Channel Capacity of Distributed Antenna Network Using Space-Time Block Coded-Joint Transmit/Receive Diversity

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***Abstract*—Distributed antenna network (DAN) is a promising wireless network to solve the problems arising from shadowing and path lesses as well as frequency-selective fading. Many anntennas are spatially distributed around each base station (BS) so that with a high probability, some antennas can always be visible from a mobile station (MS). Recently, we proposed a 2-dimensional water-filling (2D-WF) transmit diversity for single-carrier (SC) DAN downlink transmission. An MS having single receive antenna was considered. In this paper, we extend the 2D-WF transmit diversity to the case of MS having multiple receive antennas to implement frequency-domain space-time block coded-joint transmit/receive diversity (FD-STBC-JTRD). The channel capacity distribution is evaluated by Monte-Carlo numerical computation method. It is shown that the use of 2 receive antennas maximize the downlink channel capacity while the use of around 5 distributed transmit antennas is sufficient.**

Keywords-component; Distributed antenna network, space-time block coded-joint transmit/receive diversity, transmit weight, channel capacity

I. INTRODUCTION

In broadband wireless systems, the received signal is suffered from frequency-selective fading in addition to the shadowing and path losses [1]. The distributed antenna network (DAN) or the distributed antenna system (DAS) [2]-[6] is a promising wireless network to mitigate the negative impacts of fading, shadowing loss, and path loss. In DAN, many antennas connected by optical fiber cables with a base station (BS) are spatially distributed so that with a high probability, some antennas can always be visible from a mobile station (MS). There are two ways to utilize DAN: transmit/receive diversity [6]-[11] and spatial multiplexing [12]-[14]. In this study, we consider the single-carrier (SC) DAN downlink transmit/receive diversity.

Recently, we proposed a 2-dimensional water-filling (2D-WF) transmit diversity for SC DAN downlink transmission [6]. The proposed 2D-WF transmit diversity allocates the transmit power in both transmit antenna dimension and frequency dimension: power allocation across frequencies based on water filling theory [15] and across transmit antennas based on maximal ratio transmission (MRT) [10]. It was shown that the 2D-WF transmit diversity can achieve higher channel capacity than MRT. However, the proposed 2D-WF transmit diversity has only assumed an MS having single receive antenna. Receive antenna diversity has not been considered yet.

In this paper, we extend the proposed 2D-WF transmit diversity to the case of MS having multiple receive antennas to implement the frequency-domain space-time block coded-joint transmit/receive diversity (FD-STBC-JTRD) [8]. FD-STBC-JTRD can use an arbitrary number of transmit antennas while the number of receive antennas is limited. The FD-STBC-JTRD requires the channel state information (CSI) at the transmitter side only, hence the complexity problem of mobile terminals can be alleviated. In this paper, we derive the 2D-WF transmit weight for SC DAN using FD-STBC-JTRD. The channel capacity distribution is evaluated by Monte-Carlo numerical computation method.

The remainder of this paper is organized as follows. In Sect. II, we present the downlink transmit/receive diversity model using FD-STBC-JTRD. In Sect. III, we develop the downlink channel capacity expression. By using the Lagrange multiplier method, we derive the 2D-WF transmit diversity weight. In Sect. IV, we evaluate, by Monte Carlo numerical computation method, the distribution of the downlink channel capacity.

II. SC-DAN DOWNLINK TRANSMISSION

A. SC-DAN model

We consider an SC-DAN in which transmit antennas are uniformly distributed over a service area as shown in Fig. 1. In this paper, single user case is considered. Distributed antennas are ranked in the ascending order of the path loss plus shadowing loss and N_t antennas which have the smallest path loss plus shadowing loss are selected for downlink FD-STBC-JTRD transmissions. At the MS receiver, N_r receive antennas are used.

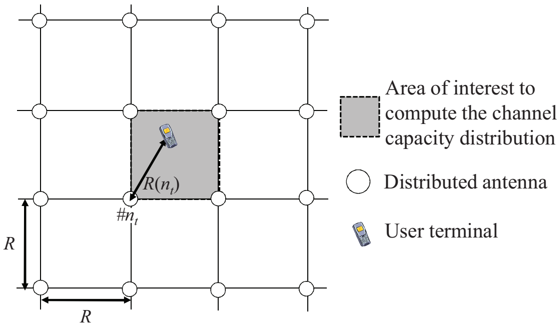

Figure 1. SC-DAN model.

B. FD-STBC-JTRD encoding/decoding

Downlink FD-STBC-JTRD transmission system model using N_t transmit and N_r receive antennas is illustrated in Fig. 2.

A sequence of J data blocks $\{d_j(t);\ t=0\sim N_c-1\}$, $j=0\sim J-1$, to be transmitted are encoded using complex-valued transmit weights into N_t parallel sequences of Q blocks each, where N_c denotes the size of fast Fourier transform (FFT). A combination of J and Q is shown in Table 1 for $N_r=1\sim4$. As shown in [7,8], FD-STBC-JTRD achieves an $N_t\times N_r$-order maximal ratio combining (MRC) diversity. The FD-STBC-JTRD encoding requires the CSI between each transmit/receive antenna pair. However, the FD-STBC-JTRD decoding requires no CSI since only addition, subtraction, and conjugate operations are necessary.

C. Encoding

The data symbol block is represented in vector form as $\mathbf{d}_j=[d_j(0),\cdots,d_j(t),\cdots,d_j(N_c-1)]^T$, $j=0\sim J-1$. Before STBC-JTRD encoding, N_c-point FFT is applied to transform $\mathbf{d}_j$ into the frequency-domain signal vector $\mathbf{D}_j=[D_j(0),\cdots,D_j(k),\cdots,D_j(N_c-1)]^T$ as

$$\mathbf{D}_j=\mathbf{F}\mathbf{d}_j, \qquad (1)$$

where

$$\mathbf{F}=\frac{1}{\sqrt{N_c}}\begin{bmatrix}1 & 1 & \cdots & 1\\ 1 & e^{-j2\pi\frac{1\times1}{N_c}} & \cdots & e^{-j2\pi\frac{1\times(N_c-1)}{N_c}}\\ \vdots & \vdots & \ddots & \vdots\\ 1 & e^{-j2\pi\frac{(N_c-1)\times1}{N_c}} & \cdots & e^{-j2\pi\frac{(N_c-1)\times(N_c-1)}{N_c}}\end{bmatrix} \qquad (2)$$

is an $N_c\times N_c$ FFT matrix.

FD-STBC-JTRD encoding [8] is carried out as follows. First, $\mathbf{D}_j$ is encoded into $\tilde{\mathbf{D}}_q(N_r)$, $q=0\sim Q-1$, as

$$\begin{aligned}
&\tilde{\mathbf{D}}_0(1)=\mathbf{D}_0 && \text{for } N_r=1\\
&\tilde{\mathbf{D}}_0(2)=\begin{bmatrix}\mathbf{D}_0\\ \mathbf{D}_1\end{bmatrix},\ \tilde{\mathbf{D}}_1(2)=\begin{bmatrix}-\mathbf{D}_1^*\\ \mathbf{D}_0^*\end{bmatrix} && \text{for } N_r=2\\
&\begin{cases}\tilde{\mathbf{D}}_0(3)=\begin{bmatrix}\mathbf{D}_0\\ \mathbf{D}_1\\ \mathbf{D}_2\end{bmatrix},\ \tilde{\mathbf{D}}_1(3)=\begin{bmatrix}-\mathbf{D}_1^*\\ \mathbf{D}_0^*\\ \mathbf{0}\end{bmatrix}\\ \tilde{\mathbf{D}}_2(3)=\begin{bmatrix}-\mathbf{D}_2^*\\ \mathbf{0}\\ \mathbf{D}_0^*\end{bmatrix},\ \tilde{\mathbf{D}}_3(3)=\begin{bmatrix}\mathbf{0}\\ -\mathbf{D}_2^*\\ \mathbf{D}_1^*\end{bmatrix}\end{cases} && \text{for } N_r=3\\
&\begin{cases}\tilde{\mathbf{D}}_0(4)=\begin{bmatrix}\mathbf{D}_0\\ \mathbf{D}_1\\ \mathbf{D}_2\\ \mathbf{0}\end{bmatrix},\ \tilde{\mathbf{D}}_1(4)=\begin{bmatrix}-\mathbf{D}_1^*\\ \mathbf{D}_0^*\\ \mathbf{0}\\ \mathbf{D}_2\end{bmatrix}\\ \tilde{\mathbf{D}}_2(4)=\begin{bmatrix}-\mathbf{D}_2^*\\ \mathbf{0}\\ \mathbf{D}_0^*\\ -\mathbf{D}_1\end{bmatrix},\ \tilde{\mathbf{D}}_3(4)=\begin{bmatrix}\mathbf{0}\\ -\mathbf{D}_2^*\\ \mathbf{D}_1^*\\ \mathbf{D}_0\end{bmatrix}\end{cases} && \text{for } N_r=4
\end{aligned} \qquad (3)$$

Then, the 2D-WF transmit weight matrix $\mathbf{W}$ is multiplied to $\tilde{\mathbf{D}}_q(N_r)$ to obtain $\mathbf{S}_q(n_t)=[S_q(n_t,0),\cdots,S_q(n_t,k),\cdots,S_q(n_t,N_c-1)]^T$, $q=0\sim Q$-1 and $n_t=0\sim N_t$-1, for STBC-JTRD. As a result, the FD-STBC-JTRD encoding can be expressed as

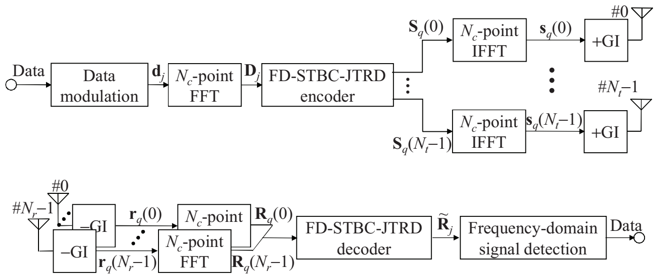

Figure 2. Downlink FD-STBC-JTRD transmission system model.

Table 1 J, Q and R for N_r=1~4.

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

$$\mathbf{S}_q=\begin{bmatrix}\mathbf{S}_q(0)\\ \vdots\\ \mathbf{S}_q(n_t)\\ \vdots\\ \mathbf{S}_q(N_t-1)\end{bmatrix}=C(N_r)\mathbf{W}\tilde{\mathbf{D}}_q(N_r), \qquad (4)$$

where $C(N_r)$ is the power normalization factor given by

$$C^2(N_r)=\frac{N_c}{\sum_{n_t=0}^{N_t-1}\sum_{n_r=0}^{N_r-1}\sum_{k=0}^{N_c-1}|W(n_t,n_r,k)|^2} \qquad (5)$$

and $\mathbf{W}$ is the 2D-WF transmit weight matrix expressed as

$$\mathbf{W}=\begin{bmatrix}\mathbf{W}(0,0) & \cdots & \mathbf{W}(0,n_r) & \cdots & \mathbf{W}(0,N_r-1)\\ \vdots & & \vdots & & \vdots\\ \mathbf{W}(n_t,0) & \cdots & \mathbf{W}(n_t,n_r) & \cdots & \mathbf{W}(n_t,N_r-1)\\ \vdots & & \vdots & & \vdots\\ \mathbf{W}(N_t-1,0) & \cdots & \mathbf{W}(N_t-1,n_r) & \cdots & \mathbf{W}(N_t-1,N_r-1)\end{bmatrix} \qquad (6)$$

with $\mathbf{W}(n_t,n_r)=diag[W(n_t,n_r,0),\cdots,W(n_t,n_r,k),\cdots,W(n_t,n_r,N_c-1)]$, which will be derived in Sect.III.

$\mathbf{S}_q(n_t)$ is transformed by N_c-point inverse FFT (IFFT) into the time-domain signal block $\mathbf{s}_q(n_t)=[s_q(n_t,0),\cdots,s_q(n_t,t),\cdots,s_q(n_t,N_c-1)]^T$ as

$$\mathbf{s}_q(n_t)=\mathbf{F}^H\mathbf{S}_q(n_t), \qquad (7)$$

where $(.)^H$ is the Hermitian transpose operation. Before transmission, the last N_g samples 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.

D. Channel model

The broadband channel is characterized by the distant-dependent path loss, log-normally distributed shadowing loss, and frequency-selective fading. The received power $P_r(n_t)$ for an MS whose distance from the n_tth distributed antenna is $R(n_t)$ can be modeled as [1]

$$P_r(n_t)=\left(P_t(n_t)\cdot R^{-\alpha}\right)\cdot d(n_t)^{-\alpha}\cdot 10^{-\frac{\eta(n_t)}{10}}, \quad (8)$$

where $P_t(n_t)$ is the transmit power of the n_tth distributed antenna, α is the path loss exponent, and $\eta(n_t)$ is the shadowing loss in dB having zero-mean and standard variation σ. $d(n_t)=R(n_t)/R$ is the distance normalized by the distributed antenna spacing R. Letting $\widetilde{P}_t(n_t)=P_t(n_t)\cdot R^{-\alpha}$ and $\Omega(n_t)=d(n_t)^{-\alpha}\cdot 10^{-\eta(n_t)/10}$, Eq. (8) is rewritten as

$$P_r(n_t)=\widetilde{P}_t(n_t)\cdot\Omega(n_t). \quad (9)$$

Assuming that the frequency-selective channel is composed of L distinct paths, the channel impulse response $h(n_t,n_r,\tau)$ between the n_tth distributed antenna and the n_rth receive antenna of MS is expressed as

$$h(n_t,n_r,\tau)=\sum_{l=0}^{L-1}h_l(n_t,n_r)\delta(\tau-\tau_l), \quad (10)$$

where $h_l(n_t,n_r)$ and τ_l are respectively the complex-valued path gain and the time delay of the lth path with $E[\sum_{l=0}^{L-1}|h_l(n_t,n_r)|^2]=\Omega(n_t)$.

E. Decoding

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

$$\mathbf{r}_q=\begin{bmatrix}\mathbf{r}_q(0)\\ \vdots\\ \mathbf{r}_q(n_r)\\ \vdots\\ \mathbf{r}_q(N_r-1)\end{bmatrix}=\sqrt{\frac{2E_s}{T_s}}\mathbf{h}\mathbf{s}_q+\mathbf{n}_q, \quad (11)$$

where $\mathbf{r}_q(n_r)=[r_q(n_r,0),\cdots,r_q(n_r,t),\cdots,r_q(n_r,N_c-1)]^T$, E_s is the normalized total average transmit symbol energy with $E_s=P_tT_s$, P_t is the total average transmit power, and T_s is the symbol duration. $\mathbf{h}$ is the $N_rN_c\times N_tN_c$ channel impulse response matrix given as

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

with

$$\mathbf{h}(n_t,n_r)=\begin{bmatrix}h_0(n_t,n_r) & & & h_{L-1}(n_t,n_r) & \cdots & h_1(n_t,n_r)\\ h_1(n_t,n_r) & \ddots & & & \ddots & \vdots\\ \vdots & \ddots & h_0(n_t,n_r) & & \mathbf{0} & h_{L-1}(n_t,n_r)\\ h_{L-1}(n_t,n_r) & & h_1(n_t,n_r) & h_0(n_t,n_r) & & \\ & \ddots & \vdots & h_1(n_t,n_r) & \ddots & \\ \mathbf{0} & & h_{L-1}(n_t,n_r) & \vdots & \ddots & h_0(n_t,n_r)\end{bmatrix}, \quad (13)$$

and $\mathbf{n}_q=[\mathbf{n}_q(0),\cdots,\mathbf{n}_q(n_r),\cdots,\mathbf{n}_q(N_r-1)]^T$ is the noise vector, where $\mathbf{n}_q(n_r)=[n_q(n_r,0),\cdots,n_q(n_r,t),\cdots,n_q(n_r,N_c-1)]$ and $n_q(n_r,t)$ 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 $\mathbf{r}_q$ is transformed by N_c-point FFT into the frequency-domain signal $\mathbf{R}_q$ as

$$\mathbf{R}_q=\begin{bmatrix}\mathbf{R}_q(0)\\ \vdots\\ \mathbf{R}_q(n_r)\\ \vdots\\ \mathbf{R}_q(N_r-1)\end{bmatrix}=\begin{bmatrix}\mathbf{F} & & \mathbf{0}\\ & \ddots & \\ \mathbf{0} & & \mathbf{F}\end{bmatrix}\mathbf{r}_q$$
$$=C(N_r)\sqrt{\frac{2E_s}{T_s}}\begin{bmatrix}\mathbf{F} & & \mathbf{0}\\ & \ddots & \\ \mathbf{0} & & \mathbf{F}\end{bmatrix}\mathbf{h}\begin{bmatrix}\mathbf{F}^H & & \mathbf{0}\\ & \ddots & \\ \mathbf{0} & & \mathbf{F}^H\end{bmatrix}\mathbf{W}\widetilde{\mathbf{D}}_q(N_r)+\begin{bmatrix}\mathbf{N}_q(0)\\ \vdots\\ \mathbf{N}_q(n_r)\\ \vdots\\ \mathbf{N}_q(N_r-1)\end{bmatrix}, \quad (14)$$

where $\mathbf{R}_q(n_r)=[R_q(n_r,0),\cdots,R_q(n_r,k),\cdots,R_q(n_r,N_c-1)]^T$ and $\mathbf{N}_q(n_r)=\mathbf{F}\mathbf{n}_q=[N_q(n_r,0),\cdots,N_q(n_r,k),\cdots,N_q(n_r,N_c-1)]^T$. $\mathbf{N}_q(n_r)$ is the frequency-domain noise vector. Since $\mathbf{h}(n_t,n_r)$ in Eq. (12) is a circulant matrix, the eigenvalue decomposition using $\mathbf{F}$ can be applied [16] and we obtain

$$\mathbf{F}\mathbf{h}(n_t,n_r)\mathbf{F}^H=\begin{bmatrix}H(n_t,n_r,0) & & & & \mathbf{0}\\ & \ddots & & & \\ & & H(n_t,n_r,k) & & \\ & & & \ddots & \\ \mathbf{0} & & & & H(n_t,n_r,N_c-1)\end{bmatrix}\equiv\mathbf{H}(n_t,n_r), \quad (15)$$

where $H(n_t,n_r,k)=\sum_{l=0}^{L-1}h_l(n_t,n_r)\exp(-j2\pi k\tau_l/N_c)$. By introducing

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

Eq. (14) can be rewritten as

$$\mathbf{R}_q=C(N_r)\sqrt{\frac{2E_s}{T_s}}\mathbf{H}\mathbf{W}\widetilde{\mathbf{D}}_q(N_r)+\mathbf{N}_q. \quad (17)$$

FD-STBC-JTRD decoding is done on $\mathbf{R}_q$ to obtain $\tilde{\mathbf{R}}_q = [R_q(0),\cdots,R_q(N_c-1)]^T$ as

$$\tilde{\mathbf{R}}_0 = \mathbf{R}_0(0) \quad \text{for } N_r = 1$$

$$\begin{cases}\tilde{\mathbf{R}}_0 = \mathbf{R}_0(0)+\mathbf{R}_1^*(1)\\ \tilde{\mathbf{R}}_1 = \mathbf{R}_0(1)-\mathbf{R}_1^*(0)\end{cases} \quad \text{for } N_r = 2$$

$$\begin{cases}\tilde{\mathbf{R}}_0 = \mathbf{R}_0(0)+\mathbf{R}_1^*(1)+\mathbf{R}_2^*(2)\\ \tilde{\mathbf{R}}_1 = \mathbf{R}_0(1)-\mathbf{R}_1^*(0)+\mathbf{R}_3^*(2)\\ \tilde{\mathbf{R}}_2 = \mathbf{R}_0(2)-\mathbf{R}_2^*(0)-\mathbf{R}_3^*(1)\end{cases} \quad \text{for } N_r = 3 . \quad (18)$$

$$\begin{cases}\tilde{\mathbf{R}}_0 = \mathbf{R}_0(0)+\mathbf{R}_1^*(1)+\mathbf{R}_2^*(2)+\mathbf{R}_3(3)\\ \tilde{\mathbf{R}}_1 = \mathbf{R}_0(1)-\mathbf{R}_1^*(0)-\mathbf{R}_2(3)+\mathbf{R}_3^*(2)\\ \tilde{\mathbf{R}}_2 = \mathbf{R}_0(2)+\mathbf{R}_1(3)-\mathbf{R}_2^*(0)-\mathbf{R}_3^*(1)\end{cases} \quad \text{for } N_r = 4$$

Now, we assumed $W(n_t,n_r,k)=H^*(n_t,n_r,k)\cdot \tilde{W}(k)$ because an $N_t \times N_r$-order MRC diversity can be achieved, where $\tilde{W}(k)$ is the transmit weight in the frequency dimension. $\tilde{\mathbf{R}}_q$ is given by

$$\tilde{\mathbf{R}}_q = \sqrt{\frac{2E_s}{T_s}}\tilde{\mathbf{H}}\mathbf{d}_q + \sum_{n_r=0}^{N_r-1}\mathbf{N}_q(n_r), \quad (19)$$

where $\tilde{\mathbf{H}} = C(N_r)\left(\sum_{n_t=0}^{N_t-1}\sum_{n_r=0}^{N_r-1}\mathbf{H}(n_t,n_r)\mathbf{W}(n_t,n_r)\right)\mathbf{F}$ is the equivalent channel matrix.

In Sect.III, we will develop the channel capacity expression using Eq. (19) and find the 2D-WF transmit diversity weight matrix **W** that maximizes the channel capacity.

III. 2D-WF Transmit Weight For Multiple Receive Antenna Case

According to Ref. [13] which deals with the channel capacity of MIMO multiplexing, the channel capacity of SC-DAN downlink transmit diversity is given by

$$\begin{aligned}C &= \frac{R}{N_c}\log_2\left[\det\left(\mathbf{I}+\frac{E_s}{N_rN_0}\tilde{\mathbf{H}}\tilde{\mathbf{H}}^H\right)\right]\\ &= \frac{R}{N_c}\sum_{k=0}^{N_c-1}\log_2\left(1+\frac{E_s}{N_rN_0}\left|C^2(N_r)\sum_{n_t=0}^{N_t-1}\sum_{n_r=0}^{N_r-1}H(n_t,n_r,k)W(n_t,n_r,k)\right|^2\right)\end{aligned} . \quad (20)$$

Below, we describe a set of transmit weights that maximizes the channel capacity. The maximization problem can be written as

$$\begin{aligned}&\mathbf{W} = \max_{\{W(n_t,n_r,k)\}} .C\\ &\text{s.t.} \sum_{n_t=0}^{N_t-1}\sum_{n_r=0}^{N_r-1}\sum_{k=0}^{N_c-1}|W(n_t,n_r,k)|^2 = N_c \quad (\text{i.e, } C^2(N_r)=1)\end{aligned} . \quad (21)$$

In this paper, we derive the transmit weight that can maximize the upper-bound of Eq. (20). Using Cauchy-Schwarz inequality [17], Eq. (20) can be upper-bounded as

$$C \le \frac{R}{N_c}\sum_{k=0}^{N_c-1}\log_2\left(1+\frac{E_s}{N_rN_0}\sum_{n_t=0}^{N_t-1}\sum_{n_r=0}^{N_r-1}|H(n_t,n_r,k)|^2\sum_{n_t=0}^{N_t-1}\sum_{n_r=0}^{N_r-1}|W(n_t,n_r,k)|^2\right). \quad (22)$$

In Eq. (22), the equality holds if and only if

$$\begin{aligned}&\frac{W(0,0,k)}{H^*(0,0,k)} = \frac{W(1,0,k)}{H^*(1,0,k)} = \cdots = \frac{W(0,1,k)}{H^*(0,1,k)}\\ &= \cdots = \frac{W(n_t,n_r,k)}{H^*(n_t,n_r,k)} = \cdots = \frac{W(N_t-1,N_r-1,k)}{H^*(N_t-1,N_r-1,,k)}\end{aligned} . \quad (23)$$

The maximization problem of the capacity upper bound under the power constraint (i.e., $C^2(N_r)=1$) becomes a concave optimization problem under the total transmit power constraint [18]. This can be solved as [19,20] (for the sake of brevity, the derivation is omitted)

$$W(n_t,n_r,k) = \frac{H^*(n_t,n_r,k)}{\sqrt{\sum_{n_t=0}^{N_t-1}\sum_{n_r=0}^{N_r-1}|H(n_t,n_r,k)|^2}} \times \sqrt{\max\left\{\left(\varphi_{2D} - \frac{(E_s/N_rN_0)^{-1}}{\sum_{n_t=0}^{N_t-1}\sum_{n_r=0}^{N_r-1}|H(n_t,n_r,k)|^2}\right),0\right\}}, \quad (24)$$

where φ_{2D} is chosen so as to satisfy $\sum_{n_t=0}^{N_t-1}\sum_{n_r=0}^{N_r-1}\sum_{k=0}^{N_c-1}|W(n_t,n_r,k)|^2 = N_c$ (total transmit power constraint). The 2D-WF transmit diversity weight is a product of two terms; one is with respect to transmit/receive antennas and the other is with respect to the frequency. Using the 2D-WF transmit diversity weight, the power allocation is done across frequencies based on the WF theory and across transmit antennas based on the maximal ratio transmission (MRT). In this paper, we call the transmit diversity weight of Eq. (24) as the 2D-WF transmit diversity weight. When N_r=1, Eq. (24) reduces to the 2D-WF transmit diversity weight derived in [6].

IV. Numerical Evaluation

A. *Numerical evaluation condition*

The numerical evaluation condition is summarized in Table 2. The distribution of channel capacity is evaluated by Monte-Carlo numerical computation method. 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., $E[|h_l(n_t,n_r)|^2]=(1/L)\Omega(n_t)$). Ideal channel estimation is assumed. The user location is uniformly distributed over the area of interest.

Table 2 Numerical evaluation condition

Fading type	Block Rayleigh fading
Power delay profile	Uniform
Path loss exponent, α	3.5
Shadowing loss standard variation, σ	7.0 (dB)
No. of paths, L	16
Time delay τ_l	l (l = 0 ~ L−1)
No. of transmit antennas, N_t	1~10
No. of receive antennas, N_r	1~4
FFT size N_c	256
Normalized total average transmit E_s/N_0	10 (dB)
Channel estimation	Ideal

B. Comparison of number of transmit/receive antennas

Figure 3 plots the cumulative distribution function (CDF) of the channel capacity achievable with the 2D-WF transmit diversity weight for various combinations of N_t and N_r. From Fig. 3, we obtained the 10% channel capacity $C_{10\%}$ (below which the channel capacity falls with 10% probability). $C_{10\%}$ is plotted in Fig. 4 as a function of N_t with N_r as a parameter.

It can be seen from Fig. 4 that increasing N_t consistently increases the transmit diversity gain and hence, the channel capacity irrespective of N_r; however, the use of N_t of around 5 is sufficient. As for N_r, N_r=2 is found to maximize the channel capacity. This is a consequence of trade-off between diversity gain and coding rate. Increasing N_r increases the receive diversity gain, but, the coding rate reduces to 3/4 when N_r increases from 2 to 3 (see Table 1).

V. Conclusions

In this paper, we derived the 2D-WF transmit diversity weight for the SC-DAN downlink using FD-STBC-JTRD. Multiple transmit and receive antennas were considered. We evaluated the CDF of channel capacity by Monte-Carlo numerical computation method and showed that N_r=2 can achieve the highest channel capacity while the use of N_t of around 5 is sufficient.

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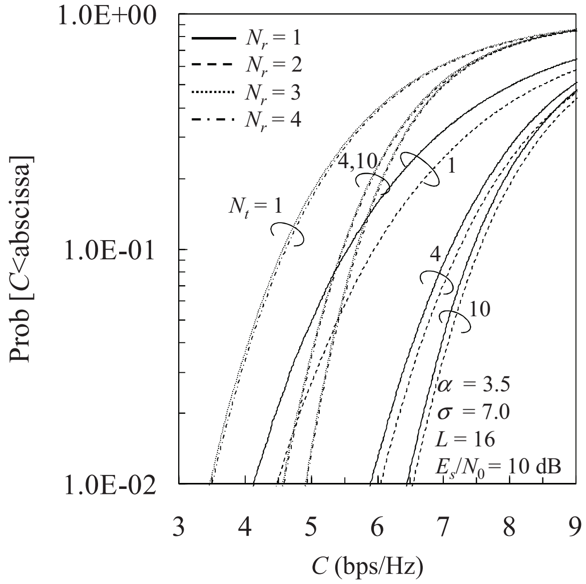

Figure 3. CDF of channel capacity.

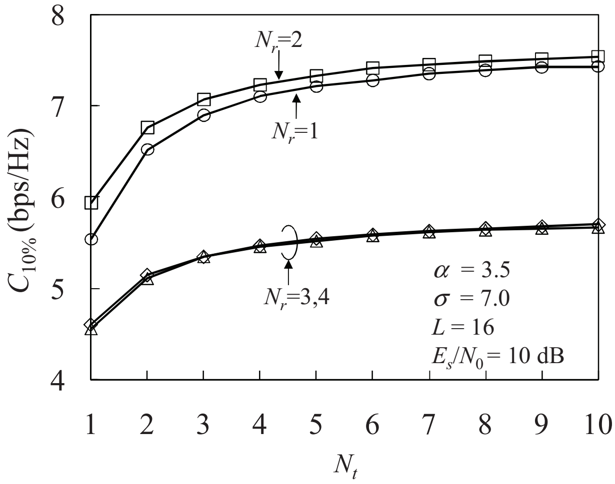

Figure 4. Impact of N_t and N_r on $C_{10\%}$.