Transmit-Nulling SDMA for Coexistence with Fixed Wireless Service

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Abstract

This paper proposes a systematic design for a precoding codebook for a transmit-nulling space-division multiple access (TN-SDMA) sharing spectrum with existing fixed wireless service (FWS). Based on an estimated direction angle of a victim FWS system, an interfering transmitter adaptively constructs a codebook, forming a transmit null in the direction angle, while satisfying orthogonal beamforming constraints. Sum throughput results indicate that the throughput loss of TN-SDMA relative to a practical SDMA, called *per user unitary and rate control* (PU^{2}RC), is lower at larger number of transmission antennas, lower signal-to-noise ratio, or a smaller number of users. In particular, a small loss (12% throughput loss) is provided for practical system parameters. Spectrum sharing results confirm that TN-SDMA efficiently shares spectrum with FWS systems by reducing protection distance to more than 66 %. Although a TN-SDMA system always has lower throughput compared to PU^{2}RC in non-coexistence scenarios; it offers an intriguing opportunity to re-use a spectrum already allocated to an FWS.

Key words : SDMA, Orthogonal Beamforming, Transmit-Nulling, Spectrum Sharing.

I. Introduction

The 3,400~3,600 band for International Mobile Telecommunications (IMT)-Advanced was already allocated, or is under consideration, for fixed wireless services (FWS), such as fixed services (FS), fixed satellite services (FSS), or fixed wireless access (FWA) in many countries around the world [1]. Moreover, an IMT-Advanced system requires higher data rates: approximately 100 Mbps and 1 Gbps with high-mobility and 100 MHz for low-mobility bandwidth, respectively [2]. Space-division multiple access (SDMA), which is capable of achieving very high throughput, is a candidate for Long-Term Evolution (LTE)-Advanced and IMT-Advanced standards [3]. This paper considers in particular an SDMA approach that allows spectrum sharing with existing FWS systems.

Spectrum sharing among wireless communication systems is made possible by adjusting system resources, such as transmission power, frequency, time of transmission, etc. Multiple antenna arrays using null-steering [4] can protect existing systems without additional radio resources in frequency or time. For a base station (BS) using null-steering, no downlink throughput gain is obtained, due to its focusing on the mitigation of interference toward a victim system. Meanwhile, a codebook-based orthogonal beamforming SDMA, which is proposed for 3GPP-LTE standards [5] under the name per-user unitary rate control (PU^{2}RC), focuses only on throughput improvement, based on the reduction of inter-user interferences in homogeneous systems. However, multi-user MIMO systems coexisting with FWS requires a suppression of the interference between heterogeneous systems, as well as the inter-user interferences in homogeneous systems.

Combining null-steering and orthogonal beamforming provides a clue for achieving both high data rates and spectrum sharing. A part of the spatial transmission resources provided by multiple antennas is employed for spectrum sharing instead of data transmission. This concept is embodied by using a multi-user precoder satisfying both null-steering and orthogonality constraints, where a precoding matrix comprises mutually $N-1$ (N denotes the number of transmission antennas) orthonormal vectors that are orthogonal to an array-steering vector in the direction of a victim system. The Gram-Schmidt process can be a general method for designing a precoding matrix satisfying both constraints, but it has drawbacks in the usage of a codebook-based SDMA[1]) where both transmitter and receiver always use an identical codebook. Given an arbitrary initial basis (linear independent set) containing an array-steering vector in the direction

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1) We focus on codebook-based SDMA because of its strength; that is, link adaptation and user scheduling facilitated by more exact SINR estimation in comparison with zero-forcing SDMA [6].

of a victim system, the Gram-Schmidt process constructs a precoding matrix (orthogonal basis). Therefore, using identical initial bases at both transmitter and receiver is needed for an identical codebook construction at both sides. To do this, a transmitter calculates initial bases according to the array-steering vector and sends it to a receiver. However, the procedure results in additional downlink overhead.

To overcome the drawbacks of the Gram-Schmidt process, we propose a systematic design for transmit-nulling SDMA (TN-SDMA) precoder codebooks, which forms a transmit null in the direction of victim FWS systems, and satisfies orthogonal beamforming constraints. The design ensures low complexity, as the precoding matrix is given as a product of a linear transformation matrix and a column-reduced discrete Fourier transform (DFT) matrix. Moreover, unlike in the Gram-Schmidt process, both transmitter and receiver independently construct identical codebooks at both ends by sharing the proposed systematic design rule, based only on the direction angle of victim FWS systems, which takes fewer overhead bits than does the Gram-Schmidt process.

II. System Model and Algorithms

A downlink system consisting of a BS (or transmitter) with N transmission antennas and K users with one receiving antenna, operating in the spectrum owned by a FWS, is considered. We assume homogeneous users and a flat Rayleigh fading channel from the BS to the kth user. Letting $\mathbf{x} \in \mathbb{C}^{N \times 1}$ be a transmission symbol vector, the received signal of the kth user is given by

$$y_k = \mathbf{h}_k \mathbf{x} + z_k, \tag{1}$$

where $\mathbf{h}_k \in \mathbb{C}^{1 \times N}$ is a channel gain vector with zero mean unit variance, and z_k is an additive noise with unit variance complex in a Gaussian noise vector. We assume that channel gain vector $\mathbf{h}_k$ has uncorrelated complex Gaussian entries. The investigation of a correlated channel model is left to future work. Unlike $\mathbf{h}_k$, however, we assume a highly-correlated channel from a BS to a FWS based on a high probability of line-of-sight between the two. It facilitates mitigating interference of FWS by constructing a transmit null at the BS.

An SDMA system that constructs $N-1$ orthonormal beams and transmits to $N-1$ scheduled users via precoding vector $\{\mathbf{f}_m\}_{m=1,\cdots,N-1}$. is considered. The precoded transmission signal is then

$$\mathbf{x} = \mathbf{F}\mathbf{s} = \sum_{m=1}^{N-1} \mathbf{f}_m s_m, \tag{2}$$

where $\mathbf{F} = [\mathbf{f}_1 \, \mathbf{f}_2 \cdots \mathbf{f}_{N-1}] \in \mathbb{C}^{N \times N-1}$ and $\mathbf{s} = [s_1 \, s_2 \cdots s_{N-1}]^T$ is a un-coded symbol vector with $\mathrm{E}\{\|s\|^2\} = P$. The total transmission power P is equally allocated over $N-1$ scheduled users. Note that P also denotes a signal-to-noise ratio (SNR), since we assume an additive noise with unit variance. The $N-1$ precoding vectors are selected within a TN-SDMA codebook, $\mathcal{F}(\hat{\phi})$, using the beam- and user-selection algorithm described in Section 3.2. In order to support SDMA and suppress interference against victim FWS systems, the codebook $\mathcal{F}(\hat{\phi})$ was designed based on the direction angle of victim FWS $\hat{\phi}$ (DOV), which is described in following section.

2-1 Systematic Codebook Design

In this section, a systematic design of a codebook for TN-SDMA is proposed to not only support SDMA, but also to suppress interference against a victim FWS. Based on the assumption that a BS is already aware of DOV$\hat{\phi}$, both the BS and users independently construct an identical codebook at both ends by sharing a codebook design rule and DOV$\hat{\phi}$. The BS obtains the DOV by adopting a popular spatial-spectrum estimation direction-finding method [7], or from a database with information concerning DOV. DOVs are then sent to all K users via a downlink control channel.

Our design objective is the construction of a codebook that produces a transmit null in DOV $\hat{\phi}$ while remaining orthogonal beamforming. Starting from the G sets of the DFT matrix, $\mathcal{E} = \{\mathbf{E}_g\}_{g=1,\cdots,G}$, a TN-SDMA codebook $\mathcal{F}(\hat{\phi}) = \{\mathbf{F}_g(\hat{\phi})\}_{g=1,\cdots,G}$ is designed. Here, G denotes the number of precoding matrices in the codebook $\mathcal{F}(\hat{\phi})$. The m precoding vector of matrix $\mathbf{E}_g$ is given by [8]

$$\mathbf{e}_{g,m} = \frac{1}{\sqrt{N}}\left[1 \; e^{j\frac{2\pi}{N}\left(\frac{g-1}{G}+m-1\right)} \cdots e^{j\frac{2\pi}{N}(N-1)\left(\frac{g-1}{G}+m-1\right)}\right]^T, \tag{3}$$

where $(\cdot)^T$ represents a transpose matrix operation. When considering uniform linear antenna arrays at a transmitter with spacings of d, the transmission gain of $\mathbf{e}_{g,m}$ in a direction ϕ is given by

$$\Omega(\mathbf{e}_{g,m},\phi) = |\mathbf{e}_{g,m}^T \mathbf{v}(\phi)|^2 = \left|\frac{1}{N}\sum_{k=0}^{N-1} e^{j2\pi k\left(\frac{1}{N}\left(\frac{g-1}{G}+m-1\right)+\frac{d}{\lambda}\sin\phi\right)}\right|^2, \tag{4}$$

where λ is the wavelength. $\mathbf{v}(\phi)$ is the array steering vector at ϕ, which is given by

$$\mathbf{v}(\phi) = \frac{1}{\sqrt{N}}\left[1 \quad e^{j2\pi\frac{d}{\lambda}\sin\phi} \cdots e^{j2\pi(N-1)\frac{d}{\lambda}\sin\phi}\right]^T, \tag{5}$$

The null points ϕ satisfying $\Omega(\mathbf{e}_{g,m}, \phi) = 0$ have the property described by the following Lemma.

Lemma 1: There exists at least one common null-point $\phi_g^{(n)}$, such that $\mathbf{e}_{g,m}^{\mathrm{T}} \mathbf{v}(\phi_g^{(n)}) = 0$ for all $m \in \mathcal{M}_n$, where $N \geq 2$, $-\pi/2 \leq \phi_g^{(n)} < \pi/2$, $n \in \{1, \cdots, N\}$, and $\mathcal{M}_n = \{1, \cdots, N\} \setminus \{n\}$.

Proof: $\{\mathbf{e}_{g,m}\}_{m=1,\cdots,N}$ is an orthonormal basis of a N-dimensional complex vector space $\mathbb{C}^N$ ($N \geq 2$). It then becomes clear that vector $\mathbf{e}_{g,n}$ is orthogonal to $N-1$ vectors $\{\mathbf{e}_{g,m}\}_{m \in \mathcal{M}_n}$ (e.g., $\mathbf{e}_{g,m}^{\mathrm{T}} \mathbf{e}_{g,n} = 0$ for $m \in \mathcal{M}_n$). At this point, it is shown that at least one $\phi_g^{(n)}$ exists such that $\mathbf{v}(\phi_g^{(n)}) = \mathbf{e}_{g,n}$. The complex exponential function $f(\phi) = e^{j\phi}$ has a period of 2π; therefore, $2\pi(N-1)\sin\phi_g^{(n)} d/\lambda = \frac{2\pi}{N}(N-1)((g-1)/G+m-1) + 2k\pi$ for any integer k to satisfy $\mathbf{v}(\phi_g^{(n)}) = \mathbf{e}_{g,n}$. The above equation is rewritten as $\sin\phi_g^{(n)} = \frac{\lambda}{dN}((g-1)/G+m-1) + k'$ for any integer k'. Then, without a loss of generality, there is at least one k' such that $\left| \frac{\lambda}{dN}((g-1)/G+m-1) + k' \right| \leq 1$, which indicates that there is at least one $\phi_g^{(n)}$ such that $\mathbf{v}(\phi_g^{(n)}) = \mathbf{e}_{g,n}$, $-\pi/2 \leq \phi_g^{(n)} < \pi/2$.

Remark 1: Lemma 1 states that, for one-column-reduced DFT matrix comprising mutually orthogonal column vectors, all beams that column vectors generate have at least one common null-point.

Let us define a matrix $\mathbf{G}_g^{(n)} \in \mathbb{C}^{N \times N-1}$ as the nth column-reduced matrix of $\mathbf{E}_g$, which consists of $N-1$ column vectors of $\mathbf{E}_g$, $\{\mathbf{e}_{g,m}\}_{m \in \mathcal{M}_n}$. Then, from Lemma 1, $\mathbf{G}_g^{(n)}$ has a transmission gain of zero at direction $\phi_g^{(n)}$ as follows:

$$\Omega(\mathbf{G}_g^{(n)}, \phi_g^{(n)}) = \sum_{m \in \mathcal{M}_n} |\mathbf{e}_{g,m}^{\mathrm{T}} \mathbf{v}(\phi_g^{(n)})|^2 = 0. \tag{6}$$

Constructing $\mathbf{F}_g(\hat{\phi})$—which forms a transmit null at DOV $\hat{\phi}$—from $\mathbf{G}_g^{(n)}$ demands adaptive steering of the null at a direction of $\phi_g^{(n)}$ to the $\hat{\phi}$. Here, since there are multiple numbers of $\phi_g^{(n)}$ for a specific g and (n) as stated in Lemma 1, one null-point ϕ_g minimizing $|\phi_g^{(n)} - \hat{\phi}|$, among $\phi_g^{(n)}$, is used for the null steering. Based on $\mathbf{G}_g^{(n)}$ and ϕ_g, the precoding matrix $\mathbf{F}_g(\hat{\phi})$ is finally constructed as shown in following theorem.

Theorem 1: (Orthogonal Beamforming Matrix with Transmit Null) Let $\alpha_g = \hat{\phi} - \phi_g$; TN-SDMA precoding matrix $\mathbf{F}_g(\hat{\phi})$ such that $\mathbf{F}_g^{\dagger}(\hat{\phi}) \mathbf{F}_g(\hat{\phi}) = \mathbf{I}$ (orthogonal beam-

forming) and $G(\mathbf{F}_g(\hat{\phi}), \hat{\phi}) = 0$ (transmit null construction at $\hat{\phi}$) is given as

$$\mathbf{F}_g(\hat{\phi}) = \mathbf{R}_g \mathbf{G}_g^{(n)}, \tag{7}$$

where $(\cdot)^{\dagger}$ denotes a conjugate transpose of a matrix, $\mathbf{I}$ is an identity matrix, and $\mathbf{R}_g = diag(\mathbf{r}_g)$ is an $N \times N$ diagonal matrix containing the vector of diagonal entries $\mathbf{r}_g$ given by

$$\mathbf{r}_g = \left[1 \quad e^{-j2\pi \frac{d}{\lambda}\cos\phi_g \sin\alpha_g} \left(e^{-j2\pi \frac{d}{\lambda}\sin\phi_g} \right)^{\cos\alpha_g - 1} \cdots \right.$$
$$\left. e^{-j2\pi(N-1)\frac{d}{\lambda}\cos\phi_g \sin\alpha_g} \left(e^{-j2\pi(N-1)\frac{d}{\lambda}\sin\phi_g} \right)^{\cos\alpha_g - 1} \right]^{\mathrm{T}}. \tag{8}$$

Proof: To reduce the complexity in a derivation of $\mathbf{F}_g(\hat{\phi}) \in \mathbb{C}^{N \times N-1}$ from $\mathbf{G}_g^{(n)} \in \mathbb{C}^{N \times N-1}$, we define a linear transformation $f: \mathbf{G}_g^{(n)} \to \mathbf{F}_g(\hat{\phi})$. Then, f is represented as a matrix where $\mathbf{R}_g \in \mathbb{C}^{N \times N}$ (e.g., $\mathbf{F}_g(\hat{\phi}) = \mathbf{R}_g \mathbf{G}_g^{(n)}$). Since $\mathbf{R}_g$ must be a unitary matrix to satisfy a given constraint $\mathbf{F}_g^{\dagger}(\hat{\phi}) \mathbf{F}_g(\hat{\phi}) = \mathbf{I}$, we assume a simple unitary matrix $\mathbf{R}_g$ given as a diagonal matrix $diag(\mathbf{r}_g)$, where the vector of diagonal entries $\mathbf{r}_g = [r_1 \cdots r_N]^{\mathrm{T}}$, such that $|r_i|^2 = 1, 1 \leq i \leq N$. Then, $\mathbf{R}_g \mathbf{G}_g^{(n)} = \mathbf{r}_g \otimes \mathbf{G}_g^{(n)}$, where $\otimes$ is the Hadamard (point-wise) product.

The transmission gain of $\mathbf{F}_g(\hat{\phi})$ in target direction $\hat{\phi}$ is then given as

$$\Omega(\mathbf{F}_g(\hat{\phi}), \hat{\phi}) = G(\mathbf{r}_g \otimes \mathbf{G}_g^{(n)}, \phi_g + \alpha_g) =$$
$$\sum_{m \in \mathcal{M}_n} \left| (\mathbf{r}_g \otimes \mathbf{e}_{g,m})^{\mathrm{T}} \mathbf{v}(\phi_g + \alpha_g) \right|^2$$
$$= \sum_{m \in \mathcal{M}_n} \left| \frac{1}{N} \sum_{k=0}^{N-1} r_{k+1} e^{j2\pi k \left(\frac{1}{N}\left(\frac{g-1}{G}+m-1 \right) + \frac{d}{\lambda}\sin(\phi_g + \alpha_g) \right)} \right|^2$$
$$\overset{(a)}{=} \sum_{m \in \mathcal{M}_n} \left| \frac{1}{N} \sum_{k=0}^{N-1} [r_{k+1} e^{jkq\cos\phi_g \sin\alpha_g} \left(e^{jkq\sin\phi_g} \right)^{\cos\alpha_g - 1} \right.$$
$$\left. e^{j\frac{2\pi k}{N}\left(\frac{g-1}{G}+m-1 \right)} e^{jkq\sin\phi_g}] \right|^2, \tag{9}$$

where $q = 2\pi d/\lambda$, $\mathcal{M}_n = \{1, \cdots, N\} \setminus \{n\}$, and (a) is obtained by using the addition formula of trigonometric functions ($\sin(x+y) = \sin x \cos y + \cos x \sin y$). Here, let $r_{k+1} = e^{-jkq\cos\hat{\phi}\sin\alpha_g} \left(e^{-jkq\sin\hat{\phi}} \right)^{\cos\alpha_g - 1}$, and (9) is rewritten as

$$\Omega(\mathbf{F}_g(\hat{\phi}), \hat{\phi}) = \sum_{m \in \mathcal{M}_n} \left| \frac{1}{N} \sum_{k=0}^{N-1} e^{j\frac{2\pi k}{N}\left(\frac{g-1}{G}+m-1\right)} e^{jkq\sin\phi_g} \right|^2$$

$$= \Omega(\mathbf{G}_g^{(n)}, \phi_g)$$

$$\overset{(a)}{=} 0, \tag{10}$$

where (a) is obtained from (6). From (10), when matrix $\mathbf{R}_g$ is a diagonal matrix with the vector of diagonal entries $\mathbf{r}_g$ given as

$$\mathbf{r}_g = \left[1 \quad e^{-j2\pi\frac{d}{\lambda}\cos\phi_g\sin\alpha_g}\left(e^{-j2\pi\frac{d}{\lambda}\sin\phi_g}\right)^{\cos\alpha_g-1} \quad \cdots \right.$$

$$\left. e^{-j2\pi(N-1)\frac{d}{\lambda}\cos\phi_g\sin\alpha_g}\left(e^{-j2\pi(N-1)\frac{d}{\lambda}\sin\phi_g}\right)^{\cos\alpha_g-1} \right]^T, \tag{11}$$

the desired matrix $\mathbf{F}_g(\hat{\phi}) = \mathbf{R}_g\mathbf{G}_g^{(n)}$ constructs a transmit null in the target direction $\hat{\phi}$. Moreover, since each diagonal entry of $\mathbf{R}_g$ has an absolute value of 1 from (11), $\mathbf{R}_g$ is a unitary matrix (e.g., the desired matrix $\mathbf{F}_g(\hat{\phi})$) and satisfies orthogonal beamforming constraints.

Remark 2. The method in Theorem 1 ensures low complexity due to a simple matrix product. The linear transform matrix $\mathbf{R}_g$ matches the common null-point ϕ_g to DOV $\hat{\phi}$ while preserving orthogonality of the column vectors in the column-reduced matrix $\mathbf{G}_g^{(n)}$. Note that $\mathbf{R}_g$ preserves orthogonality, in contrast to conventional null-steering [4].

2-2 Multi-user Scheduling with Limited Feedback

Consider the TN-SDMA codebook $\mathcal{F}(\hat{\phi}) = \{\mathbf{F}_g(\hat{\phi})\}_{g=1,\cdots,G}$, where each precoding matrix consists of $N-1$ column vectors such as $\mathbf{F}_g = [\mathbf{f}_{g,1},\cdots,\mathbf{f}_{g,N-1}] \in \mathbb{C}^{N \times N-1}$. Each precoding matrix within a codebook leads to a different active receiver set that maximizes sum throughput by using the corresponding precoding matrix. This leads to a different sum throughput result. Therefore, to find optimal precoding matrix and its active receiver set, all precoding matrices within a codebook and their active receiver sets should be considered. This requires feedback from G sets of $N-1$ signal-to-interference-and-noise ratios (SINRs), $\{\gamma_{k,m}(\mathbf{F}_g)\}_{g=1,\cdots,G,m=1,\cdots,N-1}$, from each user. Although this SINR feedback makes it possible to find an optimal precoding matrix, the feedback is prohibitive to practical implementation. For limited feedback, each

receiver can send back only a set of $N-1$ SINRs, $\{\gamma_{k,m}(\mathbf{F}_{g_k})\}_{m=1,\cdots,N-1}$ by selecting a precoding matrix $\mathbf{F}_{g_k}$ within the codebook, where $g_k \in \{1,\cdots,G\}$.

Based on the assumption that a kth user has a perfect receiver CSI $\mathbf{h}_k$ and the precoder codebook, the kth user chooses a precoding matrix as follows:

$$\mathbf{F}_{g_k} = \arg\max_{\mathbf{F}_g \in \mathcal{F}(\hat{\phi})} \max_{m=\{1,\cdots,N-1\}} \gamma_{k,m}(\mathbf{F}_g) \tag{12}$$

where $\gamma_{k,m}(\mathbf{F}_g)$ is the SINR on the mth stream precoded by $\mathbf{F}_g$ of the kth user, which is given as

$$\gamma_{k,m} = \frac{\frac{P}{N-1}|\mathbf{h}_k\mathbf{f}_{g,m}|^2}{1 + \frac{P}{N-1}\sum_{j=1,j\neq m}^{N-1}|\mathbf{h}_k\mathbf{f}_{g,j}|^2}. \tag{13}$$

Since codebook $\mathcal{F}(\hat{\phi})$ is known *a priori* to both BS and the user, the index g_k of the selected precoding matrix is sent back to the BS. It requires feedback bits of $\lceil \log G \rceil$, where $\lceil x \rceil$ is the smallest integer that is larger than or equal to x. Additionally, the kth user feeds back to the BS $N-1$ SINRs $\{\gamma_{k,m}(\mathbf{F}_{g_k})\}_{m=1,\cdots,N-1}$.

According to their selected precoding matrix, all K users fall into G groups defined by

$$\mathcal{S}_g = \{1 \le k \le K \mid g_k = g\}, 1 \le g \le G. \tag{14}$$

Within each group, the BS attempts to assign $\mathbf{f}_{g,m}$ to a user $k_{g,m}^*$ with the highest SINR, given as $\gamma_{g,m}^* = \max_{k \in \mathcal{S}_g} \gamma_{k,m}(\mathbf{F}_{g_k})$. Finally, the g^* th precoding matrix $\mathbf{F}_{g^*}$ (the precoding matrix $\mathbf{N}$ in (2) is $\mathbf{F}_{g^*}$) used for transmission is determined to maximize instantaneous sum throughput, as follows:

$$g^* = \arg\max_{g=1,\cdots,G} \sum_{m=1}^{N-1} \log_2\left(1 + \gamma_{g,m}^*\right) \tag{15}$$

Thereby, the scheduled users, who are specified by indices $\{k_{g^*,m}^*\}_{m=1,\cdots,N-1}$, share the g^* th precoding matrix, which makes feedback information on SINRs from users valid, and thus enables the BS to exactly predict user SINRs. The average sum throughput is given by

$$T = \mathbb{E}\left[\sum_{m=1}^{N-1} \log_2\left(1 + \gamma_{g^*,m}^*\right)\right] \tag{16}$$

Ⅲ. Performance Evaluation

3-1 Sum Throughput

We compare the sum throughput of TN-SDMA and PU^{2}RC for SNR P, as well as the number of users K, precoding matrices G, and BS antennas N. The sum throughput of TN-SDMA and PU^{2}RC with codebook $\mathcal{E}$ given in (3) is numerically calculated based on (16). All results in Figs. 1, 2, and 3 show the throughput loss of TN-SDMA relative to PU^{2}RC. The reason for this is that, in a TN-SDMA scheme, a fraction of available spatial degrees of freedom is dedicated for interference mitigation, and the rest is used for spatial multiplexing, while PU^{2}RC exploits available spatial degrees of freedom for sending data streams only.

A larger codebook (e.g., larger $G(N-1)$) for TN-SDMA (or larger GN for PU^{2}RC) leads to higher channel quantization resolution, which increases the probability of selected weight vectors to a large channel gain, and eventually induces larger fluctuations in the SINR at each link. In contrast, a larger codebook reduces the effective diversity order of multi-user scheduling for each group S_g defined in (14). This trade-off yields a favorable impact by a larger G or N on the sum throughput for a relatively large number of users, but has a detrimental effect for a smaller number of users, as shown in Figs. 1 and 2. In Figs. 1 and 2, a decreasing number of users K reduces the throughput loss of TN-SDMA relative to PU^{2}RC for a fixed G and N. The reason for this is that, for a fixed codebook size, low K reduces multi-user diversity gain, and the throughput loss given from one less fewer data stream decreases. Fig. 2 shows that the throughput loss decreases with increasing N. This is because the ratio of the number of data streams in PU^{2}RC and TN-SDMA, $N/(N-1)$, decreases as increasing N. In Fig. 3, a higher average SNR P or higher
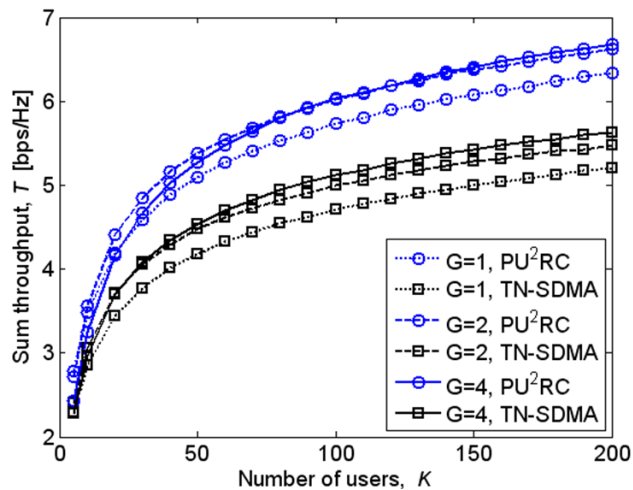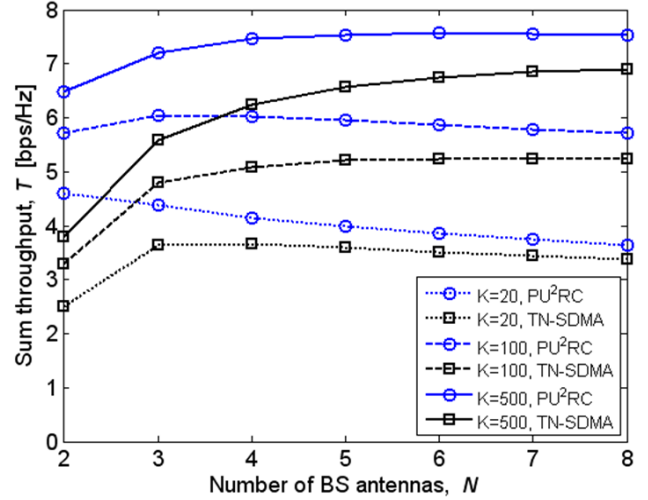

Fig. 2. Sum throughput versus the number of BS antennas for K=20, 100, 500, and SNR P=5 dB.
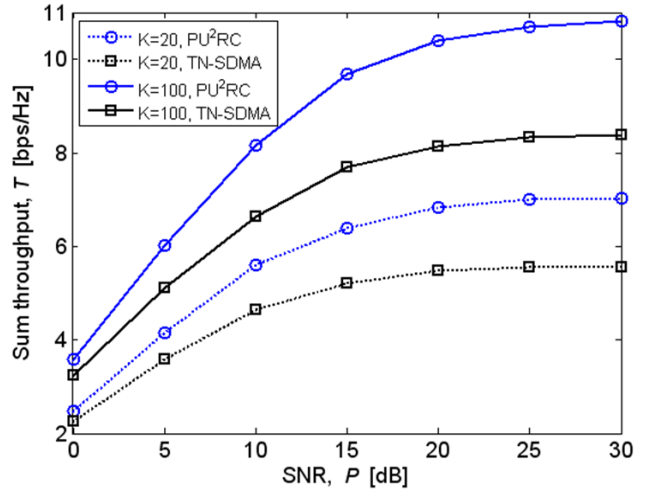

Fig. 3. Sum throughput versus SNR for G=4, N=4, and K=20, 100.

number of users K increases throughput loss. This can be explained by the fact that PU^{2}RC sends one more data stream than TN-SDMA, and throughput gain given from an additional data stream increases with increasing P and K. Note that the throughput loss eventually converges to an upper limit as P increases. The reason for this convergence is that, at high SNR, both of the user SINR of TN-SDMA and PU^{2}RC in (13) depend on instant channel gain $|\mathbf{h}_k \mathbf{f}_{g,m}|^2$, but not on SNR P, (e.g., (13) is modified as $\gamma_{k,m} \approx |\mathbf{h}_k \mathbf{f}_{g,m}|^2 \big/ \sum_{j=1, j \neq m}^{N-1} |\mathbf{h}_k \mathbf{f}_{g,j}|^2$ for high P).

The results indicate that the throughput loss of TN-SDMA relative to PU^{2}RC is smaller at larger N, lower P, or smaller K. In particular, a small loss (12 % through-

Fig. 1. Sum throughput T versus the number of users K for various numbers of precoding matrices G=1, 2, 4, and transmit antennas N=4, SNR P=5 dB.

put loss) is provided for practical system parameters of K=20, P=5 dB, and N=4. Note that, although TN-SDMA always has a lower throughput compared to PU^{2}RC in non-coexistence scenarios, it offers an opportunity for re-using a spectrum already allocated to a FWS, thereby providing a higher system capacity than PU^{2}RC does in a coexistence scenario. The spectrum sharing performance of TN-SDMA will be discussed in the remaining sections.

3-2 Interference Mitigation

In order to analyze the performance of interference mitigation using a TN-SDMA codebook, we must evaluate the protection distance for the IMT-Advanced BS with TN-SDMA (interfering system) and FWS (victim system) in a co-channel (or coexistence) scenario. An FSS earth station (ES), which is a typical FWS, was chosen to be a victim system for the evaluation. Frequency sharing between different radio systems is feasible when the interference power in a victim system is lower than the permissible interference power. A protection distance, which is a well-known measure for spectrum sharing, is defined as a minimum distance (between interfering and victim systems) satisfying the feasibility condition.

3-2-1 Evaluation Methodology

When the shielding effects of terrain and artificial objects are considered, the interference region decreases and is non-uniformly distributed, as shown in Fig. 4, due to the irregular distribution of terrain and objects.

■ IMT-Advanced BS
△ Evaluation region confined to ΔR and ΔA
▨ Entire interference region due to IMT-Advanced BS
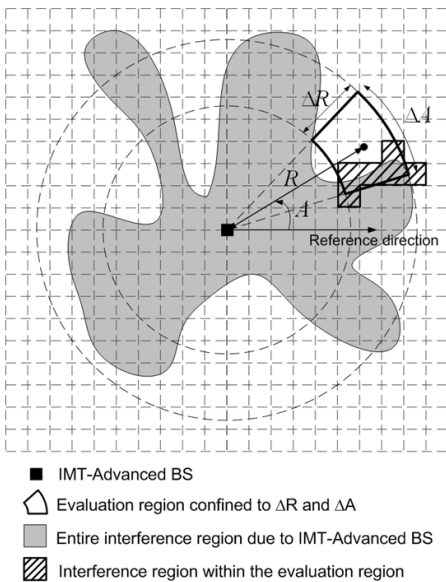▨ Interference region within the evaluation region

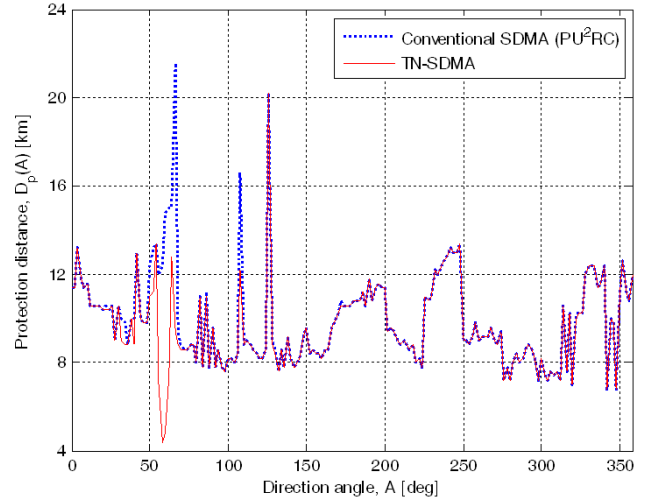Fig. 4. Protection distance evaluation in real radio environment.

Fig. 5. Protection distance comparison of TN-SDMA and conventional SDMA schemes in real radio environments (N=4, G=4).

Fig. 4 also shows protection distance evaluation in real radio environments. Interference power from an interfering system (IMT-Advanced BS) is evaluated using ray tracing [12] at the center point of each grid square. The sample points at which interference power exceeds the maximum permissible interference power comprise the entire interference region. The area ratio $\Upsilon(R,A)$ for the distance from the interfering system R and azimuth angle A is defined as $\Upsilon(R,A) = S_i/S_e$, where S_e denotes the area of the evaluation region confined to ΔR and ΔA. S_i represents the area of the interference region within the evaluation region. For a specific angle A, the protection distance $D_p(A)$ is defined as the minimum R such that $\Upsilon(R,A) < \Upsilon_{th}$, where Υ_{th} is the threshold of the area ratio.

3-2-2 Results

The system parameters are given in Table 1. We consider a squared rural area of 30×30 km. A single IMT-Advanced BS with 3 sectors is located at the center of the map, and constructs null at A=60°. Interference power from the BS with TN-SDMA is evaluated using ray tracing at the center point of each grid square of 100×100 m, (e.g., at 90,000 points). Fig. 5 compares the protection distances of TN-SDMA and conventional SDMA (without interference mitigation technique) for ΔA=1°, ΔR=200 m, and Υ_{th}=0.05. Due to the shielding effects by geographic objects, the protection distance varies by the different direction angle A. When TN-SDMA precoding is used, the protection distance is significantly reduced to 4.4 km (33 % of the protection distance of 13 km of conventional SDMA) around 60°,

Table 1. Simulation parameters.

System	Parameter	Value
FSS ES	Center frequency of operation (MHz)	3,500
	Bandwidth (MHz)	5
	Maximum antenna gain (dBi)	41
	Antenna height (m)	5
	Antenna type	Dish(diameter: 3.8 m)
	Antenna pattern	The ratio between the antenna diameter and the wavelength is less than or equal to 100 [10]
	Maximum permissible interference power for $I/N = -12.2$ dB [11] (dBW/5 MHz)	-153.8
IMT-Advanced BS with TN-SDMA	Center frequency of operation (MHz)	3,500
	Transmit power (dBW)	13
	Number of transmit antennas	4
	Antenna pattern	3-sector pattern [9]
	Antenna height (m)	15
	Bandwidth (MHz)	10
	Maximum antenna gain (dBi)	12

rather than other directions; the reduction of received power in unwanted directions is minimized. Therefore, TN-SDMA provides spectrum sharing with FSS ES at a minimal cost of IMT-Advanced spectrum resources.

3-2-3 Practicality

TN-SDMA tends to impose a restriction on system configuration in terms of the number of transmission antennas. This method requires the number of transmission antennas to be larger than the number of different DOVs at which FWS are located. However, FWS systems are not contiguous with each other. Therefore, TN-SDMA is a practical solution for spectrum sharing between IMT-Advanced and FWS. Some mobile terminals near FWS will cause interference to the FWS when considering the IMT-Advanced uplink. Since the DOV at a user is not constant due to user mobility, the user must continuously estimate the DOV; this is a heavy burden on the user. Additionally, when each user linked to a BS utilizes a different precoder forming a null in their own DOV, the BS must simultaneously employ different precoders. Therefore, using the non-overlapping spectrum with FWS is preferable to TN-SDMA for an uplink interference scenario.

Ⅳ. Conclusions

This paper has proposed a systematic design for the TN-SDMA precoder codebook, where the codebook forms a transmit null in the direction of victim systems and satisfies orthogonal beamforming constraints. The throughput loss of TN-SDMA relative to PU^{2}RC is smaller at larger N, lower P, or smaller K. Particularly, a small loss (12 % throughput loss) is provided for practical system parameters of $K=20$, $P=5$, and $N=4$. TN-SDMA precoding reduces protection distances by more than 66 % in comparison with conventional SDMA, without transmit-nulling, e.g. PU^{2}RC and ZF-SDMA. These results motivate TN-SDMA in IMT-Advanced systems coexisting with FWS.

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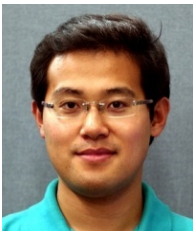
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