

[Poster Presentation] Optimum Number of Transmit Antennas for STTD in MC-CDMA Distributed Antenna Network

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Abstract Distributed antenna network (DAN) can mitigate the problems arising from of the propagation path loss and shadowing loss and achieve better transmission performance over the entire cell area compared to centralized antenna network (CAN). Many transmit antennas are spatially distributed in DAN while they are co-located at the same base station in CAN. The use of space time block coded transmit diversity (STTD) can further improve performance. As the multi-access scheme, multicarrier code division multiple access (MC-CDMA) is attractive since it can alleviate a sophisticated resource allocation while achieving the frequency diversity gain. In this paper, the STTD downlink transmission is considered for MC-CDMA DAN. In STTD downlink, higher spatial antenna diversity gain is obtained by increasing the number of transmit antennas. However, the STTD code rate reduces. Furthermore, since the propagation time delays of signals transmitted from some transmit antennas may exceed the cyclic prefix (CP) length of MC-CDMA, the inter-block interference (IBI) is produced. Therefore, there exist the optimum numbers of transmit antennas to be involved in STTD downlink. In this paper, we derive an expression for the received SINR after de-spreading and evaluate, by computer simulation, the STTD downlink capacity to discuss about the optimum number of transmit antennas.

Keywords Distributed antenna network, MC-CDMA, transmit antenna diversity.

1. Introduction

In next generation wireless communications systems, very high-speed and high-capacity data transmission is demanded. However, broadband wireless transmission suffers from propagation path loss, shadowing loss and frequency-selective fading [1]. Distributed antenna network (DAN), in which many antennas are spatially distributed in the cell, is an effective method to solve this problem [2-3]. Since some antennas are located near the user with high probability, better transmission performance can be obtained entire the cell compared to centralized antenna network (CAN), where many antennas are co-located at the base station.

Which multi-access technique should be used is an important issue. Frequency-division multi access (FDMA) and time-division multi access (TDMA) allocate the orthogonal channels to different users [4]. The channel allocation for FDMA and TDMA must be carefully done by taking into account the signal-to-interference plus noise power ratios (SINRs) of all active users and becomes sophisticated [5]. On the other hand, the code division multiple access (CDMA) [4] can alleviate the channel allocation problem. Therefore, CDMA can be a good candidate as the multiple access for DAN. There are two types of CDMA: multi-carrier (MC)-CDMA and DS-CDMA. In this paper, we consider MC-CDMA DAN. DS-CDMA DAN is under study by Inoshita et al. [6].

Even replacing CAN by DAN, the transmission

performance is still affected by propagation path loss, shadowing loss and frequency-selective fading. The use of transmit and receive diversity can further improve the performance [7]. Among various diversity techniques, the transmit diversity using space-time block coding (STBC) [8-9] is attractive. In this paper, the transmit diversity using STBC is called space-time transmit diversity (STTD). We have studied an application of STTD to single-carrier DAN [10].

In this paper, as our initial study, we study the STTD downlink of MC-CDMA DAN. In STTD downlink, the antenna diversity gain increases as the number of transmit antennas increases while the STTD code rate decreases with increasing the number of transmit antennas. Therefore, there exists the optimum number of transmit antennas that maximizes the downlink capacity in STTD downlink. Furthermore, in DAN, the antenna-to-mobile terminal (MT) distance is different for a different transmit antennas since antennas are spatially distributed. Thus, if too many transmit antennas are involved in the STTD downlink, the propagation time delay associated with some transmit antennas may exceed the cyclic prefix (CP) length and as a consequence, the downlink capacity of DAN degrades due to the inter-block interference (IBI) caused by the longer antenna-to-MT distance difference than the CP length. Therefore, the optimum numbers of transmit antennas may be different for DAN and CAN.

In this paper, we derive an expression of the received

SINR after de-spreading and evaluate, by computer simulation, the STTD downlink capacity to discuss about the optimum number of transmit antennas.

The remainder of this paper is organized as follows. Section 2 presents MC-CDMA network model. The STTD downlink transmission is described in Sect. 3. Section 4 discusses computer simulation results and finally, Sect. 5 offers some conclusions and future work.

2. STTD Downlink for MC-CDMA DAN

2.1 Cellular structure

In this paper, we consider multi-cell and multi-user environment. Fig. 1 illustrates the cellular structure considered in this paper. The total number of antennas in cell is set to $N_{total}=7$. The center cell is the cell of interest and there are 6 co-channel interference (CCI) cells around the cell of interest. In DAN, many antennas are spatially distributed and each distributed antenna covers regular hexagon small cell with radius of $D_{cell}'=D_{cell}/\sqrt{7}$, where D_{cell} is cell radius in CAN. Each distributed antenna is connected with signal processing center (SPC) by optical fiber. In this paper, the transmission between distributed antenna and signal processing center is assumed to be ideal. On the other hand, in CAN, N_{total} antennas are co-located at the center of cell.

There are U users in each cell and each user equips with single antennas. Each user selects N_t distributed antennas as transmit antennas for STDD downlink.

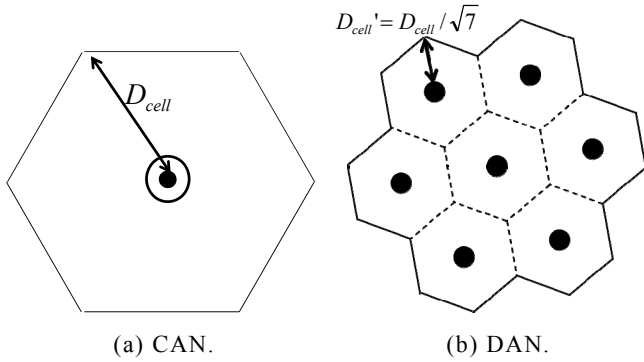


Fig. 1 Network model.

2.2 Propagation channel

Broadband wireless channel is characterized by propagation path loss, shadowing loss and frequency selective fading [1]. Assuming frequency-domain fading is composed of L -delay paths having different delay time, the channel impulse response, $\tilde{h}_{c,u}^{c',m(u')\rightarrow u}(\tau)$, of the link between the $m(u')$ -th transmit antenna for the u' -th user in c' -th cell and the u -th user in the c -th cell can be expressed as

$$\tilde{h}_{c,u}^{c',m(u')\rightarrow u}(\tau) = \sum_{l=0}^{L-1} \tilde{h}_{c,u,l}^{c',m(u')\rightarrow u} \cdot \delta(\tau - \tau_{c,u,l}^{c',m(u')\rightarrow u}), \quad (1)$$

where $\tilde{h}_{c,u,l}^{c',m(u')\rightarrow u}$ is the l -th complex valued path gain, including the impact of propagation path loss and shadowing loss, of the link between the $m(u')$ -th transmit

antenna for the u' -th user in the c' -th cell and the u -th user in the c -th cell. It can be expressed as

$$\tilde{h}_{c,u,l}^{c',m(u')\rightarrow u} = \sqrt{D_{c,u}^{c',m(u')\rightarrow u} \cdot 10^{-\frac{\eta_{c,u}^{c',m(u')\rightarrow u}}{10}}} \cdot g_{c,u,l}^{c',m(u')\rightarrow u}, \quad (2)$$

where $D_{c,u}^{c',m(u')\rightarrow u}$ is the distance between the $m(u')$ -th transmit antenna for the u' -th user in c' -th cell and the u -th user in the c -th cell. α denotes the path loss exponent and $\eta_{c,u}^{c',m(u')\rightarrow u}$ is the shadowing loss in dB between the $m(u')$ -th transmit antenna for the u' -th user in the c' -th cell and the u -th user in the c -th cell. $g_{c,u,l}^{c',m(u')\rightarrow u}$ is the l -th complex valued path gain with $E[\sum_{l=0}^{L-1} |g_{c,u,l}^{c',m(u')\rightarrow u}|^2] = 1$ of the link between the $m(u')$ -th transmit antenna for the u' -th user in the c' -th cell and the u -th user in the c -th cell, where $E[\cdot]$ denotes ensemble average operation. $\tau_{c,u,l}^{c',m(u')\rightarrow u}$ is delay time of the l -th path of the link between the $m(u')$ -th transmit antenna for the u' -th user in the c' -th cell and the u -th user in the c -th cell.

The instantaneous received power, $P_{c,u}^{c',m(u')\rightarrow u}$, at the u -th user in the c -th cell from the $m(u')$ -th transmit antenna for the u' -th user in the c' -th cell can be expressed as

$$\begin{aligned} P_{c,u}^{c',m(u')\rightarrow u} &= P_t^{c',m(u')} \cdot \left| \sum_{l=0}^{L-1} \tilde{h}_{c,u,l}^{c',m(u')\rightarrow u} \right|^2 \\ &= P_t^{c',m(u')} \cdot D_{c,u}^{c',m(u')\rightarrow u - \alpha} \cdot 10^{-\frac{\eta_{c,u}^{c',m(u')\rightarrow u}}{10}} \cdot \left| \sum_{l=0}^{L-1} g_{c,u,l}^{c',m(u')\rightarrow u} \right|^2, \end{aligned} \quad (3)$$

where $P_t^{c',m(u')}$ is the transmit power of the $m(u')$ -th transmit antenna for the u' -th user in the c' -th cell. (3) can be rewritten as

$$P_{c,u}^{c',m(u')\rightarrow u} = P_t^{c',m(u')} \cdot d_{c,u}^{c',m(u')\rightarrow u - \alpha} \cdot 10^{-\frac{\eta_{c,u}^{c',m(u')\rightarrow u}}{10}} \cdot \left| \sum_{l=0}^{L-1} g_{c,u,l}^{c',m(u')\rightarrow u} \right|^2, \quad (4)$$

where $P_t^{c',m(u')} = p_t^{c',m(u')} D_{cell}^{-\alpha}$ is the normalized transmit power of the $m(u')$ -th transmit antenna for the u' -th user in the c' -th cell. $d_{c,u}^{c',m(u')\rightarrow u} = D_{c,u}^{c',m(u')\rightarrow u} / D_{cell}$ is the normalized distance between the $m(u')$ -th transmit antenna for the u' -th user in the c' -th cell and the u -th user in the c -th cell.

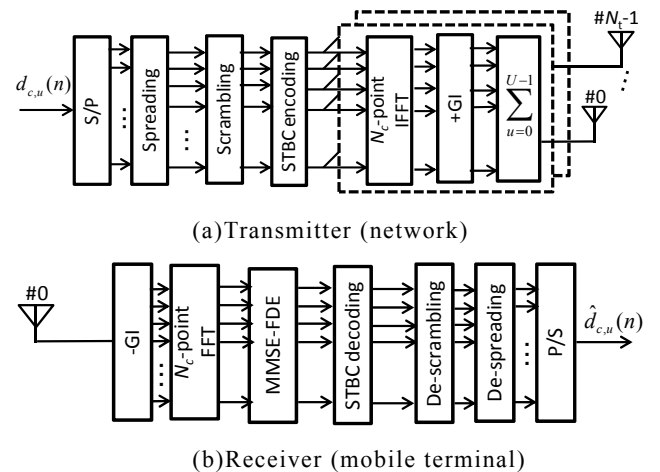


Fig. 2 STTD downlink transmitter/receiver structures.

2.3 STTD downlink

The transmitter/receiver structure of STTD downlink in MC-CDMA DAN is illustrated Fig. 2. The STTD code rate, R , is functions of the number of transmit antennas as shown in Table 1.

Table 1 STTD code rate.

| Number N_t of transmit antennas | Code rate |
|-----------------------------------|-----------|
| 1 | 1 |
| 2 | 1 |
| 3 | 3/4 |
| 4 | 3/4 |

Throughout this paper, the chip-spaced discrete-time signal representation is used. Due to the page limitation, we describe the case of $N_t=2$.

In the following, we focus on STTD downlink transmission for the u th user in the c -th cell. At the transmitter in the c -th cell, J data modulated symbol blocks of N_c/SF symbols each are spread by multiplying a user-specific spreading code $\{c_u^{SF}(k); k=0\sim(N_c-1)\}$, where N_c denotes the number of subcarriers per block and SF is spreading factor, respectively. The j -th MC-CDMA transmit signal $\{D_{c,u,j}(k); k=0\sim(N_c-1)\}$ is given as

$$D_{c,u,j}(k) = c_c^{scr}(k)c_u^{SF}(k \bmod SF)d_{c,u,j}(\lfloor k/SF \rfloor), \quad (5)$$

where $\lfloor x \rfloor$ denotes the largest integer which is smaller than or equal to x . Then, $N_t \times Q$ coded frequency-domain signals are generated by applying STBC encoding to J MC-CDMA transmit signals. This processing is done at each subcarrier. $N_t \times Q$ coded frequency-domain transmit signal matrix $\{\mathbf{S}_{std}^{c,u}(k); k=0\sim(N_c-1)\}$ at the k -th subcarrier is given as

$$\mathbf{S}_{std}^{c,u}(k) = \sqrt{\frac{2P_t}{N_t}} \begin{pmatrix} D_{c,u,0}(k) & -D_{c,u,1}^*(k) \\ D_{c,u,1}(k) & D_{c,u,0}^*(k) \end{pmatrix}. \quad (6)$$

The frequency-domain transmit signal is transformed into time-domain signal by N_c -point inverse fast Fourier transform (IFFT). After inserting CP into the guard interval at the beginning each block, the transmitter multiplexes U transmit signals and transmit it to U users.

At the u th user, after CP removal, the received signal is transformed back to the frequency-domain received signal by N_c -point FFT. The frequency-domain received signal vector of the u -th user in c -th cell $\mathbf{R}_{c,u}(k) = [R_{c,u,0}(k), \dots, R_{c,u,q}(k), \dots, R_{c,u,Q-1}(k)]^T$ can be expressed as

$$\begin{aligned} \mathbf{R}_{c,u}(k) &= \mathbf{H}_{c,u}(k)\mathbf{S}_{std}^{c,u}(k) + \sum_{\substack{u'=0 \\ u' \neq u}}^{U-1} \mathbf{H}_{c,u'}(k)\mathbf{S}_{std}^{c,u'}(k) \\ &+ \sum_{\substack{c'=0 \\ c' \neq c}}^{CELL-1} \sum_{u'=0}^{U-1} \mathbf{H}_{c',u'}(k)\mathbf{S}_{std}^{c',u'}(k) + \mathbf{V}_{c,u}(k) + \mathbf{N}_{c,u}(k) \end{aligned}, \quad (7)$$

where $\mathbf{H}_{c,u'}(k) = [H_{c,u'}^{0(u) \rightarrow u'}(k), \dots, H_{c,u'}^{m(u) \rightarrow u'}(k), \dots, H_{c,u'}^{N_t-1(u) \rightarrow u'}(k)]$ is $N_t \times 1$ channel transfer function matrix of the link between the transmit antennas for the u' -th user in the c' -th cell and the u -th user in the c -th cell.

$\mathbf{V}_{c,u}(k) = [V_{c,u,0}(k), \dots, V_{c,u,q}(k), \dots, V_{c,u,Q-1}(k)]^T$ is the IBI vector of the u -th user in the c -th cell given as

$$\begin{aligned} &V_{c,u,q}(k) \\ &= \sum_{l=0}^{N_c-1} \left[\sum_{l=0}^{L-1} h_{c,u,l}^{m(u) \rightarrow u} \begin{Bmatrix} \tilde{s}_{c,u}(t - \tau_l - \Delta_{c,u}) \\ -\tilde{s}_{c,u}((t - \tau_l - \Delta_{c,u}) \bmod N_c) \end{Bmatrix} \right] \\ &\quad \times (u_0(t) - u_0(t - \tau_l - \Delta_{c,u})) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{aligned}, \quad (8)$$

where $\Delta_{c,u}$ denotes the propagation time delay difference between the transmit antennas. $\tilde{s}_{c,u}(t)$ is given as

$$\tilde{s}_{c,u}(t) = \begin{cases} s_{c,u,q}(t) & \text{if } 0 \leq t \leq N_c - 1 \\ s_{c,u,q}(t + N_c) & \text{if } -N_g \leq t < 0 \\ s_{c,u,q-1}(t + N_c + N_g) & \text{if } -(N_c + N_g) \leq t \leq -N_g \end{cases}. \quad (9)$$

$u_0(t)$ is the unit step function given as

$$u_0(t) = \begin{cases} 1 & (t \geq 0) \\ 0 & (t < 0) \end{cases}. \quad (10)$$

If the maximum propagation time delay difference exceeds the CP length ($N_g < (\tau_l + \Delta_{c,u})$), (8) can be rewritten as

$$\begin{aligned} &V_{c,u,q}(k) \\ &= \sum_{l=0}^{N_c-1} \left[\sum_{l=0}^{L-1} h_{c,u,l}^{m(u) \rightarrow u} \begin{Bmatrix} s_{c,u,q-1}(t - \tau_l - \Delta_{c,u} + N_c + N_g) \\ -s_{c,u,q}(t - \tau_l - \Delta_{c,u} + N_c) \end{Bmatrix} \right] \\ &\quad \times (u_0(t) - u_0(t - \tau_l - \Delta_{c,u} + N_g)) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{aligned}. \quad (11)$$

$\mathbf{N}_{c,u}(k) = [N_{c,u,0}(k), \dots, N_{c,u,q}(k), \dots, N_{c,u,Q-1}(k)]^T$ is the noise vector and $N_{c,u,q}(k)$ is the zero mean complex-valued additive white Gaussian noise (AWGN) having variance $2N_0/T_c$ being N_0 and T_c as the single-sided power spectrum density of AWGN and chip duration, respectively. Then, minimum mean square error based frequency-domain equalization (MMSE-FDE) and STBC decoding are performed. The j ($j=0, 1$)th decoded frequency-domain signal, $\{\hat{D}_{c,u,j}(k); k=0\sim(N_c-1)\}$, of the u -th user in the c -th cell is given as

$$\begin{pmatrix} \hat{D}_{c,u,0}(k) \\ \hat{D}_{c,u,1}(k) \end{pmatrix} = \begin{pmatrix} W_{c,u}^{0(u) \rightarrow u}(k)R_{c,u,0} + W_{c,u}^{1(u) \rightarrow u}(k)R_{c,u,1}^* \\ W_{c,u}^{1(u) \rightarrow u}(k)R_{c,u,0} - W_{c,u}^{0(u) \rightarrow u}(k)R_{c,u,1}^* \end{pmatrix}, \quad (12)$$

where $W_{c,u}^{(u) \rightarrow u}(k)$ is MMSE-FDE weight. MMSE-FDE weight is designed so as to the mean square error (MSE) between the transmit signal, $D_{c,u,j}(k)$, before STBC encoding and receive signal, $\hat{D}_{c,u,j}(k)$, after STBC decoding. The MMSE-FDE weight is given as

$$W_{c,u}^{m(u) \rightarrow u}(k) = \frac{H_{c,u}^{(m(u) \rightarrow u)^*}(k)}{\left[\begin{aligned} & \sum_{m=0}^{N_t-1} \left| H_{c,u}^{m(u) \rightarrow u}(k) \right|^2 \\ & + \sum_{m=0}^{N_t-1} \sum_{\substack{u'=0 \\ u' \neq u}}^{U-1} \left| H_{c,u'}^{m(u') \rightarrow u}(k) \right|^2 \\ & + \sum_{m=0}^{N_t-1} \sum_{\substack{c'=0 \\ c' \neq c}}^{CELL-1} \sum_{u'=0}^{U-1} \left| H_{c',u'}^{m(u') \rightarrow u}(k) \right|^2 \\ & + \sum_{l=0}^{L-1} \left| h_{c,u,l} \right|^2 (\tau_l + \Delta_{c,u}) + \frac{U_m N_t}{SF} \left(\frac{E_c}{N_0} \right)^{-1} \end{aligned} \right]}, \quad (13)$$

where U_m and E_c are the number of users who have selected the m -th transmit antenna and the transmit chip energy ($E_c = P_t T_c$), respectively, and P_t and T_c denote the transmit signal power and the chip length, respectively.

2.4 STTD downlink capacity expression

We derive an expression for the received SINR after de-spreading to compute the STTD downlink capacity (bps/Hz) by applying the Shannon capacity formula [4]. The STTD downlink capacity is computed using

$$C = R \frac{SF}{N_c} \sum_{n=0}^{N_c/SF-1} \log_2(1 + \gamma(n)) \quad (\text{bps/Hz}), \quad (14)$$

where $\gamma(n)$ is the received SINR after de-spreading at the n -th symbol. When the maximum propagation time delay difference exceeds CP length, $\gamma(n)$ can be expressed as

$$\gamma(n) = \frac{\frac{2\Gamma}{N_t} \left| \frac{1}{SF} \sum_{k=nSF}^{(n+1)SF-1} W_{c,u}^{m(u) \rightarrow u}(k) H_{c,u}^{m(u) \rightarrow u}(k) \right|^2}{X + Y + Z + \frac{1}{N_c} \sum_{k=0}^{N_c-1} \sum_{m(u)=0}^{N_t-1} \left| W_{c,u}^{m(u) \rightarrow u}(k) \right|^2}, \quad (15)$$

where $\Gamma = E_c / N_0$. X , Y and Z can be respectively expressed as

$$X = \frac{\Gamma}{N_t} \sum_{\substack{u'=0 \\ u' \neq u}}^{U-1} \left| \frac{1}{SF} \sum_{k=nSF}^{(n+1)SF-1} c_{u'}^{SF}(k) c_u^{SF}(k) W_{c,u}^{m(u) \rightarrow u}(k) H_{c,u}^{m(u') \rightarrow u'}(k) \right|^2, \quad (16a)$$

$$Y = \frac{\Gamma}{N_t} \sum_{\substack{c'=0 \\ c' \neq c}}^{CELL-1} \sum_{u'=0}^{U-1} \left| \frac{1}{SF} \sum_{k=nSF}^{(n+1)SF-1} c_{c'}^{scr}(k) c_c^{scr}(k) W_{c,u}^{m(u) \rightarrow u}(k) H_{c',u'}^{m(u') \rightarrow u'}(k) \right|^2, \quad (16b)$$

and

$$Z = \left(\frac{\Gamma}{N_t} \frac{1}{N_c} \sum_{m(u)=0}^{N_t-1} \sum_{l=0}^{L-1} \left| h_{c,u,l}^{m(u) \rightarrow u} \right|^2 \right) \times \frac{1}{SF} \sum_{k=nSF}^{(n+1)SF-1} \left| W_{c,u}^{m(u) \rightarrow u}(k) \right|^2 (l + \Delta_{c,u} - N_g) \quad (16c)$$

The first term in the denominator in (15) is the contribution of the interference from users in the cell of interest, called as multi-access interference (MUI). The second term is the contribution of co-channel interference (CCI). The third term is the contribution of IBI caused by the antenna-to-MT distance difference. It is seen from (15) that, when the maximum propagation time delay difference

exceeds CP length, the received SINR degrades due to IBI caused by the antenna-to-MT distance difference. In this paper, we discuss how IBI affects the optimum number of transmit antennas by computer simulation.

3. Computer simulation

3.1 The simulation condition

The simulation conditions are summarized in Table 2. In this paper, the number N_c of subcarrier and GI length, N_g , are respectively set to $N_c=1024$ and $N_g=128$. Long PN sequence is used as the spreading code and spreading factor SF is set to $SF=16$. The channel is assumed to be a frequency-selective channel having chip-spaced $L=16$ paths and uniform power delay profile. The propagation time delay is approximated by an integer number of chip duration. The path loss exponent, α , and shadowing loss standard deviation, δ , are assumed to be $\alpha=3.5$ and $\delta=7.0$ (dB), respectively. In this paper, we assume that the cell radius corresponds to the propagation distance of microwave during 500 chip durations. For example, when the chip rate is 100Mcps, the cell radius corresponds to 1500m. The number of users in the cell is set to $U=4$.

In this paper, we evaluate the cumulative distribution function (CDF) of the STTD downlink capacity by Monte Carlo simulation and define the $x\%$ value of CDF as the $x\%$ outage STTD downlink capacity.

Table 2 Simulation conditions.

| | |
|-----------------------------------|---|
| Number of subcarriers | $N_c=1024$ (chips) |
| Guard interval length | $N_g=128$ (chips) |
| Spreading factor | $SF=16$ |
| Spreading code | Long PN code |
| Power delay profile | $L=16$ -path uniform power delay profile |
| Fading | Frequency-selective block Rayleigh fading |
| Delay time | $\tau_l=l + \Delta$ (chips) ($l=0 \sim L-1$) |
| Path loss exponent | $\alpha=3.5$ |
| Shadowing loss standard deviation | $\delta=7.0$ (dB) |

3.2 Downlink capacity

The 10% and 90% outage STTD downlink capacities of MC-CDMA DAN are plotted in Figs. 3 and 4, respectively. For comparison, the STTD downlink capacities of MC-CDMA DAN without IBI (i.e., the IBI is removed perfectly) and of MC-CDMA CAN are also plotted in Figs. 3 and 4. It is shown from Fig. 3 that the 10% outage STTD downlink capacity improves by increasing the number of transmit antennas from $N_t=1$ to $N_t=2$. However, the 10% outage STTD downlink capacity decreases as the number of transmit antennas increases beyond $N_t=2$. The reason for this is explained as follows. Larger antenna diversity gain can be obtained by increasing N_t from 1 to 2 without reducing the STTD code rate R . Therefore, the STTD

downlink capacity increases. However, if we increase N_t beyond 2, R decreases from 1 to 3/4. As consequence, the STTD downlink capacity decreases. Therefore, the 10% outage STTD downlink capacity is maximized when $N_t=2$. It is seen from Fig. 3 is that $N_t=2$ is optimal for both CAN and DAN. However, the optimum N_t to maximize the 90% outage STTD downlink capacity is 1 for DAN while it is 2 for CAN. In CAN, the antenna diversity gain increases with increasing the number of transmit antennas because transmit antennas are co-located at the same base station. On the other hand, in DAN, when user is located near one distributed antenna, the other diversity transmit antennas are far from the user. Therefore, sufficient antenna diversity gain cannot be obtained even if N_t increases. Furthermore, there exists the IBI due to longer propagation time delay than the CP length. Therefore, the optimum N_t to maximize the 90% outage STTD downlink capacity becomes $N_t=1$ for DAN unlike CAN.

The transmit antennas election criterion affects the STTD downlink capacity. It is seen from Figs.3 and 4 that, in DAN, the instantaneous received power based transmit antenna selection achieves higher STTD downlink capacity than the distance based transmit antenna selection when $N_t=2$. However, when $N_t>2$, the latter selection achieves higher STTD downlink capacity than the former. The reason for this can be explained as follows. When using the instantaneous received power based transmit antenna selection, the user sometimes selects far distributed transmit antennas due to the shadowing and frequency-selective fading. Accordingly, IBI becomes stronger compared to the case when using the distance based transmit antenna selection. When the number of transmit antennas is small, IBI is weaker than both MUI and CCI. Therefore, the instantaneous received power based transmit antenna selection achieves higher STTD downlink capacity than the distance based transmit antenna selection. On the other hand, when the number of transmit antennas is large, IBI is stronger than both MUI and CCI. Therefore, the distance based transmit antenna selection achieves higher STTD downlink capacity than the instantaneous received power based transmit antenna selection.

It can be clearly understood by comparing the STTD downlink capacities of MC-CDMA DAN and MC-CDMA DAN without IBI in Figs. 3 and 4 that IBI reduces the STTD downlink capacity. When $N_t=2$ and the instantaneous received signal power based transmit antenna selection is used, the presence of IBI reduces the 10% (90%) outage STTD downlink capacity by about 8% (10%). Therefore, some techniques are necessary to suppress IBI caused by the antenna-to-MT distance difference. However, it can be said that DAN can still improve the STTD downlink capacity compared to CAN even if IBI caused by the antenna-to-MT distance difference exists. When $N_t=2$ and the instantaneous received signal power based transmit antenna selection is used, DAN can achieve about 2.3 (1.9) times higher 10%

(90%) outage STTD downlink capacity than CAN.

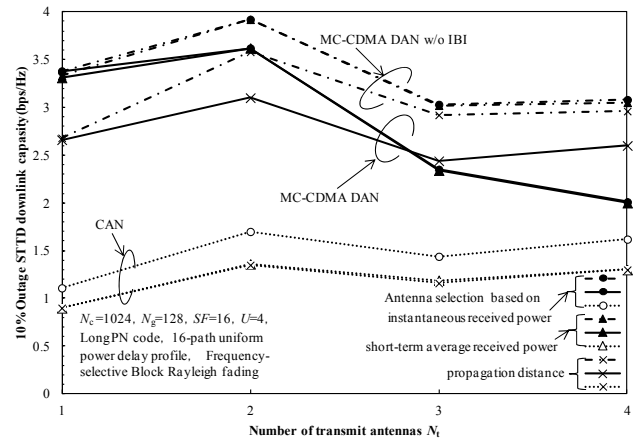


Fig. 3 10% outage STTD downlink capacity.

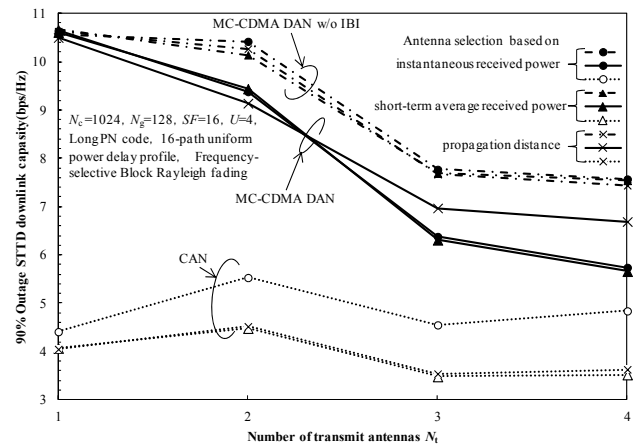


Fig. 4 90% outage STTD downlink capacity.

4. Conclusion

In this paper, we derived the received SINR after de-spreading, and evaluated the STTD downlink capacity to discuss about the optimum number of transmit antennas to be involved in STTD downlink of MC-CDMA DAN. It was found that the optimum number of transmit antennas for the 10% outage STTD downlink capacity is $N_t=2$ for both CAN and DAN. On the other hand, the optimum number of transmit antennas for the 90% outage STTD downlink capacity is $N_t=2$ for CAN while it is $N_t=1$ for DAN. It was also found that the presence of IBI reduces the 10% (90%) outage STTD downlink capacity by about 8% (10%). However, DAN still achieves higher STTD downlink capacity than CAN even in the presence of the IBI caused by the antenna-to-MT distance difference longer than the CP length. Suppressing the IBI is left as our future study.

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