Downlink Transmit Diversity For Broadband Single-carrier Distributed Antenna Network

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***Abstract*— In this paper, a 2-dimensional water-filling (2D-WF) based transmit weight that maximizes the channel capacity is theoretically derived for the downlink transmit diversity of single-carrier distributed antenna network (SC DAN) in a frequency-selective channel. The 2D-WF transmit weight allocates the transmit power in both transmit antenna dimension and frequency dimension: power allocation across frequencies based on WF theory and across transmit antennas based on maximal ratio transmission (MRT). The cumulative distribution function (CDF) of the achievable channel capacity by 2D-WF transmit diversity is evaluated by the Monte-Carlo numerical computation method. Channel capacities achievable with 2D-WF, MRT, and 1D WF transmit weight are compared. It is shown that the 2D-WF transmit weight provides the highest channel capacity among three transmit weights.**

Keywords-component; Distributed antenna network, transmit diversity, channel capacity, frequency-selective channel

I. INTRODUCTION

In broadband wireless systems, the received signal suffers from shadowing and path losses as well as frequency-selective fading [1]. The distributed antenna network (or system) (DAN or DAS) [2]-[7] is a promising wireless network to solve the problems arising from shadowing and path losses as well as frequency-selective fading. In DAN, many antennas 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). There are two ways to utilize DAN: transmit/receive diversity [6]-[11] and spatial multiplexing [12]-[15]. In this paper, we consider the downlink transmit diversity of the single-carrier (SC) DAN.

In our previous paper [7], we assumed a frequency-nonselective channel and investigated the channel capacity distribution of the DAN downlink using the maximal ratio transmit (MRT) diversity. The MRT diversity [10] is the optimal diversity in a frequency-nonselective channel, but it is not the optimal in a frequency-selective channel. In this paper, we theoretically derive a two-dimensional water-filling (2D-WF) based transmit diversity that can maximize the channel capacity in a frequency-selective channel. Then, we investigate, by the Monte Carlo numerical computation method, the cumulative distribution function (CDF) of the SC DAN downlink channel capacity.

The reminder of this paper is organized as follows. In Sect. II, we present the downlink transmit diversity model. In Sect. III, we develop a downlink capacity expression and then, by using the Lagrange multiplier method, derive the 2D-WF transmit weight that maximizes the channel capacity. In Sect. IV, we evaluate, by the Monte Carlo numerical computation method, the CDF 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. The SC DAN downlink transmit diversity model is illustrated in Fig. 2. N_t antennas close to an MS are selected for transmit diversity. A single user is assumed.

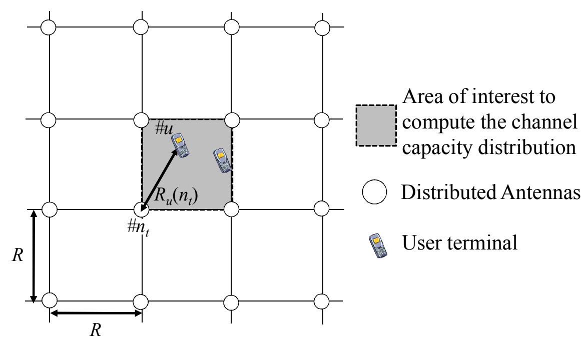

Figure 1. Transmit antenna allocation.

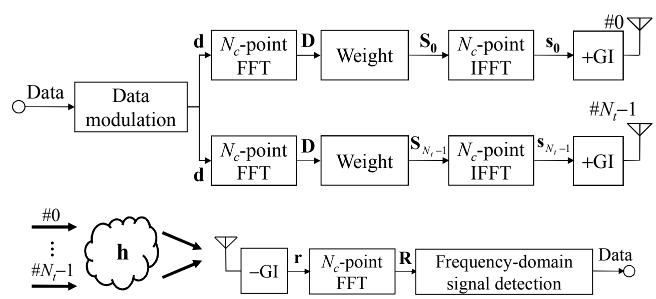

Figure 2. SC DAN downlink transmit diversity model.

B. Tramsmit signal

An information bit sequence is transformed into the data-modulated symbol sequence, which is then divided into a sequence of signal blocks of N_c symbols each, where N_c is the size of fast Fourier transform (FFT). In this paper, without loss of generality, one block of transmission, i.e., $\{d(t); t=0\sim N_c-1\}$, from N_t distributed antennas is considered. The transmit data symbol block is represented by $\mathbf{d}=[d(0),\cdots, d(N_c-1)]^T$.

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At the nth transmit antenna (n=0~N_t-1), N_c-point FFT is applied to transform **d** into the frequency-domain signal vector $\mathbf{D}$=$[D(0),\cdots,D(k),\cdots,D(N_c\text{-}1)]^T$ as

$$\mathbf{D}=\mathbf{Fd}\ , \tag{1}$$

where

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

is the FFT matrix of size $N_c\times N_c$. An $N_c\times N_c$ diagonal transmit weight matrix $\mathbf{W}_n$=diag$[W_n(0),...,W_n(N_c-1)]$ is multiplied to **D** to obtain $\mathbf{S}_n=[S_n(0),\cdots,S_n(N_c\text{-}1)]^T$ as

$$\mathbf{S}_n=\mathbf{W}_n\mathbf{D},\ n=0\sim N_t-1\ . \tag{3}$$

N_c-point inverse FFT (IFFT) is applied to transform $\mathbf{S}_n$ into the time-domain transmit signal block $\mathbf{s}_n=[s_n(0),\cdots,s_n(N_c\text{-}1)]^T$ as

$$\mathbf{s}_n=\mathbf{F}^H\mathbf{S}_n\ , \tag{4}$$

where $(.)^H$ is the Hermitian transpose operation. The last N_g samples in the block are copied as a cyclic prefix (CP) and inserted into the guard interval (GI) placed at the beginning of each block before transmission.

C. 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 $P_{r,n}$ for an MS whose distance from the nth distributed antenna is R_n can be modeled as [1]

$$\begin{aligned}P_{r,n}&=P_{t,n}\cdot R_n^{-\alpha}\cdot 10^{-\frac{\eta_n}{10}}=\left(P_{t,n}\cdot R^{-\alpha}\right)\cdot\left(\frac{R_n^{-\alpha}}{R^{-\alpha}}\right)\cdot 10^{-\frac{\eta_n}{10}}\\ &=\left(P_{t,n}\cdot R^{-\alpha}\right)\cdot\left(d_n^{-\alpha}\right)\cdot 10^{-\frac{\eta_n}{10}}\end{aligned}\ , \tag{5}$$

where $P_{t,n}$ is the transmit power, α is the path loss exponent and η_n is the shadowing loss (dB) with zero-mean and standard variation σ. Letting $E_{s,n}$ be $\left(P_{t,n}\cdot R^{-\alpha}\right)\cdot T_s$, where T_s being the symbol duration, Eq. (5) is rewritten as

$$P_{r,n}=\frac{E_{s,n}}{T_s}\cdot d_n^{-\alpha}\cdot 10^{-\frac{\eta_n}{10}}\ . \tag{6}$$

Assuming that the frequency-selective channel is composed of L distinct paths, the channel impulse response $h_n(\tau)$ between the nth distributed antenna and the MS receive antenna is given by

$$h_n(\tau)=\sum_{l=0}^{L-1}h_{n,l}\delta(\tau-\tau_l)\ , \tag{7}$$

where $h_{n,l}=\sqrt{\Omega_n}\cdot\widetilde{h}_{n,l}$ and τ_l are respectively the complex-valued path gain and the time delay of the lth path between the nth transmit antenna and the MS with $\widetilde{h}_{n,l}$ being the complexed-valued path gain with $E[\sum_{l=0}^{L-1}|\widetilde{h}_{n,l}|^2]=1$ and Ω_n denoting the path loss plus shadowing loss, given by

$$\Omega_n=d_n^{-\alpha}\cdot 10^{-\frac{\eta_n}{10}}\ . \tag{8}$$

D. Received signal

The GI-removed received signal block $\mathbf{r}=[r(0),\cdots,r(N_c-1)]^T$ can be expressed using the matrix form as

$$\mathbf{r}=\sqrt{\frac{2E_s}{T_s}}\mathbf{hs}+\mathbf{n}\ , \tag{9}$$

where $E_s=\sum_{n=0}^{N_t-1}E_{s,n}$, $\mathbf{s}=[\mathbf{s}_0,\cdots,\mathbf{s}_{N_t-1}]^T$ and **h** is the $N_c\times N_tN_c$ channel impulse response matrix given as

$$\mathbf{h}=[\mathbf{h}_0\ \ \cdots\ \ \mathbf{h}_{N_t-1}] \tag{10}$$

with

$$\mathbf{h}_n=\begin{bmatrix}h_{n,0} & & & h_{n,L-1} & \vdots\\ \vdots & h_{n,0} & \mathbf{0} & & h_{n,L-1}\\ h_{n,L-1} & \vdots & \ddots & & \\ & h_{n,L-1} & & h_{n,0} & \\ & & \ddots & \vdots & h_{n,0}\\ \mathbf{0} & & & h_{n,L-1} \ \vdots & h_{n,0}\end{bmatrix}, \tag{11}$$

and $\mathbf{n}=[n(0),\cdots,n(N_c-1)]^T$ is the noise vector. The tth element, $n(t)$, of **n** is the zero-mean additive white Gaussian noise (AWGN) having variance $2N_0/T_s$ with N_0 being the one-sided noise power spectrum density.

The received signal block **r** is transformed by N_c-point FFT into the frequency-domain signal $\mathbf{R}=[R(0),\cdots,R(N_c-1)]^T$ as

$$\mathbf{R}=\mathbf{Fr}=\sqrt{\frac{2E_s}{T_s}}\mathbf{Fhs}+\mathbf{Fn}=\sqrt{\frac{2E_s}{T_s}}\mathbf{FhF}^H\mathbf{WFd}+\mathbf{Fn}\ , \tag{12}$$

where $\mathbf{W}=[\mathbf{W}_0,\cdots,\mathbf{W}_{N_t-1}]^T$. In III-B, we derive the channel capacity using Eq. (12) and find **W** that maximizes the channel capacity.

III. Transmit Weight

A. Conventional Transmit Weight

MRT transmit weight which can maximize the received signal-to-noise power ratio (SNR) is given by [10]

$$W_n(k)=\frac{H_n(k)^*}{\sqrt{\frac{1}{N_c}\sum_{n'=0}^{N_t-1}\sum_{k'=0}^{N_c-1}|H_{n'}(k')|^2}}\qquad(\text{MRT})\ . \tag{13}$$

The transmit diversity weight based on the water filling theorem [16] can also be used. This transmit weight is called one-dimensional water filling (1D-WF) transmit weight in this paper. The 1D-WF transmit weight is derived by applying the water filling theorem to an array of N_tN_c channel gains, $[H_0(0),\cdot\cdot,H_0(N_c-1),H_1(0),\cdot\cdot,H_1(N_c-1),\cdot\cdot,H_{N_t-1}(0),\cdot\cdot,H_{N_t-1}(N_c-1)]$.

The 1D-WF transmit weight $W_n(k)$ at the kth frequency for the nth distributed antenna is given by

$$W_n(k)=\frac{H_n(k)^*}{|H_n(k)|}\left[\max\left\{\varphi_{1D}\left(\frac{E_s}{N_0},\{H_n(k)\}\right)-\left(\frac{E_s}{N_0}\right)^{-1}\frac{1}{|H_n(k)|^2},0\right\}\right]^{\frac{1}{2}} \quad (14)$$

(1D–WaterFilling)

where $\varphi_{1D}(E_s/N_0,\{H_n(k)\})$ is chosen so that $\sum_{n=0}^{N_t-1}\sum_{k=0}^{N_c-1}|W_n(k)|^2=N_c$.

B. 2D-WF transmit weight

Since the channel impulse response matrix $\mathbf{h}_n$ is a circulant matrix, the eigenvalue decomposition using $\mathbf{F}$ can be applied [17]. We have

$$\mathbf{F}\mathbf{h}_n\mathbf{F}^H=\begin{bmatrix}H_n(0) & & & \mathbf{0}\\ & H_n(1) & & \\ & & \ddots & \\ \mathbf{0} & & & H_n(N_c-1)\end{bmatrix}\equiv\mathbf{H}_n. \quad (15)$$

where

$$H_n(k)=\sum_{l=0}^{L-1}h_{n,l}\exp(-j2\pi k\tau_l/N_c),\ k=0\sim N_c-1. \quad (16)$$

Using Eq. (15), Eq. (12) can be rewritten as

$$\begin{aligned}\mathbf{R}&=\sqrt{\frac{2E_s}{T_s}}\mathbf{F}\mathbf{h}\mathbf{F}^H\mathbf{W}\mathbf{F}\mathbf{d}+\mathbf{F}\mathbf{n}=\sqrt{\frac{2E_s}{T_s}}\mathbf{H}\mathbf{W}\mathbf{F}\mathbf{d}+\mathbf{N}\\&=\sqrt{\frac{2E_s}{T_s}}\widetilde{\mathbf{H}}\mathbf{d}+\mathbf{N}\end{aligned}, \quad (17)$$

where $\mathbf{H}=[\mathbf{H}_0,\cdots,\mathbf{H}_{N_r-1}]$. $\widetilde{\mathbf{H}}=\mathbf{H}\mathbf{W}\mathbf{F}$ and $\mathbf{N}=[N(0),\cdots,N(N_c-1)]$ are respectively the equivalent channel matrix and the frequency-domain noise vector.

According to Ref. [14] which deals with the channel capacity of MIMO multiplexing, the channel capacity of SC DAN downlink transmit diversity is given by

$$\begin{aligned}C&=\frac{1}{N_c}\log_2\left[\det\left(\mathbf{I}+\frac{E_s}{N_0}\widetilde{\mathbf{H}}\widetilde{\mathbf{H}}^H\right)\right]\\&=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\log_2\left(1+\frac{E_s}{N_0}\left|\sum_{n=0}^{N_t-1}H_n(k)W_n(k)\right|^2\right)\end{aligned}. \quad (18)$$

We want to find the set of transmit weights that maximizes the channel capacity. The maximization problem can be written as

$$\begin{aligned}&\max.C\{W_n(k)\}=\sum_{k=0}^{N_c-1}\log_2\left(1+\frac{E_s}{N_0}\left|\sum_{n=0}^{N_t-1}H_n(k)W_n(k)\right|^2\right)\\&\text{s.t.}\sum_{n=0}^{N_t-1}\sum_{k=0}^{N_c-1}|W_n(k)|^2=N_c\end{aligned}. \quad (19)$$

It is quite difficult to solve Eq. (19). In this paper, we derive the transmit weight that can maximize the upper bound of Eq. (18). Using Cauchy-Schwarz inequality [18], Eq. (19) can be upper bounded as

$$C\le\sum_{k=0}^{N_c-1}\log_2\left(1+\frac{E_s}{N_0}\sum_{n=0}^{N_t-1}|H_n(k)|^2\sum_{n=0}^{N_t-1}|W_n(k)|^2\right). \quad (20)$$

In Eq. (20), the equality holds if and only if

$$\frac{W_0(k)}{H_0(k)^*}=\cdots=\frac{W_n(k)}{H_n(k)^*}=\cdots=\frac{W_{N_t-1}(k)}{H_{N_t-1}(k)^*}, \quad (21)$$

Equation (19) is a concave optimization problem under the total transmit power constraint [19]. This can be solved as [20,21] (for the sake of brevity, the derivation is omitted).

$$W_n(k)=\frac{H_n(k)^*}{\sqrt{\sum_{n=0}^{N_t-1}|H_n(k)|^2}}\left[\max\left\{\varphi_{2D}\left(\frac{E_s}{N_0},\{H_n(k)\}\right)-\left(\frac{E_s}{N_0}\right)^{-1}\frac{1}{\sum_{n=0}^{N_t-1}|H_n(k)|^2},0\right\}\right]^{\frac{1}{2}}, \quad (22)$$

where $\varphi_{2D}(E_s/N_0,\{H_n(k)\})$ is chosen so that $\sum_{n=0}^{N_t-1}\sum_{k=0}^{N_c-1}|W_n(k)|^2=N_c$. In this paper, we call the transmit weight of Eq. (22) as the 2D-WF transmit weight.

C. Discussion

When N_t=1, Eq. (22) reduces to

$$W_n(k)=\frac{H_n(k)^*}{|H_n(k)|}\left[\max\left\{\varphi_{2D}\left(\frac{E_s}{N_0},\{H_n(k)\}\right)-\left(\frac{E_s}{N_0}\right)^{-1}\frac{1}{|H_n(k)|^2},0\right\}\right]^{\frac{1}{2}}, \quad (23)$$

which is identical to the 1D-WF transmit weight given by Eq. (17). This suggests that the power allocation is done based on the WF theorem across N_c frequencies. On the other hand, when N_t>1 assuming frequency-nonselective channel (i.e., L=1), $H_n(k)=H_n$ for k=0~N_c–1 and therefore, Eq. (22) reduces to

$$W_n(k)=\frac{H_n(k)^*}{\sqrt{\sum_{n=0}^{N_t-1}|H_n(k)|^2}}, \quad (24)$$

which is identical to the MRT transmit weight given by Eq. (16). This shows that the transmit power allocation is done based on the MRT strategy across N_t transmit antennas.

As a consequence, the 2D-WF transmit weight allocates the transmit power in both transmit antenna dimension and frequency dimension: power allocation based on the WF theorem across N_c frequencies and based on the MRT strategy across N_t transmit antennas.

IV. Numerical Evaluation

A. Numerical evaluation condition

The numerical evaluation condition is summarized in Table 1. 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. Ideal channel estimation is assumed. The user location is uniformly distributed over the area of interest. Independent shadowing losses and independent fading variations among different transmit antennas are assumed.

Table 1 Numerical computation 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	$\alpha = 3.5$
Shadowing loss standard deviation	$\sigma = 7.0$ (dB)
No. of selected antennas	$N_t = 1,2,3,\ldots,10$
FFT size	$N_c = 256$
Normalized transmit E_s/N_0	$E_s/N_0 = 10$ (dB)
Channel estimation	Ideal

B. *Comparison of transmit weight*

Figure 3 shows the cumulative distribution function (CDF) of channel capacity with N_t as a parameter for 2D-WF transmit weight, 1D-WF transmit weight and MRT transmit weight. From Fig. 3, we obtained the 10% channel capacity $C_{10\%}$ (below which the channel capacity falls with 10% probability), the 90% channel capacity $C_{90\%}$ (which the channel capacity exceeds with 10% probability), and the ergodic capacity E[C]. They are plotted in Fig. 4 as a function of N_t.

It can be seen from Fig. 4 that the 2D-WF transmit weight provides the highest channel capacity among three transmit weights. As N_t increases, $C_{10\%}$ and E[C] increases (see Fig. 4(a) and Fig. 4(c)); however, $C_{90\%}$ is almost constant except for the 1D-WF transmit weight (see Fig. 4(b)). The reason for this is discussed below.

Due to increased space diversity as N_t increases, the probability that the capacity drops is reduced. This contributes to increase the value of $C_{10\%}$. This happens if the MS is far from all antennas. On the other hand, if the MS is closer to one of distributed antennas, higher capacity can be obtained. How the transmit diversity can increase the capacity in such a case can be discussed using $C_{90\%}$. Since the MS is close to one of distributed antennas, most of transmit power is allocated to that antenna and hence, the diversity gain does not increase that much even if N_t is increased. It is interesting to notice that as N_t increases, $C_{90\%}$ reduces when the 1D-WF transmit weight is used. This is because that when 1D-WF is used the transmit power of the nearest antenna decreases [Appendix A]. Using 2D-WF, as N_t increases, $C_{10\%}$ increases while $C_{90\%}$ is almost constant, resulting in the increased E[C].

V. CONCLUSIONS

In this paper, we derive theoretically the 2D-WF transmit weight for the SC DAN downlink transmit diversity in a frequency-selective channel. The 2D-WF transmit weight allocates the transmit power in both transmit antenna dimension and frequency dimension. We evaluated the CDF of channel capacity by Monte-Carlo numerical computation method and compared the channel capacities achievable by the 2D-WF, 1D-WF, and MRT transmit weights. It was shown that the 2D-WF transmit weight can achieve the highest channel capacity among three transmit weights.

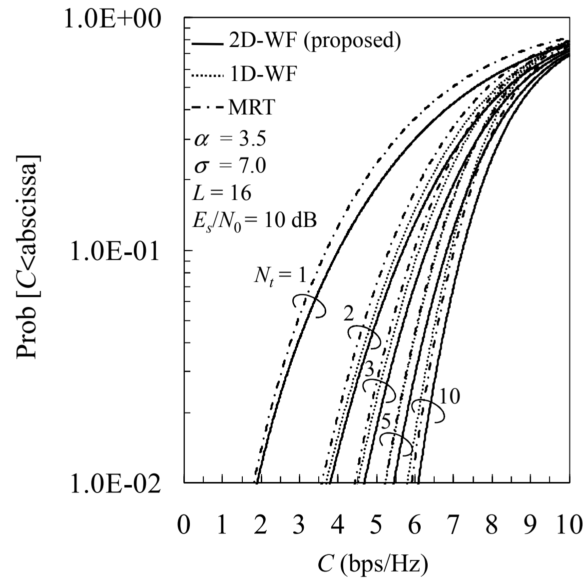
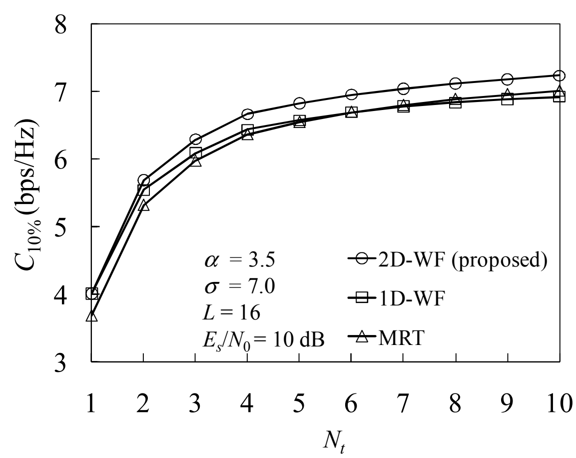
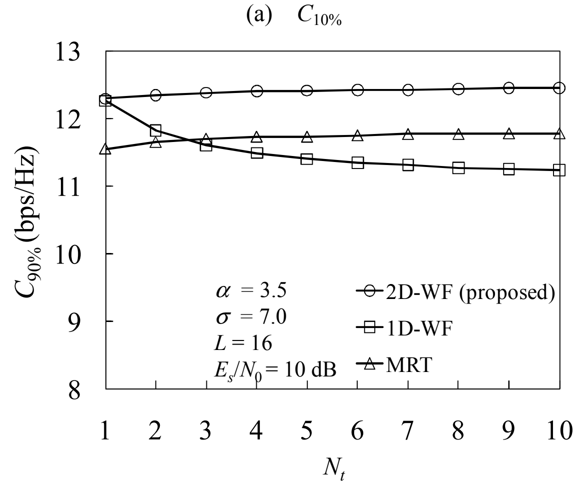


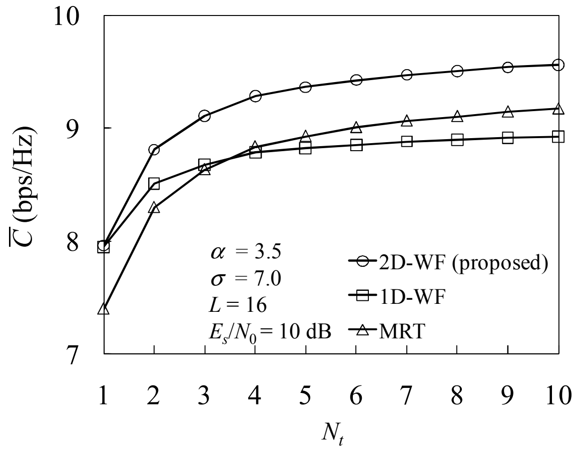
Figure 3. CDF of channel capacity.



(a) $C_{10\%}$



(b) $C_{90\%}$



(c) E[C]

Figure 4. N_t vs Capacity.

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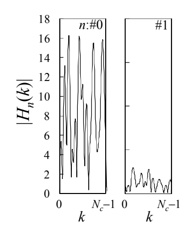
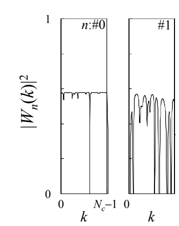
Appendix A: Transmit weight of 1D-WF

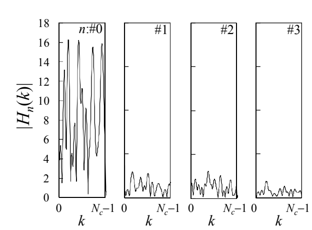
Figure 5 shows the channel transfer function and the 1D-WF transmit weight when the user is located near the transmit antenna #0. The channel gains for the transmit antennas #1, #2 and #3 are lower than that for the transmit antenna #0. It can be understood from Fig. (a-2) and Fig. (b-2), as N_t increases from 2 to 4, the transmit power of antenna #0 decreases.



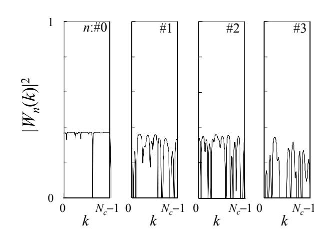
(a-1) Channel transfer functions,



(a-2) 1D-WF with N_t=2,



(b-1) Channel transfer functions,



(b-2) 1D-WF with N_t=4,

Figure 5. 1D-WF weight.