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Uplink Capacity of A Cellular System Using Multi-user Single-carrier MIMO Multiplexing Combined with Frequency-domain Equalization and Transmit Power Control

Takahiro Chiba · Kazuaki Takeda · Kazuki Takeda · Fumiyuki Adachi

Abstract Multi-user single-carrier multiple-input multiple-output (MU SC-MIMO) multiplexing can increase the uplink capacity of a cellular system without expanding the signal bandwidth. It is practically important to make clear an extent to which the MU SC-MIMO multiplexing combined with frequency-domain equalization (FDE) and transmit power control (TPC) can increase the uplink capacity in the presence of the co-channel interference (CCI). Since the theoretical analysis is quite difficult, we resort to the computer simulation to investigate the uplink capacity. In this paper, frequency-domain zero-forcing detection (ZFD) and frequency-domain minimum mean square error detection (MMSED) are considered for MU signal detection. It is shown that ZFD and MMSED provide almost the same uplink capacity and that an advantage of fast TPC over slow TPC diminishes. As a result, MU SC-MIMO using computationally efficient ZFD can be used together with slow TPC instead of using MMSED. With 8 receive antennas and slow TPC, MU SC-MIMO multiplexing using ZFD can achieve about 1.5 times higher uplink capacity than SU SC-SIMO diversity.

Keywords Multi-user MIMO · Single-carrier · Frequency-domain equalization · Cellular system · Uplink-capacity

1 Introduction

Broadband data services are demanded in next-generation mobile communication systems [1]. Broadband wireless channel consists of many propagation paths having different time delays, resulting in frequency-selective fading [2]. Recently, broadband single-carrier (SC) transmission has been attracting a lot of attention because of its low peak-to-average power ratio (PAPR) property and because the frequency diversity gain can be obtained by the use of frequency-domain equalization (FDE) [3,4]. In cellular systems, the same carrier frequency

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is reused in spatially separated different cells to efficiently utilize the limited bandwidth [5]. However, only reusing the same carrier frequency is not sufficient for offering broadband services in cellular systems with the limited bandwidth.

To increase the uplink capacity with the limited bandwidth, multi-user multiple-input multiple-output (MU MIMO) multiplexing [6–9] can be used. MU SC-MIMO multiplexing is promising for the uplink applications owing to its low PAPR property [10,11]. Several signal detection schemes exist combined with FDE for MU SC-MIMO multiplexing, e.g., frequency-domain zero-forcing detection (ZFD) [12] and frequency-domain minimum mean square error detection (MMSED) [13]. In a cellular system, a user who is located close to the cell edge suffers from large power loss due to shadowing loss and path loss as well as fading [14]. To eliminate the power loss problem, transmit power control (TPC) can be used [15]; however, in return, this increases the co-channel interference (CCI) to other cells.

Many studies on the information theoretic capacity and the sum capacity achievable by MU SC-MIMO multiplexing in a cellular system can be found in the literature, e.g., [16–19]. However, to the best of authors' knowledge, it is not clear to which extent the MU SC-MIMO multiplexing combined with FDE and TPC can increase the uplink capacity in a cellular system.

In this paper, since the theoretical analysis is quite difficult, we investigate by the computer simulation the uplink capacity of a broadband cellular system using MU SC-MIMO multiplexing combined with FDE and TPC. The remainder of this paper is organized as follows. Section 2 presents the MU SC-MIMO cellular system model. MU SC-MIMO multiplexing combined with FDE and TPC is described in Sect. 3. Frequency-domain ZFD and frequency-domain MMSED are considered as the signal detection. In Sect. 4, the uplink capacity of MU SC-MIMO multiplexing is discussed. Section 5 concludes this paper.

2 MU SC-MIMO Cellular System Model

In a cellular system, the available channels are divided into a number of channel groups and each channel group is reused in different cells [5]. The number N of different channel groups is called the cluster size. The smaller the cluster size becomes, the more channels can be assigned to each cell. In this paper, it is assumed that U users in each cell are simultaneously transmitting their data by using the same carrier frequency. Figure 1 illustrates the uplink CCI model for MU SC-MIMO multiplexing. In this paper, only 6 nearest co-channel cells ($c = 1 \sim 6$) surrounding the desired cell ($c = 0$) are considered since they are the predominant cause of CCI which degrades the uplink capacity. The CCI power becomes weaker by

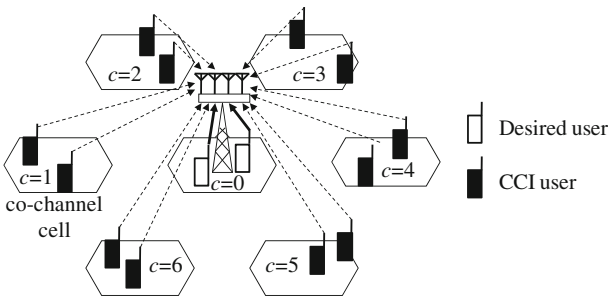
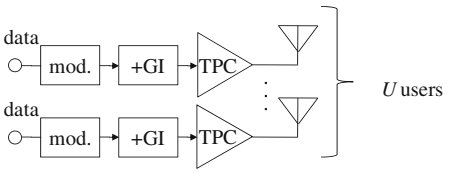
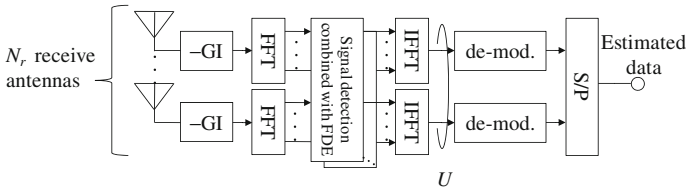


Fig. 1 CCI model



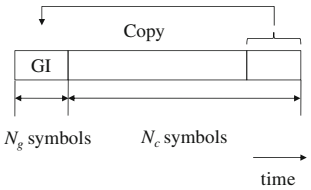
(a) Mobile station transmitter



(b) Base station receiver

Fig. 2 Uplink transmission system model

Fig. 3 Transmit symbol block



increasing the distance between the co-channel cells or the cluster size N; however, in return, this reduces the number of channels in one channel group which can be assigned to each cell, resulting in the reduced uplink capacity.

In this paper, the normalized transmit power and distance are introduced. By denoting the transmit power by P, the received signal power $P(r)$ (assuming no shadowing loss and fading) for a user who is located at the distance r from its communicating base station (BS) is given by $P(r) = P \times r^{-\alpha} = (P \cdot R^{-\alpha}) \times (r/R)^{-\alpha}$, where α is the distance-dependent path loss exponent and R is the cell radius. When the TPC with its target P_{tpc} is used and ignoring the shadowing loss and fading, the transmit signal power is given by $P(r) = P_{\text{tpc}} \times r^{\alpha} = (P_{\text{tpc}} \cdot R^{\alpha}) \times (r/R)^{\alpha}$. For the simplicity purpose in this paper, $P \cdot R^{-\alpha}$, $P_{\text{tpc}} \cdot R^{\alpha}$, and r/R are simply represented by P, P_{tpc}, and r $(0 < r \leq 1)$, respectively. P, P_{tpc}, and r are called the normalized transmit power, the normalized TPC target, and the normalized distance, respectively.

The uplink transmission in the desired cell ($c = 0$) is considered. Figure 2 illustrates the uplink transmission system model of MU SC-MIMO multiplexing with FDE. It is assumed that each user has the single transmit antenna and a base station (BS) of each cell has N_r $(\geq U)$ receive antennas. At a mobile transmitter, each user's binary data sequence is data-modulated and divided into a sequence of blocks of N_c symbols each. The last N_g symbols in each block are copied and inserted as a cyclic prefix into the guard interval (GI), placed at the beginning of each block, to form a block of $N_g + N_c$ symbols (see Fig. 3) [3].

The transmit timing offset plus maximum propagation time delay among U users of the $c = 0$-th cell is assumed to be less than the GI length. It should be noted that the CCIs from 6 co-channel cells ($c = 1 \sim 6$) are time-asynchronous to the signals of the desired $c = 0$-th cell. At the BS receiver of the $c = 0$-th cell, a superposition of U users' transmitted signals

corrupted by the CCI is received by N_r receive antennas. After the removal of the GI from the received signal, N_c-point fast Fourier transform (FFT) is applied to transform the GI-removed received signal into the frequency-domain signal. After performing signal detection combined with FDE, each user's frequency-domain signals are transformed back into time-domain signal block by using N_c-point inverse FFT (IFFT) and then, data demodulation is done.

3 Signal Detection Combined with FDE and TPC

The uplink capacity depends on the propagation environment. There have been many works on the realistic propagation modeling of MIMO systems [20,21]. However, the realistic propagation model is too complex to draw a general conclusion. In this paper, we use the following simple and popular channel model [16,17,22,23]:

(a) Assuming non line of sight (NLoS) channel between each user and BS, the channel is characterized by the distance-dependent path-loss, fading loss, and frequency-selective Rayleigh fading.
(b) The shadowing loss associated with a different user is uncorrelated. The frequency-selective Rayleigh fading associated with a different user is also uncorrelated.
(c) Assuming broadband transmissions, the variations in the path-loss, shadowing loss and fading stay constant during the transmission of one signal block of $N_g + N_c$ symbols, but, fading changes block by block due to the user mobility.

Without loss of generality, a transmission of one block of N_c symbols from each user is considered. Throughout the paper, symbol-spaced discrete-time signal representation is used. First, we give the received signal representation and then, the MU SC-MIMO signal detection combined with FDE and TPC is presented. In the following, for the sake of simplicity, we omit the insertion and removal of the GI.

3.1 Received Signal

The t-th ($t = 0 \sim N_c - 1$) transmit signal from u-th user in the c-th cell is denoted by $\sqrt{2P_{0(c)}}s_{0(c)}(t)$, where $s_{u(c)}(t)$ and $P_{u(c)}$are the data-modulated symbol and the transmit power, respectively. The transmit signals from U users are represented using the vector form as $\mathbf{s}_c(t)=\left[\sqrt{2P_{0(c)}}s_{0(c)}(t), \sqrt{2P_{1(c)}}s_{1(c)}(t), \ldots, \sqrt{2P_{u(c)}}s_{u(c)}(t), \ldots, \sqrt{2P_{U-1(c)}}s_{U-1(c)}(t)\right]^T$, where $(\cdot)^T$ denotes the transposition. Each user's transmitted signal passes through a different frequency-selective fading channel. The channel impulse response between the u-th user in the c-th cell and the m-th receive antenna of the BS of the $c = 0$-th cell is given by

$$\tilde{h}_{m(0),u(c)}(t) = \sqrt{r_{0,u(c)}^{-\alpha} \cdot 10^{-\frac{\eta_{0,u(c)}}{10}}} \sum_{l=0}^{L-1} h_{m(0),u(c)}^{(l)} \cdot \delta(t - \tau_{u(c),l}), \tag{1}$$

where $r_{0,u(c)}$, $\eta_{0,u(c)}$, and α denote the normalized distance between the u-th user in the c-th cell and the BS of the $c = 0$-th cell, the shadowing loss in dB, and the path loss exponent, respectively. In Eq. (1), the variable $l(= 0 \sim L - 1)$ represents the path index and $h_{m(0),u(c)}^{(l)}$ and $\tau_{u(c),l}$ are respectively the complex-valued path gain with $\sum_{l=0}^{L-1} E\left[\left|h_{m(0),u(c)}^{(l)}\right|^2\right]=1$ ($E[\cdot]$ denotes the ensemble average operation) and the time delay between the u-th user in

the c-th cell and the BS of the $c = 0$-th cell. Using the matrix representation, the received signal vector of the $c = 0$-th cell can be expressed as

$$\begin{aligned}\mathbf{y}(t) &= \left[y_{0(0)}(t), \ldots, y_{m(0)}(t), \ldots, y_{N_r-1(0)}(t)\right]^T \\ &= \sum_{l=0}^{L-1} \tilde{\mathbf{h}}_0^{(l)} \cdot \mathbf{s}_0\left((t-\tau_{u(0),l}) \bmod N_c\right) + \sum_{c=1}^{6}\sum_{l=0}^{L-1} \tilde{\mathbf{h}}_c^{(l)} \cdot \mathbf{i}_c\left(t-\tau_{u(c),l}\right) + \mathbf{n}(t), \quad (2)\end{aligned}$$

where $y_{m(0)}(t),\ m = 0 \sim N_r - 1,\ t = 0 \sim N_c - 1$, represents the t-th received signal of the m-th receive antenna at the BS of the $c = 0$-th cell and the first and second terms are the desired signal and the CCI, respectively. $\tilde{\mathbf{h}}_c^{(l)}$ is the path gain matrix of size $N_r \times U$ of the l-th path, the $(m(0), u(c))$-th element of which is given by $\tilde{h}_{m(0),u(c)}^{(l)} = \sqrt{r_{0,u(c)}^{-\alpha} \cdot 10^{-\frac{\eta_{0,u(c)}}{10}}} h_{m(0),u(c)}^{(l)}$. $\mathbf{i}_c(t) = \left[\sqrt{2P_{0(c)}} i_{0(c)}(t), \ldots, \sqrt{2P_{u(c)}} i_{u(c)}(t), \ldots, \sqrt{2P_{U-1(c)}} i_{U-1(c)}(t)\right]^T$ is the CCI vector of size $U \times 1$, which represents the CCI from U users in the c-th cell, and $\mathbf{n}(t) = \left[n_{0(0)}(t), \ldots, n_{m(0)}(t), \ldots, n_{N_r-1(0)}(t)\right]^T$ is the noise vector of size $N_r \times 1$, where $\left\{n_{m(0)}(t);\ m(0) = 0 \sim N_r - 1\right\}$ are the independent zero-mean complex Gaussian variables with variance $2\sigma_n^2$ due to additive white Gaussian noise (AWGN).

The received signal vector $\mathbf{y}_0(t)$ is transformed into the frequency-domain signal vector $\mathbf{Y}_0(k)$ by using an N_c-point FFT. $\mathbf{Y}_0(k)$ can be expressed as

$$\mathbf{Y}_0(k) = \sum_{t=0}^{N_c-1} \mathbf{y}_0(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) = \tilde{\mathbf{H}}_0(k)\mathbf{S}_0(k) + \mathbf{I}_0(k) + \mathbf{\Pi}_0(k), \quad (3)$$

where $\tilde{\mathbf{H}}_c(k) = \left[\tilde{\mathbf{H}}_{0(c)}(k), \ldots, \tilde{\mathbf{H}}_{u(c)}(k), \ldots, \tilde{\mathbf{H}}_{U-1(c)}(k)\right]$ with $\tilde{\mathbf{H}}_{u(c)}(k) = \left[\tilde{H}_{0(0),u(c)}(k), \ldots, \tilde{H}_{m(0),u(c)}(k), \ldots, \tilde{H}_{N_r-1(0),u(c)}(k)\right]^T$, $\mathbf{S}_0(k) = [S_{0(0)}(k), \ldots, S_{u(0)}(k), \ldots, S_{U-1(0)}(k)]^T$, $\mathbf{I}_0(k) = [I_{0(0)}(k), \ldots, I_{m(0)}(k), \ldots, I_{N_r-1(0)}(k)]^T$, and $\Pi_0(k) = \left[\Pi_{0(0)}(k), \ldots, \Pi_{m(0)}(k), \ldots, \Pi_{N_r-1(0)}(k)\right]^T$ denote the channel gain matrix of size $N_r \times U$, the frequency-domain signal vector of size $U \times 1$, the CCI vector of size $N_r \times 1$, and the noise vector of size $N_r \times 1$, respectively. The $m(0)$-th element $\tilde{H}_{m(0),u(c)}(k)$ of $\tilde{\mathbf{H}}_{u(c)}(k)$ is given by

$$\tilde{H}_{m(0),u(c)}(k) = \sum_{l=0}^{L-1} \tilde{h}_{m(0),u(c)}^{(l)} \exp\left(-j2\pi k \frac{\tau_{u(c),l}}{N_c}\right) = \sqrt{r_{u(c)}^{-\alpha} \cdot 10^{-\frac{\eta_{0,u(c)}}{10}}} \cdot H_{m(0),u(c)}(k), \quad (4)$$

where $H_{m(0),u(c)}(k)$ denotes the channel gain between the u-th user in the c-th cell and the m-th receive antenna of the $c = 0$-th cell BS. $S_{u(0)}(k)$, $I_{m(0)}(k)$, and $\Pi_{m(0)}(k)$ are given as

$$\begin{cases} S_{u(0)}(k) = \sqrt{2P_{u(0)}} \sum_{t=0}^{N_c-1} s_{u(0)}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ I_{m(0)}(k) = \sum_{c=1}^{6} \sqrt{2P_{u(c)}} \sum_{t=0}^{N_c-1} \sum_{u(c)=0}^{U-1} \sum_{l=0}^{L-1} \tilde{h}_{m(0),u(c)}^{(l)} \cdot i_{u(c)}\left(t-\tau_{u(c),l}\right) \exp\left(-j2\pi k \frac{t}{N_c}\right). \quad (5) \\ \Pi_{m(0)}(k) = \sum_{t=0}^{N_c-1} n_{m(0)}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{cases}$$

3.2 Signal Detection

ZFD and MMSED [12,13] are modified to be used for the MU SC-MIMO signal detection in the frequency-domain. The detector output vector is expressed using the matrix form as

$$\tilde{\mathbf{S}}_0(k) = \left[\tilde{S}_{0(0)}(k), \ldots, \tilde{S}_{u(0)}(k), \ldots, \tilde{S}_{U-1(0)}(k)\right]^T = \mathbf{W}_0(k)\mathbf{Y}_0(k) \text{ for } k = 0 \sim N_c - 1, \tag{6}$$

where $\mathbf{W}_0(k) = \left[\mathbf{w}_{0(0)}(k), \ldots, \mathbf{w}_{m(0)}(k), \ldots, \mathbf{w}_{N_r-1(0)}(k)\right]$ represents the weight matrix of size $U \times N_r$ for the signal detection combined with FDE. ZFD perfectly restores the frequency-nonselective channel if the perfect knowledge of channel state information is available, while MMSED minimizes the mean square error (MSE) between the transmitted signal and the equalizer output and thus, partially restores the frequency-nonselective channel. The weight matrix $\mathbf{W}_0(k)$ is given by

$$\mathbf{W}_0(k) = \begin{cases} \left(\tilde{\mathbf{H}}_0^H(k)\tilde{\mathbf{H}}_0(k)\right)^{-1}\tilde{\mathbf{H}}_0^H(k) & \text{for ZFD} \\ \tilde{\mathbf{H}}_0^H(k)\left(\tilde{\mathbf{H}}_0(k)\tilde{\mathbf{H}}_0^H(k) + \frac{\sigma_I^2+\sigma_n^2}{P}\mathbf{I}_{N_r}\right)^{-1} & \text{for MMSE} \end{cases}, \tag{7}$$

where $(\cdot)^H$, σ_I^2, and σ_n^2 denote the Hermitian transpose, the average received interference power, and the noise power, respectively. P is the normalized transmit signal power which is same for all users in the case of no TPC, while it is the normalized TPC target P_{tpc} in the case of TPC. The CCI power σ_I^2 and the noise power σ_n^2 are given by

$$\begin{cases} \sigma_I^2 = \frac{1}{2}E\left[\left|I_{m(0)}(k)\right|^2\right] \\ \sigma_n^2 = \frac{1}{2}E\left[\left|\Pi_{m(0)}(k)\right|^2\right] \end{cases} \text{for } m\,(0) = 0 \sim N_r - 1 \tag{8}$$

which must be estimated, but in this paper, they are assumed to be perfectly known at the receiver.

The detector output signal block $\left\{\tilde{S}_{u(0)}(k); k = 0 \sim N_c - 1\right\}$ of the $u(0)$-th user in the $c = 0$-th cell is transformed by using an N_c-point IFFT to the time-domain signal block $\left\{\tilde{s}_{u(0)}(t); t = 0 \sim N_c - 1\right\}$ with

$$\tilde{s}_{u(0)}(t) = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\tilde{S}_{u(0)}(k)\exp\left(j2\pi t\frac{k}{N_c}\right), \tag{9}$$

which is the decision variable associated with the t-th symbol transmitted from the $u(0)$-th user. The symbol decision is done as

$$\hat{s}_{u(0)}(t) = \arg\min_{s_{u(0)}\in D}\left|\tilde{s}_{u(0)}(t) - s_{u(0)}(t)\right|, \tag{10}$$

where D is the set of candidate symbols.

3.3 Transmit Power Control

The received signal power varies according to the distance-dependent path loss, shadowing loss, and fading. The signals transmitted from users close to the cell edge are weak and therefore, the bit error rate (BER) performance may degrade. To remedy this performance degradation, TPC can be used. However, in return, the use of TPC may increase the CCI to

Table 1 Simulation condition

Transmitter	Data modulation		QPSK
	Number of users per cell		$U = 1 \sim 6$
	FFT block size		$N_c = 256$
	Transmit power control	Type	Slow and fast
		Target E_b/N_0	∞
Channel	Path-loss exponent		$\alpha = 3.5$
	Standard deviation of shadowing loss		$\sigma = 7\,\text{dB}$
	Fading	Type	Block Rayleigh fading ($L = 16$)
		Power delay profile	Uniform
		Time delays	$\tau_l = lT_s, l = 0 \sim L-1$
Receiver	Number of receive antennas		$N_r = 4, 6, 8$
	Signal detection		ZFD and MMSED
	Channel estimation		Ideal
Required quality	Required BER		$\text{BER}_{\text{req}} = 10^{-3}$
	Allowable outage probability		$Q = 0.1$

other cells. To which extent the MU SC-MIMO multiplexing combined with FDE and TPC can increase the uplink capacity is a practically important problem.

We consider the slow TPC and fast TPC. Slow TPC keeps the local average received signal power (averaged over the fading statistics) always at the prescribed target value, while fast TPC keeps the instantaneous received signal power always at the prescribed target value. The transmit signal power $P_{u(c)}$ of the u-th user in the c-th cell is given as

$$P_{u(c)} = \begin{cases} \dfrac{P_{\text{tpc}}}{r_{c,u(c)}^{-\alpha} \cdot 10^{-\frac{\eta_{c,u(c)}}{10}}} & \text{for slow TPC} \\ \dfrac{P_{\text{tpc}}}{r_{c,u(c)}^{-\alpha} \cdot 10^{-\frac{\eta_{c,u(c)}}{10}} \cdot \frac{1}{N_c \cdot N_r} \sum_{k=0}^{N_c-1} \sum_{m=0}^{N_r-1} \left| H_{m(c),u(c)}(k) \right|^2} & \text{for fast TPC} \end{cases}, \tag{11}$$

where P_{tpc}, $r_{c,u(c)}^{-\alpha}$, and $\eta_{c,u(c)}$ represent the normalized TPC target, the distance-dependent path loss, and the shadowing loss in dB between the u-th user and its communicating BS of the c-th cell, respectively.

4 Simulation Results

4.1 Simulation Procedure

Table 1 summarizes the simulation condition. The channel is assumed to be a frequency-selective block Rayleigh fading having a symbol-spaced $L = 16$-path uniform power delay profile, i.e., $E\left[\left|h_{m(0),u(c)}^{(l)}\right|^2\right] = 1/L$ for all l. It is assumed that the path loss exponent is $\alpha = 3.5$ and the shadowing loss follows the log-normal distribution with standard deviation $\sigma = 7$dB. In this paper, we consider the interference-limited channel condition (i.e., $\sigma_n^2 \to 0$).

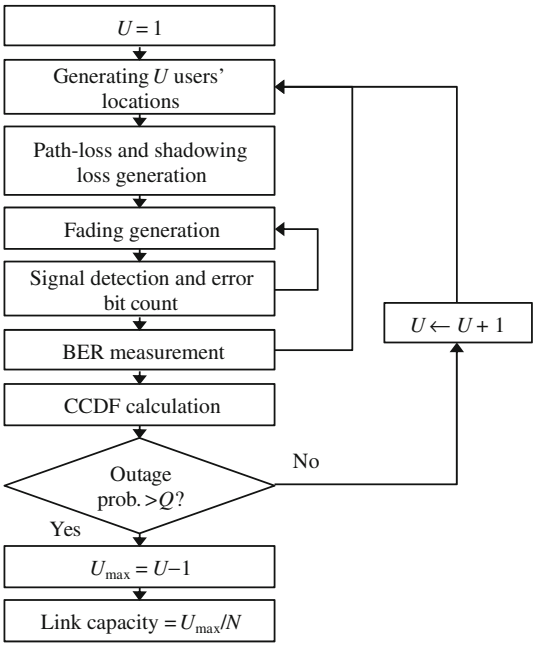


Fig. 4 Computer simulation procedure

The computer simulation procedure is illustrated in Fig. 4. First, U users' locations (starting from $U = 1$) are randomly generated in each cell. The path loss and the shadowing loss are generated for each user. Then, $L = 16$ path gains associated with each user are generated. The local average BERs for U users in the $c = 0$-th cell are measured. As stated in Sect. 3, we are assuming block fading in which fading stay constant during the transmission of one signal block of $N_g + N_c$ symbols, but, changes block by block due to the user mobility. In the computer simulation, first we measure the local average BER which is the BER averaged over the fading statistics. The local average BER changes according to the change in user location. The local average BER measurement is repeated a sufficient number of times by changing the user locations to find the complementary cumulative distribution function (CCDF) of the local average BER. The outage probability is the probability of the local average BER exceeding the required BER. If the outage probability is less than the allowable outage probability, the number U of users is incremented by one. We introduce the normalized uplink capacity defined as the maximum number $U_{\max}$ of supportable users normalized by the cluster size N. In this paper, the required BER, $\mathrm{BER}_{\mathrm{req}}$, and the allowable outage probability Q are set as $\mathrm{BER}_{\mathrm{req}} = 10^{-3}$ and $Q = 0.1$, respectively.

As the benchmark comparison, we also evaluate the BER performance achievable using single-user (SU) SC-SIMO diversity (using MMSE-FDE) [24]. The output vector of the SU SC-SIMO diversity combiner, corresponding to Eq. (6), is given as

$$\tilde{S}_{u(0)}(k) = \mathbf{Y}_0^T(k) \cdot \mathbf{w}_{u(0)}(k), \tag{12}$$

where $\mathbf{w}_{u(0)}(k)$ is the antenna weight vector given as

$$\mathbf{w}_{u(0)}(k) = \frac{\tilde{\mathbf{H}}^*_{u(0)}(k)}{\left\|\tilde{\mathbf{H}}_{u(0)}(k)\right\|^2 + \frac{\sigma_I^2 + \sigma_n^2}{P}} \tag{13}$$

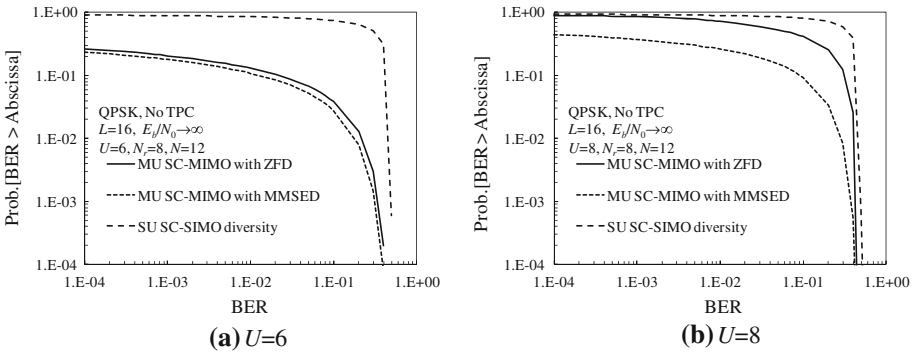


Fig. 5 CCDFs of local average BER.

with $||\cdot||$ representing the vector norm operation and P is the same as in Eq. (7). Note again that we are assuming the interference-limited condition, $\sigma_n^2 \to 0$ in Eq. (13).

4.2 BER Distribution

Figure 5 illustrates the CCDF of the local average BER for the case of $N_r = 8$ (the number of receive antennas), $L = 16$ (the number of paths), and $N = 12$ (the cluster size). No TPC is assumed. ZF has a problem of noise plus CCI enhancement resulting from the restoration of the channel frequency-nonselectivity. However, it can be seen from Fig. 5a that when $U = 6$ (the number of users), ZFD provides almost the same outage probability as MMSED. The reason for this is explained below. The received SINR in a Rayleigh fading channel is chi-square distributed with the degree of freedom of $2(N_r - U + 1)$ for both ZFD and MMSED [25,26] and hence, both ZFD and MMSED achieve the same $(N_r - U + 1)$-th diversity order [27]. When U is smaller than N_r, diversity order of higher than one is achieved. In this case, ZF can mitigate the CCI enhancement problem and hence, ZF approaches MMSED. On the other hand, when $U = N_r = 8$ (see Fig. 5b), both ZFD and MMSED have a diversity order of one. In this case, ZF suffers from the CCI enhancement problem and therefore, ZF gives the higher outage probability than MMSED. On the other hand, SU SC-SIMO diversity is the single-user detection and hence, its BER performance significantly degrades due to large intra-cell interference in addition to the CCI from surrounding co-channel cells. This leads to always higher outage probability than those of ZFD and MMSED.

Below, we discuss how much MU SC-MIMO multiplexing can increase the normalized uplink capacity (defined by $U_{\max}/N$) compared to SU SC-SIMO diversity. We also discuss the impacts of the number N_r of receive antennas and TPC.

4.3 Uplink Capacity

Figure 6 plots the normalized uplink capacity as a function of the cluster size N. No TPC is assumed. SU SC-SIMO diversity can always accommodate single user only ($U_{\max} = 1$). It can be seen from Fig. 6a that when $N_r = 4$, the normalized uplink capacity can be maximized at $N = 12$ for MU SC-MIMO by using either ZFD or MMSED and it is $U_{\max}/N = 0.166$. On the other hand, the normalized uplink capacity is maximized at $N = 4$ for SU SC-SIMO diversity and it is $U_{\max}/N = 0.143$ which is only slightly lower than MU SC-MIMO. When N_r is increased to 8 (see Fig. 6b), the normalized uplink capacity is maximized at $N = 9$ for MU SC-MIMO multiplexing by using either ZFD or MMSED and it is $U_{\max}/N = 0.333$

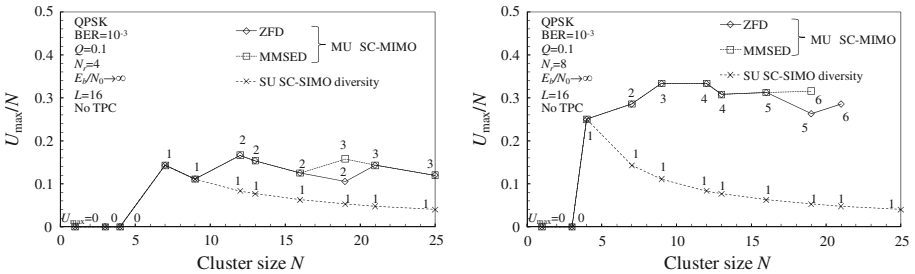


Fig. 6 Normalized uplink capacity

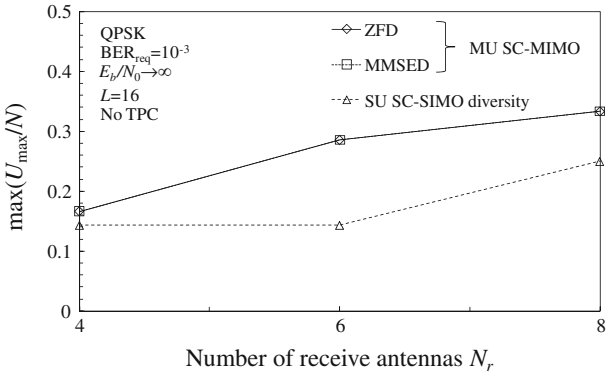


Fig. 7 Impact of N_r

which is about 1.3 times higher than that of SU SC-SIMO diversity (the maximum value of $U_{\max}/N$ is 0.25 at $N = 4$).

There is a trade-off relationship between the number of users and the diversity order (remember that the diversity order of MU SC-MIMO multiplexing is given by $N_r - U + 1$). When $U = N_r$, both ZFD and MMSED have the diversity order of one and therefore, the cluster size N must be large enough to sufficiently reduce the CCI, leading to the decreased uplink capacity. The maximum normalized uplink capacity is obtained from Fig. 6 which plots $U_{\max}/N$ curves as a function of N. Figure 7 illustrates the impact of N_r (the number of receive antennas) on the normalized uplink capacity of MU SC-MIMO multiplexing. Increasing N_r is quite effective to increase the uplink capacity; MU SC-MIMO multiplexing can increase the uplink capacity almost two times by increasing N_r from 4 to 8. ZFD and MMSED can achieve almost the same uplink capacity irrespective of N_r. This is because, as discussed in Sect. 4.2, both ZFD and MMSED can achieve the same $N_r - U + 1$-th diversity order as far as $U < N_r$.

4.4 Impact of TPC

Figure 8 compares the effects of slow TPC and fast TPC. Both TPC provide almost the same maximum uplink capacity. This is because the block averaged received signal power is given by $\frac{1}{N_c \cdot N_r} \sum_{k=0}^{N_c-1} \sum_{m=0}^{N_r-1} \left| H_{m(c),u(c)}(k) \right|^2$ and its variation is very shallow in a strong frequency-selective channel (since $L = 16$). This can be understood from Fig. 9 which

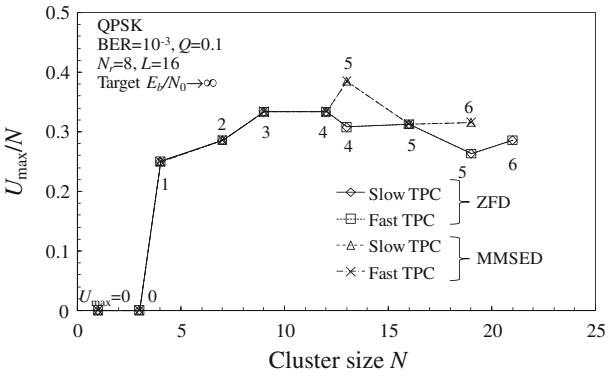


Fig. 8 Comparison of slow TPC and fast TPC

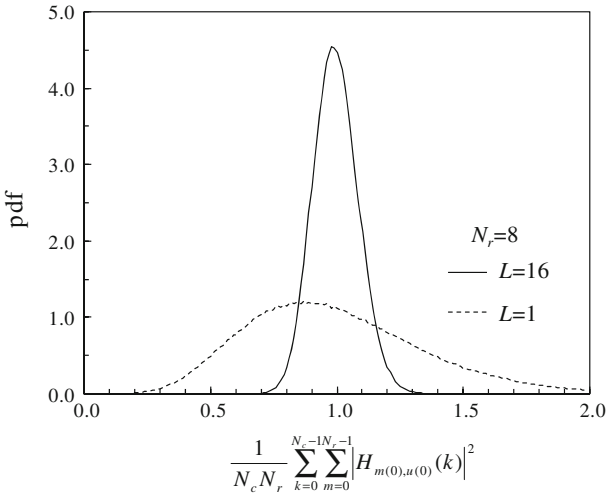


Fig. 9 Pdf of block averaged received signal power

illustrates the probability distribution function (pdf) of the block averaged received signal power. An advantage of fast TPC diminishes and slow TPC can be used.

Figure 10 compares the normalized uplink capacities of MU SC-MIMO multiplexing and SU SC-SIMO diversity both using slow TPC. MU SC-MIMO multiplexing using ZF can achieve almost the same maximum uplink capacity as that using MMSED which is $U_{\max}/N = 0.385$ at $N = 13$ and is about 1.5 times higher than SU SC-SIMO diversity ($U_{\max}/N = 0.25$ at $N = 4$).

5 Conclusions

In this paper, we investigated the uplink capacity of MU SC-MIMO multiplexing combined with FDE and TPC in a broadband cellular system. ZFD and MMSED were considered as the signal detection method. The effects of slow and fast TPC were compared to show that ZFD and MMSED provide almost the same uplink capacity in a strong frequency-selective channel. An advantage of fast TPC diminishes and slow TPC can be used since the block averaged

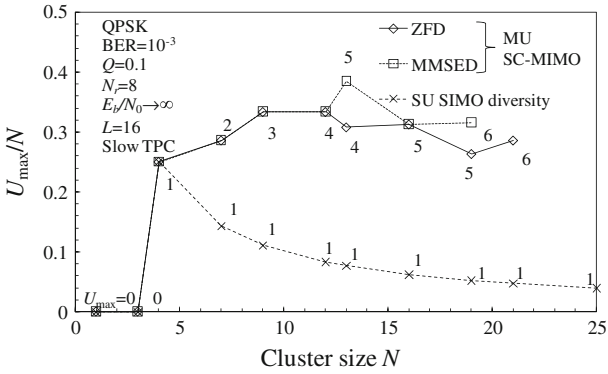


Fig. 10 Normalized uplink capacity when slow TPC is used

received signal power variation is very shallow in a strong frequency-selective channel. This suggests that in a broadband cellular system, MU SC-MIMO multiplexing using computationally efficient ZFD can be used together with slow TPC. With $N_r = 8$ and slow TPC, MU SC-MIMO multiplexing using ZFD can achieve about 1.5 times higher uplink capacity than SU SC-SIMO diversity.

In the future broadband wireless systems, the cell size will be reduced to solve the transmit power problem and hence, the propagation channel may become a line of sight (LoS) channel similar to the indoor environment [28]. The investigation of uplink capacity in such a LoS environment is quite important. This is left as an important future study.

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