공학석사 학위논문

Effective Delay Spread Reduction Precoding for Low Latency MISO-OFDM Systems

저지연 다중 안테나 직교 주파수 분할 다중화 시스템을 위한 유효 지연 확산 감소 프리코딩

2015 년 2 월

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Abstract

Effective Delay Spread Reduction Precoding for Low Latency MISO-OFDM Systems

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These days, many issues are being actively discussed for the next generation wireless communication system. One of the important issues is latency reduction to support latency-sensitive applications such as virtual reality games, remote controlled robots, emergency alerts in vehicle communications. To meet the tight latency requirements for such applications, the symbol duration should be remarkably reduced than that of current orthogonal frequency division multiplexing (OFDM) systems which make use of cyclic prefix (CP) to mitigate inter-symbol-interference (ISI). However, as symbol duration decreases, the CP overhead occupies a larger portion of resources, and impairs the performance of the OFDM system. On the other hand, if we decrease CP duration to reduce overhead, ISI due to delay spread would also degrade the system performance. Hence, we not only need to reduce CP duration for efficient low latency communications, but also need to mitigate ISI induced from the reduced CP duration.

In this thesis, we propose a novel precoding method in MISO-OFDM systems to reduce effective delay spread of channels for CP overhead reduction. We formulate the precoding vector selection problem which aims to minimize effective delay spread as well as signal to noise ratio (SNR) impairment. In addition, to reduce the computational complexity of the problem, we suggest an efficient precoding vector selection algorithm which make use of the gradient descent method.

Simulation results show that the proposed scheme remarkably reduces the effective delay spread of channels while maintaining tolerable SNR degradation. We show that the performance of the proposed scheme improves as the base station (BS) has more antennas. Future work may include the extension of the proposed scheme to massive MIMO systems in which the BS has a massive number of antennas. In massive MIMO systems, we can make use of the excess degree of freedom (DoF) to reduce effective delay spread of channels while multiuser interference (MUI) is eliminated.

Keywords: low latency, MIMO, precoding, delay spread, OFDM, inter-symbol interference

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Chapter 1

Introduction

Recently, many issues are being actively discussed for the next generation mobile communication system (5G). One of the important issues is about latency reduction to support latency-sensitive applications such as virtual reality games, remote controlled robots, emergency alerts for vehicle communications, and so on. For such latency-sensitive applications, end-to-end latency should be smaller than 1ms, and the PHY layer latency should be less than 100 μs for one-way communication [1], [2]. To meet the extremely tight latency requirement, the symbol duration should be far smaller than that of current orthogonal frequency division multiplexing (OFDM) systems which make use of cyclic prefix (CP) to combat inter-symbol interference (ISI) arising from delay spread. Although the CP overhead is negligible in the case of long symbol duration, it would capture a larger portion of time resources as we decrease symbol duration without reduction in CP duration. However, if we decrease CP duration, ISI arising from delay spread would also severely degrade the system performance. Hence, for efficient low latency communications, we not only need to reduce CP duration, but also need to mitigate ISI induced from the reduced CP duration.

There have been a lot of approaches to mitigate ISI. In [3], zero-forcing (ZF) and minimum mean square error (MMSE) equalizers were described. However, they require heavy computational loads at mobile stations (MSs). Moreover, they are unsuitable for low latency communications because they mitigate ISI by designing a proper impulse response based on channel state information (CSI), and the impulse response may generate large processing delay. In [4] and [5], a pre-equalization method was proposed for multiple-input multiple-output (MIMO), and it does not require high computational loads at MSs. However, the fundamentals are similar to the aforementioned equalization methods except that it requires computational loads at the base station (BS). Therefore it is also unsuitable for low latency communications due to large processing delay.

In this thesis, we propose a novel ISI mitigation scheme which makes use of precoding vector selection in multiple-input single-output OFDM (MISO-OFDM) systems. The proposed scheme is suitable for low latency communications in that the spatial domain precoding at the BS does not require heavy computational loads at MSs and large processing delay. We formulate the precoding vector selection problem which minimizes effective delay spread of a wireless link, and hence, mitigates ISI while minimizing the signal-to-noise ratio (SNR) degradation. We also propose a precoding vector selection algorithm which reduces the required computational complexity.

The remainder of the thesis is organized as follows. Chapter 2 describes the channel model and the system model. In Chapter 3, we formulate a optimization problem and propose a precoding algorithm to reduce effective delay spread for low latency MISO-OFDM systems. Simulation results show the performance of the proposed algorithm with varying parameters in Chapter 4. Conclusion and future works are presented in Chapter 5.

Chapter 2

System Model

2.1 Spatial channel model

According to [6], we model the spatial channel for MIMO systems which is described in Figure 1 and parameters of the spatial channel model can be shown in Table 1. For MIMO spatial channel models, the channel coefficients of the n_r-th receive antenna and the n_t-th transmit antenna for the p-th multipath component (p=1,...,P) are given by

$$h_{n_r,n_t,p}(t)=\sqrt{\frac{P_p\sigma_{SF}}{S}}\sum_{s=1}^{S}\left(\begin{array}{l}\sqrt{G_{BS}\left(\theta_{p,s,AoD}\right)}\exp\left(j\left[kd_{n_t}\sin\left(\theta_{p,s,AoD}\right)+\Phi_{p,s}\right]\right)\\ \times\sqrt{G_{MS}\left(\theta_{p,s,AoA}\right)}\exp\left(jkd_{n_r}\sin\left(\theta_{p,s,AoA}\right)\right)\\ \times\exp\left(jk\|\mathbf{v}\|\cos\left(\theta_{p,s,AoA}-\theta_v\right)t\right)\end{array}\right) \tag{1}$$

where P_p denotes the power of the p-th path, σ_{SF} denotes the lognormal shadow fading, S denotes the number of subpaths per-path, $\theta_{p,s,AoD}$ and $\theta_{p,s,AoA}$ denote the AoD and AoA for the s-th subpath of the p-th path, respectively, $G_{BS}\left(\theta_{p,s,AoD}\right)$ and $G_{MS}\left(\theta_{p,s,AoA}\right)$ denote the BS and MS antenna gains of each array element, respectively, k denotes the wave number $2\pi/\lambda$ where λ is the carrier wavelength in meters, d_{n_t} and d_{n_r} denote the distances in meters from BS and MS antenna element n_t and n_r from the reference antenna, respectively, $\Phi_{p,s}$ denotes the phase of the s-th subpath of the p-th path, $\|\mathbf{v}\|$ denotes the magnitude of the MS velocity vector and θ_v denotes the angle of the MS velocity vector.

For the sake of simplicity, we establish a MISO-OFDM system which has a single receive antenna at the MS. Also, without loss of generality, we assume that the velocity of the MS is zero. Hence, the channel coefficient for p-th path can be expressed as

$$h_{n_t,p}(t)=\sqrt{\frac{P_p\sigma_{SF}}{S}}\sum_{s=1}^{S}\left(\sqrt{G_{MS}G_{BS}\left(\theta_{p,s,AoD}\right)}\exp\left(j\left[kd_{n_t}\sin\left(\theta_{p,s,AoD}\right)+\Phi_{p,s}\right]\right)\right).$$

In [7], the delay spread of multipath channel is described as

$$T_d=\max_{i,j}\left|\tau_i-\tau_j\right| \tag{2}$$

where τ_i denotes propagation delay of i-th delay bin and it dictates frequency coherence of multipath channel. For a metric of frequency coherence, coherence bandwidth is used and typically expressed as

$$W_c=\frac{1}{2T_d} \tag{3}$$

When the delay spread of multipath channel is large, the coherence bandwidth is small and vice versa. So, delay spread and coherence bandwidth stand for metrics in time and frequency domain of multipath channel, and they have a reciprocal relationship.

Since the effective delay spread is related to frequency response of the effective channel, we propose an effective delay spread reduction precoding algorithm which makes the frequency flat response of the effective channel in Chapter 3.

For the metric of delay spread, the root-mean-square (RMS) delay spread of the channel response is shown in [8] and it can be expressed as

$$\tau_{\mathrm{rms}}=\frac{\sum_i\left(\tau_i-\mu_\tau\right)^2P_h\left(\tau_i\right)}{\sum_i P_h\left(\tau_i\right)} \tag{4}$$

where $P_h\left(\tau_i\right)$ denotes the power delay profile (PDP) of i-th delay bin which is given by $P_h\left(\tau_i\right)=\left|h\left(\tau_i\right)\right|^2$ for the time-invariant channel and μ_τ is the mean delay spread which is given by

$$\mu_\tau=\frac{\sum_i \tau_i P_h\left(\tau_i\right)}{\sum_i P_h\left(\tau_i\right)}. \tag{5}$$

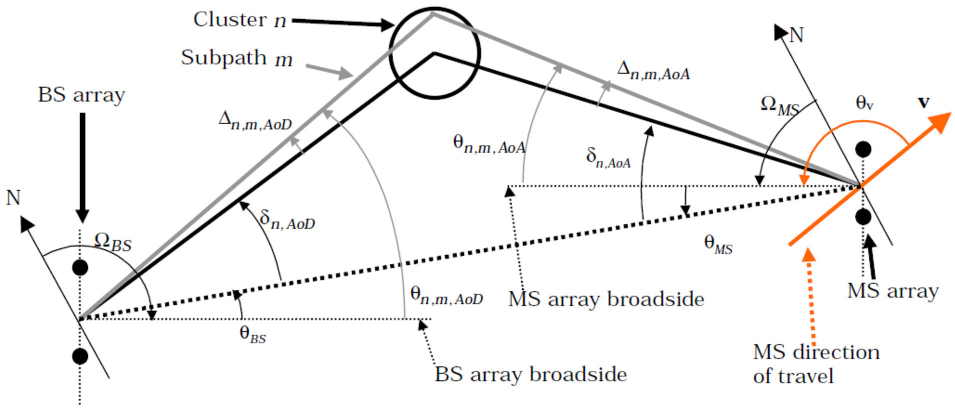

Figure 1. 3GPP spatial channel model

Parameter	Definition
Ω_{BS}	BS antenna array orientation
θ_{BS}	LOS AoD direction between the BS and MS with respect to the broadside of the BS array
$\delta_{n,AoD}$	AoD for the *n*-th path with respect to the LOS AoD
$\Delta_{n,m,AoD}$	Offset for the m-th subpath of the n-th path with respect to $\delta_{n,AoD}$
$\theta_{n,m,AoD}$	Absolute AoD for the m-th subpath of the n-th path at the BS with respect to the BS broadside
Ω_{MS}	MS antenna array orientation
θ_{MS}	Angle between the BS-MS LOS and the MS broadside
$\delta_{n,AoA}$	AoA for the n-th path with respect to the LOS AoA
$\Delta_{n,m,AoA}$	Offset for the m-th subpath of the n-th path with respect to $\delta_{n,AoA}$
$\theta_{n,m,AoA}$	Absolute AoA for the m-th subpath of the n-th path at the MS with respect to the BS broadside.
$\mathbf{v}$	MS velocity vector
θ_v	Angle of the velocity vector with respect to the MS broadside

Table 1. Spatial channel model parameters

2.2 MISO-OFDM systems

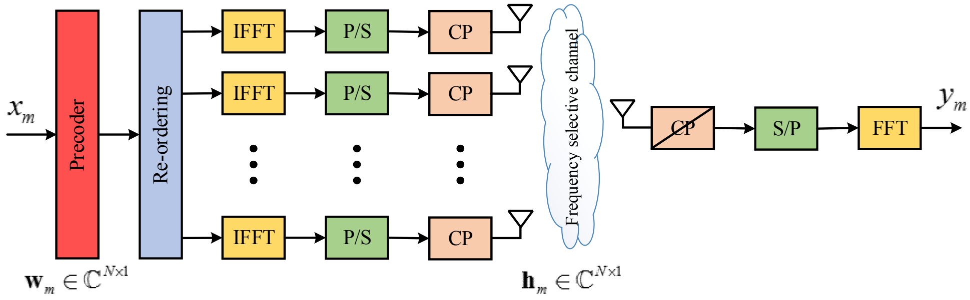

Figure 2. MISO-OFDM systems

We consider an OFDM-based single user multiple-input single-output (MISO-OFDM) system as depicted in Figure 2. BS and MS are equipped with N transmit antennas and a single receive antenna. The received signal at the MS is expressed as

$$y_m = \mathbf{h}_m^H \mathbf{w}_m x_m + n_m = \sum_{n=1}^{N} h_{mn} w_{mn} x_m + n_m \tag{6}$$

where x_m denotes the input symbol transmitted to the MS, $\mathbf{h}_m$ and $\mathbf{w}_m$ denote a $N\times 1$ channel vector and a $N\times 1$ precoding vector of the m-th subcarrier respectively. The effective channel response of the m-th subcarrier is given by

$$h_{\text{eff},m} = \mathbf{h}_m^H \mathbf{w}_m \in \mathbb{C} \tag{7}$$

and the effective channel response vector over subcarriers is expressed as

$$\mathbf{h}_{\text{eff}} = \left[\mathbf{h}_1^H \mathbf{w}_1 \quad \mathbf{h}_2^H \mathbf{w}_2 \quad \cdots \quad \mathbf{h}_M^H \mathbf{w}_M\right] \in \mathbb{C}^M \tag{8}$$

where M denotes the total number of OFDM tones,

Then, the channel matrix of the system can be expressed as

$$\mathbf{H} = \left[\mathbf{h}_1 \quad \mathbf{h}_2 \quad \cdots \quad \mathbf{h}_M\right] \in \mathbb{C}^{N\times M} \tag{9}$$

where each column vector corresponds to the channel response vector of each subcarrier. In the same way, the precoding matrix of the system can be represented as

$$\mathbf{W} = \begin{bmatrix} \mathbf{w}_1 & \mathbf{w}_2 & \cdots & \mathbf{w}_M \end{bmatrix} \in \mathbb{C}^{N \times M} \tag{10}$$

where each column vector corresponds to the precoding vector of each subcarrier.

Chapter 3

Proposed Precoding Algorithm

3.1 Relationship between precoding and delay spread

In conventional multiple antenna systems, we can adjust weights of antenna elements and it aims sum rate maximization, but delay spread reduction is not considered so far. Precoding using multiple antenna systems can achieve effective delay spread reduction that is essential for low latency communications.

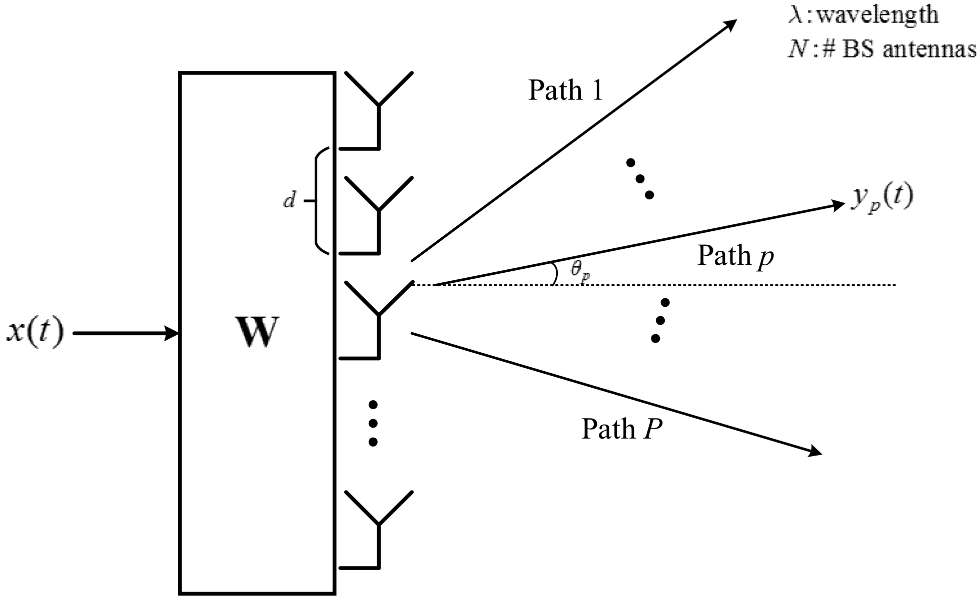

Figure 3. Multiple antenna system

Multiple antenna system and output of paths are depicted at Figure 3 as shown in [9]. The output of the path p can be expressed as

$$
\begin{aligned}
y_p(t) &= \sum_{n=1}^{N} w_n x(t) e^{j(n-1)2\pi\frac{d\sin\theta_p}{\lambda}} \\
&= \sum_{n=1}^{N} \underbrace{\left\{ w_n e^{j(n-1)2\pi\frac{d\sin\theta_p}{\lambda}} \right\}}_{\text{Weight for } n\text{-th antenna element}} x(t)
\end{aligned} \tag{11}
$$

where $x(t)$ denotes the input of the precoder, w_n denotes the weight of the n-th antenna element, N denotes the number of antenna elements, P denotes the number of paths, λ denotes the wavelength, d denotes the distance between adjacent antenna elements, and θ_p denotes the angle of the p-th path comprised with the normal direction of the antenna array.

As expressed above, the output of the path p is a weighted sum of antenna elements. By properly adjusting the weight vector of the antenna array, the output of the particular path can be changed and we are able to make the dominant delay paths suppressed. Hence, the delay spread of the effective channel response can be reduced by proper precoding vector selection.

In this chapter, we formulate a precoding vector selection problem for effective delay spread reduction. On the property described in section 2.1, we can reduce delay spread when the frequency response of the effective channel be flat. So, we utilize the frequency response of the channel as a criterion of the proposed precoding algorithm for delay spread reduction and our problem purposes the frequency response of the effective channel to be flat. This approach is depicted in Figure 4 and Figure 5. However, when the dominant delay path component is suppressed, the effective channel gain will be degraded. So, our problem also aims to maintain the channel gain with a tolerance.

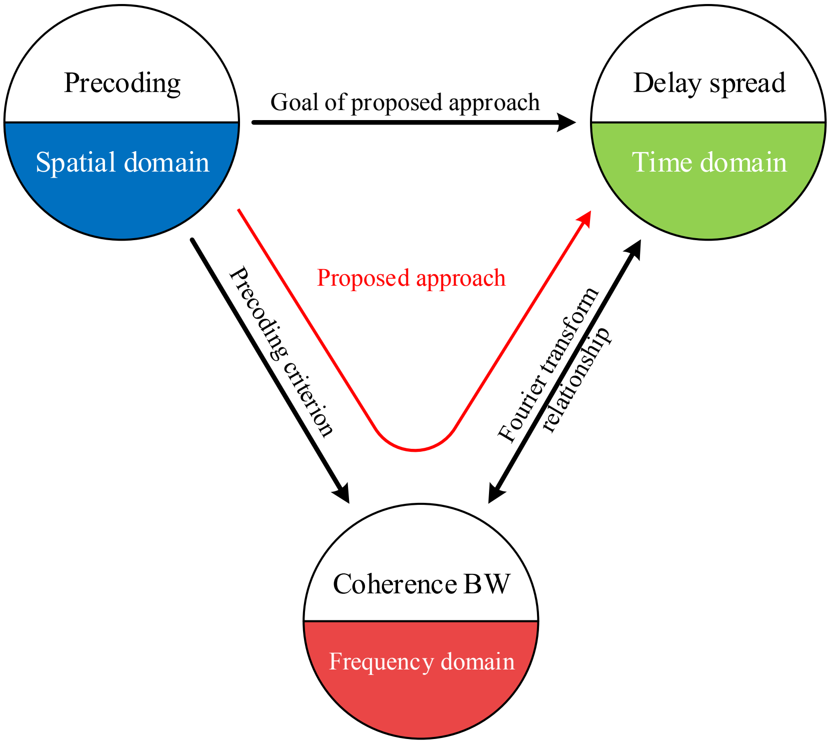

Figure 4. Relation among domains in the proposed approach

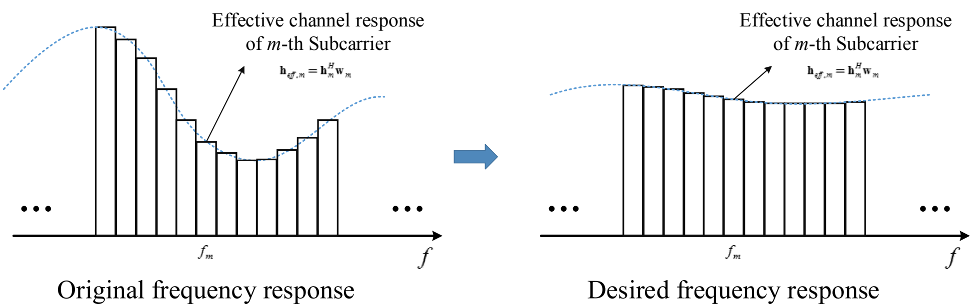

Figure 5. Purpose of the delay spread reduction precoding problem

3.2 Criterions for effective delay spread reduction precoding

As we discussed in section 3.1, we can formulate precoding vector selection problem from two point of views. One is delay spread reduction and another is channel gain degradation tolerance.

3.2.1 Delay spread reduction

The ideal channel response with no delay spread has a constant magnitude with linear phase [10]. For the m-th subcarrier, the ideal channel response can be described as

$$h_{\text{ideal},m} = ae^{-j2\pi f_m \tau} \tag{12}$$

where a denotes the constant desired magnitude, τ denotes the constant desired delay, and f_m denotes the frequency of m-th subcarrier. To reduce delay spread, the frequency response of the effective channel should be flat and this means that the difference between the frequency response of the effective channel and the ideal channel response should be minimized. Hence, we formulate an objective function to minimize the aforementioned difference and it can be expressed as follows.

$$J\left(\mathbf{W}, a, \tau\right) = \sum_{m=1}^{M} \left| h_{\text{eff},m} - h_{\text{ideal},m} \right|^2 = \sum_{m=1}^{M} \left| \mathbf{h}_m^H \mathbf{w}_m - ae^{-j2\pi f_m \tau} \right|^2 . \tag{13}$$

The optimization problem with this objective function should be solved with respect to **W**, a, and τ. The objective function is convex with respect to **W** and a, but not to τ. The proof for convexity is shown as follows. To prove the convexity with respect to **W**, a, and τ, we partially differentiate the objective function with respect to each of them.

$$
\begin{aligned}
\nabla_{\mathbf{w}_m} J &= 2\frac{\partial J}{\partial \mathbf{w}_m^*} = 2\frac{\partial}{\partial \mathbf{w}_m^*}\left(\mathbf{w}_m^H \mathbf{h}_m - a e^{j2\pi f_m \tau}\right)\left(\mathbf{h}_m^H \mathbf{w}_m - a e^{-j2\pi f_m \tau}\right) \\
&= 2\frac{\partial}{\partial \mathbf{w}_m^*}\left(\mathbf{w}_m^H \mathbf{h}_m \mathbf{h}_m^H \mathbf{w}_m - a e^{j2\pi f_m \tau}\mathbf{h}_m^H \mathbf{w}_m - a e^{-j2\pi f_m \tau}\mathbf{w}_m^H \mathbf{h}_m + a^2\right) \\
&= 2\left(\mathbf{h}_m \mathbf{h}_m^H \mathbf{w}_m - a e^{-j2\pi f_m \tau}\mathbf{h}_m\right) \\
&\left(\because J \text{ is a real function of a complex vector } \mathbf{w}_m,\ \nabla_{\mathbf{w}_m} J = 2\frac{\partial J}{\partial \mathbf{w}_m^*}\right)
\end{aligned}
\tag{14}
$$

$$
\begin{aligned}
&\nabla_{\mathbf{w}_m}^2 J = 2\mathbf{h}_m \mathbf{h}_m^H \geq 0 \\
&\left(\because \text{Let } A = \mathbf{h}_m \mathbf{h}_m^H,\ x^H A x = x^H \mathbf{h}_m \mathbf{h}_m^H x = \left(\mathbf{h}_m^H x\right)^H \left(\mathbf{h}_m^H x\right) \geq 0\right)
\end{aligned}
\tag{15}
$$

$$
\begin{aligned}
\frac{\partial J}{\partial a} &= \frac{\partial}{\partial a}\sum_{m=1}^{M}\left(\mathbf{w}_m^H \mathbf{h}_m - a e^{j2\pi f_m \tau}\right)\left(\mathbf{h}_m^H \mathbf{w}_m - a e^{-j2\pi f_m \tau}\right) \\
&= \frac{\partial}{\partial a}\sum_{m=1}^{M}\left(\mathbf{w}_m^H \mathbf{h}_m \mathbf{h}_m^H \mathbf{w}_m - a e^{j2\pi f_m \tau}\mathbf{h}_m^H \mathbf{w}_m - a e^{-j2\pi f_m \tau}\mathbf{w}_m^H \mathbf{h}_m + a^2\right) \\
&= \sum_{m=1}^{M}\left(-e^{j2\pi f_m \tau}\mathbf{h}_m^H \mathbf{w}_m - e^{-j2\pi f_m \tau}\mathbf{w}_m^H \mathbf{h}_m + 2a\right) \\
&= 2aM - \sum_{m=1}^{M}\left(e^{j2\pi f_m \tau}\mathbf{h}_m^H \mathbf{w}_m + e^{-j2\pi f_m \tau}\mathbf{w}_m^H \mathbf{h}_m\right)
\end{aligned}
\tag{16}
$$

$$
\frac{\partial^2 J}{\partial a^2} = 2M \geq 0 \tag{17}
$$

$$
\begin{aligned}
\frac{\partial J}{\partial \tau} &= \frac{\partial}{\partial \tau}\sum_{m=1}^{M}\left(\mathbf{w}_m^H \mathbf{h}_m - a e^{j2\pi f_m \tau}\right)\left(\mathbf{h}_m^H \mathbf{w}_m - a e^{-j2\pi f_m \tau}\right) \\
&= \frac{\partial}{\partial \tau}\sum_{m=1}^{M}\left(\mathbf{w}_m^H \mathbf{h}_m \mathbf{h}_m^H \mathbf{w}_m - a e^{j2\pi f_m \tau}\mathbf{h}_m^H \mathbf{w}_m - a e^{-j2\pi f_m \tau}\mathbf{w}_m^H \mathbf{h}_m + a^2\right) \\
&= \sum_{m=1}^{M}\left(-j2\pi f_m a e^{j2\pi f_m \tau}\mathbf{h}_m^H \mathbf{w}_m + j2\pi f_m a e^{-j2\pi f_m \tau}\mathbf{w}_m^H \mathbf{h}_m\right)
\end{aligned}
\tag{18}
$$

$$
\begin{aligned}
\frac{\partial^2 J}{\partial \tau^2} &= \sum_{m=1}^{M}\left\{\left(2\pi f_m\right)^2 a e^{j2\pi f_m \tau}\mathbf{h}_m^H \mathbf{w}_m + \left(2\pi f_m\right)^2 a e^{-j2\pi f_m \tau}\mathbf{w}_m^H \mathbf{h}_m\right\} \\
&= \sum_{m=1}^{M}\left(2\pi f_m\right)^2 a\left\{e^{j2\pi f_m \tau}\mathbf{h}_m^H \mathbf{w}_m + e^{-j2\pi f_m \tau}\mathbf{w}_m^H \mathbf{h}_m\right\}
\end{aligned}
\tag{19}
$$

Since we cannot determine whether the last equation (19) is positive or not, the objective function is not convex with respect to τ. Hence, the objective function is difficult to solve. However, this objective function can be reduced to find the precoding vector and the magnitude of desired frequency response without the desired delay. The equivalent objective function can be obtained as follows.

$$\begin{aligned}\sum_{m=1}^{M}\left|\mathbf{h}_m^H\mathbf{w}_m - ae^{-j2\pi f_m\tau}\right|^2 &= \sum_{m=1}^{M}\left|e^{-j2\pi f_m\tau}\right|^2\left|\mathbf{h}_m^H\mathbf{w}_m e^{j2\pi f_m\tau} - a\right|^2 \\ &= \sum_{m=1}^{M}\left|\mathbf{h}_m^H\mathbf{w}_m e^{j2\pi f_m\tau} - a\right|^2 \quad \left(\because \left|e^{-j2\pi f_m\tau}\right|^2 = 1\right) \quad (20) \\ &= \sum_{m=1}^{M}\left|\mathbf{h}_m^H\tilde{\mathbf{w}}_m - a\right|^2 \quad \text{where } \tilde{\mathbf{w}}_m = \mathbf{w}_m e^{j2\pi f_m\tau}\end{aligned}$$

$\tilde{\mathbf{w}}_m$ denotes the precoding vector with phase compensation, and this equation means that the precoding vector which is obtained by the original objective function is equivalent to the precoding vector with phase compensation. Hence, we just solve the equivalent problem with respect to **W** and *a*. The equivalent objective function is expressed as

$$J_{\text{eq}} = \sum_{m=1}^{M}\left|\mathbf{h}_m^H\tilde{\mathbf{w}}_m - a\right|^2 . \quad (21)$$

The equivalent objective function is convex with respect to $\tilde{\mathbf{W}}$ and *a*, and the proof of convexity is shown as follows.

$$\begin{aligned}\nabla_{\tilde{\mathbf{w}}_m} J_{\text{eq}} &= 2\frac{\partial J_{\text{eq}}}{\partial \tilde{\mathbf{w}}_m^*} = 2\frac{\partial}{\partial \tilde{\mathbf{w}}_m^*}\left(\tilde{\mathbf{w}}_m^H\mathbf{h}_m - a\right)\left(\mathbf{h}_m^H\tilde{\mathbf{w}}_m - a\right) \\ &= 2\frac{\partial}{\partial \tilde{\mathbf{w}}_m^*}\left(\tilde{\mathbf{w}}_m^H\mathbf{h}_m\mathbf{h}_m^H\tilde{\mathbf{w}}_m - a\mathbf{h}_m^H\tilde{\mathbf{w}}_m - a\tilde{\mathbf{w}}_m^H\mathbf{h}_m + a^2\right) \quad (22) \\ &= 2\left(\mathbf{h}_m\mathbf{h}_m^H\tilde{\mathbf{w}}_m - a\mathbf{h}_m\right)\end{aligned}$$

$$\begin{aligned}&\nabla_{\tilde{\mathbf{w}}_m}^2 J_{\text{eq}} = 2\mathbf{h}_m\mathbf{h}_m^H \geq 0 \\ &\left(\because \text{Let } A{=}\mathbf{h}_m\mathbf{h}_m^H,\ x^H A x = x^H\mathbf{h}_m\mathbf{h}_m^H x = \left(\mathbf{h}_m^H x\right)^H\left(\mathbf{h}_m^H x\right) \geq 0\right)\end{aligned} \quad (23)$$

$$
\begin{aligned}
\frac{\partial J_{\mathrm{eq}}}{\partial a} &= \frac{\partial}{\partial a}\sum_{m=1}^{M}\left(\tilde{\mathbf{w}}_m^H\mathbf{h}_m - a\right)\left(\mathbf{h}_m^H\tilde{\mathbf{w}}_m - a\right) \\
&= \frac{\partial}{\partial a}\sum_{m=1}^{M}\left(\tilde{\mathbf{w}}_m^H\mathbf{h}_m\mathbf{h}_m^H\tilde{\mathbf{w}}_m - a\mathbf{h}_m^H\tilde{\mathbf{w}}_m - a\tilde{\mathbf{w}}_m^H\mathbf{h}_m + a^2\right) \\
&= \sum_{m=1}^{M}\left(-\mathbf{h}_m^H\tilde{\mathbf{w}}_m - \tilde{\mathbf{w}}_m^H\mathbf{h}_m + 2a\right) \\
&= 2aM - \sum_{m=1}^{M}\left(\mathbf{h}_m^H\tilde{\mathbf{w}}_m + \tilde{\mathbf{w}}_m^H\mathbf{h}_m\right)
\end{aligned}
\tag{24}
$$

$$
\frac{\partial^2 J_{\mathrm{eq}}}{\partial a^2} = 2M \geq 0 \tag{25}
$$

3.2.2 Channel gain degradation tolerance

To reduce effective delay spread using precoding vector selection, the effective channel gain can be degraded compared to maximum ratio transmission (MRT). Therefore, the optimization problem should be constrained to prevent a large average channel gain degradation. The constraint can be described as

$$
\sum_{m=1}^{M}\left|\mathbf{h}_m^H\tilde{\mathbf{w}}_m\right|^2 \geq (1-\delta)\sum_{m=1}^{M}\left\|\mathbf{h}_m\right\|^2 \tag{26}
$$

where $\left\|\mathbf{h}_m\right\|^2$ and $\left|\mathbf{h}_m^H\tilde{\mathbf{w}}_m\right|^2$ denotes the channel gain of m-th subcarrier with MRT and the proposed precoding algorithm, respectively, and $\delta \in [0,1]$ denotes the tolerance of the channel gain degradation which is an operator's parameter. When δ closes to 0, the channel gain is preserved closed to MRT, on the other hand, the channel gain is not considered to solve the problem as δ goes to 1.

3.3 Problem formulation

From the subsection 3.2.1 and 3.2.2, the precoding vector selection problem can be formulated as

$$\begin{aligned} &\underset{\mathbf{W},a}{\text{minimize}}\ J_{\text{eq}} = \sum_{m=1}^{M} \left| \mathbf{h}_m^H \tilde{\mathbf{w}}_m - a \right|^2 \\ &\text{subject to} \sum_{m=1}^{M} \left| \mathbf{h}_m^H \tilde{\mathbf{w}}_m \right|^2 \ge (1-\delta) \sum_{m=1}^{M} \left\| \mathbf{h}_m \right\|^2, \\ &\qquad\qquad \left\| \tilde{\mathbf{w}}_m \right\| = 1,\ \forall m. \end{aligned} \tag{27}$$

The last constraint is a normalization constraint. Unfortunately, the above problem is not a convex optimization problem due to the first constraint. Hence, we apply the interior-point method with initial point value to solve the problem. After applying the logarithmic barrier method which is a kind of interior-point methods, the equivalent optimization problem can be described as

$$\begin{aligned} &\underset{\mathbf{W},a}{\text{minimize}}\ f_t(\mathbf{W}, a) = t \sum_{m=1}^{M} \left| \mathbf{h}_m^H \tilde{\mathbf{w}}_m - a \right|^2 \\ &\qquad\qquad - \log\left(\sum_{m=1}^{M} \left| \mathbf{h}_m^H \tilde{\mathbf{w}}_m \right|^2 - (1-\delta) \sum_{m=1}^{M} \left\| \mathbf{h}_m \right\|^2 \right) \\ &\text{subject to}\ \left\| \tilde{\mathbf{w}}_m \right\| = 1,\ \forall m. \end{aligned} \tag{28}$$

The whole procedure of the proposed precoding vector selection algorithm is shown in Table 2.

Algorithm 1. Proposed effective delay spread reduction precoding algorithm

Initialize $\tilde{\mathbf{W}}$

while until J_{eq} converges **do**

(Log-barrier method)

while until t grows sufficiently large **do**

$$f_t(\mathbf{W}, a) = t\sum_{m=1}^{M}\left|\mathbf{h}_m^H \tilde{\mathbf{w}}_m - a\right|^2 - \log\left(\sum_{m=1}^{M}\left|\mathbf{h}_m^H \tilde{\mathbf{w}}_m\right|^2 - (1-\delta)\sum_{m=1}^{M}\left\|\mathbf{h}_m\right\|^2\right)$$

(Gradient descent method)

while until $\tilde{\mathbf{W}}$ converges **do**

$$\hat{\mathbf{W}}_i = \tilde{\mathbf{W}}_{i-1} - \mu\nabla_{\mathbf{W}} f_t$$

$$\tilde{\mathbf{w}}_m = \frac{\hat{\mathbf{w}}_m}{\left\|\hat{\mathbf{w}}_m\right\|}$$

$$a_i = a_{i-1} - \mu\nabla_a f_t$$

end while

end while

end while

Table 2. Proposed effective delay spread reduction precoding algorithm

Chapter 4

Simulation Results

In this chapter, the proposed precoding algorithm is evaluated and compared with conventional precoding techniques referred to as maximum ratio transmission (MRT). Simulation result shows averaged results of 100 independent channel realization except for an instance realization case.

4.1 Simulation parameters

In simulations, the system parameters are set to default values which are shown in Table 3. Other parameters related to channel and systems are basically set as in 3GPP spatial channel model (SCM) [6] and LTE [11] scenario except for the subcarrier bandwidth related to the symbol duration. The symbol duration is set to the half of current LTE OFDM system.

Parameter	Variable	Value
# BS antennas	N	4
Sampling frequency	f_s	1.92 MHz
Subcarrier bandwidth	Δf	30 kHz
Channel gain tolerance	δ	0.5

Table 3. Simulation parameters

4.2 Channel response of an instance realization

Figure 6~Figure 7, Table 4 and Table 5 depict the results for an instance channel realization. The original channel response means the effective channel response at the receiver without precoding. It has 868.03ns RMS delay spread while the effective channel with MRT has 506.79ns RMS delay spread. This delay spread reduction effect in MRT is caused by spatial diversity of antenna array which occurs channel hardening effect. As we expected, channel gain in MRT is improved compared to original channel response. As depicted in Figure 6, the proposed algorithm makes the effective channel response flat, and channel gain degradation is regulated by the tolerance constraint. As shown in Figure 7, the delay spread is effectively reduced. The RMS delay spread of the effective channel response with proposed algorithm is 157.28ns which is less than one third of MRT RMS delay spread. However, the channel gain of the effective channel response with the proposed precoding is degraded about 1.3dB. It was an expected result because the fraction of delay components is eliminated by the proposed precoding algorithm.

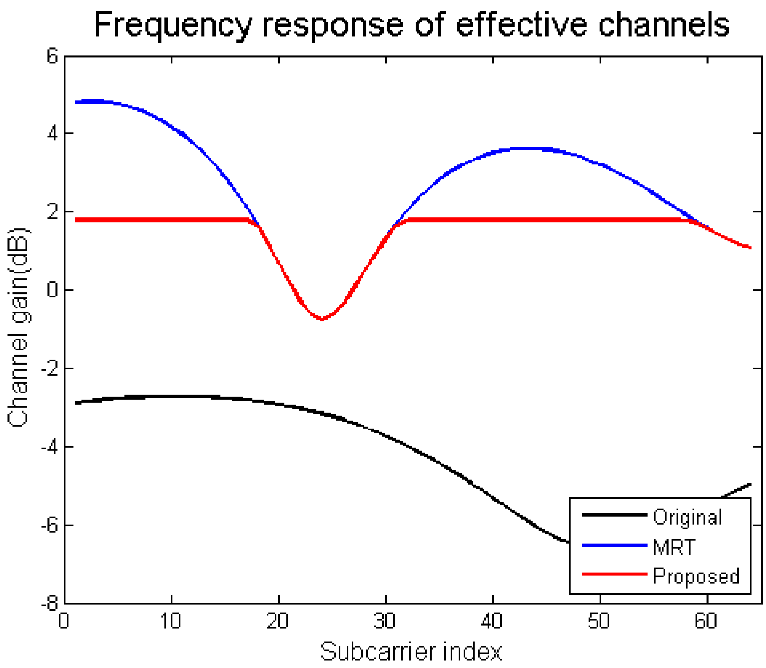

Figure 6. Frequency responses of the effective channel responses

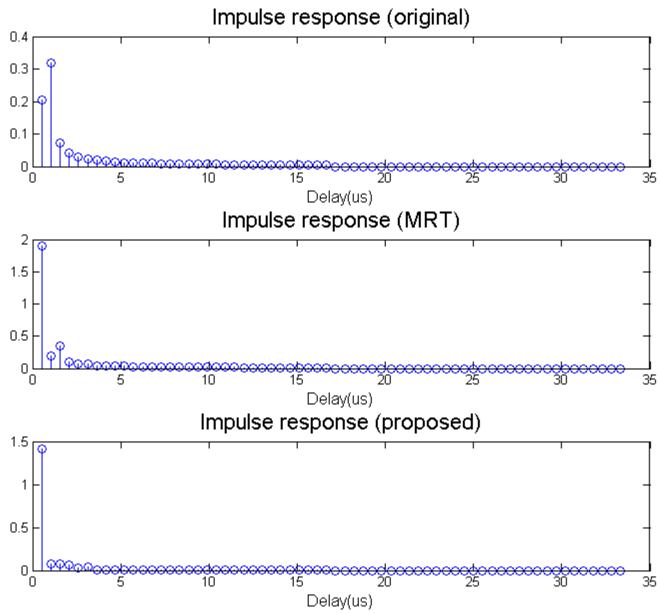

Figure 7. Impulse response of the effective channel responses

Average channel gain	
Original	-4.1167 dB
MRT	2.7960 dB
Proposed	1.4908 dB

Table 4. Average channel gain of the effective channel responses

RMS delay spread	
Original	868.03 ns
MRT	506.79 ns
Proposed	157.28 ns

Table 5. RMS delay spread of the effective channel responses

4.3 Delay spread and average channel gain

For 100 individual channel realizations, the simulation results are shown in Figure 8 and Figure 9. Delay spread is described in empirical cumulative distribution function (empirical CDF). As we expected, the proposed precoding algorithm reduces the delay spread of the effective channel response compared to MRT precoding. Channel gain is degraded in the proposed precoding algorithm compared to MRT precoding. The degradation is about 1.47dB that means about 29% channel gain reduction with 4 BS antennas. Therefore, the proposed precoding algorithm utilizes the channel gain as an expense of the effective delay spread reduction.

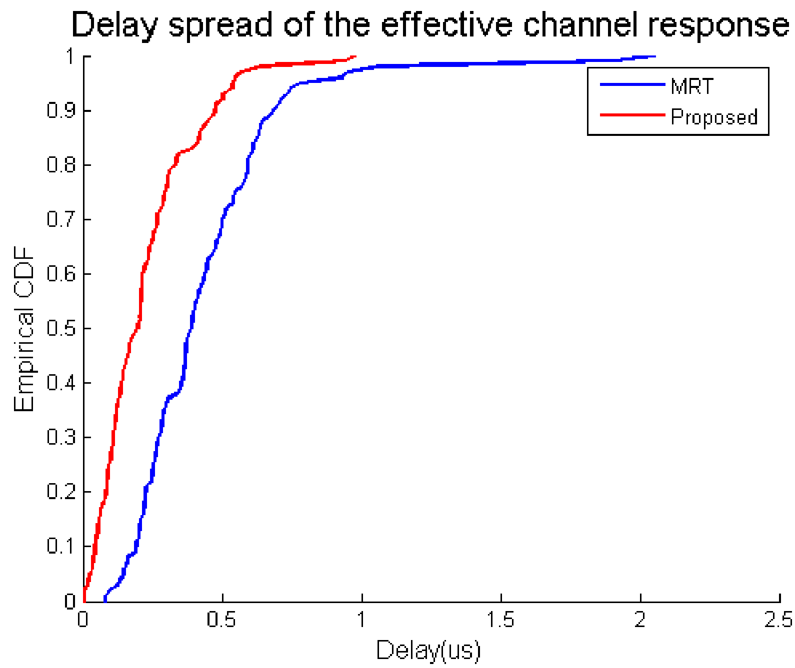

Figure 8. Delay spread with N=4, δ=0.5

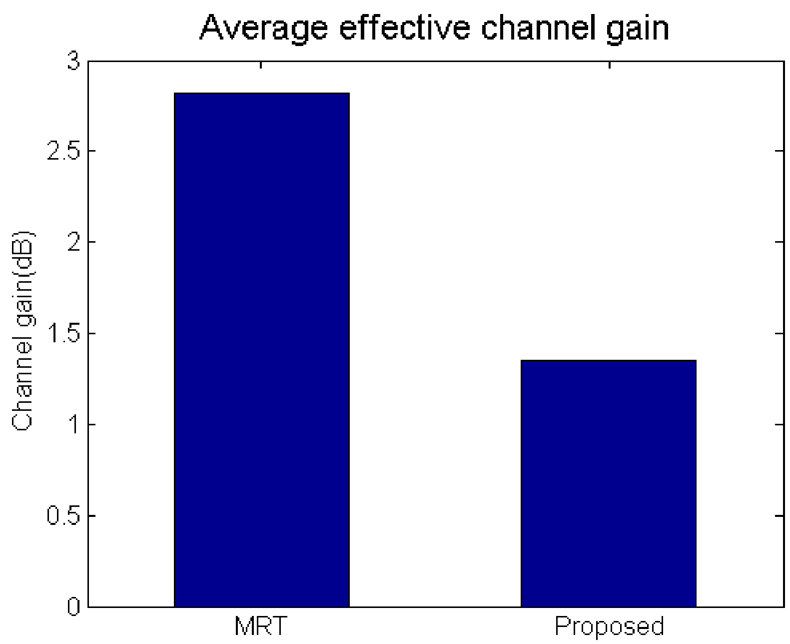

Figure 9. Average effective channel gain with N=4, δ=0.5

4.4 Delay spread and average channel gain with varying *N*

In this subsection, we simulate the proposed algorithm with varying the number of BS antennas from N=2 to 8. When the number of BS antennas increases, the delay spreads of MRT and the proposed algorithm decrease as depicted in Figure 10. This phenomenon was reported in [12] and the reason is that the spatial diversity of antenna array causes the channel hardening effect. Average channel gain performance is shown in Figure 11. Since the array gain is appeared in MRT, the channel gain of MRT increases as the number of BS antennas increases. The proposed algorithm improves the channel gain which has a similar tendency to MRT when the number of BS antennas increases. Difference of the effective channel gains of MRT and the proposed algorithm is maintained when the number of antennas increases. For instance, the degradations are 1.2416dB, 1.2488dB, 1.2620dB, 1.2807dB with N=2, 4, 6, 8, respectively.

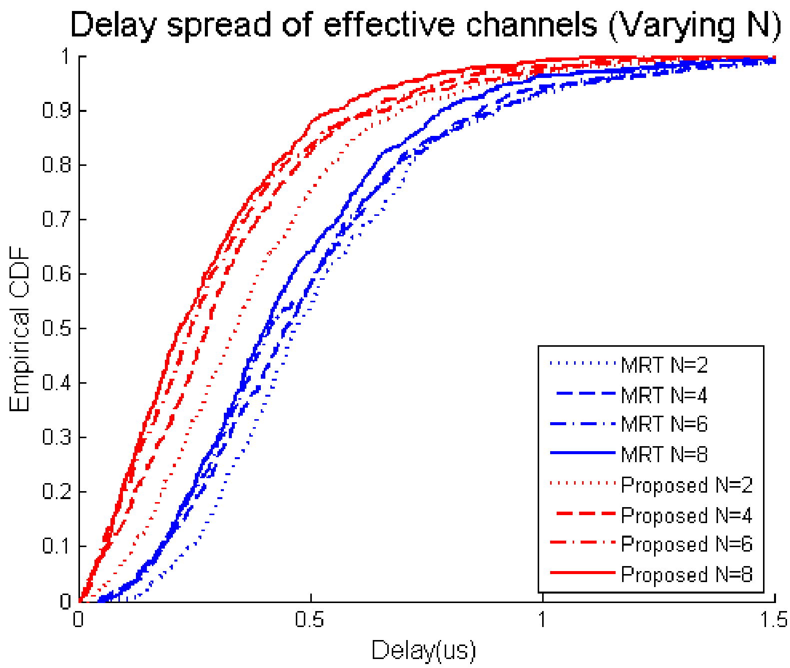

Figure 10. Delay spread with varying N, δ=0.5

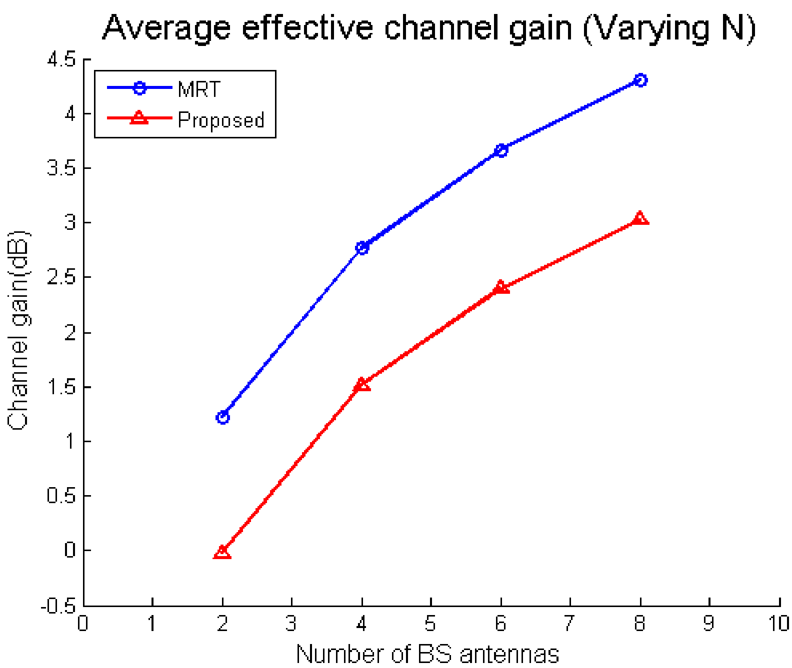

Figure 11. Average effective channel gain with varying N, δ=0.5

4.5 Delay spread and average channel gain with varying δ

We carry a simulation to describe delay spread and channel gain of the effective channels with varying the channel gain degradation tolerance, δ, which regulates the performance of the proposed algorithm. The results are shown in Figure 12 and Figure 13. When the tolerance becomes close to 1, the delay spread of the effective channel goes to zero as we expected. From the result in terms of channel gain performance, the channel gain tends to degrade as the tolerance becomes larger. On the other hand, the delay spread is closed to MRT and the channel gain is maintained as MRT when the tolerance goes to 0. So, the channel gain degradation tolerance δ is considered to operator's parameter. When the communication system requires low delay spread, δ is established as high as close to 1. In contrast, δ is set to lower than before if the system requires high channel gain performance rather than low delay spread.

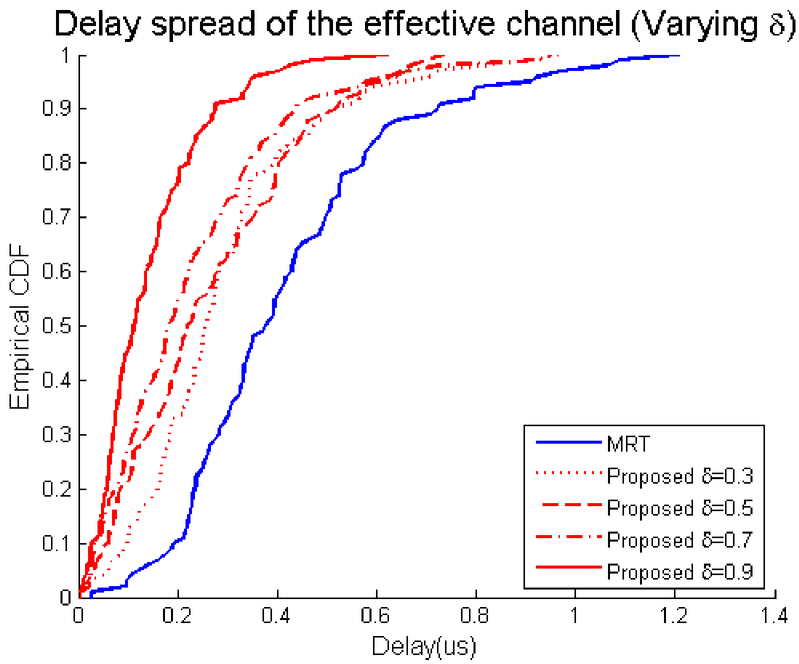

Figure 12. Delay spread with N=4, varying δ

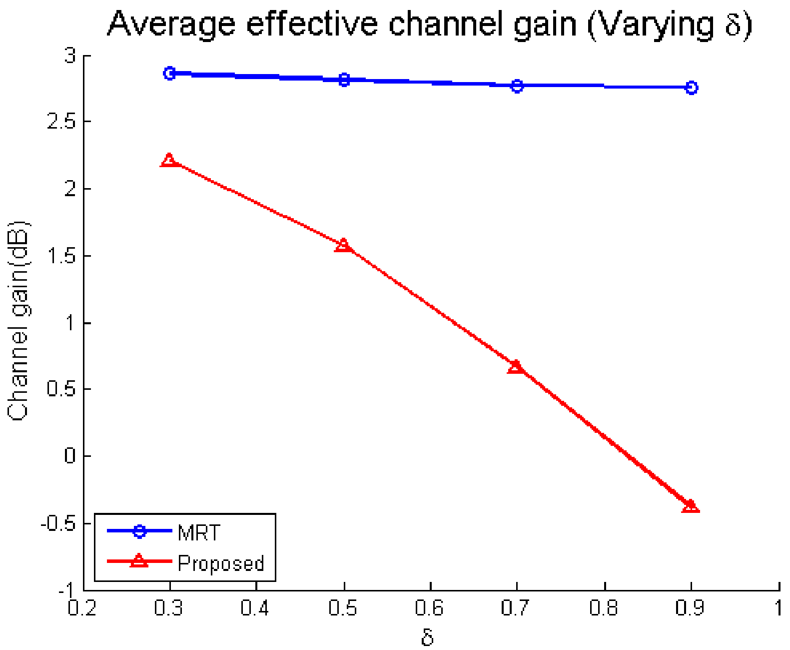

Figure 13. Average effective channel gain with N=4, varying δ

Chapter 5

Conclusion and Future Work

In this thesis, we have proposed an effective delay spread reduction precoding algorithm for the future wireless system which requires extremely low latency. To satisfy low latency requirements, MISO-OFDM systems operate with short symbol duration and those systems suffer from large cyclic prefix overhead. To solve this problem, we have proposed a precoding algorithm to reduce effective delay spread. As shown in simulation results, the proposed algorithm effectively reduces the effective delay spread with smaller number of antennas compared to MRT while maintaining the tolerable channel gain performance. The proposed algorithm have shown the tradeoff between delay spread and channel gain performance and the algorithm can regulate the performance by a tolerance parameter.

Since the channel gain performance is degraded, the throughput of the system with the proposed algorithm would be reduced. However, the delay spread reduction provides the possibility of cyclic prefix overhead reduction in OFDM systems and the throughput maximization problem utilizing the tradeoff between delay spread and channel gain can be formulated in future work.

Future work may include the proposed algorithm for multiuser MIMO systems with large-scale base station antennas referred to as massive MIMO systems. In massive MIMO systems, there exists excess degrees of freedom (DoF), and then we can utilize zero forcing beamforming (ZFBF) with other achievement by exploiting excess DoF [13]. In this point of view, the proposed algorithm can reduce delay spread while multiuser interference (MUI) nulling by exploiting excess degrees of freedom (DoF). Also, the performance of the proposed algorithm is consistent with the philosophy of massive MIMO systems since the performance improves as the number of antennas increases.

Bibliography

[1] G. Wunder *et al.*, "5GNOW: non-orthogonal, asynchronous waveforms for future mobile applications," *Communications Magazine, IEEE* , vol.52, no.2, pp.97,105, February 2014.

[2] G. P. Fettweis, "The Tactile Internet: Applications and Challenges," *Vehicular Technology Magazine, IEEE* , vol.9, no.1, pp.64,70, March 2014

[3] J. Proakis, M. Salehi, *Digital Communications*, 5th ed., McGraw-Hill Science, 2007.

[4] L. -U Choi, R. D. Murch, "A transmit MIMO scheme with frequency domain pre-equalization for wireless frequency selective channels," *Wireless Communications, IEEE Transactions on* , vol.3, no.3, pp.929,938, May 2004.

[5] Z. Yu, K. B. Letaief, "Frequency domain pre-equalization with transmit precoding for MIMO broadcast wireless channels," *Selected Areas in Communications, IEEE Journal on* , vol.26, no.2, pp.389,400, February 2008.

[6] 3GPP TR 25.996, "Universal Mobile Telecommunications Systems (UMTS); Spatial channel model for Multiple Input Multiple Output (MIMO) simulations".

[7] D. Tse, P. Viswanath, *Fundamentals of Wireless Communication*, Cambridge University Press, 2005.

[8] A. Goldsmith, *Wireless Communications*, Cambridge University Press, 2005.

[9] R. A. Monzingo, T. W. Miller, *Introduction to adaptive arrays*, SciTech Publishing, 1980.

[10] S. Haykin, B. Van Veen, *Signals and Systems*, 2nd ed., John Wiley & Sons, 2002.

[11] 3GPP TS 36.211, "Evolved Universal Terrestrial Radio Access (E-UTRA); Physical channels and modulation".

[12] S. Payami, F. Tufvesson, in "Delay spread properties in a measured massive MIMO system at 2.6 GHz," *Personal Indoor and Mobile Radio Communications (PIMRC), 2013 IEEE 24th International Symposium on* , vol., no., pp.53,57, 8-11 Sept. 2013.

[13] E. Larsson, O. Edfors, F. Tufvesson, T. Marzetta, “Massive MIMO for next generation wireless systems,” *Communications Magazine, IEEE* , vol.52, no.2, pp.186,195, February 2014.

초 록

최근, 차세대 무선 통신 시스템을 위한 많은 기술들이 활발히 논의되고 있고, 그 중 저지연 통신 기술은 가상 현실 게임, 원격 제어 로봇, 차량 비상 경고 등의 지연 시간에 민감한 응용 분야를 지원하기 위한 중요한 이슈로 부각되고 있다. 저지연 통신을 위해서는 통신 시스템의 심볼 주기가 현재의 직교 주파수 분할 다중화 (orthogonal frequency division multiplexing, OFDM) 방식에서의 심볼 주기보다 훨씬 짧아져야 한다. 그러나, 심볼 주기를 줄이게 되면, 직교 주파수 분할 다중화 방식 시스템에서 심볼 간 간섭을 대처하기 위해 존재하는 cyclic prefix (CP)의 오버헤드가 많은 시간 자원을 차지하게 되고, 저지연 통신 시스템에서 효율을 떨어뜨리는 요인이 된다. 한편, 지연 확산을 감소시키지 않고 CP의 길이를 줄이면 심볼 간 간섭에 의해 시스템의 성능이 저하되는 원인이 된다. 따라서, 저지연 통신을 위해 CP 길이를 줄이면서도 이로 인해 발생하는 ISI를 줄일 수 있어야 한다.

본 학위 논문에서는 저지연 다중 안테나 직교 주파수 다중화 방식 시스템에서 CP 길이를 줄이기 위해 유효 지연 확산을 감소시키는 프리코딩 (precoding) 기법을 제안한다. 우선, 지연 확산을 감소시키면서도 신호 대 잡음 비의 손실을 일정 수준 보장하기 위한 최적화 문제를 설계하고, 최적화 문제의 복잡도를 감소시키기 위해 경사 하강법 (gradient descent method)을 적용한 효율적인 프리코딩 알고리즘을 제안한다.

시뮬레이션 결과에서 제안된 알고리즘이 유효 채널 이득을 최대한 유지하면서 지연을 효과적으로 감소시킨다는 것을 확인할 수 있고,

안테나가 증가함에 따라 성능이 향상함을 알 수 있다. 향후, 제안된 알고리즘은 다중 사용자 다중 안테나 환경으로 확장될 수 있으며, 이 때, 최근 이슈가 되고 있는 거대 다중 안테나 (massive MIMO) 시스템이 활용될 수 있다. 거대 다중 안테나 시스템에서 제안된 알고리즘을 활용하면 여분의 자유도 (excess degrees of freedom)를 활용해 사용자 간 간섭을 제거하면서 지연 확산도 감소시킬 수 있으리라 기대된다.

주요어: 저지연, 다중 안테나, 프리코딩, 지연 확산, 직교 주파수 분할 다중화 방식, 심볼 간 간섭

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