns of optimal digital precoder/combiner are chosen when $rank(\mathbf{H}) \geq N_s$. However, the rate of analog beamforming is still unaffected as it does not require any information about RF chains for calculating its achievable rate.

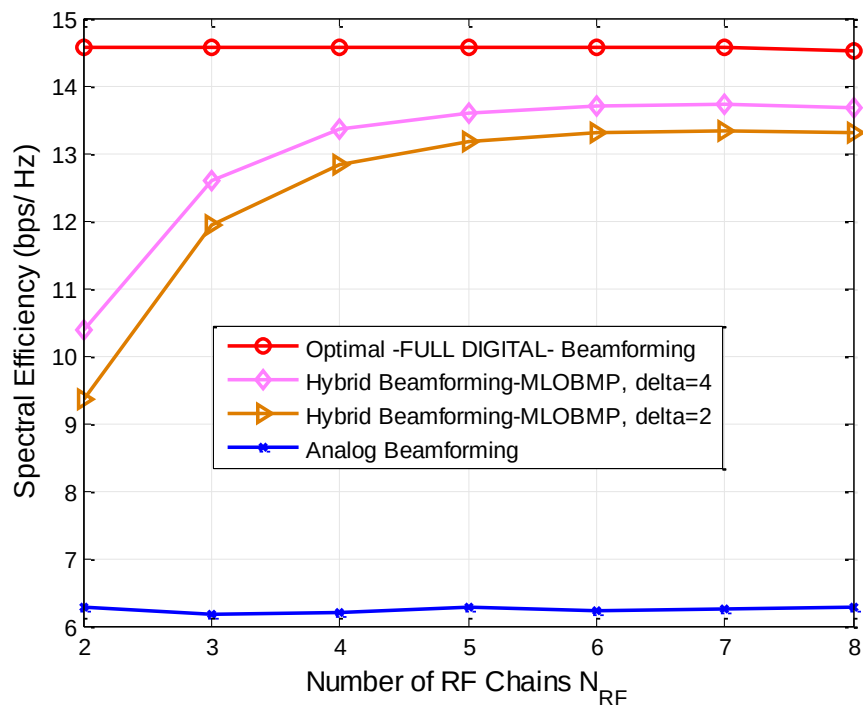


Figure 3.15 Achievable rate of MLOBMP algorithm given $N_s = 2$ and $SNR = 0dB$

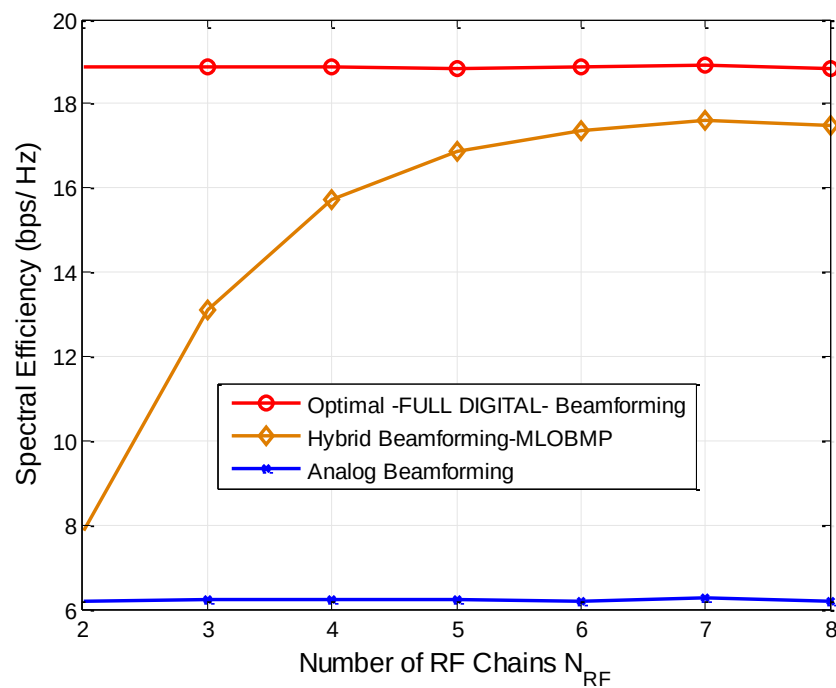


Figure 3.16 Achievable rate of MLOBMP algorithm given $N_s = 3$ and $SNR = 0dB$

3.4.2.3 Energy Efficiency versus SNR

For evaluating the performance, in terms of power consumption, the energy efficiency in (bits/Hz/J) of fully connected structure is defined by Yu, Shen, Zhang, & Lataief (2016) as the achievable rate divided by the total power consumption, that is

$$E = \frac{R}{P_{com} + N_{RF}^t P_{RF} + N_t P_{PA} + N_{PS} P_{PS}}, \quad (3.31)$$

where P_{com} is the common power of Tx, P_{RF} , P_{PA} and P_{PS} are the powers of each RF chain, power amplifier and phase shifter, respectively. Figure 3.17 and Figure 3.18 illustrate the achievable energy efficiency versus SNR. The power of power amplifier P_{PA} is set to 100mW. The achievable rate increases linearly as the number of RF chain increases, however, as expected, analog beamforming remains almost unchanged over the investigated range of SNR.

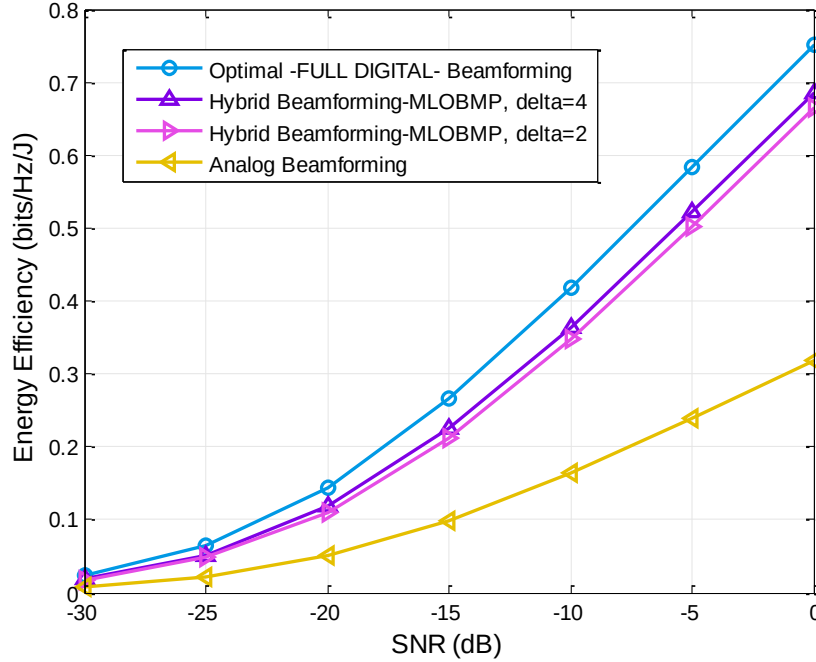


Figure 3.17 Energy efficiency of MLOBMP algorithm given $N_s = 2$

and $N_{RF}^t = N_{RF}^r = N_{RF} = 4$

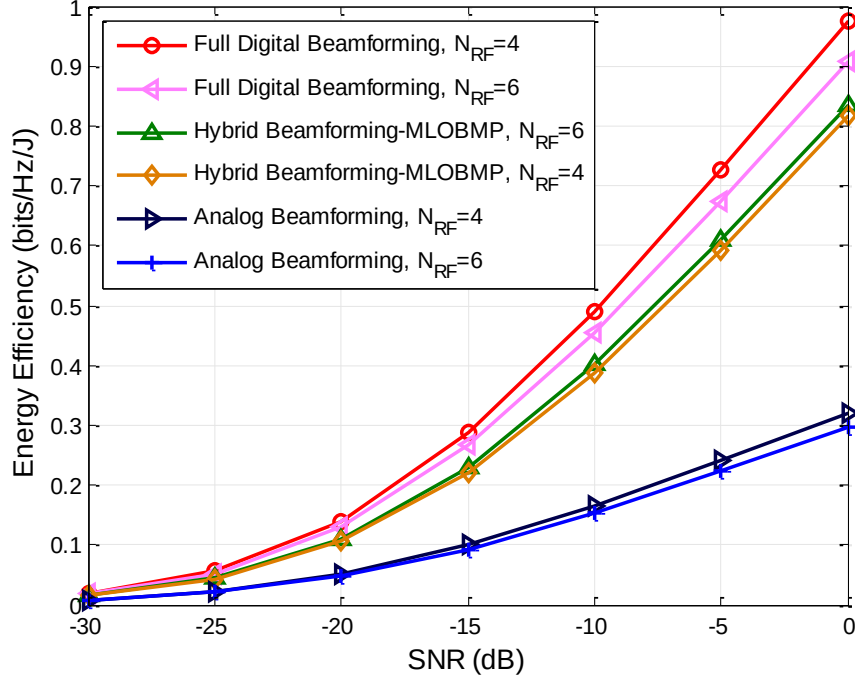


Figure 3.18 Energy efficiency of MLOBMP algorithm given $N_s = 3$

3.4.2.4 Energy Efficiency versus RF Chains

In this subsection, energy efficiency has been examined with respect to the number of RF chains. Figure 3.19 shows the energy efficiency of both optimal and hybrid precoder assuming single data stream transmission for $P_{PA} = 100mW$ and $300mW$, respectively. Since the number of phase shifters are proportional to N_t and N_{RF}^t , the power consumption rose significantly as a result to corresponding increase in N_{RF}^t and hence degrading the energy efficiency as can be concluded from equation (3.31). In addition, as can be noted from Figure 3.19, for any increase in SNR, there will be a corresponding increase in energy efficiency of the optimal and hybrid precoder, as the system rate is proportional to SNR, this change can be noted when $SNR = 0dB$.

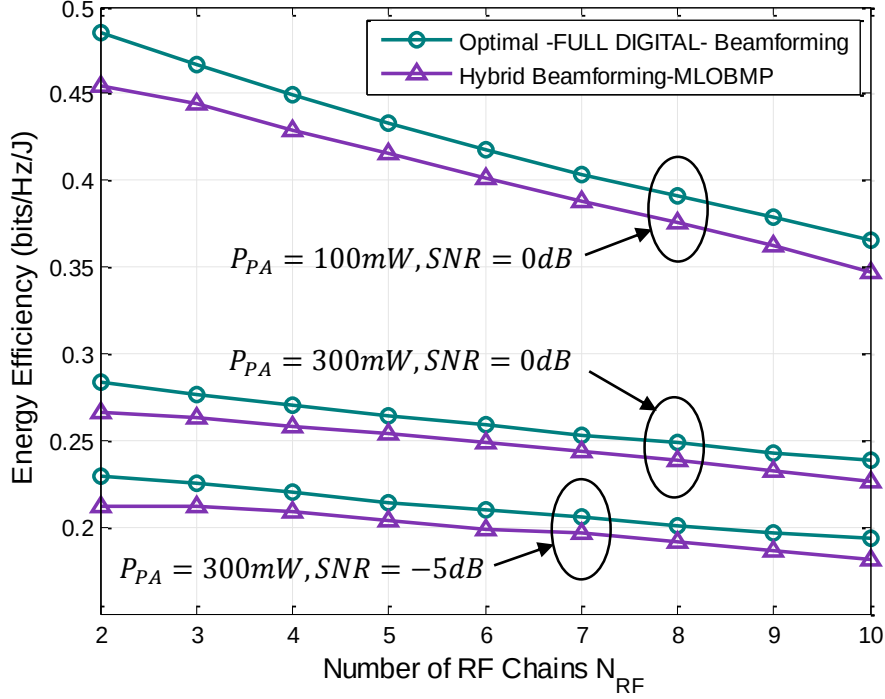


Figure 3.19 Energy efficiency of MLOBMP algorithm given $N_s = 1$

3.5 Conclusion

This chapter proposed MLOBMP method for the hybrid precoding/combining design in single user mmWave MIMO systems. First, the proposed RF-codebook design was introduced as the orthogonal candidate matrix for the hybrid precoding algorithm and proved its ability to steer the BS signal in certain $N_t(N_r)$ directions. Then, the design of ML fixed weight beamforming was implemented for ULA antenna response vectors and it was shown that with the existing random noise, this beamforming method achieves above 90% for AoD approximation.

Second, the proposed hybrid precoding algorithm was introduced based on layered-RF codebook design and ML fixed weight beamforming. The algorithm steps were outlined and proved that the iterative process for calculating RF precoder and matrix inversion process for calculating the baseband precoder was completely eliminated.

Numerical results have been demonstrated based on spectral efficiency and energy efficiency. The effect of various parameters has been studied, including number of data streams N_s , number of RF chains N_{RF} , number of RF codebook layers, power amplifier power consumption P_{PA} and SNR. Starting with the spectral efficiency with respect to SNR, two cases of δ RF codebook layers and three cases of N_{RF} RF chains were considered with fixed N_s data streams, and showed that the maximum rate can be achieved at $\delta = 4$ and $N_{RF} \geq 2N_s$. Then, the spectral efficiency has been tested again with fixed SNR and a set of N_{RF} RF chains, and showed that the spectral efficiency has been stabilized at certain value of N_{RF} RF chain regardless of N_s data streams while this change was not affecting the analog beamforming rate. Compared to the spectral efficiency testing scenario where the fully connected structure provides more degrees of freedom in the RF domain, the power consumption scenario, on the other hand, results in dramatically decrease of energy efficiency. The energy efficiency has been investigated over SNR range with different N_s data streams and N_{RF} RF chains, and it showed that the corresponding increase with N_{RF} RF chains. Finally, it has been shown that the energy efficiency of fully connected structure is high with a small number of RF chains and thus lower total power consumption.

Chapter 4

Beamspace-Based Hybrid Precoding with ML Fixed Weight Beamforming

- 4.1 System Model
- 4.2 Channel Model
- 4.3 Hybrid Precoding Digital Combining Design
- 4.4 Simulation Results and Performance Evaluation
- 4.5 Conclusion

Chapter 4

Beamspace-Based Hybrid Precoding with ML Fixed Weight Beamforming

The fabrication cost of high frequency mixed signal components makes the full digital precoding not suitable for hardware implementation since it requires a large number of RF chains in mmWave systems with massive antennas. For mmWave practical implementations, full digital MMSE combiner at receiver is considered more efficient as it minimizes the mean square error between transceiver signals. Due to sparse representation of mmWave channels, the number of RF chains for hybrid precoding/combining structure is lower than the number of antennas and can achieve same performance with full digital precoder. However, acquiring optimal full digital precoder is unpractical in hybrid precoding design, hence, a low dimensional BSVD is needed to simplify the hybrid precoder design problem.

In this chapter, the system model is presented in section 4.1 and modified channel model (discussed in chapter 3) is in section 4.2, a low complexity optimal digital precoder BSVD is proposed in section 4.3.2 to achieve lower power consumption for BS and MS based on the CSI from the estimated beamspace channel. Then, in section 4.3.3, given the reduced dimensional BSVD precoder with RF candidate beamforming codebook, the extension of MLOBMP algorithm is proposed to prove that it can reduce the computational complexity of hybrid precoders calculations. Finally, simulation results and conclusion of this chapter will be given in section 4.4 and section 4.5, respectively.

4.1 System Model

Consider a single user mmWave-MIMO system with fully-connected precoding structure shown in Figure 4.1, where N_s independent data streams are sent from the BS with N_t transmit antennas to a single MS with N_r receive antennas. The number of RF chains at Tx denoted as N_{RF}^t such that $N_s \leq N_{RF}^t \leq N_t$.

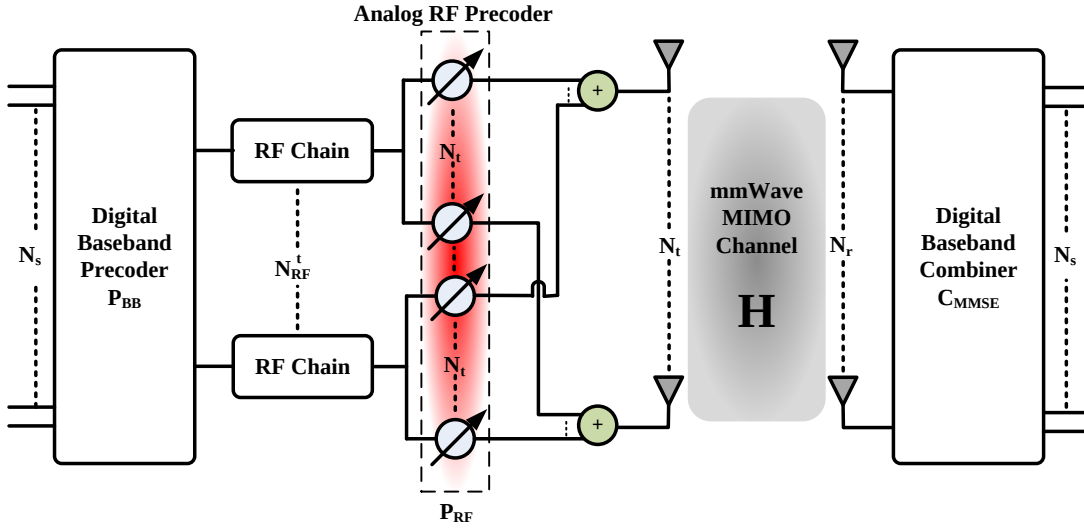


Figure 4.1 Fully-connected structure of single user mmWave system with hybrid precoding implemented using phase shifters and full digital combining

The hybrid precoder at BS which is split into a $N_{RF}^t \times N_s$ digital baseband precoder $\mathbf{P}_{BB}$ and an $N_t \times N_{RF}^t$ analog RF precoder $\mathbf{P}_{RF}$ which is subject to the transmit power constraints $\|\mathbf{P}_{RF}\mathbf{P}_{BB}\|_F^2 = N_s$. The MS, on the other hand, is with $N_r \times N_s$ baseband full digital MMSE combiner $\mathbf{C}_{MMSE}$ that can be approximated using linear estimation methods presented by Kailath, Sayed, & Hassibi (2000) as

$$\mathbf{C}_{MMSE}^H = \frac{1}{\sqrt{\rho}} \left(\mathbf{P}_{BB}^H \mathbf{P}_{RF}^H \mathbf{H}^H \mathbf{H} \mathbf{P}_{RF} \mathbf{P}_{BB} + \frac{\sigma_n^2 N_s}{\rho} \mathbf{I}_{N_s} \right)^{-1} \mathbf{P}_{BB}^H \mathbf{P}_{RF}^H \mathbf{H}^H, \quad (4.1)$$

where the average received power is denoted as ρ .

The $N_r \times 1$ narrowband post processed received signal is expressed as

$$\mathbf{y} = \sqrt{\rho} \mathbf{C}_{MMSE}^H \mathbf{H} \mathbf{P}_{RF} \mathbf{P}_{BB} \mathbf{s} + \mathbf{C}_{MMSE}^H \mathbf{n}, \quad (4.2)$$

where $\mathbf{H}$ is $N_r \times N_t$ channel matrix with $\|\mathbf{H}\|_F^2 = N_t N_r$. The independent and identically distributed $\mathcal{CN}(0, \sigma_n^2)$ noise is denoted as $\mathbf{n}$.

The achievable spectral efficiency with hybrid precoders and full digital combiner can be written as follows

$$R = \log_2 \left| \mathbf{I}_{N_s} + \frac{\rho}{N_s} \mathbf{R}_n^{-1} \mathbf{C}_{MMSE}^H \mathbf{H} \mathbf{P}_{RF} \mathbf{P}_{BB} \mathbf{P}_{BB}^H \mathbf{P}_{RF}^H \mathbf{H}^H \mathbf{C}_{MMSE} \right|, \quad (4.3)$$

where $\mathbf{R}_n = \sigma_n^2 \mathbf{C}_{MMSE}^H \mathbf{C}_{MMSE}$ is the noise covariance matrix after combining and ρ/σ_n^2 is the received SNR.

4.2 Channel Model

Consider 2D clustered channel modeled by Saleh & Valenzuela (1987), with N_{cl} scattering clusters and N_{ray} rays in each cluster as

$$\mathbf{H} = \sqrt{\frac{N_t N_r}{N_c N_{ray}}} \sum_{i=1}^{N_{cl}} \sum_{l=1}^{N_{ray}} \alpha_{il} \Lambda_r(\phi_{il}^r, \theta_{il}^r) \Lambda_t(\phi_{il}^t, \theta_{il}^t) \mathbf{a}_r(\phi_{il}^r, \theta_{il}^r) \mathbf{a}_t(\phi_{il}^t, \theta_{il}^t)^H, \quad (4.4)$$

where $\alpha_{il} \mathcal{CN}(0, \sigma_{\alpha,i}^2)$ is the complex gain of the l^{th} ray in the i^{th} cluster. $\mathbf{a}_r(\phi_{il}^r, \theta_{il}^r)$ and $\mathbf{a}_t(\phi_{il}^t, \theta_{il}^t)$ are AoA and AoD array response vectors with $\phi_{il}^\zeta(\theta_{il}^\zeta), \forall \zeta \in \{t, r\}$ azimuth (elevation) directions. Furthermore, all angles follow a uniform distribution over their mean cluster angles and angular spread. The functions $\Lambda_r(\phi_{il}^r, \theta_{il}^r)$ and $\Lambda_t(\phi_{il}^t, \theta_{il}^t)$ denote the Rx and Tx directional antenna gain given by

$$\Lambda_{\zeta}(\phi_{il}^{\zeta}, \theta_{il}^{\zeta}) = \begin{cases} 1 & \forall \phi_{il}^{\zeta} \in [\phi_{min}^{\zeta}, \phi_{max}^{\zeta}], \forall \theta_{il}^{\zeta} \in [\theta_{min}^{\zeta}, \theta_{max}^{\zeta}], \\ 0 & elsewhere, \end{cases} \quad (4.5)$$

where the unity gain is assumed over the sector $(\phi_l^{\zeta}, \phi_l^{\zeta})$ and $(\theta_l^{\zeta}, \theta_l^{\zeta})$.

For the case of two dimensional $\sqrt{N_{\zeta}} \times \sqrt{N_{\zeta}}$ UPA at BS and MS, the array response vector is given by

$$\mathbf{a}_{\zeta}^{UPA}(\phi_l^{\zeta}, \theta_l^{\zeta}) = \frac{1}{\sqrt{N_{\zeta}}} \left[1, e^{j\frac{2\pi}{\lambda}d\{n_y \sin(\theta) \sin(\phi) + n_z \cos(\theta)\}}, \dots, e^{j\frac{2\pi}{\lambda}d\{(\sqrt{N_{\zeta}}-1) \sin(\theta) \sin(\phi) + (\sqrt{N_{\zeta}}-1) \cos(\theta)\}} \right]^T, \quad (4.6)$$

where $0 \leq n_y \leq \sqrt{N_{\zeta}} - 1$ and $0 \leq n_z \leq \sqrt{N_{\zeta}} - 1$ indicates antenna elements indices at yz plane (Balanis, 2005). Figure 4.2 illustrates 4×4 2D UPA antenna elements in 3D plane.

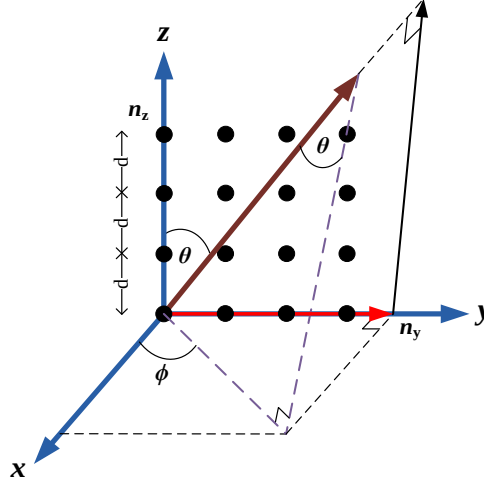


Figure 4.2 Antenna array geometry of UPA in 3D plane

Considering the UPA channel model in mmWave beamforming design is for the following benefits:

- i. Requires smaller antenna dimensions because of higher array gain.
- ii. Extra elevation domain beamforming, i.e. three dimensional beamforming.

- iii. More reliable antenna packing in reasonable array size.

4.3 Hybrid Precoding Digital Combining Design

The objective of this design is to maximize the achievable rate among three possible solutions ($\mathbf{P}_{RF}$, $\mathbf{P}_{BB}$, $\mathbf{C}_{MMSE}$). The optimization problem for hybrid precoder design was discussed in chapter 3.

In next coming subsections, the previously discussed ML fixed weight beamforming that approximates AoDs will be presented for the design of hybrid precoding algorithm based on layered-RF codebook for UPA antenna response vectors. In addition, the design of low complexity BSVD algorithm for approximating optimal unconstrained digital precoder of the spatial domain channel will be proposed.

4.3.1 Layered-RF Codebook Design

Consider a layered-RF orthogonal beamforming matrix defined as

$$\mathbf{L}(:, k_w, k_h) = \frac{1}{\sqrt{N_t}} \left[1, e^{j \frac{2\pi}{\delta N_t} \{(k_w-1)+(k_h-1)\}}, \dots, e^{j \frac{2\pi}{\delta N_t} \{(k_w-1)(\sqrt{N_t}-1)+(k_h-1)(\sqrt{N_t}-1)\}} \right]^T, \quad (4.7)$$

where $1 \leq k_w \leq \delta N_t$ and $1 \leq k_h \leq \delta N_t$. Motivated by Tse & Viswanath (2005), this RF codebook represents the basis for the space spanned by array response vectors with specific directions in 2D surface. As a result, the BS can have a more quantized direction to steer the received signals.

4.3.2 BSVD Digital Precoder Representation

It was proved by Chen, Tsai, Liu, & Wu (2017) that the compressive sensing-based channel estimator can efficiently detect the set $\mathcal{B}$ that corresponds to the lower dimensional beamspace channel, defined later, using estimated channel by Alkhateeb, Leus, & Heath (2015a).

Alternatively, assuming the perfect CSI, the hybrid precoders can be designed based on approximated optimal full digital precoder $\hat{\mathbf{P}}_{opt}$, which is denoted as the largest N_s eigenvalues of estimated channel's right singular vectors. From practical implementations point of view, the following are the benefits of acquiring BSVD for calculating $\hat{\mathbf{P}}_{opt}$:

- Conventional full-dimension SVD operations require complexity of $\mathcal{O}(N_t^3)$ which is considered high for massive MIMO in mmWave channels.
- The $N_r \times N_t$ channel entries requires a large amount of feedback.

To begin, consider the $N_r \times N_t$ beamspace equivalent channel which can be formulated as

$$\mathbf{H}_B = \mathbf{L}_{r,RF}^H \mathbf{H} \mathbf{L}_{t,RF}, \quad (4.8)$$

where $\mathbf{L}_{t,RF}(\mathbf{L}_{r,RF})$ are virtual Tx(Rx) beamforming vectors consisting of RF matrices, i.e., the layered-RF codebook defined in equation (4.7). Since these RF matrices corresponds to the Tx(Rx) candidate matrices, the entries located outside the quantized AoD set $\mathcal{B}$ are equal to zero, hence the $N_r \times L$ lower dimensional beamspace channel $\bar{\mathbf{H}}_B$ estimated by MS can be expressed as

$$\bar{\mathbf{H}}_B = \mathbf{H}_B(:, \mathcal{B}) \quad (4.9)$$

Let's define the relation between estimated beamspace channel and approximated full digital precoders. The SVD of $\bar{\mathbf{H}}_B$ is expressed as

$$\bar{\mathbf{H}}_B = \mathbf{U}_B \Sigma_B \mathbf{V}_B^H \quad (4.10)$$

where $\mathbf{V}_B$ and $\mathbf{U}_B$ are the $N_{cl}N_{ray} \times N_{cl}N_{ray}$ right and the $N_r \times N_r$ left singular matrices whereas Σ_B is the $N_r \times N_{cl}N_{ray}$ eigenvalues of $\bar{\mathbf{H}}_B$. Next, define the autocorrelation matrix of $\bar{\mathbf{H}}_B$ as

$$\mathbf{R}_B = \bar{\mathbf{H}}_B^H \mathbf{H}_B = \mathbf{V}_B \Sigma_B^H \mathbf{U}_B^H \mathbf{U}_B \Sigma_B \mathbf{V}_B^H = \mathbf{V}_B \Sigma_B^H \Sigma_B \mathbf{V}_B^H. \quad (4.11)$$

Identically, the autocorrelation matrix of $\mathbf{H}$ is expressed as

$$\mathbf{R} = \mathbf{H}^H \mathbf{H} = \mathbf{V} \Sigma^H \mathbf{U}^H \mathbf{U} \Sigma \mathbf{V}^H = \mathbf{V} \Sigma^H \Sigma \mathbf{V}^H. \quad (4.12)$$

Notably, $\mathbf{H}$ can be decomposed of $\mathbf{L}_{r,RF} \bar{\mathbf{H}}_B \mathbf{L}_{t,RF}^{(:,\mathcal{B})H}$, hence $\mathbf{R}$ can be given as

$$\mathbf{R} = \mathbf{H}^H \mathbf{H} = \mathbf{L}_{t,RF}^{(:,\mathcal{B})} \bar{\mathbf{H}}_B^H \mathbf{L}_{r,RF}^H \mathbf{L}_{r,RF} \bar{\mathbf{H}}_B \mathbf{L}_{t,RF}^{(:,\mathcal{B})H}. \quad (4.13)$$

Since $\mathbf{L}_{r,RF}$ is RF beamforming matrix, hence we have $\mathbf{L}_{r,RF}^H \mathbf{L}_{r,RF} = \mathbf{I}_{N_r}$, by substituting equation (4.11) into equation (4.12), $\mathbf{R}$ can be derived as

$$\mathbf{R} = \mathbf{L}_{t,RF}^{(:,\mathcal{B})} \bar{\mathbf{H}}_B^H \bar{\mathbf{H}}_B \mathbf{L}_{t,RF}^{(:,\mathcal{B})H} = \mathbf{L}_{t,RF}^{(:,\mathcal{B})} \mathbf{V}_B \Sigma_B^H \Sigma_B \mathbf{V}_B^H \mathbf{L}_{t,RF}^{(:,\mathcal{B})H}. \quad (4.14)$$

Note that $\mathbf{V}$ and $\mathbf{L}_{t,RF}^{(:,\mathcal{B})} \mathbf{V}_B$ are forming the column space of $\mathbf{R}$ by comparing equation (4.12) with equation (4.14). Based on this derivation, optimal full digital precoders acquisition algorithm is proposed. The algorithm for full digital precoder acquisition using BSVD is shown in Table 4.1.

Table 4.1 BSVD optimal full digital precoder approximation

Input: the estimated beamspace channel $\bar{\mathbf{H}}_B$ and the RF beamforming matrix $\mathbf{L}_{t,RF}^{(:,\mathcal{B})}$. Output: approximated optimal full digital precoders $\hat{\mathbf{P}}_{opt}$ and beamspace right singular matrix $\mathbf{V}_B$.	
Step	Process
1:	Perform SVD of $\bar{\mathbf{H}}_B = \mathbf{U}_B \Sigma_B \mathbf{V}_B^H$
2:	Form $\mathbf{V}$ as the combination of $\mathbf{L}_{t,RF}^{(:,\mathcal{B})} \mathbf{V}_B$.
3:	Select the columns of $\mathbf{V}(:, 1: N_s)$ to form $\hat{\mathbf{P}}_{opt}$.

4.3.3 BSVD-Based Hybrid Precoding (BMLOBMP)

The proposed precoding algorithm discussed in chapter 3 can avoid matrix inversion, however, still requires the knowledge about full digital precoders acquired by full dimensional SVD and with high computational complexity as well. Motivated by Chen, Tsai, Liu, & Wu (2017) and based on BSVD developed in previous section, a low complexity BMLOBMP is proposed for hybrid precoding algorithm based on ML fixed weight beamforming.

The redundant calculations resulted from $\hat{\mathbf{P}}_{opt}$ is from the matrix multiplication of $\mathbf{L}_{t,RF}^{(:,\mathcal{B})}$ and $\mathbf{V}_B^{(:,1:N_s)}$. Moreover, for $N_t \times \delta N_t$ RF candidate matrix, the correlation matrix requires $\mathcal{O}(\delta N_t^3 N_s)$ compared to $\mathcal{O}(\delta N_{RF}^t N_t^3 N_s)$ for traditional OMP to calculate the energy spreads on N_t beamforming directions. In BSVD algorithm, $\hat{\mathbf{P}}_{opt}$ is considered as the right singular vectors of $\mathbf{V}$ as shown in Table 4.1, hence the correlation matrix can be expressed as

$$\mathbf{\Psi} = \mathbf{L}_{t,RF}^H \hat{\mathbf{P}}_{opt} = \mathbf{L}_{t,RF}^H \mathbf{L}_{t,RF}^{(:,\mathcal{B})} \mathbf{V}_B^{(:,1:N_s)}. \quad (4.15)$$

For the case of RF matrix with AoDs defined within the set $\mathcal{B}$, the following condition implies

$$\mathbf{L}_{t,RF}^{(:,i)H} \mathbf{L}_{t,RF}^{(:,\mathcal{B})} = \begin{cases} 1, & \forall i \in \mathcal{B} \\ 0, & elsewhere \end{cases} \quad (4.16)$$

By substituting equation (4.16) in equation (4.15), we get

$$\mathbf{\Psi}(i, :) = \mathbf{L}_{t,RF}^{(:,i)H} \mathbf{L}_{t,RF}^{(:,\mathcal{B})} \mathbf{V}_B^{(:,1:N_s)} = \begin{cases} \mathbf{V}_B^{(k,1:N_s)}, & \forall i = \mathcal{B}(k) \\ 0, & elsewhere. \end{cases} \quad (4.17)$$

Equation (4.12) implies that as the index i lies in the estimated AoD set $\mathcal{B}$, the power spread to the estimated AoD is given by the right singular vectors $\mathbf{V}_B^{(k,1:N_s)}$ of the estimated beamspace channel $\bar{\mathbf{H}}_B$, this eliminates the

second step in Table (4.1). Steps of BMLOBMP algorithm are outlined in Table (4.2).

Table 4.2 Hybrid precoder reconstruction using BMLOBMP

Input: right singular vectors $\mathbf{V}_B^{(:,1:N_s)}$ of the beamspace channel $\bar{\mathbf{H}}_B$ and RF beamforming matrix $\mathbf{L}_{t,RF}$. Output: RF precoder $\mathbf{P}_{RF}$ and baseband precoder $\mathbf{P}_{BB}$.	
Step	Process
1:	Implement ML to estimate AoDs then generate the set $\mathcal{B}$ and select the corresponding columns of layered-RF codebook as $\mathbf{L}_{t,RF}^{(:,\mathcal{B})}$.
2:	Calculate the power spread $\mathfrak{P} = \mathbf{V}_B^{(k,1:N_s)} \mathbf{V}_B^{(k,1:N_s)H}$
3:	Initiate an empty set $\mathcal{S}$
4:	for $i = 1$ to N_{RF}^t do:
5:	Find the largest value of diagonal: $k = \arg \max_{n=1,2,\dots,N_{cl}N_{ray}} \{\mathfrak{P}(n)\}$
6:	Concatenate the set: $\mathcal{S} = [\mathcal{S} k]$ and delete the corresponding row: $\mathfrak{P}(k) = 0$
7:	end for.
8:	Build $\mathbf{P}_{RF}$ as $\mathcal{B}_{\mathcal{S}}$ columns of $\mathbf{L}$
9:	Select $\mathcal{S}$ rows and $1:N_s$ columns of $\mathbf{V}_B$ to form $\mathbf{P}_{BB}$
10:	Perform normalization to $\mathbf{P}_{BB}$

Starting with performing AoDs estimation using ML fixed weight beamforming and selecting the corresponding RF codebook matrix columns in step (1), the power spread at N_t directions in step (2) is given by

$$\mathbf{\Theta} = \mathbf{\Psi}(l, :) \mathbf{\Psi}(l, :)^H = \begin{cases} \mathbf{V}_B^{(k, 1:N_s)} \mathbf{V}_B^{(k, 1:N_s)H}, & \forall l = \mathcal{B}(k) \\ 0, & \text{elsewhere.} \end{cases} \quad (4.18)$$

Then from step (4) to (7), the set $\mathcal{S}$ is determined and the RF precoder is formed as

$$\mathbf{P}_{RF} = \mathbf{L}_{t,RF}^{(:, \mathcal{B}_S)}, \quad (4.19)$$

and the baseband precoders can be formulated as

$$\mathbf{P}_{BB} = \mathbf{\Psi}^{(\mathcal{S}, :)} = \mathbf{L}_{t,RF}^{(:, \mathcal{S})H} \hat{\mathbf{P}}_{opt} = \mathbf{L}_{t,RF}^{(:, \mathcal{S})H} \mathbf{L}_{t,RF}^{(:, \mathcal{B})} \mathbf{V}_B^{(:, 1:N_s)} = \mathbf{V}_B^{(\mathcal{S}, 1:N_s)} \quad (4.20)$$

This finding implies that the baseband precoder $\mathbf{P}_{BB}$ is designed by selecting the $\mathcal{S}$ rows of right singular vectors $\mathbf{V}_B^{(:, 1:N_s)}$ of the beamspace channel $\bar{\mathbf{H}}_B$.

4.4 Simulation Results and Performance Evaluation

In this section, numerical simulation results and the performance analysis of the proposed BMLOBMP will be presented.

4.4.1 System Configuration

A single user MIMO system is considered where single and multiple data streams are sent from Tx with $N_t = 64$ to Rx with $N_r = 16$. BS is equipped with ULA, while ULA(UPA) is adopted at MS. Both ULA and UPA are with d antenna spacing of 0.5λ . The channel parameters are given by $N_{cl} = 2$ clusters and $N_{ray} = 4$ rays per cluster. The azimuth (elevation) angles AoD (AoA) are uniformly distributed between $(\pi/2, 3\pi/2)$. The system settings for path loss calculations are given by $f_c = 28 \text{ GHz}$ carrier frequency, $BW = 100 \text{ MHz}$ bandwidth, $d = 50 \text{ m}$ Tx-Rx separation distance, $d_0 = 5 \text{ m}$ close-in distance and $n_{PL} = 3$ path loss exponent. The setting of the power consumption is set same as in chapter 3 and all

simulation results are averaged over 1000 realizations. Table 4.3 summarizes parameters of the system configuration.

Table 4.3 System parameters used in the simulation

Simulation Parameter	Value
BS antenna: N_t	64
MS antenna: N_r	16
antenna element's spacing: d	0.5λ (mm)
path clusters: N_{cl}	2
path rays in each cluster: N_{ray}	4
AoDs (AoAs) distribution	$(\pi/2, 3\pi/2)$
carrier frequency: f_c	28 GHz
Bandwidth: BW	100 MHz
Tx-Rx separation distance: d	50 (m)
close-in distance: d_0	5 (m)
common power of the transmitter: P_{com}	10 (W)
path loss exponent: n_{PL}	3
power of each RF chain: P_{RF}	100 (mW)
Power of phase shifter: P_{PS}	10 (mW)
channel realizations (iterations)	1000

4.4.2 System Performance

Assume all numerical calculations based on hybrid precoding at BS and full digital MMSE combiner given by equation (4.1) at MS where the hybrid precoders term $\mathbf{P}_{RF}\mathbf{P}_{BB}$ is replaced with full digital precoder $\mathbf{P}_{opt}(\hat{\mathbf{P}}_{opt})$. The ML beamforming is used to approximate the AoDs. The approximated full digital precoder is acquired by the BSVD digital precoder algorithm knowing beamspace channel $\bar{\mathbf{H}}_B$. Both BS and MS are

adopt δ layers of RF beamforming candidate matrix with resolution of $\delta N_t(\delta N_r)$.

4.4.2.1 Spectral Efficiency versus SNR

In Figure 4.3, the comparison of spectral efficiency for hybrid precoder is shown using proposed BSVD approximated digital precoder. Here the BS and MS communicate via $N_s = 1, 2$ and 3 data streams with the setting for RF chains is following the condition of $N_{RF} \geq 2N_s$ to examine the comparable algorithm performance in this special case. The total average transmitted power is settled to 7 dB. ULA is assumed to be adopted at BS and MS. As shown in the figure, the optimal rates for the full dimension SVD and the beamspace-based hybrid precoder are following each other and increasing as the number of streams increased due to higher gain achieved by number of RF chains. The two curves show that BSVD technique has less approximation error. This result satisfies the mathematical derivations in section 4.3.2.

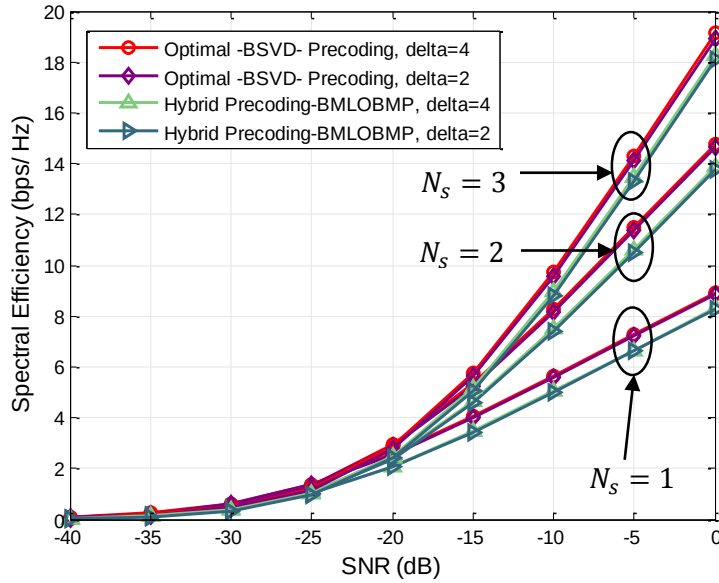


Figure 4.3 Achievable rate of proposed MLOBMP with BSVD assuming ULA at BS and MS

Next, the scenario of adopting ULA at BS and UPA at MS is considered in Figure 4.4 where the BS communicates with the MS via 3 data streams and full dimensional SVD precoding with BSVD-based hybrid precoding. The graph shows the performance of $\approx 93\%$ of unconstrained full digital optimal precoder for the case of 6 RF chains compared to $\approx 88\%$ when $N_{RF} = 4$, this increase in performance will add more complexity to the hardware design. Analog beamforming rate, on the other hand, is still not affected due to different number of N_{RF} RF chains. In Figure 4.5, the curves of SVD and beamspace-based SVD show the accuracy of BSVD (the green line) in acquiring the full digital optimal precoder with low complexity for different number of streams and RF chains.

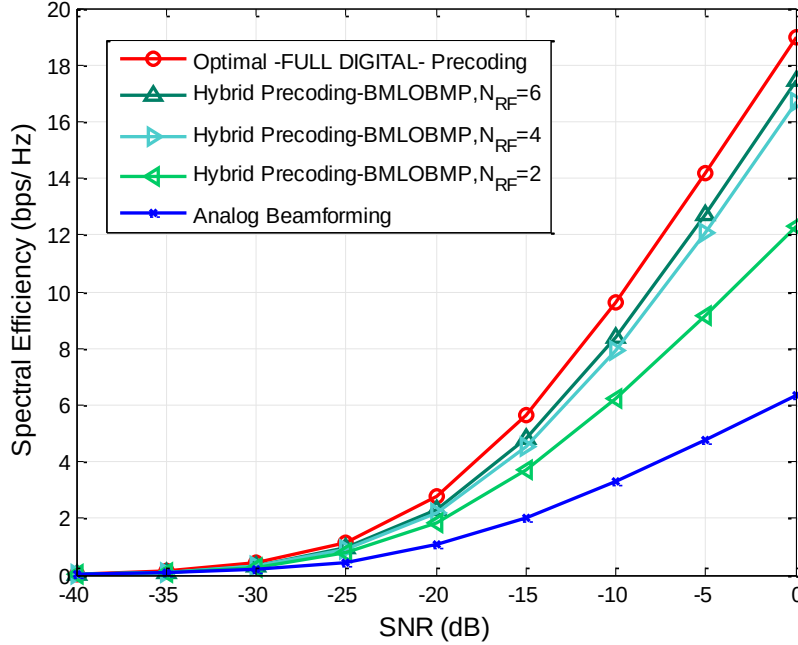


Figure 4.4 Achievable rate of proposed MLOBMP with full dimension SVD assuming ULA(UPA) at BS(MS) and $N_s = 3$

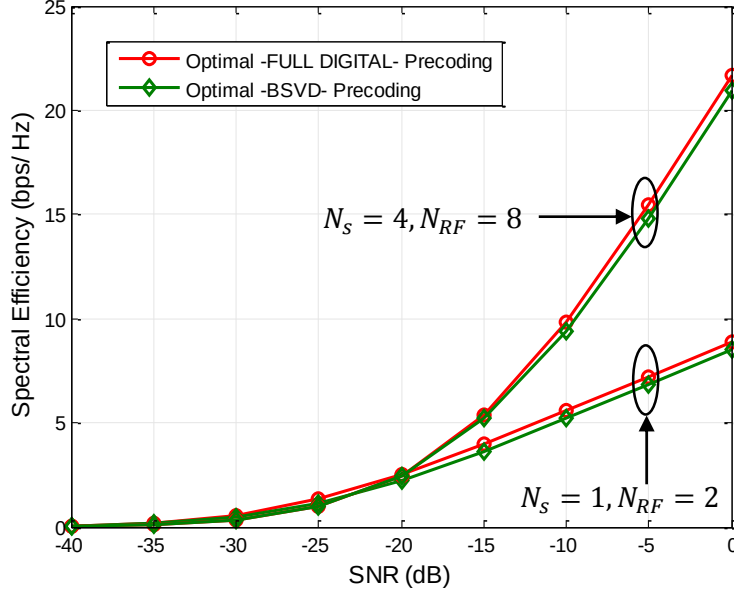


Figure 4.5 Achievable rate of full dimension SVD and proposed BSVD assuming ULA(UPA) at BS(MS)

4.4.2.2 Spectral Efficiency versus RF Chains

Starting with ULA antenna response vectors setting, Figure 4.6 and Figure 4.7 compare the performance of the proposed MLOBMP hybrid precoding when the number of RF chains varies from $N_{RF} = 2$ to $N_{RF} = 8$. The SNR is set to 0 dB. It can be observed that as the number of RF chains increases, the spectral efficiency of the hybrid precoder approximates the optimal full digital precoder and beamspace-based SVD precoder as can be predicted from last simulations. Furthermore, the case of four layers of RF codebook gives better performance than when $\delta = 2$, as can be noted from Figure 4.6. The hybrid rate started to saturate at $N_{RF} \geq 2N_s$ for different N_s data streams. The analog beamforming rate is fixed among the full investigated RF chains range as it does not require information about the number of streams N_s and RF chains N_{RF} . Figure 4.8 shows that the performance loss gap between two optimal cases when BS and MS adopt ULA and UPA, respectively, is very small and hence

validates the accuracy of BSVD precoder ($\hat{\mathbf{P}}_{opt}$) for approximating the optimal precoder $\mathbf{P}_{opt}$.

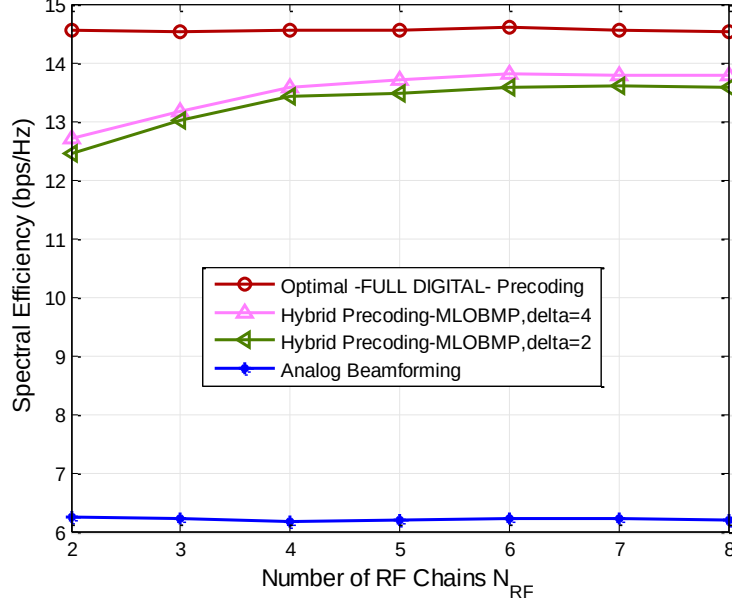


Figure 4.6 Achievable rate vs. N_{RF} of proposed MLOBMP with full dimension SVD assuming ULA at BS and MS, $N_s = 2$

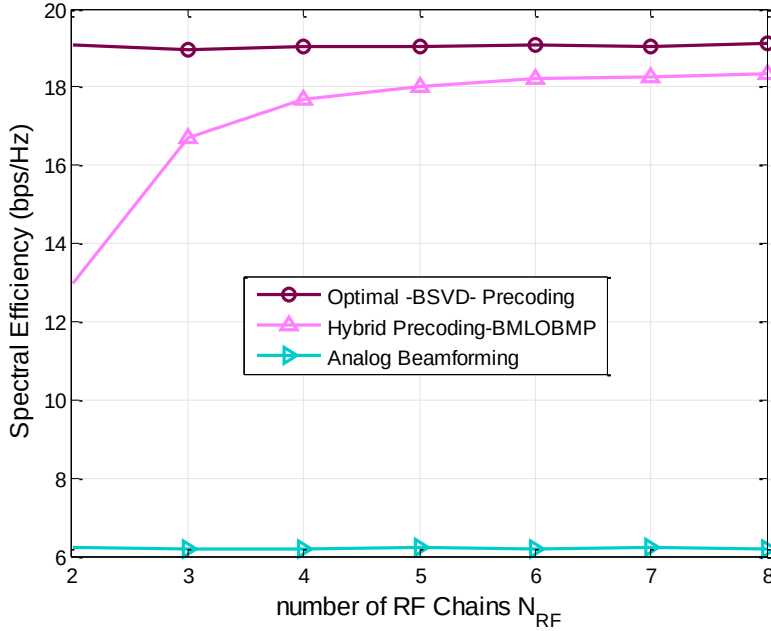


Figure 4.7 Achievable rate vs. N_{RF} of proposed BMLOBMP with BSVD optimal precoder assuming ULA at BS and MS, $N_s = 3$

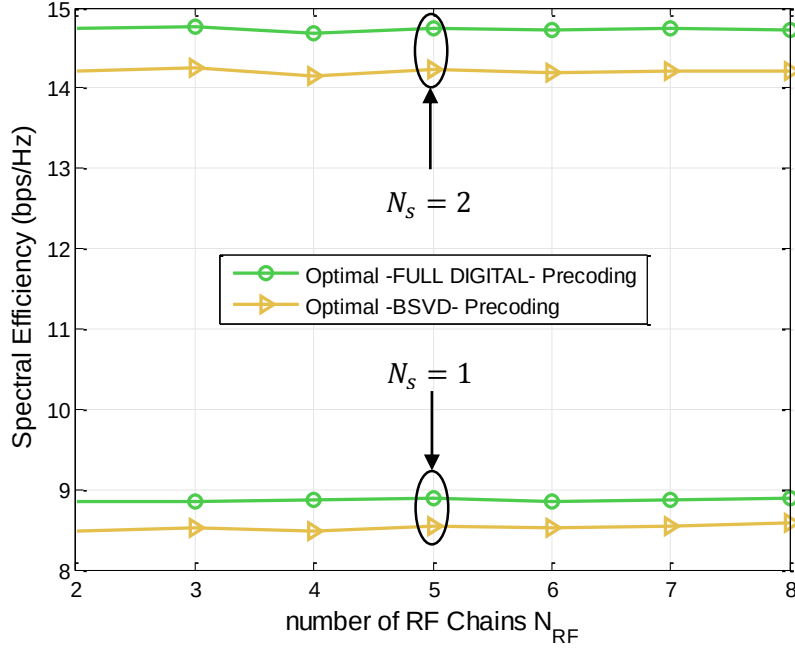


Figure 4.8 Achievable rate vs. N_{RF} of the proposed BSVD and full dimension SVD assuming ULA(UPA) at BS(MS)

4.4.2.3 Energy Efficiency versus SNR

In terms of energy efficiency in bits/Hz/J which was formulated in chapter 3, Figure 4.9 and Figure 4.10 compare the BSVD precoding with the proposed MLOBMP hybrid precoding for different numbers of RF chains and power amplifiers. The increase of power amplifier, as in Figure 4.9, shows a corresponding decrease in optimal and hybrid rates for this figures setups especially at higher SNR. Then, in Figure 4.10, the power amplifier is fixed to 100mW and the optimal/hybrid rates are investigated over a different number of RF chains. The hybrid rate was high when the number of RF chains was $N_{RF}^t = 6$ which achieves $\approx 92\%$ of system performance, for that reason the same assumption was implemented in this figure as started in this section ($N_{RF}^t \geq 2N_s$).

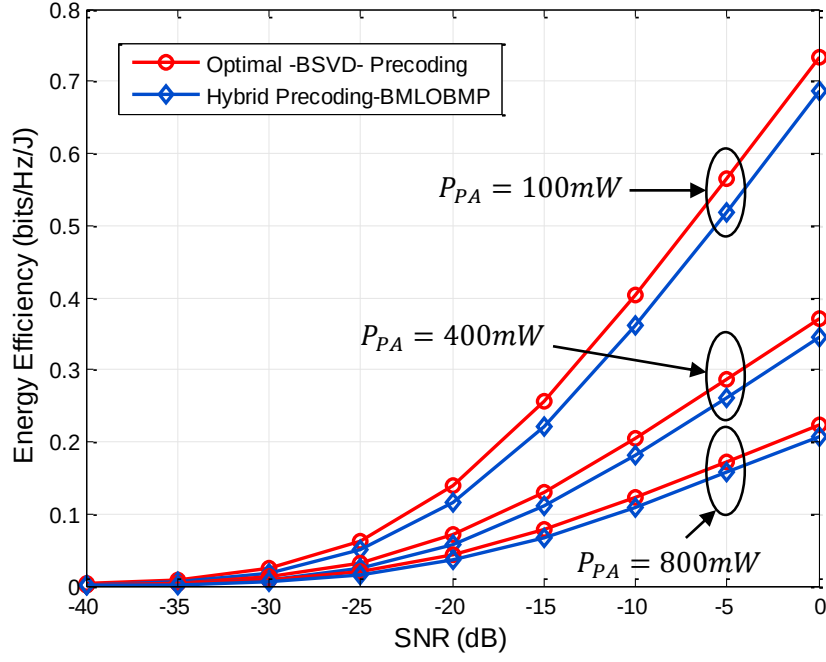


Figure 4.9 Energy efficiency of proposed MLOBMP with BSVD assuming ULA(UPA) at BS(MS), $N_s = 2$ and $N_{RF} = 4$

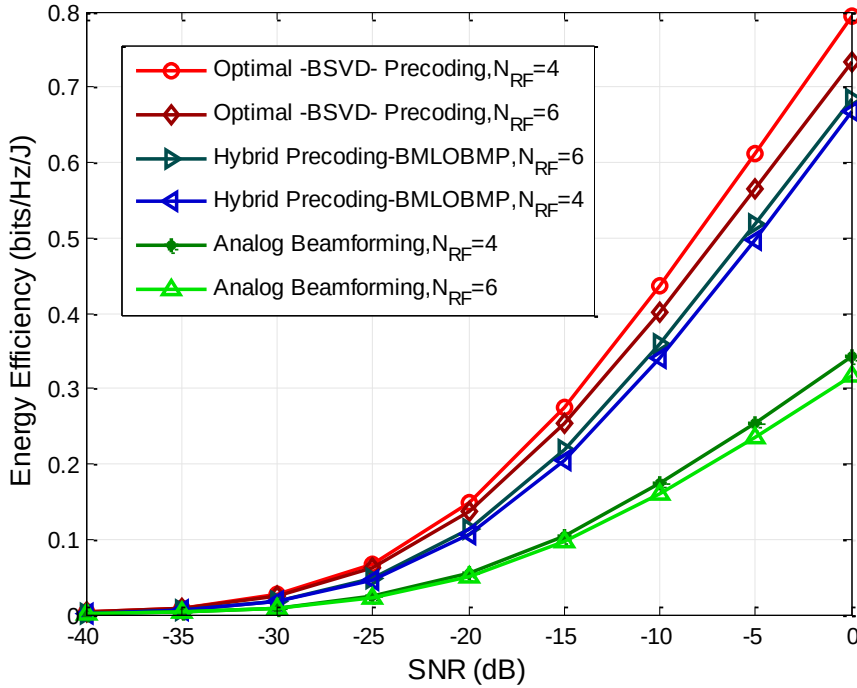


Figure 4.10 Energy efficiency of proposed MLOBMP with BSVD assuming ULA(UPA) at BS(MS), $N_s = 2$

4.4.2.4 Energy Efficiency versus RF Chains

In Figure 4.11, the energy efficiency is plotted as a function of the number of RF chains N_{RF} in the range between 3 to 10 for different number of codebook layers. The performance of the hybrid precoder started to follow the optimal case at $N_{RF} \geq 4$ after it peaked at $N_{RF} = 3$ and decreased at higher RF chains. Moreover, better performance was reported when using more layers of RF codebook since it has more quantized points for selecting the sets of virtual array response vectors. Then, the energy efficiency has been tested again for different power amplifiers and SNR sets in Figure 4.12. The two optimal rates (full-digital SVD and beamspace-based SVD) show less performance loss for all settings of P_{PA} and SNR . In all design cases, the performance is decreased because the energy efficiency is defined as a ratio of spectral efficiency to the total power consumption.

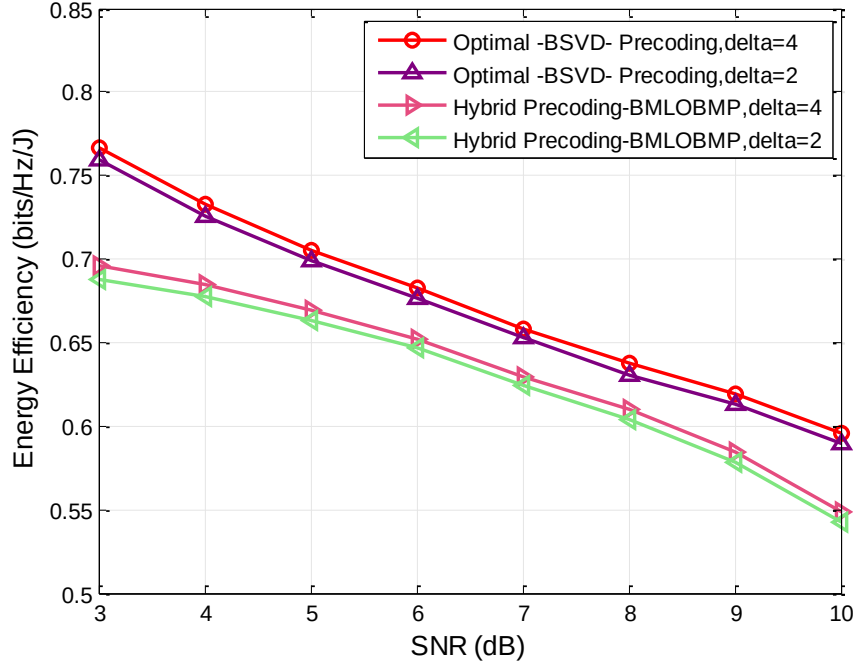


Figure 4.11 Energy efficiency vs. N_{RF}^t of proposed MLOBMP with BSVD assuming ULA(UPA) at BS(MS) and $N_s = 2$

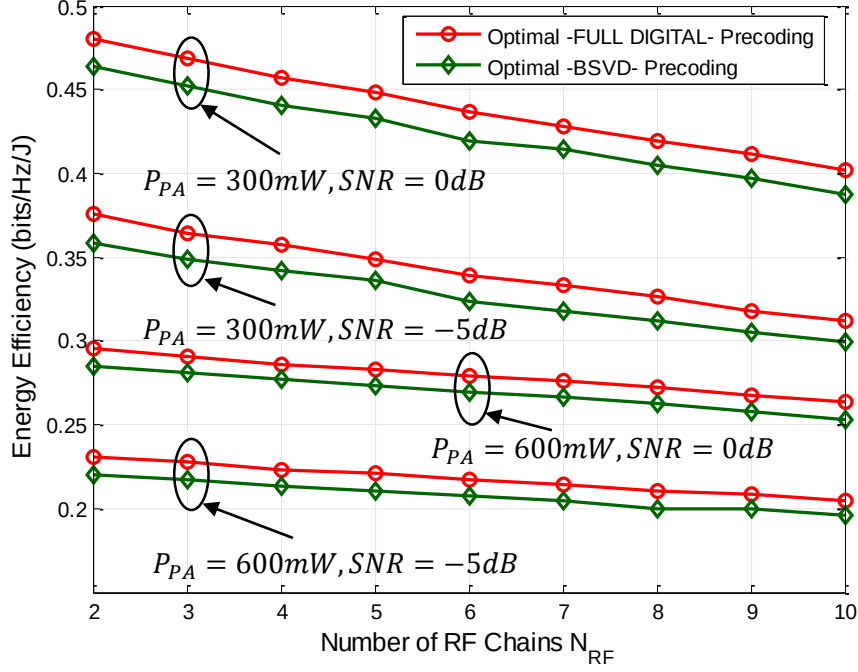


Figure 4.12 Energy efficiency vs. N_{RF}^t of full dimension SVD and BSVD assuming ULA(UPA) at BS(MS) and $N_s = 2$

4.5 Conclusion

In this chapter, a low complexity BMLOBMP hybrid precoding algorithm was proposed for single user mmWave MIMO system. Starting by introducing the modified layered-RF codebook design for UPA antenna response vector as a candidate matrix for the hybrid design problem. Then, a BSVD algorithm was used to acquire the estimated low dimensional full digital precoder was implemented based on perfect CSI of the modeled mmWave channel to reduce the hybrid precoder's computational complexity.

By combining layered-RF codebook, ML fixed weight beamforming and BSVD, a hybrid precoding algorithm was introduced to further simplify the design complexity of the existing MLOBMP discussed in chapter 3.

The modified BMLOBMP algorithm was designed to combine the merits of BSVD and layered-RF codebook, and implement a hybrid precoder with beamSpace-based CSI rather than implicitly in the spatial domain.

The proposed algorithm demonstrates the performance analysis in terms of spectral efficiency and energy efficiency for comparison between optimal and hybrid cases. Different number of data streams, number of RF chains, codebook layers and power amplifier were considered. First, the design of ULA at BS and MS for hybrid precoder with full digital MMSE combiner were evaluated for $N_s = 1, 2$ and 3, and it showed the constant gap with near optimal performance achievements compared to full dimensional SVD and BSVD optimal precoders. Further, the rates of two optimal precoders, i.e., full dimensional SVD and BSVD, were noted near each other among the investigated range of SNR. Another demonstration of achievable rate was considered to explore the effect of number of RF chains on the system performance, and it was reached a specific steady condition at $N_{RF} \geq 2N_s$. In terms of energy efficiency, the performance of the proposed fully connected structure showed the decrease of efficiency over the number of RF chain range. Second, the same performance testing scenarios were conducted when ULA were adopted at BS and UPA were adopted at MS. The analog beamforming design was also shown in some exhibits with no change were noted due to independently effect of various studied parameters changing in the design.

Chapter 5

Conclusion and Future Work

5.1 Conclusion

5.2 Future work

Chapter 5

Conclusion and Future Work

This thesis proposed modified precoding and combining algorithms for a single user mmWave MIMO system based on Maximum Likelihood (ML) AoD estimation and orthogonal beamforming layered-RF codebook. In this chapter, conclusions of numerical results and discussions of thesis outcomes are presented in section 5.1, followed by suggestions of possible extensions and future work in section 5.2.

5.1 Conclusion

The main objective of this thesis is to introduce a low complexity hybrid precoding algorithms with predefined rotated layered-RF codebook with ML estimation of AoD. The purpose is to avoid the iterative process and matrix inversion for calculating the hybrid precoder while achieving same system performance compared to the state-of-art algorithms. In chapter 1, an introduction of mmWave wireless communications has been presented. In chapter 2, an extensive survey on mmWave MIMO systems was conducted starting with different mmWave propagation channel models in section 2.1 including spatial characteristics and beamspace system representation which are used in the beamforming design. Then, MIMO architectures for mmWave with the corresponding analog processing of hybrid beamforming have been discussed in section 2.2 including switches, lenses and phase shifters that can be used for analog RF precoder/combiner structure design. Hybrid precoding/combining techniques have been discussed in section 2.3 for single user mmWave MIMO system. The hybrid precoding design problem has been presented as an optimization problem targeting to minimize the difference in

performance between hybrid and optimal precoders. Complexity of the design for hybrid precoders has been discussed and a comparison of previous research on conventional OMP were studied.

Two cases of multi user MIMO system has been compared in terms of efficiency rate discussing the accuracy of hybrid efficiency rate with respect to single user case.

In chapter 3, quantized AoD were calculated with high accuracy using ML fixed weight beamforming technique. The algorithm has been implemented and low estimation errors are observed. Therefore, ML beamforming is proposed to be part of hybrid precoder and combiner enhanced models.

A modified layered-RF beamforming codebook has been introduced for efficiently approximate the corresponding RF candidate matrix for MLOBMP algorithm. In section 3.3, hybrid precoder has been modeled using orthogonal beamforming matrix and a modified version of the algorithm. In this modified algorithm, the iterative process and matrix inversion for calculating RF and baseband precoders are eliminated, which results in reduced hardware complexity implementation.

In section 3.4, performance analysis shows that this low complexity design can achieves almost same spectral efficiency with the state-of-art hybrid precoding/combining algorithms. Also, it has been shown that the accuracy of hybrid precoder algorithm became higher as it is proportional to the number of RF chains, however, with higher power consumption and RF complexity are required. It has been concluded from the presented results that the power consumption scaled linearly with the number of RF chains. Moreover, the proposed design with fully connected structure has an advantage of being more efficient in terms of achievable data rate. On

the other hand, system performance becomes degraded when considering the energy efficiency.

Instead of dealing with full dimension SVD for calculating unconstrained optimal precoder that has been discussed in chapter 3. In chapter 4, a modified low complexity algorithm BSVD has been proposed. The proposed algorithm is applied for acquiring full digital precoder using estimated beamspace channel for MLOBMP hybrid precoder reconstruction process. In this chapter, full digital MMSE combiner has been considered to be implemented with UPA antenna adopted at MS. After, presenting the system model in section 4.1. Channel model has been discussed in section 4.2. The corresponding design for layered-RF codebook and BMLOBMP hybrid precoding algorithm has been defined in section 4.3. In section 4.4, it has been shown that the performance rate for BSVD was in the acceptable range compared with the optimal full digital SVD at higher SNR for different settings of antenna arrays, number of data streams, number of RF codebook layers and number of RF chains. Furthermore, the proposed BSVD can approximate the achievable rate of the full SVD optimal precoder of almost 96%, which validated the accuracy of BSVD design for MLOBMP hybrid precoding algorithm.

Also, it has been shown that the performance of the proposed algorithm increased as the number of RF chains increased for a fixed SNR.

By comparing the energy efficiency of BSVD and full dimension SVD with single stream downlink transmission, less performance difference with optimum precoder has been reported when adopting planar arrays at MS.

5.2 Future work

Recommendation about future work may include the following

- Modifying the ML fixed weight beamforming AoA/AoD estimation with different range of angles in azimuth and elevation directions for the case of planar arrays (UPA).
- Extending the existing work to partially connected structure to overcome the decrease in energy efficiency.
- Studying the effect of different carrier frequencies (38 GHz, 60 GHz and 72 GHz) on the propagation path loss calculations as compared to 28 GHz which has been used in this thesis.
- Proposing other types of orthogonal RF beamforming matrices as the candidate matrix for hybrid precoder design.
- Applying different fixed weight beamforming techniques as well as adaptive weights beamforming taking the interfering angles noise into consideration.
- Modeling more realistic mmWave channel and different codebook designs, i.e., hierarchical codebook, for the proposed precoding algorithm.
- Considering multi user mmWave MIMO system including inter-cell and intra-cell interference for cellular networks simulations.

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المخلص

تعتبر الموجة الملليمترية للاتصالات اللاسلكية تكنولوجيا حديثة تسمح للجيل الخامس بتحقيق معدلات بيانات جيا ببت لكل ثانية للتطبيقات النقاله في المستقبل. الهدف الرئيسي من هذه الأطروحة هو تصميم نظام تشكيل الحزمة الثنائية المنخفض التعقيد مع الأخذ في الاعتبار الشروط المتعلقة بالتنفيذ العملي للموجة الملليمترية.

في هذه الأطروحة، تتم مراجعة وحدتي ما قبل الترميز وما بعد الترميز، ويقترح إجراء عمليات تنفيذ معدلة حديثا. كما تدرس الخصائص المكانية لنموذج القناة في بنية متعددة المدخلات متعددة المخرجات لاستخدامها في نموذج النظام. يتم عرض مشكلة التصميم كإعادة بناء متفرقة وحلها باستخدام خوارزمية متعامدة الأساس متطابقة التتبع. يستخدم الحد الأقصى من احتمال تشكيل الحزمة الثابتة الوزن جنباً إلى جنب مع خوارزمية متعامدة الأساس متطابقة التتبع لتقدير اتجاه زاوية انطلاق الإشارة. يتم تبسيط تصميم وحدة ما قبل الترميز الثنائي من خلال الجمع بين تصميم أقصى احتمال لتشكيل الشعاع وقاعدة البيانات ذات التردد الكهرومغناطيسي والحزمة المكانية ذات الترميز الرقمي الكامل.

ويتم تقديم أداء التصميم لوحدي ما قبل وما بعد الترميز الثنائية عن طريق تقييم الطيف وكفاءة الطاقة. يتم التحقق من أداء النظام بتغيير معاملات النظام التالية: عدد البيانات المرسله وروابط التردد الكهرومغناطيسي وطبقات قاعدة البيانات ذات التردد الكهرومغناطيسي ونسبة الإشارة إلى الضوضاء وقيم استهلاك الطاقة. وتبين نتائج المحاكاة قرب الكفاءة المثلى (أعلى من تسعين في المائة) مقارنة بالحزمة الرقمية الكاملة غير المقيدة لأعلى معدلات إشارة إلى الضوضاء وروابط التردد الكهرومغناطيسي. تتم مقارنة النتائج من حيث اختلاف درجة صعوبة التنفيذ العملي ويظهر أنه مع التصميم المعدلة يتم تحقيق انخفاض فعال في العملية التكرارية. أيضاً، هناك حاجة إلى حساب أقل لمصفوفة الانعكاس بالمقارنة مع آخر ما توصلت له الخوارزميات الحديثة.

وعلاوة على ذلك، يجري التحقق من تأثير استخدام مصفوفة تخطيط موحدة في جهة المستخدم. وتبين الأطروحة أن اكتساب الحزمة المكانية ذات الترميز الرقمي الكامل مع الحد الأقصى المرجح لتشكيل الحزمة الثابتة هو التنفيذ الأمثل لتحقيق مشابه لأفضل وحدة ما قبل الترميز الرقمية الكاملة للموجة الملليمترية.