Investigating to the CFO Phenomena for MC-CDMA Systems Combining with Multi-Dimension Receiver over Selective Fading

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***Abstract*—The scenario assumed that the CFO (carrier frequency offset), which is caused by the ICI (inter-carrier interference), exists in the environments of short-term fading, a multi-dimension combining (M-D combining) receiver for an MC-CDMA (multi-carrier coded-division multiple-access) system is proposed in this paper. The system performance is validated by the evaluation with assumption of the working environment state is in the frequency selective fading environments. In order to obtain the validation of the accuracy of the derivation and the proof of the proposed schemes the numerical results are illustrated in this paper.**

***Keywords*—MC-CDMA system, ICI (inter-carrier interference), multi-dimension combining (M-D combining), MIP**

1. INTRODUCTION

Recently, the wideband radio system combining with the application of multicarrier modulating and CDMA (coded-division multiple-access) schemes has been considered interesting in the cellular wireless communications. It is the reason that DS (direct sequence) waveform exist the broader bandwidth for combating the ISI (inter-symbol interference) caused by the multipath fading during the transmission. Generally, the Rake receiver provides a correlator for each carrier and the signal at the output of the correlator is combined with some tapped delay lines [1, 2]. The receiver provides a correlator for each carrier and at the output of the correlators combined with MRC diversity. In wireless mobile communications the performance of the antenna diversity is always an attractive area of interesting to study [3]. In additions, for the purpose of increasing the ability of the resolve and combination for multiple paths the wideband spread. It is known that the correlation phenomenon will impact the performance of the diversity combining schemes [5]. The researchers Kang and Yao analyzed the performance of an MC-CDMA system over frequency-selective Nakagami-m fading channels with correlated and independent subcarriers [7]. Yang and Hanzo evaluated the performance of an MC-DS-CDMA to consider the correlation presents in the fading of the various subcarriers [9]. Moreover, the CFO (carrier frequency offset) phenomenon, which is caused by the mismatch in frequency generated from the oscillator between the transmitter and the receiver, i.e. the estimation of the receiver goes wrong, induces the ICI (intercarrier interference) which will abolish the orthogonality of the transmitted data over an MC-CDMA systems [11]. Recently, Xiang and Hanzo proposed the exact closed-form for the average BER calculation of OFDM system in the presence of both CFO and phase estimation error in frequency-selective fading channels [15].

In this paper we propose the investigation takes account of the CFO phenomena into the performance evaluation for an MC-CDMA system with multi-dimension combining receiver, called as M-D combining receiver, and which is considered working in frequency selective fading environments. The proposed M-D combining receiver is definitely proved can mitigate the system performance of an MC-CDMA system of an MC-CDMA system degraded by the effect of CFO. In order to validate the accuracy of the determined results some of the parameters, i. g., the resolvable multipath number, the number of antenna with combining receiver, the fading parameter, the power decay factor MIP (multipath intensity profile), and the correlation characteristic between the antennas, are all considered in the numerical analysis section.

The paper is organized as follows. In section II the system model of an asynchronous MC-CDMA is described. A closed-form expression for the pdf (probability density function) of the decision static is shown in section III. In section IV illustrates the closed-form expression for the average BER with the impact of correlated and uncorrelated fading among spatially separated receiver fingers on system performance the numerical results are presented in section V. Finally, there is a simple conclusion drawn in section VI.

2. SYSTEM MODELS

2.1. Transmitter Model

Assuming that there exist K simultaneous users with N subcarriers within a single cell in the system model, any effect of correlation among users will be ignored by assuming that the number of users is uniform within the distribution. A signal data symbol is replicated into N parallel copies. A signature sequence chip with a spreading code of length L is used for BPSK (binary phase shift keying) to modulate each of the N subcarriers of the kth user, where the subcarrier has frequency I/T_b Hz, and where I is an integer [3].

$$S_k(t)=\sqrt{\frac{2P}{N}}\sum_{m'=0}^{M-1}\sum_{n=0}^{N-1}c_k^n b_k^{m'} P_{T_b}(t-nT_b) \; \mathrm{Re}\left\{e^{-j[2\pi f_c t+2\pi F(N-1)t/T_b]}\right\} \tag{1}$$

where both c_k^n and $b_k^{m'}$ belong to $\{-1,\ 1\}$, P is the power of the data bit, M denotes the number of the data bit, N expresses the number of subcarriers, and the sequences $c_k^0,...,c_k^{N-1}$ and $b_k^0,...,b_k^{M-1}$ represent the signature sequence and the data bit of the k-th user, respectively. The term $P_{T_b}(t)$ is defined as the unit amplitude pulse that is non-zero in the interval of $[0,\ T_b]$, $\omega_n=2\pi(f_c+nI/T_b)$ is the angular frequency of the nth subcarrier, where f_c indicates the carrier frequency, T_b is the symbol duration.

2.2. Channel Model

The k-th user's a_t-th receiver antenna has the low-pass channel impulse response expressed as

$$h_{a_t}(t)=\sum_{l=0}^{L_R-1}\alpha_{a_t,k}^l e^{j\varphi_{a_t,k}^l}\delta(\tau-\tau_{a_t,k}^l) \tag{2}$$

where the path phases and path delays are assumed to be independent and uniformly distributed over [0, 2π] and [0, T], respectively. L_R is the number of resolvable propagation paths that reach the receive antenna. Each path is characterized by its instantaneous fading amplitude $\alpha_{a_t,k}^l$, its phase shift, and its propagation delay $\tau_{a_t,k}^l$. Under the assumption of that the total time average channel gain per antenna for each user is normalized to one, that is able to be written an $\sum_{l=0}^{L_R-1}E[(\alpha_{a_t,k}^l)^2]=\sum_{l=0}^{L_R-1}\Omega_{a_t,k}^l=1$ in the real world, general, the effect of MIP phenomena is an exponential type and is given as $\Omega_{a_t,k}^l=\Omega_{a_t,k}^0 e^{-l\delta}$, $l=0,\ 1,...,\ L_R-1$, where the parameter δ indicates the rate of decay of the average path strength as a function of the path delay. Assume that the fading amplitude of the l-th path at the a_t-th antenna of the k-th user is modeled as Nakagami-m fading model, and the instantaneous power, ζ , of the l-th path, $l=0,\ 1,...,\ L_R-1$, can be easily shown follows the gamma pdf given as

$$f_{\zeta_{a_t,k}^l}(\zeta_l)=\frac{(m_{a_t,k}^l/\Omega_{a_t,k}^l)^{m_k^l}}{\Gamma(m_{a_t,k}^l)}(\zeta_l)^{m_{a_t,k}^l-1}\times\exp\{-\frac{m_{a_t,k}^l}{\Omega_{a_t,k}^l}\},\quad \zeta_l\ge 0 \tag{3}$$

where $\Omega_{a_t,k}^l=E[(\alpha_{a_t,k}^l)^2]$ is the average channel power of l-th at a_t-th antenna of the k-th user.

2.3. Receiver Model

A block diagram of the receiver of an MC-CDMA system with multiple-dimension combining scenario is shown in Fig. 1. Hereafter, the received signal at the output of the referenced user can be obtained as

$$\begin{aligned} r(t)&=\sum_{k=0}^{K-1}\alpha_k^l S_k(t+\varphi_k^l)+n(t)\\ &=\sqrt{\frac{2P}{N}}\sum_{k=0}^{K-1}\sum_{m'=0}^{M-1}\sum_{l=0}^{L_R-1}\alpha_k^l c_k^l b_k^{m'} P_{T_b}(t-nT_b+\varphi_k^l)\\ &\quad\times\cos[\omega_c t+2\pi n\, It/T_b+\varphi_k^l]+N(t) \end{aligned} \tag{4}$$

where α_k^l denotes the channel fading intensity modeled as Nakagami-m distributed, $\varphi_k^l=\phi_k^l-\theta_k^l$ represents the phase difference between the transmitter and receiver and $N(t)$ is the AWGN (additive white Gaussian noise). Assuming that acquisition has been accomplished for the user of interesting ($k=0$), due to using combining scheme, it is assumed that a perfect phase correction can be obtained, i.e., $\hat{\theta}_0^l=\theta_0^l$. With all the assumptions for Rake combined, the decision variable ξ_0 of the nth data bit for the referenced user is given by

$$\begin{aligned} \xi_0&=\int_{nT_c}^{nT_c+T_b} r(t)\varepsilon_0^l\cos(\omega_c t+\varphi_0^l)\,dt\\ &=Z_0+I_{0,SI}+I_{0,MAI}+I_{0,AWGN} \end{aligned} \tag{5}$$

where $r(t)$ is the received signal shown in (4) for a single cell, and T_c denotes the chip duration. The first term of the previous equation represents the desired signal Z_0 of the referenced user; the second term, $I_{0,SI}$, in this study the SI (self-interference) shown in the second term of (5) is assumed to be ignorable by carefully choosing the PN (pseudo noise) sequence of the MC-CDMA system in practice; $I_{0,MAI}$ is the MAI (multiple-access interference); and the last term, $I_{0,AWGN}$, is the AWGN with zero mean and $N_0 T_b/4$ variance.

Since $b_k^{m'}$ belongs to $\{-1, 1\}$, hence the average value of the desired signal, the first term shown in (5), for the referenced user, 0-th user, in a single-cell environment can be obtained by using the method of expectation operating and determined as

$$\begin{aligned}(S_0)&=E\left[\sqrt{\frac{P}{2N}}\sum_{l=0}^{L_R-1}\sum_{m'=0}^{M-1}\left(\alpha_0^l\right)^2 b_0^{m'}\right]\\&=\sqrt{\frac{P}{2N}}E\left[\sum_{l=0}^{L_R-1}\left(\alpha_0^l\right)^2\right]\end{aligned} \tag{6}$$

Once the desired values of referenced user is decided, then the variance of the total interference, $\left(\sigma_T^2\right)$, of the referenced user in an MC-CDMA system working in a single-cell can be obtained as [14]

$$\left(\sigma_T^2\right)=\frac{P}{3N}\sum_{l=0}^{L_R-1}(K-1)\Omega_0^l E\left[\left(\alpha_k^l\right)^2\right]+\frac{N_0}{4T_b}\sum_{l=0}^{L_R-1}\Omega_0^l \tag{7}$$

where the correlation between the branches has been included in the term $E[(\sum_{n=0}^{N-1}\beta_{n,k})^2]$.

2.4. The CFO consideration

There are A_t total antenna number assumed in the combining receiver and the combining subchannel follows after the LPF. The attenuation factor of the useful data is designed when $n=1$. In order to include the CFO effect in the system performance evaluation of an MC-CDMA system, the ICI coefficient caused by the nth subcarrier for $n=2,\cdots,N$ subchannel is given as

$$m_n=\left[M\right]_{1,n}=\Lambda(\varepsilon)\cdot e^{j\pi\frac{N-1}{N}(n-1)} \tag{8}$$

where

$\Lambda(\varepsilon)=\sin\left[\pi(n-1+\varepsilon)\right]/N\sin\left[\pi(n-1+\varepsilon)/N\right]$,

and ε indicates the CFO magnitude. Hence, the SNR at the combining receiver output of the MC-CDMA system can be determined by putting (6) and (7) together and obtained as

$$\Sigma_0=\left\{\Lambda(\varepsilon)\big|_{n=1}\cdot\left[\frac{2\cdot(k-1)}{N}+\frac{1}{SNR}\right]+\sum_{n=2}^{N}\left|\Lambda(\varepsilon)e^{j\pi\frac{N-1}{N}(n-1)}\right|^2\right\}^{-1} \tag{9}$$

where *SNR* in the ratio of bit energy and the noise result in each antenna.

3. Statistical Analysis of an M-D Combining Receiver

The SINR (signal-to-interference-plus-noise ratio) at the output of an M-D combining receiver is easily combined with all the path components then can be expressed as

$$\gamma=\sum_{a_t=1}^{A_t}\sum_{l=0}^{L_R-1}\gamma_{a_t}^l \tag{10}$$

where $\gamma_{a_t}^l=\Sigma_0(\alpha_{a_t}^l)^2$ is the instantaneous SINR of the n-th finger of the a_t-th antenna finger, the number of the combining receiver is indicated by L_R. The computation of statistical characteristics of the SINR is required to gain. On the other hand, when the fading figure are real and arbitrary, the pdf of γ is able to be calculated as an approximate expression and illustrated in some of the publications [4], and in [6] it can be expressed as an indefinite integral. Thereafter, by using the changing variable, let $y_{a_t}^l=m_{a_t}^l/\Sigma_0\Omega_{a_t}^l$, where $a_t=1,2,...,A_t$, and $l=0,1,...,L_R-1$. By substituting (11) into the MGF formula, and the MGF of the SINR, $\gamma_{a_t}^l$, can be expressed as

$$\begin{aligned}\varphi_{\gamma_{a_t}^n}(t)&=\int_0^\infty e^{-xt}f_{\gamma_{a_t}^n}(x)dx\\&=\frac{(y_{a_t}^n)^{m_{a_t}^n}}{\Gamma(m_{a_t}^n)}\int_0^\infty e^{-xt}x^{m_{a_t}^n-1}\exp(-y_{a_t}^n x)dx\end{aligned} \tag{11}$$

By applying the closed form definition shown in [10], the formula (11) can be computed by some of the steps as follows. Based on some of the integral formulas, for instance the results shown in [3] the closed form of integral in (11) can be easily calculated. However, the integral in (11) will be determined by using of the approach proposed in [15]. Firstly, the exponential function may be expressed as a contour integral $\exp(-x)=\int_{-i\infty}^{i\infty}\Gamma(-s)x^s ds/2\pi i$ [6, p. 43], where $i=\sqrt{-1}$. By interchanging the order of integration and substituting it into the MGF formula (11), which can be derived as

$$\varphi_{\gamma_{a_t}^n}(t)=\frac{1}{\Gamma(m_{a_t}^n)}(\frac{y_{a_t}^n}{t})^{m_{a_t}^n}\frac{1}{2\pi i}$$
$$\times\int_{-i\infty}^{i\infty}\Gamma(-s)\Gamma(m_{a_t}^n+s)(\frac{y_{a_t}^n}{t})^s ds \quad (12)$$

where the components of the multipath channel are assumed all identical and via the imaginary axis (in the complex s-plane), separating the poles of $\Gamma(m_{a_t}^n+s)$, $a_t=1,\ 2,\dots,\ A_t$, $n=0,\ 1,\dots,\ L_R-1$ from the poles of $\Gamma(-s)$. Thus following up the inverse Laplace then the pdf, $f_\gamma(\gamma)$, of the SINR can be calculated in terms of the confluent form of the multivariate Lauricella hyper-geometric function and obtained as

$$f_\gamma(\gamma)=\frac{1}{\Gamma(\sum_{a_t=1}^{A_t}\sum_{l=0}^{L_R-1}m_{a_t}^l)}\left[\prod_{a_t=1}^{A_t}\prod_{l=0}^{L_R-1}(y_{a_t}^l)^{m_{a_t}^l}\right]$$
$$\times\gamma^{\left(\sum_{a_t=1}^{A_t}\sum_{l=0}^{L_R-1}m_{a_t}^l\right)-1}\Phi_2^{(A_t\cdot L_R)}(m_1^0,m_1^1,\dots m_{A_t}^{L_R-1}; \quad (13)$$
$$\sum_{a_t=1}^{A_t}\sum_{l=0}^{L_R-1}m_{a_t}^l;-y_1^0\gamma,-y_1^1\gamma,\dots,-y_{A_t}^{L_R-1}\gamma)$$

where the multiple Barnes-type contour integral [12] has been applied in determine the previous equation, and $\Phi_2^{(n)}\left(u_1,\dots,u_n;v;w_1,\dots,w_n\right)$ is a known formula called as confluent Lauricella hyper-geometric function, which is define as [10]

$$\Phi_2^{(n)}\left(u_1,\dots,u_n;v;w_1,\dots,w_n\right)$$
$$=\sum_{i_1,\dots,i_n=0}^{\infty}\frac{w_1^{i_1}}{i_1!}\cdots\frac{w_n^{i_n}}{i_n!}\frac{(u_1)_{i_1}\dots(u_n)_{i_n}}{(v)_{i_1+\dots+i_n}} \quad (14)$$

The parameters $y_{a_t}^l$ shown in (13) are defined as equal to the ratio of the amount of fading $m_{a_t}^l$ to the corresponding average SINR, $\Sigma_0\Omega_{a_t}^l$, of the l-th combining receiver finger at the a_t-th antenna, i. e, $y_{a_t}^l=m_{a_t}^l/\Sigma_0\Omega_{a_t}^l$. For the negative exponential MIP with power decay factor, δ, from the mentioned equations shown in channel model section, the average power of the l-th path can be written as [9]

$$\Omega_{a_t}^l=\frac{e^{-l\delta}}{q(L_R,\delta)}, \quad (15)$$
$$a_t=1,\ 2,\dots,\ A_t,\quad l=0,\ 1,\dots,\ L_R-1$$

where $q(L_R,\delta)=\sum_{l=0}^{L_R-1}e^{-l\delta}=(1-e^{-L_R\delta})/(1-e^{-\delta})$, and the number of multipath, L_R, is considered for the referenced user.

4. BER ANALYSIS

4.1. Un-correlated Channels

The conditional BER of coherent BPSK (binary phase shift keying) in AWGN channel is given by [8]

$$P_{BER}(\gamma)=Q(\sqrt{2\gamma})=0.5\frac{\Gamma(0.5,\gamma)}{\sqrt{\pi}} \quad (16)$$

where $Q(\cdot)$ is the Q-function, and $\Gamma(\alpha,x)=\int_x^\infty t^{\alpha-1}e^{-t}dt$ denotes the complementary incomplete gamma function [8, (8.350-2)]. By means of the random process means the average BER, $\bar{P}_{BER}^{un-cor}$, of the M-D combining receiver with un-correlated branch for an MC-CDMA system working in Nakagami-m fading channel is given as

$$\bar{P}_{BER}^{un-cor}=\int_0^\infty P_{BER}(\gamma)\cdot f_\gamma(\gamma)d\gamma \quad (17)$$

Next, by means of the definition of the Lauricella multivariate hyper-geometric function, $F_D^{(n)}(\dots)$, which is defined as [10]

$$F_D^{(n)}\left(\beta,u_1,\dots,u_n;v;w_1,\dots,w_n\right)$$
$$=\frac{\Gamma(v)}{\Gamma(\beta)\Gamma(v-\beta)}\int_0^1 t^{\beta-1}(1-t)^{v-\beta-1}\prod_{i=1}^{n}(1-w_i t)^{-u_i}dt \quad (18)$$
$$,\ \mathrm{Re}(v)>\mathrm{Re}(\beta)>0$$

where $\left(\chi\right)_n=\Gamma(\chi+n)/\Gamma(\chi)$ is the Pochammer symbol, $|w_1|<1,\dots,\ |w_n|<1$, and the Lauricella function $F_D^{(n)}(\dots)$. Therefore, the system BER of an MC-CDMA system with un-correlated combining branch and CFO in (17) is to be determined as

$$\bar{P}_{BER}^{un-cor}=\frac{\Gamma(\frac{1}{2}+\sum_{a_t=1}^{A}\sum_{l=0}^{L_R-1}m_{a_t}^l)}{2\sqrt{\pi}\Gamma(1+\sum_{a_t=1}^{A}\sum_{l=0}^{L_r-1}m_{a_t}^l)}\left[\prod_{a_t=1}^{A}\prod_{l=0}^{L_R-1}(\frac{y_{a_t}^l}{y_{a_t}^l+1})^{m_{a_t}^l}\right]$$
$$\times F_D^{(A_t\cdot L_R)}(\frac{1}{2}+m_1^0,\ m_1^1,\dots,\ m_{A_t}^{L_R-1};\ 1+\sum_{a_t=1}^{A}\sum_{l=0}^{L_R-1}m_{a_t}^l, \quad (19)$$
$$\frac{y_1^0}{y_1^0+1},\ \frac{y_1^1}{y_1^1+1},\dots,\ \frac{y_{A_t}^{L_R-1}}{y_{A_t}^{L_R-1}+1})$$

The average BER presented in the previous equation converges for all practical values of the system parameters (fading parameters), and where the representation in (19) provides a convenient method for fast and accurate numerical computation of the multivatiate hyper-geometric function. The convergence of the Lauricella multivariate hyper-geometric function in (18) can be shown by using of the following transformation [13]

$$F_D^{(n)}\left(\beta,u_1,...,u_n;v;w_1,...,w_n\right)$$
$$=\left[\prod_{i=1}^{n}(1-w_i)^{-u_i}\right] \quad (20)$$
$$\cdot F_D^{(n)}\left(v-\beta,u_1,...,u_n;v;\frac{w_1}{w_1-1},...,\frac{w_n}{w_n-1}\right)$$

4.2. Correlated Channels

Since each path with the same delay may suffer correlated fading, which is caused by the paths with different delays arrive at each antenna after traveling different ways, among the spatially separated combining receive antennas. Thus the correlation characteristic can't be ignored in the linear combining environments. The degree of correlation depends on the distance between the antennas and their configuration [5]. A set of correlated branch, $\left\{\gamma_{a_t},\ a_t=1,\cdots,A_t\right\}$, is assumed to express the sequence of branch SINRs', and which is considered as not necessarily identically distributed, gamma varieties with parameters m_n and $\Sigma_0\Omega_{a_t}^n$, respectively, and the correlation coefficient, ρ_{i,a_t}^n, between γ_i^n and $\gamma_{a_t}^n$ can be presented as

$$\rho_{i,a_t}^n=$$
$$\rho_{a_t,i}^n=\frac{Cov(\gamma_i^n,\gamma_{a_t}^n)}{\sqrt{\mathrm{var}(\gamma_i^n)\cdot\mathrm{var}(\gamma_{a_t}^n)}},\quad 0\le\rho_{i,a_t}^n\le 1 \quad (21)$$

where $\mathrm{i},a_t=1,2,...,A_t$, and $n=0,\ 1,...,\ L_R-1$. The CHF (characteristic function) of the instantaneous SINR, $\phi_\gamma(t)$, is given as [6]

$$\phi_\gamma(t)=\prod_{l=0}^{L_R-1}\left|I_{A_t}+tA^lC^l\right|^{-m_l}$$
$$=\prod_{l=0}^{L_R-1}\prod_{a_t=1}^{A_t}[1+t\cdot(y_{a_t}^l)^{-1}\lambda_{a_t}^l]^{-m_l} \quad (22)$$

where I_{A_t} is the $A_t\times A_t$ identity matrix, $|\cdot|$ is the determinant operator, and the matrices A^l and C^l, $l=0,\ 1,...,\ L_R-1$, are $A_t\times A_t$ diagonal matrices with entries $\Sigma_0\Omega_{a_t}^l/m_l=(y_{a_t}^l)^{-1}$ and a $A_t\times A_t$ positive definite matrices, respectively, the later is defined by

$$C^l=\begin{bmatrix}1 & \left(\rho_{12}^n\right)^{0.5} & ... & \left(\rho_{1A_t}^n\right)^{.5}\\ \left(\rho_{21}^n\right)^{0.5} & 1 & ... & \left(\rho_{2A_t}^n\right)^{0.5}\\ \vdots & ... & ... & \vdots\\ \left(\rho_{21}^n\right)^{0.5} & ... & ... & 1\end{bmatrix}_{A_t\times A_t} \quad (23)$$

There are A_t eigenvalues, determined by matrix C^l, are corresponding to the values of $\lambda_{a_t}^n,\ a_t=1,\ 2,...,\ A_t$ shown in (22). Now, we can check with the case of all the values of the eigenvalues with $\lambda_{a_t}^n=1$, $a_t=1,\ 2,...,\ A_t$, $l=0,\ 1,...,\ L_R-1$, which represents the independent fading among the receive antennas. By means of the known equivalent $(1-z)^{-c}={}_0F_1[c;-;z]$ [10], and (22) can be written as

$$\phi_\gamma(t)=\prod_{a_t=1}^{A_t}\prod_{l=0}^{L_R-1}{}_1F_0(m_l;-;-\frac{\lambda_{a_t}^l t}{y_{a_t}^l}) \quad (24)$$

By the same way, the previous equation is adopted to obtain the average BER in the case of un-correlated branch, for obtaining the determination of average BER for the case of correlated. By substituting (24) into (22), the CHF then becomes as

$$\phi_\gamma(t)=(\frac{1}{2\pi i})^{A_tL_R}\overbrace{\int_{\mathbb{C}_1}\int_{\mathbb{C}_2}\cdots\int_{\mathbb{C}_{A_tL_R}}}^{A_t\cdot L_R}\left\{\prod_{a_t=1}^{A_t}\prod_{l=0}^{L_R-1}\frac{1}{\Gamma(m_l)}\left(\frac{y_{a_t}^l}{\lambda_{a_t}^l t}\right)^{m_n}\right.$$
$$\left.\times\Gamma(-s_{a_t}^l)(m_l+s_{a_t}^l)\left(\frac{a_{a_t}^l}{\lambda_{a_t}^l t}\right)^{s_{a_t}^l}\right\}\underbrace{ds_1^0ds_1^1\cdots ds_{A_t}^{L_R-1}}_{A_t\cdot L_R} \quad (25)$$

where the Barnes-Mellin contour-type integral [10, p. 43],

$${}_1F_0(g;-;h\cdot t)=\frac{\Gamma(g)}{2\pi i}\int_{-i\infty}^{i\infty}(-h\cdot t)^s\Gamma(-s)\Gamma(g+s)ds \quad (26)$$

, has been applied. It then follows that the pdf of γ, which is in the case of correlated fading among combining fingers with the same path delay in spatially separated antennas, is given by

$$f_\gamma(\gamma)=\frac{1}{\Gamma(A_t\cdot\sum_{l=0}^{L_R-1}m_l)}\left[\prod_{a_t=1}^{A_t}\prod_{l=0}^{L_R-1}\left(\frac{y_{a_t}^l}{\lambda_{a_t}^l}\right)^{m_l}\right]\Upsilon^{(A_t\cdot\sum_{l=0}^{L_R-1}m_l)-1}$$
$$\times\Phi_2^{A_t\cdot L_R}(m_1^0,\ m_1^1,...,m_{A_t}^{L_R-1}\ ;\ A_t\sum_{l=0}^{L_R-1}m_l\ ;$$
$$-\frac{y_1^0}{\lambda_1^0}\gamma,\ -\frac{y_1^1}{\lambda_1^1}\gamma,...,\ -\frac{y_{A_t}^{L_R-1}}{\lambda_2^{L_R-1}}\gamma) \quad (27)$$

where the restriction is considered with $m_{a_t}^l=m_l$ for $a_t=1,\ 2,...,\ A_t$. By comparing (27) with (18) for the independent fading case, we can obtain the outage and average BER expressions for the correlated fading case by replacing the random variable $y_{a_t}^l$ with ($y_{a_t}^l/\lambda_{a_t}^l$). Similarly, the average BER in the spatially correlated Nakagami-m fading channel, $\overline{P}_{BER}^{cor}$, for an MC-CDMA system with combining receiver accompanied by CFO becomes as

$$\overline{P}_{BER}^{cor} = \frac{\Gamma(\frac{1}{2}+A_t\cdot\sum_{l=0}^{L_R-1} m_l)}{2\sqrt{\pi}\,\Gamma(1+A_t\cdot\sum_{l=0}^{L_r-1} m_l)}\left[\prod_{a_t=1}^{A_t}\prod_{l=0}^{L_R-1}\left(\frac{y_{a_t}^l}{y_{a_t}^l+\lambda_{a_t}^l}\right)^{m_l}\right]$$
$$\times F_D^{(A_t\cdot L_R)}\left(\frac{1}{2}+m_1^0,\ m_1^1,\ldots,\ m_{a_t}^l\ ;\ 1+A_t\cdot\sum_{l=0}^{L_R-1} m_l\ ;\ \frac{y_1^0}{y_1^0+\lambda_1^0}\gamma_{th},\ \frac{y_1^1}{y_1^1+\lambda_1^1}\gamma_{th},\ldots,\ \frac{y_{A_t}^{L_R-1}}{y_{A_t}^{L_R-1}+\lambda_{A_t}^{L_R-1}}\gamma_{th}\right) \qquad (28)$$

5. NUMERICAL RESULTS AND DISCUSSION

Generally, it is known that adopts the results from a software simulating package or applies the monte caro are the most useful methods for validating the fact of a some research results. However, sometime it is difficult to implement the scenarios of the assumption for the system works in real world by using of the current software package. In this section some parameters, antenna number (A_t), CFO (ε), MIP's decay exponent (δ), subcarrier (N), and the received path number (L_R), are taken into account for comparison purpose. The system performance influenced by the phenomenon of CFO is shown in Fig. 2, in which the applied parameters are set as, $N=128$, $K=25$, $L=L_R=4$, and $\delta=1$, respectively. It is obviously understood that the more CFO values the worse system performance of the MC-CDMA system is, nevertheless how many the antenna number is. Moreover, the reason for the influence of CFO becomes constant after SNR is greater than about 10 dB, which is claimed that the interference (including CFO) will not dominate the behavior of system performance after the SNR increase an beyond a fixed level. On contrast the results in Fig. 2 to that of in Fig. 3 where the results of average BER function of user's number, K, are presented. By adopting the same parameters in Fig. 2, but the SNR=15dB is fixed now. Same as to the results shown in Fig. 2, the curves shown in Fig. 3 can be observed that it is not only the better system performance can be obtained by reducing the CFO, but the addition of a second antenna can offer considerable improvement on the average BER performance. Next, it is worth noting that the results shown in Fig. 4, in which the situation of that is being described previous with $N=128$, $A_t=2$ and $N=512$, $A_t=1$ are not the same again. It is the reason that caused by the CFO affect. Hence, in short it is able to claim that the effect of CFO definitely can be mitigated by increasing the number of antenna.

6. CONCLUSION

The effect of CFO phenomenon is inspected in the paper for an MC-CDMA system based on the OFDM techniques. It is known that the CFO occurs in the environments where the estimation of the frequency in oscillator is not complete between the transmitter and the receiver. We try to adopt much more antenna number for promoting the capability of frequency estimation so as try to reduce the effect of CFO. Moreover, the analysis of system performance for an MC-CDMA system with and without CFO under the assumptions with some different parameters are presented and discussed. It can be remarked that the system performance of an MC-CDMA system can be improved by the utilization of higher dimension of receiver antenna.

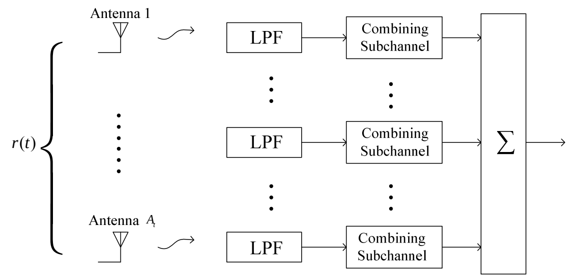

Fig.1. The receiver block diagram of M-D combining receiver for a reference user (0-th user)

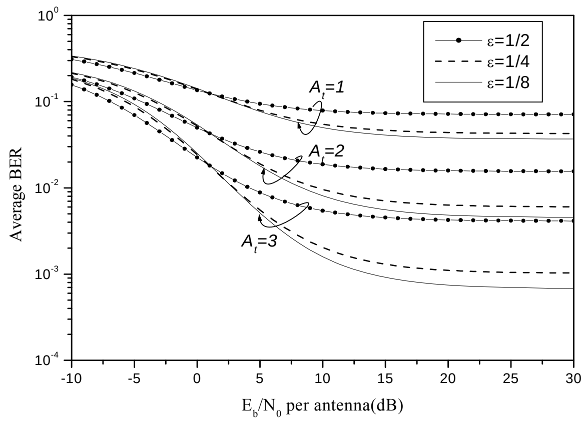

Fig. 2. The plots of BER versus SNR, E_b/N_0, N=128, k=25,L=L_R=4, δ=1, with different CFO values

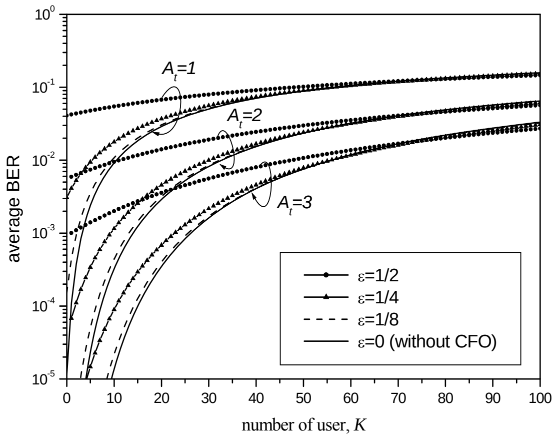

FIG. 3. THE PLOTS OF BER VERSUS NUMBER OF USER, K , N=128, L=LR= 4, Δ=1, SNR=15DB, WITH DIFFERENT CFO VALUES

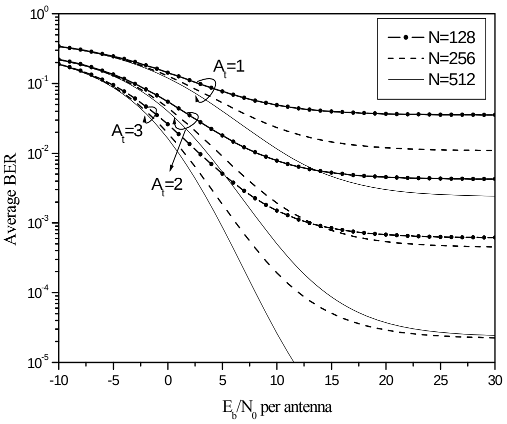

FIG. 4. THE PLOTS OF BER VERSUS SNR, E_b / N_0 , K=25, L=LR= 4, Δ=1, WITH CFO, $\varepsilon = 1/4$, AND DIFFERENT SUBCARRIER NUMBER

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