Channel Capacity of Distributed Antenna Network Using Transmit/Receive Diversity in a Multi-cell Environment

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Abstract— **Distributed antenna network (DAN) is a promising wireless network which can mitigate the impacts of distance-dependent/shadowing losses and fading. Frequency-domain space-time block coded-joint transmit/receive diversity (FD-STBC-JTRD) is an antenna diversity technique which allows an arbitrary number of transmit antennas while keeping the coding rate constant and is suitable for the downlink transmission. It obtains the frequency-diversity gain by exploiting the frequency-selective fading channel and improves the signal-to-interference plus noise power ratio (SINR) in a multi-cell environment. In this paper, we evaluate the channel capacity distribution of downlink single-carrier (SC-) DAN using FD-STBC-JTRD in a multi-cell environment. It is shown that multi-cell SC-DAN using FD-STBC-JTRD can allow the single-frequency reuse.**

Keywords; channel capacity, distributed antenna network, antenna diversity, transmit frequency-domain equalization, co-channel interference

I. INTRODUCTION

In the mobile communications, the received power significantly degrades due to path loss, shadowing loss, and fading. In the cellular networks, the signal-to-noise ratio (SNR) and signal-to-co-channel interference ratio (SIR) of the user located near the cell edge drop quite often [1]. Distributed antenna network (DAN) [2]-[4] is a promising network which can mitigate the impacts of distance-dependent/shadowing losses and fading. In DAN, many antennas connected with the signal processing center (SPC) by means of optical fiber cables or wireless links are spatially distributed so that some antennas can always be visible from a mobile terminal (MT) with a high probability. Frequency-domain space-time block coded-joint transmit/receive diversity (FD-STBC-JTRD) [5] is an antenna diversity technique combined with frequency-domain transmit equalization (FDE) [6]-[8]. It can allow an arbitrary number of transmit antennas while keeping the coding rate constant (the coding rate is determined by the number of receive antennas). Since the decoding at the MT does not require the channel state information (CSI), the complexity problem of receivers can be alleviated. Therefore, FD-STBC-JTRD is suitable for the DAN downlink application. It obtains the frequency-diversity gain by exploiting the frequency-selective fading channel and also improves the SNR and SIR (thus, the signal-to-interference plus noise power ratio (SINR)) in a multi-cell environment.

Due to its low peak-to-average power ratio (PAPR) property, the single-carrier (SC) waveform is attractive. Recently, the transmit FDE has been studied for the application to FD-STBC-JTRD [9,10]. In [9], joint water-filling and maximal ratio transmit (WF-MRT) FDE weight which maximizes the achievable capacity was derived assuming the single-cell environment. In a cellular DAN, conventional cell is replaced by a cell of DAN and the same frequency is reused in different DAN cells. Therefore, similar to present cellular system, the co-channel interference (CCI) limits the performance.

In this paper, we extend our study to the case of multi-cell environment and derive the transmit FDE for FD-STBC-JTRD by taking into account the CCI from surrounding DAN cells. We will evaluate the channel capacity distribution of multi-cell SC-DAN using FD-STBC-JTRD by Monte Carlo numerical computation method.

The remainder of this paper is organized as follows. In Sect. II, the SC-DAN downlink model is presented. Section III describes the principle of FD-STBC-JTRD. In Sect. IV, a new WF-MRT FDE that takes into account the CCI is derived by the Lagrange multiplier method [11]. In Sect. V, we discuss the channel capacity distribution. Section VI offers some conclusions.

II. SC-DAN DOWNLINK TRANSMISSION

A. SC-DAN downlink model

We consider a SC-DAN in which transmit antennas are distributed over a cell as shown in Fig. 1. In this paper, it is assumed that 7 antennas are distributed in each DAN cell. A single user having N_r antennas per cell is assumed. N_t transmit antennas are selected from these 7 antennas for the FD-STBC-JTRD downlink transmission.

Similar to the present cellular networks [1], the same frequency is reused in spatially separated DAN cells to efficiently utilize the limited frequency bandwidth. The available frequency bandwidth is divided into N sub-bands. N is called the cluster size. Figure 2 illustrates the CCI model of N=3. The center cell #0 is the cell of interest. There are 6 CCI cells in the first tier and 12 CCI cells in the second tier.

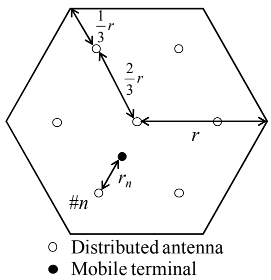

Figure 1. SC-DAN cell model.

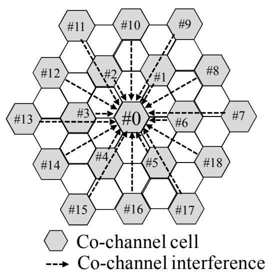

Figure 2. CCI model of N=3.

B. Channel model

The broadband channel is characterized by distant-dependent path loss, log-normally distributed shadowing loss and frequency-selective fading. The signal power $P_{\mathrm{rx},n}$ received at MT for the signal transmitted from the nth distributed antenna can be modeled as [1]

$$P_{\mathrm{rx},n} = P_{\mathrm{tx},n} \cdot r_n^{-\alpha} \cdot 10^{-\frac{\eta_n}{10}}, \tag{1}$$

where $P_{\mathrm{tx},n}$ is the transmit power, r_n is the distance between the nth distributed antenna and MT, α is the path loss exponent, and η_n is the shadowing loss in dB having zero-mean and standard deviation σ. By introducing the normalized distance $d_n=r_n/r$ and the normalized transmit power $\widetilde{P}_{\mathrm{tx},n} = P_{\mathrm{tx},n} \cdot r^{-\alpha}$ with r being the cell radius, Eq. (1) is rewritten as

$$P_{\mathrm{rx},n} = \widetilde{P}_{\mathrm{tx},n} \cdot \Omega_n, \tag{2}$$

where $\Omega_n = d_n^{-\alpha} \cdot 10^{-\frac{\eta_n}{10}}$.

Assuming that the frequency-selective channel is composed of L distinct paths, the channel impulse response $h_{m,n}(\tau)$ between the nth transmit antenna and the mth receive antenna is expressed as

$$h_{m,n}(\tau) = \sum_{l=0}^{L-1} h_{m,n}^{(l)} \delta(\tau - \tau_l), \tag{3}$$

where $h_{m,n}^{(l)}$ and τ_l are respectively the complex-valued path gain and the delay time of the lth path with $E[\sum_{l=0}^{L-1} |h_{m,n}^{(l)}|^2] = 1$. $\delta(\tau)$ is the delta function.

Antenna selection is an important issue. In this paper, N_t antennas having the largest square sum $\Omega_n \sum_{m=0}^{N_r-1} \sum_{l=0}^{L-1} |h_{m,n}^{(l)}|^2$ of the path gains are selected for FD-STBC-JTRD downlink transmission.

III. FD-STBC-JTRD ENCODING/DECODING

The downlink transmission model of SC-DAN using FD-STBC-JTRD is illustrated in Fig. 3. The data symbol sequence to be transmitted is divided into a sequence of J blocks of N_c data symbols each, where N_c denotes the size of fast Fourier transform (FFT). A sequence of J blocks is FD-STBC-JTRD encoded into N_t parallel sequences of Q blocks each (each block consists of N_c data symbols).

FD-STBC-JTRD can use an arbitrary number of transmit antennas while keeping the coding rate constant. On the other hand, as the number of receive antennas N_r increases, the coding rate R decreases [5]. A combination of J and Q is shown in Table 1.

Table 1. J, Q and R.

No. of receive antennas, N_r	No. of transmit symbol blocks, J	No. of coded blocks, Q	Coding rate, $R(=J/Q)$
1	1	1	1
2	2	2	1
3	3	4	3/4
4	3	4	3/4

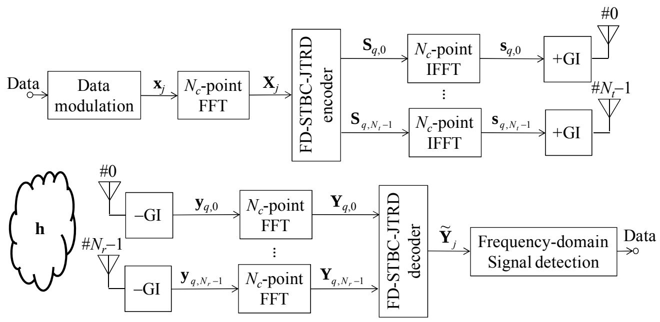

Figure 3. Downlink transmission model of SC-DAN using FD-STBC-JTRD.

A. FD-STBC-JTRD encoding

The transmit symbol block is represented in vector form as $\mathbf{x}_j=[x_j(0),\cdots, x_j(N_c-1)]^{\mathrm{T}}$, j=0~J−1. Before encoding, N_c-point FFT is applied to transform $\mathbf{x}_j$ into the frequency-domain signal vector $\mathbf{X}_j=[X_j(0),\cdots X_j(N_c-1)]^{\mathrm{T}}$ as

$$\mathbf{X}_j=\mathbf{F}\mathbf{x}_j, \tag{4}$$

where

$$\mathbf{F} = \frac{1}{\sqrt{N_c}} \begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{-j2\pi\frac{1\times 1}{N_c}} & \cdots & e^{-j2\pi\frac{1\times(N_c-1)}{N_c}} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & e^{-j2\pi\frac{(N_c-1)\times 1}{N_c}} & \cdots & e^{-j2\pi\frac{(N_c-1)\times(N_c-1)}{N_c}} \end{bmatrix} \tag{5}$$

is an $N_c \times N_c$ FFT matrix.

FD-STBC-JTRD encoding is carried out as follows. First, $\mathbf{X}_j$ is encoded into $\widetilde{\mathbf{X}}_{q,N_r}$, q=0~Q−1, as (for N_r>2, see [5])

$$\begin{aligned} &\widetilde{\mathbf{X}}_{0,1} = \mathbf{X}_0 && \text{for } N_r = 1, \\ &\widetilde{\mathbf{X}}_{0,2} = \begin{bmatrix} \mathbf{X}_0 \\ \mathbf{X}_1 \end{bmatrix}, \ \widetilde{\mathbf{X}}_{1,2} = \begin{bmatrix} -\mathbf{X}_1^* \\ \mathbf{X}_0^* \end{bmatrix} && \text{for } N_r = 2. \end{aligned} \tag{6}$$

Then, $\widetilde{\mathbf{X}}_{q,N_r}$ is multiplied by the transmit FDE weight matrix $\mathbf{W}$ to obtain the encoded signal vector $\mathbf{S}_{q,n}=[S_{q,n}(0), \cdots, S_{q,n}(N_c-1)]^{\mathrm{T}}$, q=0~Q−1 and n=0~N_t−1, which can be expressed as

$$\mathbf{S}_q = \begin{bmatrix} \mathbf{S}_{q,0} \\ \vdots \\ \mathbf{S}_{q,1} \\ \vdots \\ \mathbf{S}_{q,N_t-1} \end{bmatrix} = \beta_{N_r} \mathbf{W} \widetilde{\mathbf{X}}_{q,N_r}, \tag{7}$$

where β_{N_r} is the power normalization factor given by

$$\beta_{N_r}^2 = \frac{N_c}{\sum_{n=0}^{N_t-1} \sum_{m=0}^{N_r-1} \sum_{k=0}^{N_c-1} |W_{m,n}(k)|^2} \tag{8}$$

and $\mathbf{W}$ is the transmit weight matrix given as

$$\mathbf{W}=\begin{bmatrix}\mathbf{W}_{0,0} & \cdots & \mathbf{W}_{m,0} & \cdots & \mathbf{W}_{N_r-1,0}\\ \vdots & & \vdots & & \vdots\\ \mathbf{W}_{0,n} & \cdots & \mathbf{W}_{m,n} & \cdots & \mathbf{W}_{N_r-1,n}\\ \vdots & & \vdots & & \vdots\\ \mathbf{W}_{0,N_t-1} & \cdots & \mathbf{W}_{m,N_t-1} & \cdots & \mathbf{W}_{N_r-1,N_t-1}\end{bmatrix} \quad (9)$$

with $\mathbf{W}_{m,n}=diag[W_{m,n}(0),\cdots,W_{m,n}(N_c-1)]$, $m=0\sim N_r-1$ and $n=0\sim N_t-1$, which will be derived in Sect. IV.

$\mathbf{S}_{q,n}$ is transformed by N_c-point inverse FFT (IFFT) into the time-domain signal block $\mathbf{s}_{q,n}=[s_{q,n}(0),\cdots,s_{q,n}(N_c-1)]^{\mathrm{T}}$ as

$$\mathbf{s}_{q,n}=\sqrt{\frac{2E_s}{T_s}}\mathbf{F}^{\mathrm{H}}\mathbf{S}_{q,n}, \quad (10)$$

where $E_s=P_{\mathrm{tx}}\cdot T_s$ is the normalized total transmit symbol energy with $P_{\mathrm{tx}}=\sum_{n=0}^{N_t-1}\widetilde{P}_{\mathrm{tx},n}$ being the normalized total transmit power and T_s being the symbol duration. $(.)^{\mathrm{H}}$ is the Hermitian transpose operation. Before transmission, the last N_g symbols in each signal block are copied as a cyclic prefix (CP) and inserted into the guard interval (GI) placed at the beginning of each block.

B. FD-STBC-JTRD decoding

The GI-removed received signal block $\mathbf{y}_q$ can be expressed using the matrix form as

$$\mathbf{y}_q=\begin{bmatrix}\mathbf{y}_{q,0}\\ \vdots\\ \mathbf{y}_{q,m}\\ \vdots\\ \mathbf{y}_{q,N_r-1}\end{bmatrix}=\sqrt{\frac{2E_s}{T_s}}\mathbf{h}\mathbf{s}_q+\sqrt{\frac{2E_s}{T_s}}\sum_{i=1}^{\infty}\mathbf{h}^{(i)}\mathbf{s}_q^{(i)}+\mathbf{n}_q, \quad (11)$$

where $\mathbf{y}_{q,m}=[y_{q,m}(0),\cdots,y_{q,m}(N_c-1)]^{\mathrm{T}}$, $m=0\sim N_r-1$. The first, second, and third terms of right hand side are the desired signal, CCI, and noise, respectively. $\mathbf{h}$ is the $N_rN_c\times N_tN_c$ channel impulse response matrix given as

$$\mathbf{h}=\begin{bmatrix}\mathbf{h}_{0,0} & \cdots & \mathbf{h}_{0,n} & \cdots & \mathbf{h}_{0,N_t-1}\\ \vdots & & \vdots & & \vdots\\ \mathbf{h}_{m,0} & \cdots & \mathbf{h}_{m,n} & \cdots & \mathbf{h}_{m,N_t-1}\\ \vdots & & \vdots & & \vdots\\ \mathbf{h}_{N_r-1,0} & \cdots & \mathbf{h}_{N_r-1,n} & \cdots & \mathbf{h}_{N_r-1,N_t-1}\end{bmatrix} \quad (12)$$

with

$$\mathbf{h}_{m,n}=\sqrt{\Omega_n}\begin{bmatrix}h_{m,n}^{(0)} & & & h_{m,n}^{(L-1)} & \cdots & h_{m,n}^{(1)}\\ h_{m,n}^{(1)} & \ddots & & & \ddots & \vdots\\ \vdots & \ddots & h_{m,n}^{(0)} & & \mathbf{0} & h_{m,n}^{(L-1)}\\ h_{m,n}^{(L-1)} & & h_{m,n}^{(1)} & h_{m,n}^{(0)} & & \\ & \ddots & \vdots & h_{m,n}^{(1)} & \ddots & \\ \mathbf{0} & & h_{m,n}^{(L-1)} & \vdots & \ddots & h_{m,n}^{(0)}\end{bmatrix}. \quad (13)$$

$\mathbf{h}^{(i)}$ is the channel impulse response matrix between the transmit antennas in the ith co-channel cell and the MT receive antennas in the cell of interest. $\mathbf{n}_q=[\mathbf{n}_{q,0},\cdots,\mathbf{n}_{q,N_r-1}]^{\mathrm{T}}$ is the noise vector, where $\mathbf{n}_{q,m}=[n_{q,m}(0),\cdots,n_{q,m}(N_c-1)]^{\mathrm{T}}$ is the zero-mean complex-valued random variable having variance $2N_0/T_s$ with N_0 being the one-sided power spectrum density of additive white Gaussian noise (AWGN).

The received signal block $\mathbf{y}_q$ is transformed by N_c-point FFT into the frequency-domain signal $\mathbf{Y}_q$ as

$$\begin{aligned}\mathbf{Y}_q&=\begin{bmatrix}\mathbf{Y}_{q,0}\\ \vdots\\ \mathbf{Y}_{q,m}\\ \vdots\\ \mathbf{Y}_{q,N_r-1}\end{bmatrix}=\begin{bmatrix}\mathbf{F} & & \mathbf{0}\\ & \ddots & \\ \mathbf{0} & & \mathbf{F}\end{bmatrix}\mathbf{y}_q\\ &=\sqrt{\frac{2E_s}{T_s}}\beta_{N_r}\begin{bmatrix}\mathbf{F} & & \mathbf{0}\\ & \ddots & \\ \mathbf{0} & & \mathbf{F}\end{bmatrix}\mathbf{h}\begin{bmatrix}\mathbf{F}^{\mathrm{H}} & & \mathbf{0}\\ & \ddots & \\ \mathbf{0} & & \mathbf{F}^{\mathrm{H}}\end{bmatrix}\mathbf{W}\widetilde{\mathbf{X}}_{q,N_r}\\ &+\sqrt{\frac{2E_s}{T_s}}\sum_{i=1}^{\infty}\beta_{N_r}^{(i)}\begin{bmatrix}\mathbf{F} & & \mathbf{0}\\ & \ddots & \\ \mathbf{0} & & \mathbf{F}\end{bmatrix}\mathbf{h}^{(i)}\begin{bmatrix}\mathbf{F}^{\mathrm{H}} & & \mathbf{0}\\ & \ddots & \\ \mathbf{0} & & \mathbf{F}^{\mathrm{H}}\end{bmatrix}\mathbf{W}^{(i)}\widetilde{\mathbf{X}}_{q,N_r}^{(i)}\\ &+\begin{bmatrix}\mathbf{N}_{q,0}\\ \vdots\\ \mathbf{N}_{q,m}\\ \vdots\\ \mathbf{N}_{q,N_r-1}\end{bmatrix}\end{aligned}, \quad (14)$$

where $\mathbf{Y}_{q,m}=[Y_{q,m}(0),\cdots,Y_{q,m}(N_c-1)]^{\mathrm{T}}$ and $\mathbf{N}_{q,m}=\mathbf{F}\mathbf{n}_{q,m}=[N_{q,m}(0),\cdots,N_{q,m}(N_c-1)]^{\mathrm{T}}$ are the frequency-domain received signal and the noise vector of the mth receive antenna, respectively. Since $\mathbf{h}_{m,n}$ is a circulant matrix, the eigenvalue decomposition using $\mathbf{F}$ can be applied and we obtain

$$\begin{aligned}\mathbf{F}\mathbf{h}_{m,n}\mathbf{F}^{\mathrm{H}}&=diag\left[H_{m,n}(0),\cdots,H_{m,n}(N_c-1)\right]\\ &\equiv\mathbf{H}_{m,n}\end{aligned}, \quad (15)$$

where $H_{m,n}(k)=\sqrt{\Omega_n}\sum_{l=0}^{L-1}h_{m,n}^{(l)}\exp(-j2\pi k\tau_l/N_c)$. By introducing

$$\mathbf{H}=\begin{bmatrix}\mathbf{H}_{0,0} & \cdots & \mathbf{H}_{0,N_t-1}\\ \vdots & \ddots & \vdots\\ \mathbf{H}_{N_r-1,0} & \cdots & \mathbf{H}_{N_r-1,N_t-1}\end{bmatrix}, \quad (16)$$

Eq. (14) can be rewritten as

$$\begin{aligned}\mathbf{Y}_q&=\sqrt{\frac{2E_s}{T_s}}\beta_{N_r}\mathbf{H}\mathbf{W}\widetilde{\mathbf{X}}_{q,N_r}\\ &+\sqrt{\frac{2E_s}{T_s}}\sum_{i=1}^{\infty}\beta_{N_r}^{(i)}\mathbf{H}^{(i)}\mathbf{W}^{(i)}\widetilde{\mathbf{X}}_{q,N_r}^{(i)}+\mathbf{N}_q\end{aligned}. \quad (17)$$

FD-STBC-JTRD decoding is done on $\mathbf{Y}_q$ to obtain $\widetilde{\mathbf{Y}}_j=[Y_j(0),\cdots,Y_j(N_c-1)]^{\mathrm{T}}$ as (for $N_t>2$, see [5])

$$\begin{aligned} &\widetilde{\mathbf{Y}}_0 = \mathbf{Y}_{0,0} && \text{for } N_r = 1, \\ &\begin{cases} \widetilde{\mathbf{Y}}_0 = \mathbf{Y}_{0,0} + \mathbf{Y}^*_{1,1} \\ \widetilde{\mathbf{Y}}_1 = \mathbf{Y}_{0,1} - \mathbf{Y}^*_{1,0} \end{cases} && \text{for } N_r = 2. \end{aligned} \tag{18}$$

Letting $W_{m,n}(k) = H^*_{m,n}(k)\cdot\widetilde{W}(k)$, $\widetilde{\mathbf{Y}}_j$ can be written as

$$\begin{aligned} \widetilde{\mathbf{Y}}_j = &\sqrt{\frac{2E_s}{T_s}}\beta_{N_r}\left(\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}\mathbf{H}_{m,n}\mathbf{W}_{m,n}\right)\mathbf{F}\mathbf{x}_j \\ &+\sqrt{\frac{2E_s}{T_s}}\sum_{i=1}^{\infty}\beta^{(i)}_{N_r}\left(\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}\sum_{m'=0}^{N_r-1}\mathbf{H}^{(i)}_{m,n}\mathbf{W}^{(i)}_{m',n}\right)\mathbf{Z}^{(i)}_j + \widetilde{\mathbf{N}}_j \end{aligned}, \tag{19}$$

where $\mathbf{Z}^{(i)}_j$ is the CCI vector from the ith cell and $\widetilde{\mathbf{N}}_j$ is the frequency-domain zero-mean noise vector having variance $2N_rN_0/T_s$.

IV. Optimal Transmit Weight

Approximating the sum of CCI and Gaussian noise as a new Gaussian noise, SINR $\gamma(k)$ of the kth frequency component after FD-STBC-JTRD decoding can be given as

$$\gamma(k) = \frac{\frac{E_s}{N_0}\beta^2_{N_r}\left|\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}H_{m,n}(k)W_{m,n}(k)\right|^2}{\lambda(k)+N_r}, \tag{20}$$

where

$$\lambda(k) = \frac{E_s}{N_0}\sum_{i=1}^{\infty}\left(\beta^{(i)}_{N_r}\right)^2\left|\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}\sum_{m'=0}^{N_r-1}H^{(i)}_{m,n}(k)W^{(i)}_{m',n}(k)\right|^2 \tag{21}$$

is the normalized CCI power. The channel capacity of SC-DAN downlink using FD-STBC-JTRD is given as

$$C = \frac{R}{N}\cdot\frac{1}{N_c}\sum_{k=0}^{N_c-1}\log_2\left(1+\gamma(k)\right) \ \text{(bps/Hz/cell)}. \tag{22}$$

We want to derive the transmit FDE weight $W_{m,n}(k)$ that maximizes the channel capacity under the total transmit power constraint. The maximization problem can be written as

$$\begin{aligned} &\max_{\{W_{m,n}(k)\}} C \\ &\text{s.t. } \sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}\sum_{k=0}^{N_c-1}\left|W_{m,n}(k)\right|^2 = N_c \ \left(\text{i.e., } \beta^2_{N_r}=1\right) \end{aligned}. \tag{23}$$

However, the above maximization problem is quite difficult to solve. We use the upper-bound of Eq. (22). Using the Cauchy-Schwarz inequality [12], Eq. (22) can be upper-bounded as

$$\begin{aligned} C \le &\frac{R}{N\cdot N_c}\times \\ &\sum_{k=0}^{N_c-1}\log_2\left(1+\frac{E_s/N_0}{\lambda(k)+N_r}\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}\left|H_{m,n}(k)\right|^2\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}\left|W_{m,n}(k)\right|^2\right) \end{aligned}. \tag{24}$$

In Eq. (24), the equality holds if and only if

$$\frac{W_{0,0}(k)}{H^*_{0,0}(k)} = \cdots = \frac{W_{m,n}(k)}{H^*_{m,n}(k)} = \cdots = \frac{W_{N_r-1,N_t-1}(k)}{H^*_{N_r-1,N_t-1}(k)}. \tag{25}$$

The maximization problem can be converted into a concave optimization problem under the total transmit power constraint [11]. Following to [13], the solution is the WF-MRT FDE weight and is given as (for the sake of brevity, the derivation is omitted)

$$\begin{aligned} W_{m,n}(k) = &\frac{H^*_{m,n}(k)}{\sqrt{\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}\left|H_{m,n}(k)\right|^2}}\times \\ &\sqrt{\max\left\{\varphi - \frac{\lambda(k)+N_r}{\frac{E_s}{N_0}\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}\left|H_{m,n}(k)\right|^2}, 0\right\}} \end{aligned}, \tag{26}$$

where φ is chosen so as to satisfy $\beta^2_{N_r}=1$. The optimal weight is a product of two terms: one across transmit/receive antennas and the other across the frequencies. It can be understood from Eq. (26) that the power allocation is done across frequencies based on the water-filling theory, and across transmit antennas based on MRT.

The WF-MRT FDE weight in a single-cell environment was derived in [9] as

$$\begin{aligned} W_{m,n}(k) = &\frac{H^*_{m,n}(k)}{\sqrt{\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}\left|H_{m,n}(k)\right|^2}}\times \\ &\sqrt{\max\left\{\varphi - \frac{N_r}{\frac{E_s}{N_0}\sum_{n=0}^{N_t-1}\sum_{m=0}^{N_r-1}\left|H_{m,n}(k)\right|^2}, 0\right\}} \end{aligned}. \tag{27}$$

Comparison between Eq. (26) and Eq. (27) shows that the WF power allocation is done to maximize the SNR in a single-cell case while to maximize the SINR in a multi-cell case.

V. Numerical Evaluation

A. Numerical evaluation condition

The channel capacity distribution is evaluated by Monte Carlo numerical computation method assuming equal total transmit power for all DAN cells. A single user having N_r antennas is located randomly in each cell. The CCI from 18 co-channel cells in the first and second tiers is considered (see Fig. 2). The numerical computation condition is summarized in Table 2. The channel is assumed to be a frequency-selective block Rayleigh fading channel having a symbol-spaced L=16-path uniform power delay profile (i.e., $E[|h^{(l)}_{m,n}|^2]=(1/L)$). Ideal channel estimation is assumed.

Table 2. Numerical evaluation condition.

Fading type	Block Rayleigh fading
Power delay profile	Uniform
No. of paths	L=16
Time delay	$\tau_l=l$, $l=0\sim L-1$
Path loss exponent	α=3.5
Shadowing loss standard deviation	σ=7.0 (dB)
FFT size	N_c=256
Channel estimation	Ideal

B. *Capacity distribution*

Figure 4 plots the 5%-outage capacity (below which the channel capacity falls with 5% probability) as a function of N_t, with N_r as a parameter for N=1 (single frequency reuse) when the normalized transmit E_s/N_0=10dB. It can be seen from Fig. 4 that the channel capacity increases with N_t, but it becomes almost saturated beyond N_t=3~4. As for N_r, N_r=2 is found to maximize the capacity. This is a consequence of trade-off between the diversity gain and the coding rate. Increasing N_r increases the diversity gain, but the coding rate reduces to 3/4 when N_r increases to 3 (see Table 1). As a result, (N_t,N_r)=(4,2) is an optimal choice.

Figure 5 plots the 5%-outage capacity as a function of the cluster size N with the normalized transmit E_s/N_0 as a parameter when (N_t,N_r)=(4,2). For comparison, the 5%-outage capacities using MRT weight [14] and SNR based WF-MRT weight of Eq. (27) are also plotted. It can be seen from Fig. 5 that the 5%-outage capacity is maximized when N=1. When N=1 (strong CCI is present), the SINR based WF-MRT weight of Eq. (26) provides the highest capacity. According to the numerical results, it is plausible that the same frequency can be reused even in the same cell. A more detailed investigation is an interesting future study.

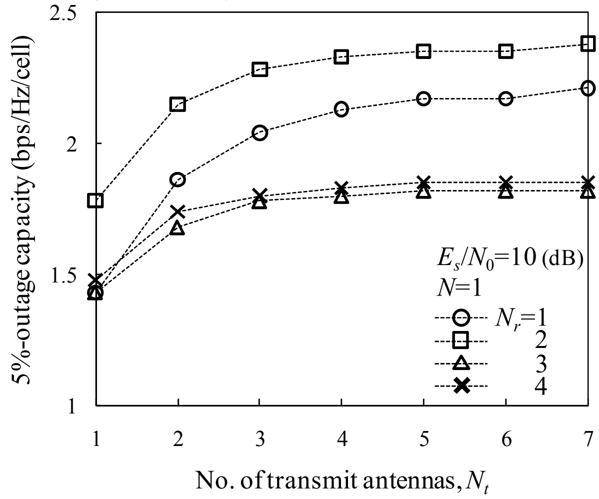

Figure 4. Impacts of N_t and N_r.

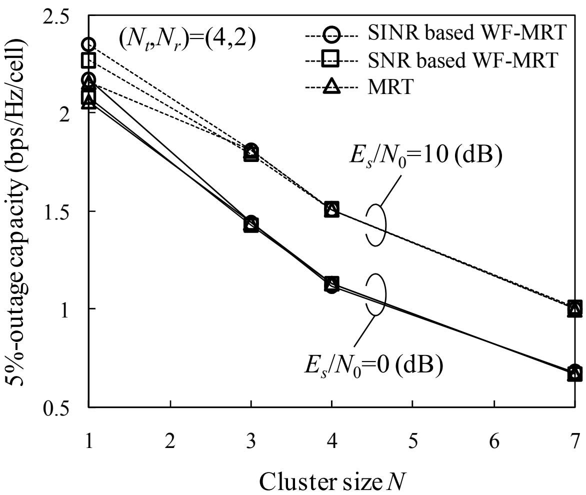

Figure 5. Impact of N.

VI. CONCLUSION

In this paper, we investigated the channel capacity distribution of multi-cell SC-DAN using FD-STBC-JTRD. It was shown that SC-DAN using FD-STBC-JTRD can allow the single frequency reuse. It is plausible that the same frequency is reused even in the same cell. A more detailed investigation of the transmission performance is left as an interesting future study.

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