Uplink Capacity of Single-carrier Frequency -interleaved Spread Spectrum Multiple Access

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***Abstract*—A single-carrier frequency-interleaved spread spectrum multiple access (SC-FI-SSMA), which we recently proposed, uses frequency-domain interleaving and MMSE-FDE to remove the uplink multi-user interference (MUI) while achieving the frequency diversity gain by assigning the orthogonal interleaving matrices to different users. However, in a multi-cell environment, the uplink capacity of SC-FI-SSMA may degrade due to the inter-cell interference. To improve the uplink bit error rate (BER) performance, transmit power control and site selection diversity techniques can be applied. In this paper, we evaluate the uplink capacity of SC-FI-SSMA to show that SC-FI-SSMA provides better uplink performance than DS-CDMA even in a multi-cell environment.**

***Index Terms*—Frequency-domain interleaving, MMSE frequency-domain equalization, SC-FI-SSMA**

I. INTRODUCTION

With the growing market of mobile wireless communications, broadband data services are demanded. A broadband wireless channel is composed of many propagation paths having different time delays, resulting in a frequency-selective fading channel [1]. For data transmissions of a few Mbps, direct sequence code division multiple access (DS-CDMA) exploits the channel frequency-selectivity by using the coherent rake combining [2], [3]. However, for much higher speed data transmissions than the present systems, the wireless channel becomes severely frequency-selective and the BER performance with rake combining degrades due to a strong inter-path interference.

FDE based on the minimum mean square error (MMSE) criterion can provide the much improved bit error rate (BER) performance of DS-CDMA downlink [4]-[7], compared with the conventional rake combining. However, the uplink BER performance significantly deteriorates due to strong multi-user interference (MUI) [8]. Recently, interleaved frequency-division multiple-access (IFDMA) or SC-FDMA was proposed [9]. IFDMA is a spread spectrum multicarrier transmission combined with FDMA and can avoid the MUI while obtaining the frequency diversity gain. In [10], chip repetition DS-CDMA was proposed that uses comb-like spectrum to avoid the spectrum overlapping among different users. Recently, we proposed single-carrier frequency-interleaved spread spectrum multiple access (SC-FI-SSMA) that uses frequency-domain interleaving and MMSE-FDE [11].

In SC-FI-SSMA, by assigning the orthogonal interleaving matrices to different users, the MUI is completely removed and better BER performance can be achieved than the conventional DS-CDMA with MMSE-FDE. Note that SC-FI-SSMA using equal space interleaving pattern is equivalent to the chip repetition DS-CDMA, while nonspread SC-FI-SSMA is corresponding to IFDMA.

So far, we have considered the single-cell environment only. In a multi-cell environment, the presence of inter-cell interference coming from other cells may degrade the uplink performance of SC-FI-SSMA. To improve the uplink capacity, transmit power control (TPC) [2] and site selection diversity (SSD) [12] can be introduced. In this paper, we evaluate, by computer simulation, the uplink capacity of SC-FI-SSMA and compare with that of DS-CDMA with MMSE-FDE.

II. TRANSMISSION SYSTEM MODEL

A. Transmitted signal representation

Fig. 1 shows the uplink transmitter/receiver for SC-FI-SSMA using frequency-domain interleaving. We assume that U_i mobile stations (MS's) (or users) are transmitting their data to the *i*th base station (BS). With SSD, only one BS is selected for communication as described in Sect. III. Hence, the number of MS's communicating with a BS may be different for a different BS.

We consider the $u(i)$th ($u(i)$=0~U_i-1) MS communicating with the *i*th BS. A binary data sequence is transformed into a data modulated symbol sequence and is divided into a sequence of blocks of N_c^i/SF_t^i symbols each, where N_c^i is the block size of discrete Fourier transform (DFT) and SF_t^i is the spreading factor for the *i*th BS. In this paper, unlike [11], DFT is used since N_c^i (or SF_t^i) is not always equal to power of two. Without loss of generality, the transmission of one block {$d^{u(i)}(n)$; n=0~N_c^i/SF_t^i-1} is considered. The data block is spread by a spreading sequence {$c^{u(i)}(t)$; t=..,–1,0,1,..} of spreading factor SF_t^i. The chip block after spreading is expressed, using vector representation, as $\mathbf{s}^{u(i)}=[s^{u(i)}(0),...,s^{u(i)}(t),\ldots,s^{u(i)}(N_c^i-1)]^{\mathrm{T}}$, where T denotes the transpose. $s^{u(i)}(t)$ can be expressed, using the equivalent lowpass representation, as

$$s^{u(i)}(t)=d^{u(i)}\left(\lfloor t/SF_t^i \rfloor\right)c^{u(i)}(t) \quad , (1)$$

where $\lfloor x \rfloor$ represents the largest integer smaller than or equal to *x*. The resulting N_c^i-chip block is decomposed by N_c^i-point DFT into frequency-domain signal

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$\mathbf{S}^{u(i)} = [S^{u(i)}(0),...,S^{u(i)}(k),\ldots,S^{u(i)}(N_c^i-1)]^{\mathrm{T}}$, where $S^{u(i)}(k)$ is the kth frequency component, given by

$$S^{u(i)}(k) = \sum_{t=0}^{N_c^i-1} s^{u(i)}(t)\exp\left(-j2\pi k\frac{t}{N_c^i}\right) .(2)$$

$\mathbf{S}^{u(i)}$ is mapped onto SF_f^i times wider bandwidth by using a frequency interleaving matrix so that frequency components of all users communicating with the ith BS do not overlap.

The overall spreading factor SF is defined as $SF = SF_t^i \times SF_f^i$ and is the same for all users. However, the choice of (SF_t^i, SF_f^i) is different for a different BS because the number U_i $(\leq SF_t^i \times SF_f^i)$ of users communicating with the ith BS is different for different i. Choosing (SF_t^i, SF_f^i) will be described in Sect. IV. Fig. 2 shows the frequency-domain interleaving for the ith BS. The frequency-interleaved block can be represented as

$$\begin{aligned}\hat{\mathbf{S}}^{u(i)} &= [\hat{S}^{u(i)}(0),...,\hat{S}^{u(i)}(k'),\ldots,\hat{S}^{u(i)}(N_c^i SF_f^i-1)]^{\mathrm{T}} \\ &= \mathbf{Q}^{u(i)}\mathbf{S}^{u(i)}\end{aligned} , (3)$$

where $\mathbf{Q}^{u(i)}$ is an $N_c^i SF_f^i$-by-N_c^i frequency-interleaving matrix. Interleaving matrices are determined so that different users' frequency components do not overlap with each other (see Fig. 2). $\mathbf{Q}^{u(i)}$ must satisfy [11]

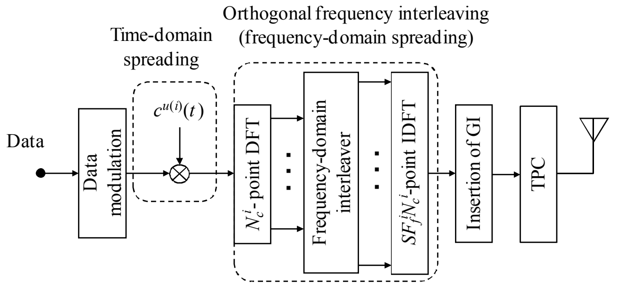
$$\left\{\mathbf{Q}^{u(i)}\right\}^T \mathbf{Q}^{u'(i')} = \begin{cases}\mathbf{I} \text{ if } i=i' \text{ and } u=u' \\ \mathbf{0} \text{ if } i=i' \text{ and } u\neq u' \\ \left\{\mathbf{Q}^{u(i)}\right\}^T \mathbf{Q}^{u'(i')} \text{ otherwise}\end{cases} , (4)$$

where $\mathbf{I}$ is an $N_c^i \times N_c^i$ identity matrix.

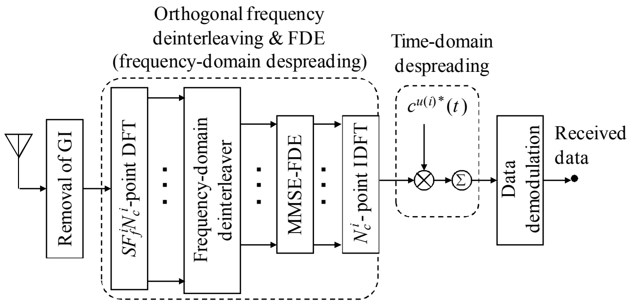
Finally, $N_c^i SF_f^i$-point inverse DFT (IDFT) is applied to obtain the SC-FI-SSMA signal block $\{\tilde{s}^{u(i)}(t');t'=0\sim(N_c^i SF_f^i-1)\}$, which can be expressed as

$$\tilde{s}^{u(i)}(t') = \frac{1}{N_c^i}\sum_{k'=0}^{N_c^i SF_f^i-1}\hat{S}^{u(i)}(k')\exp\left(j2\pi k'\frac{t'}{N_c^i SF_f^i}\right) . (5)$$

The last N_g samples of each block are copied and inserted, as a cyclic prefix, into the guard interval (GI) at the beginning of each block to form a block of $N_c^i SF_f^i + N_g$ samples.



(a) The $u(i)$th MS transmitter (communicating with the ith BS)



(b) The ith BS receiver

Fig. 1 Transmitter/receiver structure.

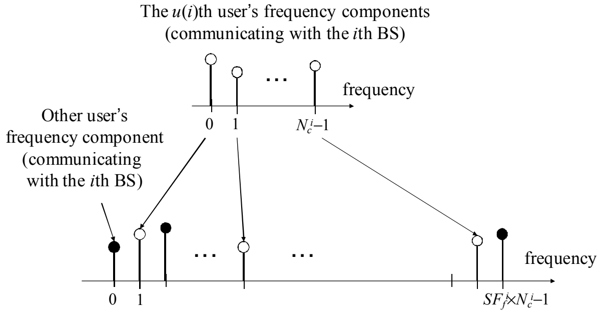


Fig. 2 Frequency-domain interleaving.

B. Received signal representation

We consider a hexagonal cell layout. Each BS is indexed by i (=0, 1, 2, …). The i=0th BS is a cell of interest. The received signal block $\{r(t');\ t'=-N_g\sim(N_c^i SF_f^i-1)\}$, at the i=0th BS is expressed as

$$r(t') = \sum_{i=0}^{\infty}\sum_{u=0}^{U_i-1}\sqrt{2P_r^{u(i)\to 0}}\sum_{l=0}^{L-1}h_l^{u(i)\to 0}\,\tilde{s}^{u(i)}(t'-\tau_l^{u(i)\to 0}) + \eta(t') , (6)$$

where $P_r^{u(i)\to 0}$ is the average received signal power from the $u(i)$th MS communicating with the ith BS and is given by

$$P_r^{u(i)\to 0} = P_t^{u(i)\to i}\cdot\left(R^{u(i)\to 0}\right)^{-\alpha}\cdot 10^{-\delta^{u(i)\to 0}/10} , (7)$$

where $P_t^{u(i)\to i}$ is the transmit signal power of the $u(i)$th MS to the ith BS, $R^{u(i)\to 0}$ is the distance between the i=0th BS and the $u(i)$th MS, α is the path loss exponent, $\delta^{u(i)\to 0}$ is the log-normally distributed shadowing loss, and $h_l^{u(i)\to 0}$ is the lth path gain, between the 0th BS and the $u(i)$th MS, with $\sum_{l=0}^{L-1} E[|h_l^{u(i)\to 0}|^2]=1$ ($E[.]$ denotes the ensemble average operation). We assume block fading so that the path gains remain constant over one block length of ($N_c^i SF_f^i$+N_g) samples. $\tau_l^{u(i)\to 0}$ is the lth path delay and the maximum time delay $\tau_{L-1}^{u(i)\to 0}$ is assumed to be shorter than the GI length for all i and $u(i)$. $\eta(t')$ is a zero-mean complex Gaussian noise process with the variance $2\sigma^2$ due to the additive white Gaussian noise (AWGN).

C. Frequency-domain deinterleaving and MMSE-FDE

The received signal block $\{ r(t') \}$ is transformed by $N_c^i SF_f^i$-point DFT into $N_c^i SF_f^i$ frequency-domain signal block represented by $\mathbf{R} = \left[R(0),..., R(k'),..., R(N_c^i SF_f^i -1)\right]^T$.

Without loss of generality, the detection of data block of the 0(0)th MS communicating with the i=0th BS is considered. $\mathbf{R}$ is deinterleaved to pick up the 0(0)th MS's frequency-domain signal block $\widetilde{\mathbf{R}}^{0(0)} = [\widetilde{R}^{0(0)}(0),...,\widetilde{R}^{0(0)}(k),...,\widetilde{R}^{0(0)}(N_c^0-1)]^{\mathrm{T}}$ as

$$\begin{aligned}\widetilde{\mathbf{R}}^{0(0)} &= \frac{1}{SF_f^0}\mathbf{Q}^{0(0)^T}\mathbf{R}\\ &= \frac{1}{SF_f^0}\sum_{i=0}^{\infty} SF_f^i \sum_{u=0}^{U_i-1}\sqrt{2P_r^{u(i)\to 0}}\left\{\mathbf{Q}^{0(0)^T}\mathbf{H}^{u(i)}\mathbf{Q}^{u(i)}\right\}\mathbf{S}^{u(i)}\\ &+\frac{1}{SF_f^0}\mathbf{Q}^{0(0)^T}\mathbf{\Pi}\end{aligned} \quad ,(8)$$

where $\mathbf{H}^{u(i)} = diag\left(H^{u(i)}(0),..,H^{u(i)}(k'),..,H^{u(i)}(N_c^i SF_f^i-1)\right)$ is ($N_c^i SF_f^i$)-by-($N_c^i SF_f^i$) channel gain matrix and $\mathbf{\Pi}$ is ($N_c^i SF_f^i$)-by-1 noise vector. $H^{u(i)}(k')$ is given by

$$H^{u(i)}(k') = \sum_{l=0}^{L-1} h_l^{u(i)\to 0}\exp\left(-j2\pi k'\frac{\tau_l^{u(i)\to 0}}{N_c^i SF_f^i}\right) \quad . (9)$$

Since the frequency-interleaving matrices $\mathbf{Q}^{u(i)}$'s are orthogonal to each other for the same i (see Eq. (4)) and $\mathbf{H}^{u(i)}$ is the diagonal matrix, $\mathbf{Q}^{0(0)^T}\mathbf{H}^{u(i)}\mathbf{Q}^{u(i)}$ satisfies

$$\mathbf{Q}^{0(0)^{\mathrm{T}}}\mathbf{H}^{u(i)}\mathbf{Q}^{u(i)} = \begin{cases}\mathbf{Q}^{0(0)^{\mathrm{T}}}\mathbf{H}^{0(0)}\mathbf{Q}^{0(0)} & \text{if } i=0 \text{ and } u=0\\ \mathbf{0} & \text{if } i=0 \text{ and } u\neq 0\\ \mathbf{Q}^{0(0)^{\mathrm{T}}}\mathbf{H}^{u(i)}\mathbf{Q}^{u(i)} & \text{otherwise}\end{cases} \quad . (10)$$

Hence, we obtain

$$\begin{aligned}\widetilde{\mathbf{R}}^{0(0)} &= \sqrt{2P_r^{0(0)\to 0}}\,\widetilde{\mathbf{H}}^{0(0)}\mathbf{S}^{0(0)}\\ &+\sum_{i=1}^{\infty}\frac{SF_f^i}{SF_f^0}\sum_{u=0}^{U_i-1}\sqrt{2P_r^{u(i)\to 0}}\,\mathbf{I}^{u(i)}+\widetilde{\mathbf{\Pi}}\end{aligned} \quad ,(11)$$

where $\widetilde{\mathbf{H}}^{0(0)} = diag\left(\widetilde{H}^{0(0)}(0),...,\widetilde{H}^{0(0)}(k),...,\widetilde{H}^{0(0)}(N_c^0-1)\right)$ and $\widetilde{\mathbf{\Pi}}^{0(0)} = \left[\widetilde{\Pi}^{0(0)}(0),...,\widetilde{\Pi}^{0(0)}(k),...,\widetilde{\Pi}^{0(0)}(N_c^0-1)\right]^T$ are the frequency-deinterleaved channel gain vector and the noise vector, given by

$$\begin{cases}\widetilde{\mathbf{H}}^{0(0)} = \mathbf{Q}^{0(0)^T}\mathbf{H}^{0(0)}\mathbf{Q}^{0(0)}\\ \widetilde{\mathbf{\Pi}} = \dfrac{1}{SF_f^0}\mathbf{Q}^{0(0)^T}\mathbf{\Pi}\end{cases} \quad . (12)$$

$\mathbf{I}^{u(i)} = [\overline{H}^{u(i)}(0)\overline{S}^{u(i)}(0),...,\overline{H}^{u(i)}(N_c^0-1)\overline{S}^{u(i)}(N_c^0-1)]^{\mathrm{T}}$ is the inter-cell interference, where $\overline{H}^{u(i)}(k)$ and $\overline{S}^{u(i)}(k)$ are the channel gain and the $u(i)$th MS's transmitted signal after frequency-domain deinterleaving. $\mathbf{I}^{u(i)}$ is given by

$$\mathbf{I}^{u(i)} = \mathbf{Q}^{0(0)^{\mathrm{T}}}\mathbf{H}^{u(i)}\mathbf{Q}^{u(i)}\mathbf{S}^{u(i)} \quad .(13)$$

From Eq. (10), $\widetilde{R}^{0(0)}(k)$ can be written as

$$\begin{aligned}\widetilde{R}^{0(0)}(k) &= \sqrt{2P_r^{0(0)\to 0}}\,\widetilde{H}^{0(0)}(k)S^{0(0)}(k)\\ &+\sum_{i=1}^{\infty}\frac{SF_f^i}{SF_f^0}\sum_{u=0}^{U_i-1}\sqrt{2P_r^{u(i)\to 0}}\,\overline{H}^{u(i)}(k)\overline{S}^{u(i)}(k)+\widetilde{\Pi}^{0(0)}(k)\end{aligned} \quad ,(14)$$

where the first term is the desired MS's signal, the second is the inter-cell interference and the third is the noise. The MUI resulting from the MS's communicating with the i=0th BS is completely removed by frequency-domain deinterleaving although the inter-cell interference still remains.

MMSE-FDE is carried out as follows:

$$\hat{R}^{0(0)}(k) = \widetilde{R}^{0(0)}(k)W^{0(0)}(k) \quad ,(15)$$

where $W^{0(0)}(k)$ is the MMSE-FDE weight and is given by

$$W^{0(0)}(k) = \frac{\left(\widetilde{H}^{0(0)}(k)\right)^*}{\left[\begin{array}{l}P_r^{0(0)\to 0}\left|\widetilde{H}^{0(0)}(k)\right|^2\\ +\displaystyle\sum_{i=1}^{\infty}\left(\frac{SF_f^i}{SF_f^0}\right)^2\sum_{u=0}^{U_i-1}P_r^{u(i)\to 0}\left|\overline{H}^{u(i)}(k)\right|^2+\hat{\sigma}^2\end{array}\right]} \quad ,(16)$$

where $\hat{\sigma}^2$ is the noise power.

The frequency-domain signal $\{\hat{R}^{0(0)}(k); k=0\sim N_c^0-1\}$ is transformed by N_c^0-point IDFT to the time-domain chip block $\{\widetilde{r}^{0(0)}(t); t=0\sim N_c^0-1\}$ as

$$\tilde{r}^{0(0)}(t) = \frac{1}{N_c^0} \sum_{k=0}^{N_c^0 - 1} \hat{R}^{0(0)}(k) \exp\left(j2\pi t \frac{k}{N_c^0} \right) .(17)$$

Finally, despreading is performed on $\tilde{r}^{0(0)}(t)$ to obtain the 0(0)th user's decision variable associated with $d^{0(0)}(n)$ as

$$\tilde{d}^{0(0)}(n) = \frac{1}{SF_t^0} \sum_{t=nSF_t^0}^{(n+1)SF_t^0 - 1} \tilde{r}^{0(0)}(t) c^{0(0)*}(t) . (18)$$

III. SITE SELECTION DIVERSITY AND TRANSMIT POWER CONTROL

A. Site selection diversity (SSD)

We assume slow SSD to reduce the impact of the inter-cell interference. In slow SSD, each MS transmits the pilot signal with equal transmit power. The surrounding BS's measure the local average received signal power using the received pilot and send its power information to radio network controller (RNC). In RNC, the local average received signal powers are sorted out in a descending order and the BS receiving with the largest local average received signal power is selected as an active BS for the communication.

B. Transmit power control

In a cellular mobile communication system, when the MS moves close to the cell edge, the received signal power gets weaker and hence, the uplink capacity may degrade. In this paper, TPC is introduced to keep the received signal power constant for all users. We assume slow TPC and fast TPC. Slow TPC regulates the distance dependant path loss and shadowing loss, while fast TPC regulates the power variation caused by multipath fading as well as the distant dependant path loss and shadowing loss. Transmit signal power $P_t^{u(i)\to i}$ of the *u*(*i*)th MS communicating with the *i*th BS is given, for slow TPC and fast TPC, by

$$P_t^{u(i)\to i} = \begin{cases} P \cdot \left(R^{u(i)\to i}\right)^{\alpha} \cdot 10^{\delta^{u(i)\to i}/10} & \text{slow TPC} \\ P \cdot \left(R^{u(i)\to i}\right)^{\alpha} \cdot 10^{\delta^{u(i)\to i}/10} & \\ \times \left(\frac{1}{N_c^i} tr\left(\mathbf{Q}^{u(i)^{\mathrm{T}}} \mathbf{H}^{u(i)\to i^{\mathrm{H}}} \mathbf{H}^{u(i)\to i} \mathbf{Q}^{u(i)} \right) \right)^{-1} & \text{fast TPC} \end{cases} , (19)$$

where P is the TPC target and $R^{u(i)\to i}$ and $\delta^{u(i)\to i}$ are the distance and the log-normally distributed shadowing loss between the *i*th BS and the *u*(*i*)th MS. $\mathbf{H}^{u(i)\to i} = diag\left(H^{u(i)\to i}(0),\ldots,H^{u(i)\to i}(N_c^i SF_f^i - 1)\right)$ is the channel gain matrix between the *i*th BS and the *u*(*i*)th MS. *tr*( ) and $(\ )^{\mathrm{H}}$ denote the trace and conjugate transpose, respectively.

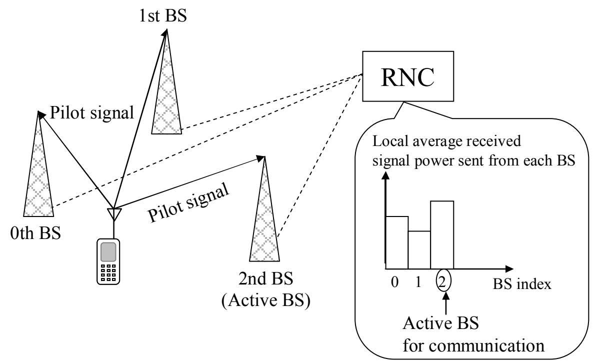


Fig. 3 SSD model.

IV. OPTIMUM SET OF (SF_t^i, SF_f^i)

When U_i MS's simultaneously communicate with the *i*th BS, (SF_t^i, SF_f^i) is set, for the given overall spreading factor *SF*, as follows. First, SF_f^i is set as SF_f^i = U_i . Using frequency-domain interleaving of $SF_f^i = U_i$, different users' frequency components do not overlap with each other and therefore, the MUI, which is the predominant cause of the performance degradation, can be completely removed. Then, SF_t^i is set as $SF_t^i = \lfloor SF/U_i \rfloor$. The time-domain spreading is effective to suppress the residual inter-chip interference (ICI) produced after MMSE-FDE as well as the inter-cell interference. The above set of (SF_t^i, SF_f^i) provides the best BER performance [11]. For example, when *SF*=32 and $U_i = 5$, $(SF_t^i, SF_f^i) = (6,5)$ is used (In [11], the case of SF_t^i and SF_f^i being power of 2 is only considered).

V. COMPUTER SIMULATION

We assume $N_c^i / SF_t^i = 8$, a DFT block size of $N_c^i SF_f^i$=256 samples and the overall spreading factor *SF*=32 for all *i*. For the case of $SF_t^i \times SF_f^i < SF$ (e.g., $(SF_t^i, SF_f^i) = (6,5)$), the zero-padding is performed after frequency-domain interleaving to keep the transmission data rate constant for all users. A GI length of N_g=32 samples and QPSK data modulation are assumed. The path loss exponent α=3.5 and the shadowing loss standard deviation of 6 dB are assumed. The channel is assumed to be a frequency-selective block Rayleigh fading channel having a sample-spaced *L*-path uniform power delay profile (i.e., $E[|h_l^{u(i)\to i}|^2] = 1/L$ for all *i*, *l* and *u*(*i*)). The time delay $\tau_l^{u(i)\to 0}$, *u*(*i*)=0~U_i–1, of the *l*th path is $\tau_l^{u(i)\to 0} = l + \bar{\tau}^{u(i)\to 0}$ samples, where $\bar{\tau}^{u(i)\to 0}$ is the *u*th user's time delay. We assume that the maximum time delay $\tau_{L-1}^{u(i)\to 0}$ is shorter than the GI length for all *i* and *u*(*i*). Perfect chip timing and ideal channel estimation are assumed. We consider an equal-space interleaving so that each user's frequency components are distributed with equal separation in the frequency-domain [11].

We consider 19 hexagonal cells (i=0~18). We consider the interference-limited environment. Computer simulation is done as follows. The locations of U users are randomly generated in each cell, and the path losses, shadowing losses, and path gains are generated for all users. Then, each BS measures the local average received signal power associated with an MS to carry out slow SSD. After selecting BS communicating with each MS, (SF_t^i, SF_f^i) is determined for the ith BS. After the MS transmit power is adjusted by TPC, the local average BER is measured for each MS in the i=0th BS. This BER measurement is repeated a sufficient number of times by changing the MS locations to obtain the distribution of the local average BER. The outage probability is defined as the probability that the local average BER exceeds the required BER (it is set to BER=10^{-2} in this paper).

Fig. 4 plots the outage probability of SC-FI-SSMA as a function of the average number of users per cell normalized by SF. For comparison, the outage probability of the conventional DS-CDMA with MMSE-FDE is also plotted (this is equivalent to SC-FI-SSMA of $(SF_t^i, SF_f^i) = (SF, 1)$). SC-FI-SSMA provides lower outage probability than DS-CDMA with MMSE-FDE, since the MUI can be completely removed by frequency-domain interleaving. On the other hand, in DS-CDMA, the large MUI is produced sometimes and hence, the outage probability significantly increases. It can also be seen that the use of slow TPC is effective to reduce the outage probability for both SC-FI-SSMA and DS-CDMA, since it can regulate the distance dependant path loss and shadowing loss. Fast TPC can regulate the fading variation as well as the path loss and shadowing loss; however, only slight reduction in the outage probability is observed. This is because the probability that $\sum_{l=0}^{L-1} |h_l^{u(i)\to i}|^2$ drops (or fades) is very small. Using fast TPC, the number of users per cell normalized by SF satisfying with the outage probability of 0.1 is about 0.14, while it is about 0.1 in DS-CDMA.

VI. CONCLUSION

In this paper, we evaluated the uplink capacity of SC-FI-SSMA that uses frequency-domain interleaving and MMSE-FDE and compared with that of the conventional DS-CDMA with MMSE-FDE. Site selection diversity and transmit power control were applied to improve the uplink capacity. It was shown by computer simulation that SC-FI-SSMA gives lower outage probability than DS-CDMA with MMSE-FDE, since the MUI can be perfectly removed. With fast TPC, the number of users per cell normalized by SF satisfying with the outage probability of 0.1 was shown to be 0.14, while it is about 0.1 for DS-CDMA. In this paper, zero-padding was used to keep the data rate constant for all users, but it reduces the total transmission data rate. To increase the total transmission data rate, the scheduling technique is effective. The scheduling in a multi-cell environment may be an important future work. Performance comparison between SC-FI-SSMA and OFDMA is also left as an interesting future work.

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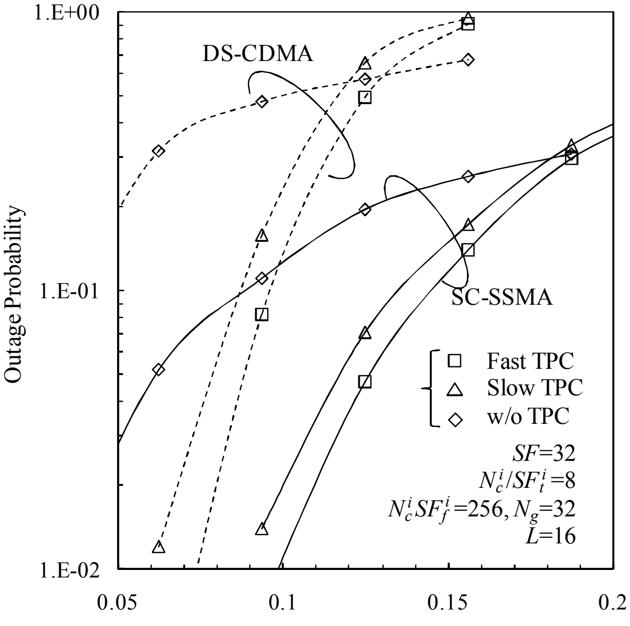

Fig. 4 Outage probability of SC-FI-SSMA