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Hybrid Beamforming for 5g and Cognitive Radios

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Abstract:
The pathloss experienced by the proposed mmwave for the 5G spectrum, is much more than the microwave signals currently used in most wireless systems. The decrease in wavelength of the mmWave, makes possible the use of large scale antenna arrays (LSA). The beamforming gain provided by these arrays helps us to overcome pathloss. Beamforming with multiple data streams is known as precoding and increases the spectral efficiency. Beamforming and precoding in 4G are done digitally at baseband for multi-antenna systems. The high cost and power consumption of digital precoding in mmWave LSA systems, however, make analog processing in the analog domain more cost efficient. The RF chain structure provides a hardware limitation to purely digital precoders. The hybrid beamforming architecture provides a solution to utilize the sparse structure of the mmWave channel and to leverage the beamforming potential of the large antenna arrays. The well-known algorithm of Orthogonal Basis Pursuit is used to approximate optimal unconstrained precoders and combiners. Cognitive radios attempt to share spectra by introducing a spectrum sensing function, so that they are able to transmit in unused portions at a given time, place and frequency. In this paper we propose an underlay cognitive transceiver design that enable the mmWave spectrum access while negating the interference of the Primary User. As such the results of applying the Hybrid beamforming architecture is shown comparatively for the presence and absence of the Primary User.

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I. INTRODUCTION

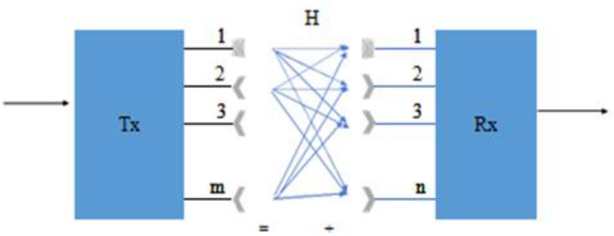

Figure 1.1: MIMO Transceiver structure

The term MIMO (Multiple-input multiple-output) refers to the usage of more than one antennas at both transmitter(Tx) and receiver(Rx). The transmission rate of a MIMO system increases linearly with respect to the number of antennas at both Tx and Rx. Additionally MIMO also enhances link reliability and improves coverage. The benefits of MIMO can be realized when the Rx as well as the Tx knows the channel. Transmit or receive channel knowledge significantly increases the benefits of MIMO[14].

Using and exploiting transmit channel side information is of great practical interest in MIMO wireless system. In this project we assume full channel knowledge and study how channel-side information (CSI) at the transmitter as well as the receiver can be used to improve link performance.

Precoding is a processing technique that uses CSI at the Tx and processes the signal before transmission. A precoder essentially functions as a multimode beamformer, optimally matching the input signal on one side to the channel on the other side. It does so by splitting the transmit signal into orthogonal spatial eigenbeams and assigns higher power along the beams where the channel is strong but lower or no power alongthe weak. Precoding design varies depending on the types of CSI and the

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performance criterion. Similarly, combining utilizes CSI at the Rx to recover the signal. Precoding and combining can be used together to bring down interference.

Figure 1.1 depicts a MIMO system with m transmit antennas and n receive antennas. The channel is given by H. The equation of the received signal is y = Hx + n where y is the $n \times 1$ received vector, x is the $m \times 1$ transmitted vector and n is the M 1 received vector.

1.1 The 5G spectrum

5G is the term used to describe the next-generation of mobile networks beyond LTE mobile networks. The fifth generation (5G) wireless communication system is to provide solutions for the exponential increase of mobile data traffic by 2020. Though research in 5G is at an initial stage, some trends have appeared in how to design 5G networks. Some of the key technologies are - large-scale antenna systems (LSAS), non-orthogonal multi-plex access, full duplex, spectrum sharing, high-frequency bands (e.g., millimeter-wave, mmWave spectrum), high-density networks, new network architecture, new waveform design, and so on. These technologies may have great potential in system performance improvement in 5G.[4]

The use of more than one antennas at both sides of the wireless channel is known as the MIMO technology. It is a key approach to achieve high spectral efficiency (SE). In this project we assume a single user. Effectively we assume a single Tx and a single Rx. The 5G spectrum allows the use of massive MIMO where the number of antenna elements at the BS reaches be in hundreds. This allows increasing the number of data streams in the link to very large values. Massive MIMO improves the link efficiency and simplifies precoding at the Tx and combining at the Rx. It also makes the channel statistically quasi-stable by channel hardening.

As we move into the 5G spectrum, massive MIMO becomes an essential asset, since the high free-space path loss at those frequencies necessitates large array gains to obtain sufficient signal-to-noise ratio (SNR), even for smaller distances like 100 m. The large number of antenna elements in massive MIMO also pose major challenge in the form of a large number of radio frequency (RF) chains required for digital precoding and combining used in 4G. These RF chains (one for each antenna element) consumes a considerableamount of energy. Another drawback is determining the channel state information (CSI) between each transmit and receive antenna, which takes up a considerable amount of spectral resources.

The proposed solution to these problems lies in the concept of hybrid beamformers, which use a combination of analog beamformers in the RF domain, together with digital beamforming in the baseband, connected to the RF with a smaller number of up/down conversion chains. The main motivation of hybrid beamforming is that the number of RF chains is only lower-limited by the number of data streams that are to be transmitted, while the beamforming gain depends on the number of antenna elements if suitable RF beamforming is done.

Fundamentally MIMO processing in 4G is performed digitally at baseband, with RF chains controlling both the signal’s phase and amplitude. This digital processing requires dedicated baseband and RF hardware for each antenna element as well as perfect CSI(Channel State Information). The high cost and power consumption of purely digital processing for mmWave hardware, with a larger number of antennas, does not justify such a transceiver architecture at present. This makes us forego digital processing and use the beamforming gain of RF or analog processing in the mmWave. The analog processing reduces the number of RF chains.

1.2 Hybrid Beamforming

As we move into the 5G spectrum of wireless communications, the channel changes. It is characterized by a high degree of pathloss, and attenuation due to rain and atmospheric moisture.

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Fortunately for this spectrum the antenna size decreases to the order of millimeters, allowing us to pack more antennas in a smaller space, making it easier to achieve beamforming gains[5]. Hybrid beamforming is a combination of both analog and digital beamforming. It takes into account the sparse structure of the channel as well as the hardware limitations. The number of RF chains which is one per antenna for purely digital beamforming is reduced, thereby reducing hardware complexity and power consumption.

1.3 Cognitive Radio

Cognitive radio (CR) is essentially a radio aware of its environment. It supports dy-namic spectrum access to address the issue of spectrum scarcity as given in [2]. Cognitive radio is a convergence of several functionalities in a smart radio thereby providing a base for useful applications such as dynamic spectrum access. Still in its initial stage, the concept of cognitive radio requires inputs from several fields on different issues. These are smart antennas, hardware architectures including software-defined radio (SDR), signal processing, networking et cetra. In addition to the hybrid beamforming mentioned above, cognitive radio too is a technology which can be utilized in the 5G spectrum. The mmWave communications assumed above is a standalone wireless MIMO system. Looking into the future, the mmWave bands are highly allocated to several services (e.g. point-to-point (P2P) or point-to-multipoint (P2MP) backhaul microwave links, satellite links et cetra). This implies an increase in traffic. This makes it essential to place constrain the interference from nearby wireless systems into account in the design of mmWave transeivers.

II. CHANNEL MODEL

As we move into the mmWave spectrum the properties and behaviour of the channel changes. The channel model used here is characterized by high free-space pathloss which is a property of mmWave channel. This leads to limited scattering of the mmWave and a lesser number of multipaths . MmWave transceivers usually have large tightly-packed antenna arrays that lead to high levels of antenna correlation. The large number of antennas enable a beamforming gain to overcome the high pathloss of the mmWave channel.

2.1 The Saleh-Valenzuela model

Saleh Valenzuela model is commonly used for modelling multipath in indoor environment. In this model, waves arriving from similar directions and delays are grouped into clusters. A mean AOA (Angle of arrival) or AOD(Angle of departure) is associated with each cluster, within a cluster there are any number of rays, and the AOAs/AODs of the rays within the same cluster are assumed to be distributed according to a certain probability density function (pdf). Figuratively the clustered channel model is shown in figure 4.1. In this thesis it is assumed that the AOAs/AODs within a cluster follow a Laplacian distribution within the cluster. In order to understand the channel model viable for the sparse scattering environment in 5G, we assume a MIMO communication link with $M_t$transmit antennas arranged in a ULA and $M_r$receive antennas in a ULA. Since we are taking a ULA, we are assuming power to be distributed only in the azimuthal plane. The channel matrix for the geometric ray tracing model, for a single cluster having N number of rays is given as

$$H = \sum_{i=1}^{N} \alpha_{il} a_r(\phi_i^r) a_t(\phi_i^t) \qquad (2.1)$$

Where N is the number of rays in a cluster, $\alpha_{il}$is the Rayleigh fading coefficient related to the i^{th} ray in the cluster, $\phi_i^r$is the angle of arrival at the Rx and $\phi_i^t$is the angle of departure at the Tx, of the i^{th} ray in the cluster based on the extended Saleh-Valenzuela model. The antenna responses at the Rx and the Tx are given by

$$a_r(\phi_i^r) = \left[1\,, e^{\phi_1(\phi_{l,i}^r)}, \dots . e^{\phi_m(\phi_{l,i}^r)}\right] \qquad (2.2)$$

$$a_t(\phi_i^t) = \left[1\,, e^{\phi_1(\phi_{l,i}^t)}, \dots . e^{\phi_m(\phi_{l,i}^t)}\right] \qquad (2.3)$$

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The traditionally used statistical distributions for fading channels used in conventional MIMO analysis or for the 4G spectrum do not work in 5G. A ray tracing approach is used here to model the clustered channel H given by

$H = A_r H_\alpha A_t^T$ (2.4)

Where the matrices A_r and A_t are given by

$$A_r = [a_r(\phi_{l,1}), a_r(\phi_{l,2}), \ldots a_r(\phi_{l,N})] \quad (2.5)$$

$$A_t = [a_t(\phi_{l,1}), a_t(\phi_{l,1}), \ldots a_t(\phi_{l,1})] \quad (2.6)$$

The above analysis gives us a simple structure of channel model H with linear arrays of antennas or ULAs. The channel used in the Hybrid beamforming system uses UPAs or the uniform planar array. These enable 3D beamforming unlike ULAs. The antenna response of a UPA is given by

$$a_{UPA}(\phi,\theta) = \frac{1}{\sqrt{N}}\left[1, \ldots, e^{jkdm(\sin(\phi)\sin(\theta)+n\cos(\theta))}, \ldots\ldots, e^{jkd(W-1)(\sin(\phi)\sin(\theta)+(H-1)\cos(\theta))}\right] \quad (2.6)$$

The y and z indices of an antenna element respectively are given by $0 < m < B$ and $0 < n < C$ and the antenna array size is $N = BC$. Planar arrays in the form of UPAs or uniform planar arrays are of great interest in 5G as they enable 3D beamforming.

The matrix H is modelled using the Saleh-Valenzuela channel model. This is a clustered channel model. The matrix H ,is assumed to be the sum of energy of propagation paths from N_{Cl} clusters each of which have N_{ray} rays.

$$H = \sqrt{\frac{N_t N_r}{N_{Cl} N_{ray}}} \sum_{i=1}^{N_{Cl}} \sum_{l=1}^{N_{ray}} \alpha_{il} \Lambda_r(\phi_{il}^r, \theta_{il}^r) \Lambda_t(\phi_{il}^t, \theta_{il}^t) a_r(\phi_{il}^r, \theta_{il}^r) a_t(\phi_{il}^t, \theta_{il}^t)^* \quad (2.7)$$

α_{il} = complex fading coefficient of the l^{th} ray in the i^{th} scattering cluster

ϕ_{il}^r = azimuth angle of arrival (AOA) at the receiver

θ_{il}^r = elevation angle of arrival (AOA) at the receiver

ϕ_{il}^t = azimuth angle of departure (AOD) at the transmitter

θ_{il}^t = elevation angle of departure at the transmitter

$\Lambda_r(\phi_{il}^r, \theta_{il}^r)$ = antenna gains at the receiver

$\Lambda_t(\phi_{il}^t, \theta_{il}^t)$ = antenna gains at the transmitter

The complex fading coefficient α_{il} is assumed to be independent identically distributed with a normal distribution represented by $CN(0; \sigma_{\alpha,i}^2)$ where $\sigma_{\alpha,i}^2$ represents the

average power of the i^{th} cluster. These cluster power satisfies the constraint $\sum_{i=1}^{N_{Cl}} \sigma_{\alpha,i}^2 = \gamma$

where γ is a normalization constant that satisfies

$$E[\|H\|_F] = \gamma$$

The antenna sectored gain at the Tx are found to be mathematically

$$\Lambda_r(\phi_{il}^t, \theta_{il}^t) = 1 \text{ for } \forall \phi_{il}^r \in (\phi_{min}^t, \phi_{max}^t), \forall \theta_{il}^r \in (\theta_{min}^t, \theta_{max}^t)$$

$$\Lambda_r(\phi_{il}^t, \theta_{il}^t) = 0 \text{ otherwise} \quad (2.8)$$

This assumes a unit power gain over the sector given by $\phi_{il}^t \in (\phi_{min}^t, \phi_{max}^t)$ and $\forall \theta_{il}^t \in (\theta_{min}^t, \theta_{max}^t)$.

The antenna sectored gain at the Rx are found to be mathematically

$$\Lambda_r(\phi_{il}^r, \theta_{il}^r) = 1 \text{ for } \forall \phi_{il}^r \in (\phi_{min}^r, \phi_{max}^r), \forall \theta_{il}^r \in (\theta_{min}^r, \theta_{max}^r)$$

$$\Lambda_r(\phi_{il}^r, \theta_{il}^r) = 0 \text{ otherwise} \quad (2.9)$$

This assumes a unit power gain over the sector given by $\phi_{il}^r \in (\phi_{min}^r, \phi_{max}^r)$ and $\forall \theta_{il}^r \in (\theta_{min}^r, \theta_{max}^r)$. The antenna array responses at the Tx and the Rx are given by $a_t(\phi_{il}^t, \theta_{il}^t)$ and $a_r(\phi_{il}^r, \theta_{il}^r)$. These functions are a function of antenna array structure and are inde-pendent of a single antenna element properties.

2.2 Motivation for Hybrid Beamforming

There is an increasing demand of wireless data traffic which increases demands on spectral efficiency (SE) and bandwidth. The LTE wireless technology currently used operates in 300 MHz to 3 GHz band. As we move into 5G wireless networks, we explore high-frequency mmWave band ranging from 3 GHz to 300 GHz. One of the ways to increase SE is MIMO technology, the use of multiple antennas at Tx and Rx.

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The 5G networks have been created to meet with the future increase in data rate and the significant increase in capacity that goes with it. Massive MIMO provides capability for enhancing capacity, SE and energy efficiency (EE). The disadvantage of the mmWave frequencies is that they suffer from high path loss. The traditional baseband digital beamforming, as done in 4G, if done in massive MIMO systems has the necessity of one individual radio frequency (RF) chain per antenna. These RF chains are high on power consumption and costly. This makes us choose hybrid beamforming which operates in the digital and as well as analog domains. Analog beamforming or beamforming in the RF domain gives us a high beamforming gain to overcome pathloss.

Following are the advantages of hybrid beamforming:

- Capacity: We predict an increased demand in the data rate and capacity. This can only be achieved by massive MIMO. Hybrid beamforming takes advantage of the sparse nature of the channel to tone down interference between users.
- The incorporation of analog or RF processing along with digital beamforming used in 4G reduces the cost and complexity of the hybrid beamforming system.[49]
- Massive MIMO or the LSAS used in hybrid beamforming reduces the transmission power, thereby increasing energy efficiency.

The large number of antennas allows the use of low cost RF amplifiers in the milliWatt range, because the total transmitted power $P_t = \frac{1}{N_t}$. This in contrast to the classical antenna arrays, which use few antennas fed from a high-power amplifier at low power efficiency.

III. HYBRID BEAMFORMING

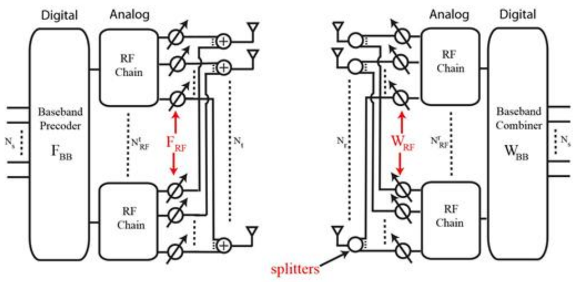

Figure 3.1: Hybrid Beamforming Architechture

3.1 Hybrid Beamforming Architechture

Hybrid beamforming uses a combination of both analog and digital precoding in order to exploit the sparse structure of the mmWave channel. A single-user mmWave system modelled according to [4] is shown in Fig 3.1. The Tx has N_t number of antennas, through which it communicates N_s data streams to the Rx. The Rx has N_r antennas. The Hybrid BF structure we have deployed has a much lower number of RF chains than the total antenna number (N_t, N_r), and is therefore more practical and cost effective as it reduces hardware complexity and power consumption. At the Tx this is done using analog BF, represented by the matrix F_{RF} where the phase of the transmitted signal on each antenna is controlled via a grid of analog phase shifters. The dimensions of this F_{RF}matrix is $N_t \times N_t^{RF}$ Digital BF, represented by the matrix F_{BB} over transceivers can then be utilized to achieve multiple data stream precoding on top of analog BF to further enhance the performance. Multistream communication is enabled at the transmitter by providing it with N_t^{RF} transmit chains such that $N_s \leq N_t^{RF} \leq N_t$. The dimensions of $F_{BB}$is therefore $N_t^{RF} \times N_s$. This explains the hardware architecture at the Tx. As we can see the number of RF chains $N_t^{RF}$is reduced, reducing hardware complexity and power consumption.

The symbol vector to be transmitted is given by s and is of the dimension $N_S \times 1$ and satisfies the constraint $[ss^*] = \frac{1}{N_s} I_{N_S}$. The transmitted signal in discrete time is given as$x = F_{RF} F_{BB} s$. The analog beamforming matrix F_{RF} is formed using phase shifters in the RF domain, and its elements are constrained to satisfy $\left(F_{RF}^{(i)} F_{RF}^{(i)*}\right) = N_t^{-1}$ (where i denotes the i^{th} diagonal element of a matrix). These phase shifters merely change the phase of the signal without altering its amplitude. This implies that all elements of F_{RF} have equal norm. We enforce the

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transmitter's total power constraint by normalizingF_{BB} such that $\|F_{RF}F_{BB}\|_F{}^2 = N_s$. Other hardware constraints are placed on the baseband precoder.

The received signal is given as $y = \sqrt{\rho}Hx + n$ (3.1)

The dimensions of the received signal are $N_r \times 1$. The channel matrix H has dimensions $N_r \times N_t$, satisfying the constraint $E[\|H\|_F] = N_r N_t$, ρrepresents the average received power, and n is vector of noise having Gaussian symbols which are independent identically distributed with zero mean and unit variance. We assume perfect timing and frequency recovery. The knowledge of the channel H is also assumed to be known instantaneously at the transmitter and the receiver.

At the Rx there are N_r^{RF} RF chains which satisfy the constraint $N_s \le N_r^{RF} \le N_r$ and with the analog phase shifters to give the received processed signal as

$$y = \sqrt{\rho}W_{BB}W_{RF}HF_{RF}F_{BB}x + W_{BB}W_{RF}n \quad (3.2)$$

The analog combining matrix is given by W_{RF} and its dimensions are given by $N_r \times N_r^{RF}$. The baseband combining matrix W_{BB} is of dimension $N_r^{RF} \times N_s$. The analog combining matrix, W_{RF} is formed using phase shifters and therefore is such that

$$\left(W_{RF}^{(i)}W_{RF}^{(i)*}\right) = N_r^{-1} \quad (3.3)$$

Assuming Gaussian symbols are transmitted over the mmWave channel, the spectral efficiency achieved is given by

$R =$
$log_2\left(\left|I_{N_s} +\right.\right.$
ρNSRn−1WBB∗WRF∗HFRFFBBFBB∗FRF∗WRF WBB (3.4)

where $R_n = \sigma_n^2 W_{BB}^* W_{RF}^* W_{RF} W_{BB}$ is the noise covariance matrix?

3.2 Spatially Sparse Precoding for The Single User

In order to find the analog and digital beamforming matrices given by $F = F_{RF}F_{BB}$ that optimize SE at the Tx and Rx, the constraint imposed by the constant amplitude phase shifters and the number of RF chains should be taken into account. The main difficulties include: i) the four matrices mandatory for the hybrid transceiver design given by $F_{RF}, F_{BB}, W_{RF}, W_{BB}$ are linkes, making the objective function of the resulting optimization non-convex, ii) the analog precoder/combiner F_{RF}, W_{RF} are characteristically realized as a phase-shifter grid, forcing constant modulus constraints on the elements of the analog precoding and combining matrices F_{RF} and W_{RF}. These are assembled using fixed modulus phase shifters, this places a non-convex constraint on the above problem. Seperating the precoding and combining optimization problem, we initially focus on the design of the hybrid precoders$F_{RF}F_{BB}$ by maximizing the mutual information achieved by transmit-ting Gaussian symbols over the mmWave channel

$$I(F_{RF}, F_{BB}) = log_2\left(\left|I_{N_s} + \frac{\rho}{N_S\sigma_n^2}HF_{RF}F_{BB}F_{BB}^*F_{RF}^*\right|\right) \quad (3.5)$$

It is assumed here that the Rx can perform optimal nearest neighbour decoding based on the received signal of N_r symbols. The optimization problem for the design of optimal precoders is given as

$\left(F_{RF}^{opt}, F_{BB}^{opt}\right) = \max_{F_{RF}, F_{BB}} log_2\left(\left|I_{N_s} +\right.\right.$
ρNSσn2HFRFFBBFBB∗FRF∗ (3.6)

Such that $F_{RF} \in \mathcal{F}_{RF}$ and $\|F_{RF}F_{BB}\|_F^2 = N_s$

where $\mathcal{F}_{RF}$is the set of feasible RF precoders, that is the set of $N_t \times N_t^{RF}$matrices having elements with constant magnitude or unity modulus with different phase. The constraint $F_{RF} \in \mathcal{F}_{RF}$, is a non-convex constraint. The objective function we assume here that is the mutual information is maximized by optimizing the hybrid precoder. This is done by reducing the distance between the optimal unconstrained precoder F_{opt} and the F_{RF}, F_{BB}. The first step is to apply the Singular Value Decomposition or SVD to the channel matrix H.

$$H = U\Sigma V \quad (3.7)$$

Rewriting (3.6)

$$I(F_{RF}, F_{BB}) = log_2\left(\left|I_{N_s} + \frac{\rho}{N_S\sigma_n^2}\Sigma^2V^*F_{RF}F_{BB}F_{BB}^*F_{RF}^*V\right|\right) \quad (3.8)$$

The objective function of mutual information maximized here by optimizing the hybrid precoder. This is done by reducing the distance between the optimal unconstrained precoderF_{opt} and the F_{RF}, F_{BB}.Partitioning the matrices Σ and V as

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$\Sigma = \begin{bmatrix} \Sigma_1 & 0 \\ 0 & \Sigma_2 \end{bmatrix}$ (3.9)

$V = \begin{bmatrix} V_1 & 0 \\ 0 & V_2 \end{bmatrix}$ (3.10)

The matrix Σ_1 is of dimensions $N_s \times N_s$, and the matrix $V_1$is of dimensions $N_t \times N_s$. The fully purely digital precoder for the matrix H is given by $F_{opt} = V_1$. Since the matrix F_{RF}, is fixed it is impossible for the hybrid precoding matrix to be written as $F_{opt} = F_{RF}F_{BB}$

The optimization problem can be simplified using a set of assumptions for the mmWave channel. The mmWave parameters denoted as (N_t ; N_r ; N_t^{RF} ; N_r^{RF}) are taken do as to make the hybrid precoders $F_{RF}F_{BB}$ to be as close to the unitary precoder $F_{opt} = V_1$ as possible. The following two approximations justify the mathematical closeness

1. The matrix given by $I_{N_s} - V_1^* F_{RF} F_{BB} F_{BB}^* F_{RF}^* V_1$ has approximately strong eigen values. For the case precoding in mmWave this can also be written as $V_1^* F_{RF} F_{BB} \approx I_{N_s}$.

2. The matrix $V_2^* F_{RF} F_{BB}$ has approx very small singular values, or we can say that $V_2^* F_{RF} F_{BB} \approx 0$.

The above approximations are assumed for the mmWave channel. The characteristics of these channel include

1. A large number of antennas N_t.

2. A number of transmit chains holding the relation $N_s \leq N_t^{RF} \leq N_t$

3. A correlated channel H.

The above approximations make us simplify the mutual information further taking help of the above defined partitions

$$V^* F_{RF} F_{BB} F_{BB}^* F_{RF}^* V = \begin{bmatrix} V_1^* F_{RF} F_{BB} F_{BB}^* F_{RF}^* V_1 & V_1^* F_{RF} F_{BB} F_{BB}^* F_{RF}^* V_2 \\ V_2^* F_{RF} F_{BB} F_{BB}^* F_{RF}^* V_1 & V_2^* F_{RF} F_{BB} F_{BB}^* F_{RF}^* V_2 \end{bmatrix} \quad (3.11)$$

This approximates the mutual information provided by $F_{RF}F_{BB}$ as

$$I(F_{RF}, F_{BB}) = log_2\left(\left|I_{rank\,(H)} + \frac{\rho}{N_S\sigma_n^2}\Sigma^2 V^* F_{RF} F_{BB} F_{BB}^* F_{RF}^* V\right|\right) \quad (3.12)$$

Applying the various approximations above gives us

$$I(F_{RF}, F_{BB}) = log_2\left(\left|I_{rank\,(H)} + \frac{\rho}{N_S\sigma_n^2}\Sigma_1^2 V_1^* F_{RF} F_{BB} F_{BB}^* F_{RF}^* V_1\right|\right) \quad (3.13)$$

Further approximations as well as the mathematical identities gives us

$$I(F_{RF}, F_{BB}) = log_2\left(\left|I_{N_s} + \frac{\rho}{N_S\sigma_n^2}\Sigma_1^2\right|\right) - tr\left(I_{N_s} - V_1^* F_{RF} F_{BB} F_{BB}^* F_{RF}^* V_1\right) \quad (3.14)$$

The second term in the above equation can be simplified using the high effective SNR approximation to be given as

$$I(F_{RF}, F_{BB}) = log_2\left(\left|I_{N_s} + \frac{\rho}{N_S\sigma_n^2}\Sigma_1^2\right|\right) - (N_s - \|V_1^* F_{RF} F_{BB}\|_F^2) \quad (3.15)$$

Maximization of the mutual information boils down to the minimization of the Euclidean distance

$$\left\|F_{opt} - F_{RF}F_{BB}\right\|_F \quad (3.16)$$

After applying the modifications and identities, the precoder design problem can be written as

$$\left(F_{RF}^{opt}, F_{BB}^{opt}\right) = \text{argmin}_{F_{RF}, F_{BB}} \left\|F_{opt} - FRFFBBF\right. \quad (3.17)$$

Such that $F_{RF} \in \mathcal{F}_{RF}$ and $\|F_{RF}F_{BB}\|_F^2 = N_s$

The above problem can be stated as finding the projection of F_{opt} onto the set of hybrid precoders of the form $F_{RF}F_{BB}$ with $F_{RF} \in \mathcal{F}_{RF}$. The minimization problem given in equation 3.17 is still non convex but is lesser in comlplexity than the original optimization problem. Minimizing the Euclidean distance as in equation (3.17) gives a sub-optimal solution for the sparse mmWave channel. Assuming the Saleh-Vanezuela channel model and exploiting its sparse nature, we make the following observations

1. Structure of optimal precoder

Singular Value Decomposition on the Channel matrix H, above gives the optimal unitary precoder, given by $F_{opt} = V_1$

The unitary matrix V has columns that form an Orthonormal basis for the channel's row space.

2. Structure of the clustered mmWave channels:

Equation (2.1) clearly shows that the array response vectors$a_t(\phi_{il}^t, \theta_{il}^t)$ also form a definite spanning set for the channel's row space. We are assuming here that $N_{Cl}N_{ray} \leq N_t$. The linearly independent array response vectors form another basis for the channel H row space. $F_{RF}$is optimized

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using the phase only constraint of the RF phase shifters. From the channel row space only those having more projections are picked leading to step by step formation of the matrix F_{RF}.

3. Connection between F_{opt} and $a_t(\phi_{il}^t, \theta_{il}^t)$:

According to the first observation stated above that the columns of the optimal precoder $F_{opt} = V_1$ are related to $a_t(\phi_{il}^t, \theta_{il}^t)$ through a linear relation. This implies that F_{opt} can be written as a linear combination of $a_t(\phi_{il}^t, \theta_{il}^t)$.

4. Vectors $a_t(\phi_{il}^t, \theta_{il}^t)$ as columns of F_{RF}:

The vectors corresponding to the array response vectors of the rays in the clustered channel model are unit magnitude vectors with only phase changing capacity which can be applied at the RF stage using analog phase shifters. The precoding algorithm assumed here takes this property of the mmWave channel and applies N_t^{RF} of the vectors $a_t(\phi_{il}^t, \theta_{il}^t)$ at the RF or analog precoding stage. With the known analog precoder, a closed-form expression of the digital precoder can then be obtained. Rewriting the optimization problem for near-optimal hybrid precoders and restricting the analog beamforming matrix to a fixed codebook consisting of array response vectors of the form $a_t(\phi_{il}^t, \theta_{il}^t)$

$(F_{RF}^{opt}, F_{BB}^{opt}) = \text{argmin}_{F_{RF}, F_{BB}} \| F_{opt} - FRFFBBF$ (3.18)

Such that

$F_{RF}^{(i)} \in \{ a_t(\phi_{il}^t, \theta_{il}^t) | 1 \le i \le N_{Cl}, 1 \le l \le N_{ray} \}$

and $\|F_{RF} F_{BB}\|_F^2 = N_s$

Considering the vectors $a_t(\phi_{il}^t, \theta_{il}^t)$ to be the basis, the hybrid precoding algorithm defined here finds the best low dimensional representation of F_{opt}. Defining our precoding algorithm as the selection of the best N_t^{RF} array response vectors and finding

$(F_{BB}^{opt}) = \text{argmin}_{F_{RF}, F_{BB}} \| F_{opt} - A_t \widetilde{F_{BB}} \|_F$

(3.19)

Such that $\| diag(\widetilde{F_{BB}} \widetilde{F^*{}_{BB}}) \|_0 = N_t^{RF}$

and $\| A_t \widetilde{F_{BB}} \|_F^2 = N_s$

The matrix $A_t = \left[a_t(\phi_{1,1}^t, \theta_{1,1}^t), \dots \dots a_t\left(\phi_{N_{Cl}N_{ray}}^t, \theta_{N_{Cl}N_{ray}}^t\right). \right]$ is a matrix whose column vectors are the individual antenna array responses of each ray in each cluster.

The sparsity constraint given in (3.21) namely $\| diag(\widetilde{F_{BB}} \widetilde{F^*{}_{BB}}) \|_0 = N_t^{RF}$ states that that the near optimal baseband precoder cannot have more that N_t^{RF} non-zero rows. This also places a constraint that out of the $N_{Cl}N_{ray}$ columns in the matrix A_t, N_t^{RF} columns are effectively selected. The design problem of the hybrid precoder for the matrices F_{RF} and F_{BB} is hence converted into a constrained matrix reconstruction problem with $N_s > 1$ number of data streams. The algorithmic solution used to determine the hybrid precoders The pseudo code of the Orthogonal matching Pursuit algorithm further referred to as OMP is given as follows:

Algorithm 1: Spatially sparse precoding via OMP

Require F_{opt}

1. F_{RF} = Empty Matrix
2. $F_{res} = F_{opt}$
3. for $i \le N_t^{RF}$ do
4. $\psi = A_t F_{res}$
5. $k = argmax_{l=1,\dots N_{Cl}N_{ray}} (\psi\psi^*)_{l,l}$
6. $F_{RF} = [F_{RF} | A_t^k]$
7. $F_{BB} = (F_{RF}^* F_{RF})^{-1} F_{RF}^* F_{opt}$
8. $F_{res} = \frac{F_{opt} - F_{RF} F_{BB}}{\|F_{opt} - F_{RF} F_{BB}\|_F}$
9. end for
10. $F_{BB} = \sqrt{N_s} \frac{F_{BB}}{\|F_{RF} F_{BB}\|_F}$
11. return F_{RF}, F_{BB}

The concept of OMP is briefly explained as follows: We start the precoding algorithm by ending the vector $a_t(\phi_{il}^t, \theta_{il}^t)$ along whose direction the optimal precoder has maximum power. The chosen vector of $a_t(\phi_{il}^t, \theta_{il}^t)$ is then appended along each iteration to the matrix of F_{RF} , which is initialized as an empty matrix before the for loop. Following this, the least squares solution to F_{BB} is found in step 7. The input of this selected vector is subsequently removed and the process is repeated for the residue of the optimal precoding matrix, F_{res} to end subsequent projections. This process is continued for about N_t^{RF} times, resulting in the choice of N_t^{RF} beamforming vectors.

3.3 Practical Millimeter Wave Receiver Design

As done for the hybrid precoders in the previous section, the practical hybrid receiver is designed with an analog and a digital combiner. Here we are making the assumption that the hybrid precoders F_{RF}, F_{BB} are fixed. The problem here is defined as the design of the hybrid combiners W_{RF}, W_{BB} minimizing the Mean Square Error or the MSE between the transmitted and the processed received signal. and is de ned as

$\left(W_{RF}^{opt}, W_{BB}^{opt}\right) = \text{argmin}_{W_{RF}, W_{BB}} \|s -$ $WRFWBByF$ (3.20)

Such that $W_{RF} \in \mathcal{W}_{RF}$

Where $\mathcal{W}_{RF}$ is the fixed codebook for analog combiners. The codebook $\mathcal{W}_{RF}$ is a set

$N_r \times N_r^{RF}$ constant gain phase only entries. The solution to the problem given in equation 3.22 is given by

$$W_{MMSE} = E[sy^*]E[yy^*]^{-1} = \frac{\sqrt{\rho}}{N_s} F_{BB}^* F_{RF}^* H^* \left(\frac{\rho}{N_s} H F_{RF} F_{BB} F_{BB}^* F_{RF}^* H^* + \sigma_n^2 I_{N_r}\right)^{-1} = \frac{1}{\sqrt{\rho}}\left(F_{BB}^* F_{RF}^* H^* H F_{RF} F_{BB} + \frac{\sigma_n^2 N_s}{\rho} I_{N_s}\right)^{-1} F_{BB}^* F_{RF}^* H^* \quad (3.21)$$

We get this equation by applying the matrix inversion Lemma. The non-convex constraint $W_{RF} \in \mathcal{W}_{RF}$ makes the problem impossible to solve algorithmically. Reformulating the mean square error problem

$E\left[\|s - W_{RF}{}^* W_{BB}{}^* y\|_F{}^2\right] = E[(\|s -$ $WRFWBByF*s-WRFWBByF=trEss*-2\mathcal{R}tr$ $Esy**WRF*WBB*+trWRF*WBB*Ess*WRF*W$ $BB*$ (3.22)

Performing a few mathematical manipulations and choosing to minimize the equivalent objective function, the MMSE estimation problem becomes

$$\left(W_{RF}^{opt}, W_{BB}^{opt}\right) = \underset{W_{RF}, W_{BB}}{\text{argmin}} \left\| E[yy^*]^{\frac{1}{2}}(W_{MMSE} - W_{RF} W_{BB}) \right\|_F \quad (3.23)$$

Such that $W_{RF} \in \mathcal{W}_{RF}$

Theoretically stating thus problem as finding the projection of the unconstrained MMSE precoder W$_{MMSE}$ onto the set of hybrid combiners of the form $W_{RF} W_{BB}$ with $W_{RF} \in \mathcal{W}_{RF}$. Similar to the sparse precoding problem stated above, the design problem of mmWave Rx is again limited by the non-convex constraint $W_{RF} \in \mathcal{W}_{RF}$ making algoithmic solution impractical. As in the hybrid precoding case above we exploit the characteristics of the mmWave channel for a suboptimal solution. Constraining W_{RF} to have the form $a_r(\phi_{il}^r, \theta_{il}^r)$ and solving the following equation

$\left(\widetilde{W}_{BB}^{opt}\right) = argmin_{W_{RF}, W_{BB}} \left\| E[yy^*]^{\frac{1}{2}} W_{MMSE} -\right.$ $Eyy*12ArWBBF$ (3.24)

Such that $\left\|diag\left(\widetilde{W_{BB}}\widetilde{W^*{}_{BB}}\right)\right\|_0 = N_r^{RF}$.

The matrix $A_r = \left[a_r\left(\phi_{1,1}^r, \theta_{1,1}^r\right), \dots \dots a_r\left(\phi_{N_{Cl}N_{ray}}^r, \theta_{N_{Cl}N_{ray}}^r\right).\right]$ is a matrix whose column vectors are the individual antenna array responses of each ray in each cluster. The same well known concept as above that is OMP is used taking A_r and W_{BB} as auxillary variables. The pseudo code for the OMP applied at the receiver is given below.

Algorthm 2: Spatially Sparse MMSE Combining via OMP

Require: W_{MMSE}

1. W_{RF} = Empty Matrix
2. $W_{res} = W_{MMSE}$
3. for $i \leq N_r^{RF}$ do

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4. $\psi = A_r^* \, E[yy^*] \, W_{res}$

5. $k = argmax_{l=1,\ldots N_{Cl} N_{ray}} \, (\psi\psi^*)_{l,l}$

6. $W_{RF} = [W_{RF} | A_r^k]$

7. $W_{BB} = (W_{RF}^* E[yy^*] W_{RF})^{-1} W_{RF}^* E[yy^*] \, W_{MMSE}$

8. $W_{res} = \frac{W_{MMSE} - W_{RF} W_{BB}}{\| W_{MMSE} - W_{RF} W_{BB} \|_F}$

9. end for

10. return W_{RF}, W_{BB}

The above methods of spatially sparse combining and precoding are assumed to be decoupled from each other. This simplifies the hybrid transceivers for mmWave system. THe number of RF chains at both ends limits the application of the array response vectors $a_r(\phi_{il}^r, \theta_{il}^r)$.

3.4 Results and Discussion

The result for a 64×16 MIMO hybrid beamforming system is shown in figure 3.2. The Spectral efficiency is plotted as a function of SNR. The number of antennas at the transmitter is $N_t = 64$ and the number of antennas at the receiver id $N_r = 16$. The number of RF chains taken here are $N_t^{RF} = N_r^{RF} = 4$. The result shows the spectral efficiency vs SNR graph for the two data streams $N_s = 1$ and $N_s = 2$. The implementation of the hybrid beamforming system closely approaches the purely digital one.

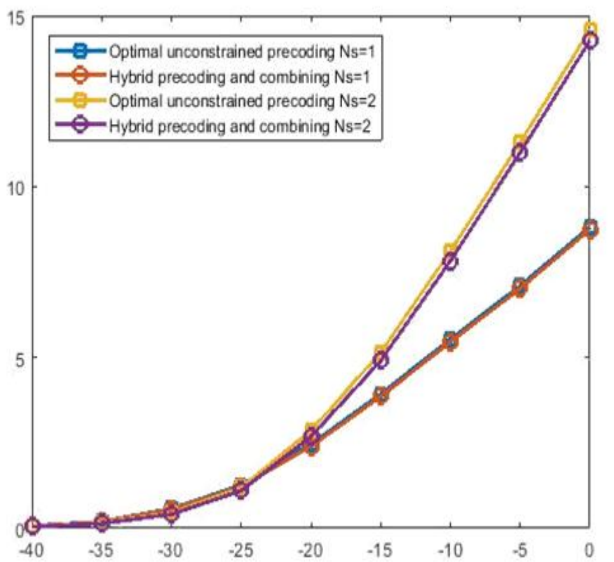

Figure 3.2 : The plot of 64 $\times$ 16 MIMO for data streams $N_s = 1$ and $N_s = 2$

IV. HYBRID TRANSCEIVER DESIGN FOR CR SYSTEM & RESULTS

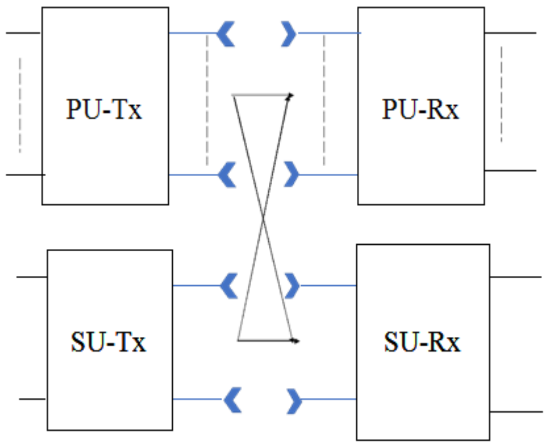

Figure 4.1: Hybrid Beamforming Architecture for CR

The hybrid beamforming architecture shown in the section above is that a standalone MIMO communications system. It has the assumption that there is no other link of mmWave communication in the vicinity, which is optimistic since the current 5G projections for traffic increase, make it compulsory for constraints to be taken on the im-posed interference. A widely used technique to address the problem of spectrum scarcity is Cognitive radio (CR). It enables dynamic spectrum access: solving the problem of limited bandwidth. Cognitive Radio is a promising technology for the future of wireless communications.

In this project we design hybrid analog-digital transceivers subject to interference constraints as in the case of traditional underlay cognitive radios. The cognitive hybrid analog/digital transceiver designs for point-to-point MIMO systems is presented to maximize the spectral efficiency subject to interference by the primary user as well (PU) as the power and hardware related (limited number of RF chains) constraints. This approach is based on the underlay cognitive radio paradigm where the Secondary User (SU) may access a spectrum area licensed to a PU, simultaneously with the latter, provided that the interference power on the PU

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transmissions is below a predefined threshold. The proposed approach derive the precoding and post-coding (combining) matrices by decoupling the transmitter receiver optimization problem which is the same as above. Initially the pre-coder is derived such that the system's mutual information is maximized subject to the constraints mentioned previously and the optimal linear post-coding matrix is derived in a Minimum Mean Square Error (MMSE) sense. This system can be successfully applied in large array systems that function in lower frequencies and under typical fading channel models in order to reduce the hardware complexity/power consumption.

The Cognitive Radio system, as modeled by [9], consists of a SU $R_s \times T_S$ MIMO trying to access a spectrum licensed to a $R_p \times T_p$ Primary User. The RF chains of the SU system is $N_t^{RF} \ll N_t$ at the Tx and $N_r^{RF} \ll N_r$. The data streams transmitted by the SU is given by $N_s < minimum(N_t, N_r)$. The hybrid precoder applied at the SU Tx is given by $F = F_{RF}F_{BB}$ where F_{RF} is the analog RF precoder of dimensions $N_t \times N_t^{RF}$, F_{BB} is the digital baseband precoder of dimensions $N_t^{RF} \times N_s$. Similarly, the hybrid combiner is given by $W = W_{RF}W_{BB}$ where W_{RF} is the analog RF postcoder of dimensions $N_r^{RF} \times N_r$. W_{BB} is the baseband combiner of the dimensions $N_s \times N_r^{RF}$. The received signal at the SU receiver is given by

$$y_s' = W^H y_s = W^H\left(H_{SS}Fx_s + \widetilde{H}_{sp}x_p + n_s\right) \quad (4.1)$$

The channels are given by

H_{SS} = Channel between SU Tx and SU Rx

H_{SP}= Channel between SU Tx and PU Rx

H_{PS} = Channel between PU Tx and SU Rx

H_{PP} = Channel between PU Tx and PU Rx

The SU transmitted symbols are represented by x_s which is a $N_s \times 1$ vector satisfying

The condition $E[x_s x_s^H] = \sigma_s^2 I_{N_S}$. $\tilde{H}_{SP} = H_{SP}F_p$ is the $R_s \times T_p$ channel matrix between the PU Tx and the SU Rx , and F_p is the PU precoding matrix. The received signal at the PU Rx is given as

$$y_p = H_{PP}F_p x_p + \widetilde{H}_{sp}x_p + n_p \quad (4.2)$$

A cognitive underlay system assumes that the SU has full knowledge of H_{SS}, H_{SP} and H_{PS}. We now aim to design the precoding and combining matrices for the SU soas to maximize the spectral efficiency R. The constraints placed here are on the total transmission power P_{max} and on the interference I_{max}. The optimization problem is defined as follows

$$(P_1): R(F, W)$$

Such that $tr(FF^H) \leq P_{max}$ and $tr(H_{PS}FF^H H_{PS}^H) \leq I_{max}$.

The spectral efficiency R is defined as $= log_2\left|I_{N_s} + R_n^{-1}W^H H_{SS}FF^H H_{SS}^H W\right|$.

The matrix R_n is the interference plus noise covariance matrix given by

$$R_n = W^H\left(H_{SP}H_{SP}^H + \sigma_n^2 I_{N_s}\right)W \quad (4.3)$$

4.1 Digital Only Solution for Optimal Precoder

The fully optimal digital solution can be found by directly solving P_1. Since this is a joint optimization problem for the precoding as well as combining matrices, it is decoupled to find first the optimal precoding matrix F_D by solving the following equation

$$(P_2): \max_{\tilde{F}} log_2\left|I_{N_s} + Q^{-\frac{1}{2}}H_{SS}\tilde{F}H_{SS}^H Q^{-\frac{1}{2}}\right| \quad (4.4)$$

Such that $tr(\tilde{F}) \leq P_{max}$ and $tr\left(H_{PS}\tilde{F}H_{PS}^H\right) \leq I_{max}$ and $\tilde{F} \succcurlyeq 0$

Where the condition $\tilde{F} \succcurlyeq 0$ denotes that $\tilde{F}$is a positive semidefinite matrix. The noise and interference covariance matrix is given by $Q = \widetilde{H}_{SP}\widetilde{H}_{Sp}^H + \sigma_n^2 I_{N_r}$. Eigen value decomposition of the matrix $\tilde{F}^* = U_F\Sigma_F U_F$. The optimal precoding solution is given by $F_D = U_F\sqrt{\Sigma_F}$. It satisfies the condition $\tilde{F} = F_D F_D^H$.

The fully digital solution for the postcoding matrix W is derived by the MMSE (Minimum Mean Square Error) is given by solving the following problem

$$\min_W \|x_s - W^H y_s\|_F^2 \quad (4.5)$$

The solution to the above problem is given by

$$W_{MMSE} = E[y_s y_s^H]^{-1}E[x_s y_s^H]^H = \left(H_{SS}F_D H_{SS}^H F_D^H + \widetilde{H}_{SP}\widetilde{H}_{Sp}^H + \sigma_n^2 I_{N_r}\right)^{-1} H_{SS}F_D \quad (4.6)$$

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4.2 Codebook Based Hybrid Transceiver Design

The assumption taken here is that the analog precoders and combiners are determined from a fixed codebook, by uniformly quantizing the azimuthal angles over the interval $[0,2\pi]$. The columns of the matrices F_{RF} and W_{RF} must satisfy ${F_{RF}}^{(i)}({W_{RF}}^{(i)}) \epsilon\, C_\phi\{a(\phi_1),..,a(\phi_{2^{N_\phi}})\}$

The precoding optimization problem to find the analog and digital precoders F_{RF} and F_{BB} can be solved by the above mentioned OMP and is defined as

(P_3): $\min_{F_{RF},F_{BB}} \|F_D - F_{RF}F_{BB}\|_F^2$ (4.7)

Such that $F_{RF}F_{BB}\, \epsilon\, S$ and $F_{RF} \in \mathcal{F}_{RF}$

Where S is the set defined by $S = \left\{A\, \epsilon\, C^{N_t \times N_t^{RF}} \,|\, \|A\|_F^2 \le P_{max}, \|H_{PS}A\|_F^2 \le I_{max}\right\}$

Here the digital precoder F_D is defined by solving the problem P_2 given above. Similarly, the combining matrices at the SU Rx can be solved as

(P_4): $\min_{F_{RF},F_{BB}} \left\|E[y_s y_s^H]^{\frac{1}{2}}(W_D - W_{RF}W_{BB})\right\|_F^2$ (4.8)

Such that $\|diag(W_{BB}W_{BB}^H)\|_0 = N_r^{RF}$

4.3 System model in the presence of 2 PU

The above analysis designs a hybrid transceiver between a SU Tx and Rx subject to the interference of a single PU. This analysis is taken further in order to design hybrid transceiver between the SU Tx and Rx in the presence of two PUs. As above the SU Tx is assumed to have Ts antennas and the SU Rx is assumed to have Rs antennas. The two PUs are assumed to have N_t antennas at the Tx and N_r antennas at the Rx. The interference constraints now change in order to acccommodate the interference from two PUs. The equation of the received signal at the SU Rx in the presence of 2 PUs is written as.

$$y_s' = W^H y_s = W^H\left(H_{SS}Fx_s + \widetilde{H}_{sp1}x_{p1} + \widetilde{H}_{sp2}x_{p2} + n_s\right) \quad (4.9)$$

A cognitive underlay system assumes that the SU has full knowledge of $H_{SS}, \widetilde{H}_{sp1}, \widetilde{H}_{sp2}$ and H_{ps}. The aim is to design the precoding and combining matrices for the SU so as to maximize the spectral efficiency R. The constraints placed here are on the total transmission power P_{max} and on the interference I_{max}. The first step is to obtain the fully digital precoder FD. The optimization problem P_2 changes again to accomodate the presence of 2 PUs and can be written as

(P_2): $\max_{\tilde{F}} log_2\left|I_{N_s} + Q^{-\frac{1}{2}}H_{SS}\tilde{F}H_{SS}^H Q^{-\frac{1}{2}}\right|$ (4.10)

Such that $tr(\tilde{F}) \le P_{max}$ and $tr(H_{P1S}\tilde{F}H_{P1S}^H +$ *HP2SFHP2SH≤Imax* and *F≽0.* Where the condition $\tilde{F} \succcurlyeq 0$ denotes that $\tilde{F}$ is a positive semidefinite matrix. The noise and interference covariance matrix is given by $Q = \widetilde{H}_{SP1}\widetilde{H}_{Sp1}^H + \widetilde{H}_{SP2}\widetilde{H}_{Sp2}^H + \sigma_n^2 I_{N_r}$, and

H_{SP1}= Channel between SU Tx and PU1 Rx

H_{SP2}= Channel between SU Tx and PU2 Rx

H_{P1S} = Channel between PU1 Tx and SU Rx

H_{P2S} = Channel between PU2 Tx and SU Rx

$\widetilde{H}_{SP1} = H_{SP1}F_{p1}$ and

$\widetilde{H}_{SP2} = H_{SP2}F_{p2}$.

Where F_{p1} and F_{p2} are the precoding matrices at the Txs of PU1 and PU2 respectively. Eigen value decomposition of the matrix $\tilde{F}^* = U_F\Sigma_F U_F$. The optimal precoding solution is given by $F_D = U_F\sqrt{\Sigma_F}$. It satisfies the condition $\tilde{F} = F_D F_D^H$. The fully digital solution for the combining matrix W is derived by the MMSE (Minimum Mean Square Error) is given by solving the following problem

$\min_W \|x_s - W^H y_s\|_F^2$ (4.10)

The solution to the above problem is given by

$$W_{MMSE} = E[y_s y_s^H]^{-1}E[x_s y_s^H]^H = \left(H_{SS}F_D H_{SS}^H F_D^H + \widetilde{H}_{SP1}\widetilde{H}_{Sp1}^H + \widetilde{H}_{SP2}\widetilde{H}_{Sp2}^H + \sigma_n^2 I_{N_r}\right)^{-1} H_{SS}F_D$$

4.4 Codebook Based Hybrid Transceiver Design

The assumption taken here is that the analog precoders and combiners are determined from a fixed codebook, by uniformly quantizing the azimuthal angles over the interval [0,2π]. The columns of the matrices F_{RF} and W_{RF} must satisfy ${F_{RF}}^{(i)}({W_{RF}}^{(i)}) \epsilon\, C_\phi\{a(\phi_1),..,a(\phi_{2^{N_\phi}})\}$. Once the

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optimal precoder F_D is known, the precoding optimization problem for F_{RF} and $F_{BB}$by the above mentioned OMP and is defined as

(P_3): $\min_{F_{RF},F_{BB}} \|F_D - F_{RF}F_{BB}\|_F^2$ (4.12)

Such that $F_{RF}F_{BB} \epsilon S$ and $F_{RF} \in \mathcal{F}_{RF}$

Where S is the set defined by $S = \left\{A \epsilon C^{N_t \times N_t^{RF}} \,|\|A\|_F^2 \leq P_{max}, \|H_{P1S}A +\right.$

HP2SAF2≤Imax

Here the digital precoder F_D is defined by solving the problem P_2given above. Similarly, the combining matrices at the SU Rx can be solved as

(P_4): $\min_{F_{RF},F_{BB}} \left\| E[y_s y_s^H]^{\frac{1}{2}}(W_D - W_{RF}W_{BB}) \right\|_F^2$

(4.13)

Such that $\|diag(W_{BB}W_{BB}^H)\|_0 = N_r^{RF}$

Results and Discussions

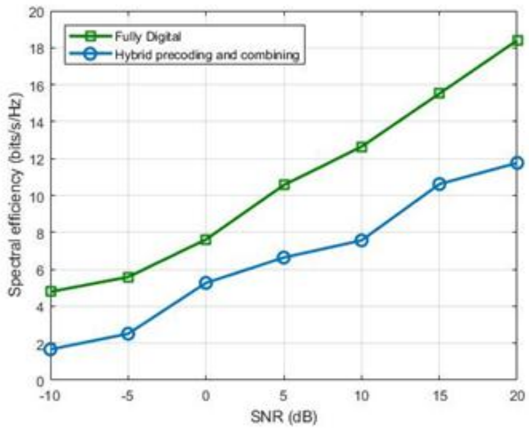

Figure 4.2 Implementation of a 64 × 16 SU CR Hybrid beamforming system with a single PU

The hybrid beamforming performance for a cognitive radio underlay system is done using 64 × 16 SU MIMO system. The SU Tx is assumed to have a ULA with N_t = 64 antennas. The SU Rx is assumed to have a ULA with N_r = 16 antennas. The PU Tx is assumed to have a ULA with N_t = 64 antennas. The PU Rx is assumed to have a ULA with N_r = 16 antennas. The PU is assumed to have N_s = 4 number of data streams. The SU is assumed to have N_s = 2 number of data streams. The number of RF chains at the Tx and Rx of the PU is $N_r^{RF} = N_t^{RF} = N_s$. The number of RF chains at Tx and Rx of the SU MIMO system is assumed to be $N_r^{RF} = N_t^{RF} = N_s$. The signalling assumed here is Gaussian. The channel is developed according to the Saleh-Valenzuela model with the number of clusters N_{Cl} = 1 and the number of rays in a cluster N_{ray} = 15.

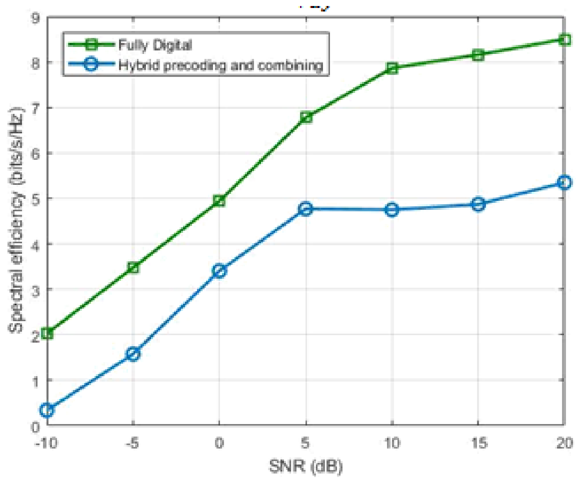

Figure 4.3 Implementation of a 16 × 4 SU CR Hybrid beamforming system with a single PU

Plotting the SE vs SNR graph for 16 × 4 SU MIMO system in the presence of a single PU for number of data streams N_s = 2. The result is shown in Figure 4.3. The number of RF chains is taken to be $N_r^{RF} = N_t^{RF} = N_s$.The number of PU is increased to two and the graph for SE vs SNR is plotted again using the same 16 × 4 SU MIMO. The SE of the SU goes down. This result is shown in Figure 4.4. It can be clearly seen by the comparision of the two graphs that the increased interference from the second PU decreases the performance of the hybrid beamforming system.

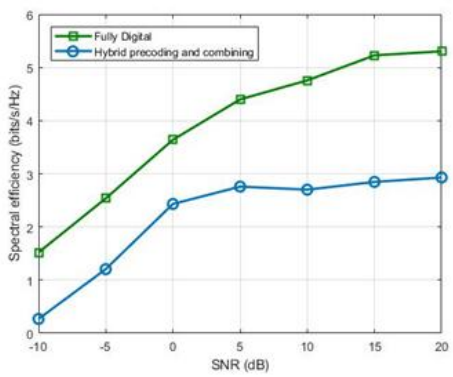

Figure 4.4 Implementation of a 16 × 4 SU CR Hybrid beamforming system with two PU

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V. CONCLUSION

The concept of MIMO is being widely used to form coarse directional beams in 4G. The nest step of Massive MIMO comes with the application of hundreds of antennas at both Tx and Rx, made possible only for the mmWave spectrum. As the data demand increases, 5G technology needs to have capacity to meet the requirements. The channel for the mmWave spectrum which promises more bandwidth, comes with its own drawbacks- namely- attenuation and high free space path loss. The physical dimensions of the antenna size and spacing of the antenna arrays comes down in accordance to the wavelength. Mmwave channel allows the arrangement of many antenna in a small physical dimension giving us a high beamforming gain. The presently used traditional baseband digital beamforming in 4g requires a distinct radio frequency (rf) chain to be attached to each antenna. As the antennas number in 5g goes higher this becomes impractical since each chain has a lot of hardware. Hybrid beamforming helps us in taking the good from both digital and baseband beamforming and going for a suboptimal but practical procedure.

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