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Making Carrier Frequency Offset an Advantage for Orthogonal Frequency Division Multiplexing

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***Abstract*—Contrary to the common belief that the carrier frequency offset (CFO) in an orthogonal frequency division multiplexing (OFDM) system would adversely impact on system performance, this paper shows that the CFO actually has the effect of linear precoding among transmitted data symbols and hence can be exploited to improve the diversity performance over frequency-selective fading channels. With both analysis and Monte Carlo simulation, it is proved that an OFDM system with CFO equal to half of the subcarrier spacing can potentially achieve the performance of diversity order four by the maximum-likelihood detection and demonstrate a 5 dB improvement using the minimum mean squared error equalization.**

Keywords—carrier frequency offset (CFO); orthogonal frequency division multiplexing (OFDM); frequency-selective fading; multipath diversity; linear precoding.

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) has been widely used in today's digital communication systems due to its effective inter-symbol interference (ISI) mitigation and simple frequency-domain channel equalization via fast Fourier transform (FFT). However, as has been commonly believed, it suffers from some major disadvantages, such as, (1) the large peak-to-average power ratio (PAPR), (2) the sensitivity to carrier frequency offset (CFO) [1], and (3) the poor frequency diversity performance in frequency-selective fading channels.

The first disadvantage is almost certain since a large PAPR in the OFDM signal waveform not only drives the dynamic range requirements for the digital-to-analog conversion (D/A) and analog-to-digital conversion (A/D) but more importantly also reduces the transmitter and receiver's power amplifier efficiency. The second disadvantage is drawn based on the observation that the CFO causes inter-carrier interference (ICI) and thus frequency synchronization/compensation is necessary. The third disadvantage is a straightforward derivative from the fact that the OFDM converts frequency-selective fading into parallel flat fading on orthogonal subcarriers so that it only achieves diversity order one and hence performs poorly in frequency-selective fading channels. Channel coding has been traditionally used to improve the diversity across frequency and time [2,3], and recently linear precoding and block spreading for OFDM systems have been introduced to improve the frequency diversity performance [4-6].

The essence of precoding for OFDM is to introduce correlations among modulated subcarriers by applying a unitary matrix to the data symbols to be transmitted to obtain different linear combinations of the original data symbols. After subcarrier mapping, the precoded data symbols are spread across the transmission frequency band. Thus, if a subcarrier experiences a deep fade after transmitting over a frequency-selective fading channel, the data symbol can be still recovered from other subcarriers so that the system performance is improved due to the increased diversity order [5]. Examining the effect of CFO from this precoding principle, we see that the so-called ICI caused by the CFO actually reflects the correlation among subcarriers, and hence it should be preserved rather than removed. With the right equalization and detection techniques, the CFO will no longer appear as a disadvantage but an advantage for OFDM.

In this paper, we illustrate how to deal with the CFO in a different way, not as an interference maker but as an effective means to combat frequency-selective fading. The conventional time-domain CFO compensation plus frequency-domain equalization approach is replaced by a new frequency-domain equalization plus interpolation approach. With both analysis and Monte Carlo simulation, we also reveal the performance lower bounds by the maximum-likelihood (ML) detection and the potential performance improvement using the more practical minimum mean squared error (MMSE) equalization.

The rest of the paper is organized as follows. In Section II, the received OFDM signal model with CFO is formulated. In Section III, the conventional and new approaches to deal with the CFO are illustrated and compared. Section IV is devoted to the theoretical analysis of the performance lower bounds by the ML detection and the performance using the MMSE equalization. Monte Carlo simulation results are given in Section V. Finally, conclusions are drawn in Section VI.

II. RECEIVED OFDM SIGNAL MODEL WITH CFO

An OFDM signal $x[n]$ is generated by performing an N-point inverse discrete Fourier transform (IDFT) on a block of data symbols $X[k]$, $0 \le k \le N-1$, after binary phase shift keying (BPSK), quadrature phase shift keying (QPSK), or any other quadrature amplitude modulation (QAM) constellation mapping of the input data bits, i.e.,

$$x[n] = \frac{1}{N}\sum_{k=0}^{N-1} X[k] e^{j\frac{2\pi}{N}kn}, \quad 0 \le n \le N-1. \tag{1}$$

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Before transmitting into a frequency-selective multipath fading channel with discrete channel impulse response $h[n]$, $n = 0,1,\cdots, L-1$, where L is the maximum multipath delay in samples, a cyclic prefix (CP) of N_{CP} samples, $N_{CP} \ge L$, is inserted in front of $x[n]$ to avoid adjacent OFDM symbol interference and turn the linear convolution of the transmitted signal with the channel into a circular one. Thus, the total number of signal samples in an OFDM symbol becomes $N + N_{CP}$, which corresponds to an OFDM symbol duration $(N + N_{CP})T$ after D/A, where T is the sampling period.

At the receiver baseband, after A/D and CP removal, the received OFDM signal can be modeled as

$$r[n] = (h[n] \otimes_N x[n]) e^{j\omega_0 n} + v[n],\ 0 \le n \le N-1, \qquad (2)$$

where $\otimes_N$ denotes the circular convolution of length N, ω_0 is a digital frequency shift due to the CFO ΔF between the transmitter and receiver's local oscillators, which can be expressed as

$$\omega_0 = 2\pi \Delta F T = \frac{2\pi}{N}\eta \qquad (3)$$

where $\eta = \Delta F N T$ is defined as the normalized carrier frequency offset with respect to the subcarrier spacing $\frac{1}{NT}$, and $v[n]$ is the additive zero-mean white Gaussian noise.

From (2), the N-point discrete Fourier transform (DFT) of $r[n]$ can be derived as

$$R[k] = H\left(e^{j\frac{2\pi}{N}(k-\eta)}\right) X\left(e^{j\frac{2\pi}{N}(k-\eta)}\right) + V[k]$$
$$= H_\eta[k] X_\eta[k] + V[k] \qquad (4)$$

where $H\left(e^{j\frac{2\pi}{N}(k-\eta)}\right)$ and $X\left(e^{j\frac{2\pi}{N}(k-\eta)}\right)$, denoted as $H_\eta[k]$ and $X_\eta[k]$ respectively for simplicity, are sampled Fourier transforms $H(e^{j\omega})$ and $X(e^{j\omega})$ of $h[n]$ and $x[n]$ respectively, and $V[k]$ is the N-point DFT of $v[n]$. When η =0, $H_\eta[k]$ becomes the N-point DFT $H[k]$ of $h[n]$, and $X_\eta[k]$ becomes the N-point DFT $X[k]$ of $x[n]$, i.e., the transmitted data symbols. According to the relationship between the Fourier transform and the discrete Fourier transform of a finite length sequence [7], $X_\eta[k]$ can be interpolated from $X[k]$ by

$$X_\eta[k] = \sum_{l=0}^{N-1} \Phi_N\left(\frac{2\pi}{N}(k-\eta-l)\right) X[l] \qquad (5)$$

where the interpolation function is defined as

$$\Phi_N(\omega) = \frac{\sin\left(\frac{\omega N}{2}\right)}{N \sin\left(\frac{\omega}{2}\right)} e^{-j\frac{\omega(N-1)}{2}}. \qquad (6)$$

III. NEW APPROACHES DEALING WITH CFO

To recover the transmitted data symbol $X[k]$, carrier frequency offset compensation is conventionally applied on $r[n]$ in the time-domain first (i.e., $r[n]e^{-j\omega_0 n}$ is calculated to shift by a frequency offset $-\eta$), and then FFT is performed to produce the product of $X[k]$ with $H[k]$. Following a simple one-tap frequency-domain equalization (dividing by $H[k]$, for example), $X[k]$ is finally recovered. The above process can be illustrated in Fig. 1 (a), (b), and (d) for N =7, $X[k]$ =1, $0 \le k \le N-1$, and η =0.5 (ignoring the noise for simplicity), where (a) and (b) show the frequency-domain representations of the received OFDM signal before and after CFO compensation respectively, and (d) shows the recovered OFDM data symbols after equalization.

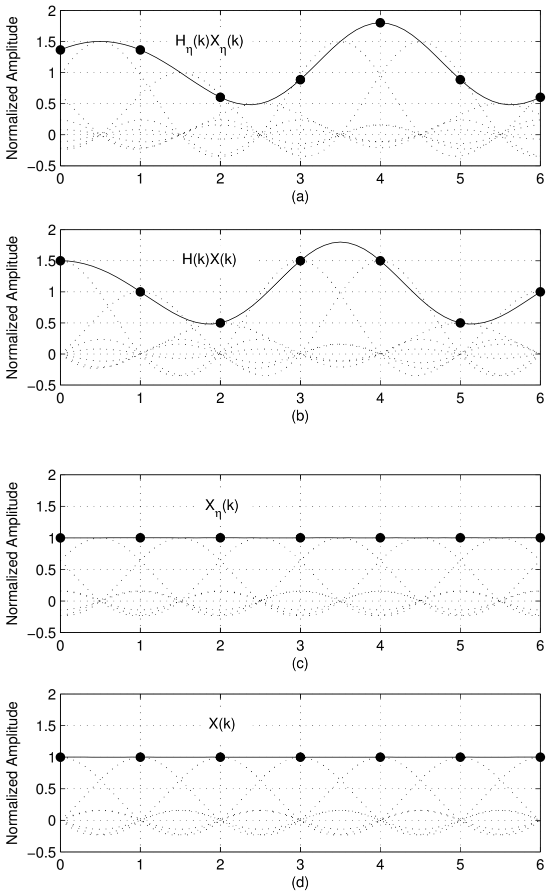

Fig. 1. Frequency-domain representations of (a) received OFDM signal with CFO, (b) received OFDM signal after CFO compensation, (c) interpolated data symbols after equalization without CFO compensation, and (d) recovered data symbols (solid dots: DFT; solid line: Fourier transform; dotted line: interpolation functions).

In a flat fading channel, the time-domain CFO compensation followed by the one-tap frequency-domain equalization works well for the OFDM system. However, in a frequency-selective multipath fading channel, this approach can only achieve the performance of diversity order one (i.e., the performance for flat fading), because the data symbols modulated on different subcarriers are independent after CFO compensation, and hence there is no correlation among subcarriers to explore for the recovery of a deep fade on a subcarrier.

In terms of introducing correlation among transmitted data symbols, the CFO now turns to be an advantage rather than a disadvantage. This can be easily seen from (5), which clearly shows that, when $\eta \neq 0$, $X_\eta[k]$ is a linear combination of $X[l]$, $0 \le l \le N-1$. Thus, instead of being compensated, the CFO should be kept. Then, new equalization and detection techniques can be developed to explore this advantage.

An example of the new approaches dealing with CFO can be illustrated using Fig. 1 (a), (c), and (d). Instead of being shifted by a frequency offset $-\eta$ via the time-domain multiplication, the DFT of the received OFDM signal $R[k] = H_\eta[k]X_\eta[k]$ (ignoring the noise for simplicity) shown in Fig. 1 (a) is first divided by $H_\eta[k]$ (i.e., zero-forcing equalization) to produce the interpolated data symbols $X_\eta[k]$ shown in Fig. 1 (c). Then, $X[k]$ is recovered, see Fig. 1 (d), by interpolating $X_\eta[k]$ using the inverse operation of (5). Compared with the conventional approach, this new approach reverses the order of CFO compensation and equalization, and the CFO compensation is replaced by interpolation in the frequency-domain.

Though the above simple zero-forcing equalization could possibly improve the diversity performance, the optimum technique will be the ML detection. In order to describe these new approaches better, we now express the received OFDM signal model in matrix form. By combining (4) and (5) together, we have

$$\mathbf{R} = \mathbf{H}_\eta \mathbf{U}_\eta \mathbf{X} + \mathbf{V} \tag{7}$$

where $\mathbf{R} = (R[k])_{N\times 1}$, $\mathbf{X} = (X[k])_{N\times 1}$, $\mathbf{V} = (V[k])_{N\times 1}$ are $N\times 1$ column vectors, $\mathbf{H}_\eta = diag(H_\eta(k))_{N\times N}$ is an $N\times N$ diagonal matrix, and $\mathbf{U}_\eta = (u_\eta(k,l))_{N\times N}$ is an $N\times N$ unitary matrix with the element at the k th row and the l th column as $u_\eta(k,l) = \Phi_N\left(\frac{2\pi}{N}(k-\eta-l)\right)$, which satisfies the property $\mathbf{U}_\eta\mathbf{U}'_\eta = \mathbf{U}'_\eta\mathbf{U}_\eta = \mathbf{I}$, where $(\cdot)'$ denotes matrix transposition and complex-conjugation operation and $\mathbf{I}$ is the identity matrix of order N.

From (7), we see that when there is a carrier frequency offset present at the receiver the OFDM system is equivalent to a precoded system with the precoding matrix $\mathbf{U}_\eta$ as defined above. The precoded symbols in $\mathbf{U}_\eta\mathbf{X}$ are then mapped onto subcarriers equally spaced across the transmission band and experience the channel fading represented by $\mathbf{H}_\eta$.

Assuming perfect channel knowledge at the receiver, the ML estimate $\hat{\mathbf{X}}$ of the data vector $\mathbf{X}$ can be obtained by minimizing the quantity

$$\left(\mathbf{R} - \mathbf{H}_\eta \mathbf{U}_\eta \hat{\mathbf{X}}\right)'\left(\mathbf{R} - \mathbf{H}_\eta \mathbf{U}_\eta \hat{\mathbf{X}}\right) \tag{8}$$

through exhaust search from all possible date vectors.

Due to the complexity of the optimum ML detection, a linear equalization is more practical, since it can simply use a one-tap equalizer for each subcarrier in the frequency-domain, as has been seen in the previous example. The equalization and detection process can be generally described as follows. Let $C[k]$ denote the one-tap equalizer coefficient to be applied to $R[k]$ on the subcarrier k and $\mathbf{C} = diag(C[k])_{N\times N}$ denote an $N\times N$ diagonal matrix with diagonal elements $C[k]$, $k = 0,1,\cdots,N-1$. First, applying $\mathbf{C}$ to $\mathbf{R}$ produces the equalized precoded data vector $\mathbf{CR}$. Second, multiplying $\mathbf{U}'_\eta$ to remove the precoding yields the decision variable vector $\mathbf{d} = \mathbf{U}'_\eta\mathbf{CR}$. Finally, an estimate $\hat{\mathbf{X}}$ of the transmitted data vector is obtained after hard decision.

IV. BER Analysis in Presence of CFO

A. BER lower bound of ML detection

Since (8) can be expanded as

$$\mathbf{R}'\mathbf{R} - 2\,\mathrm{Re}\left\{\mathbf{R}'\mathbf{H}_\eta\mathbf{U}_\eta\hat{\mathbf{X}}\right\} + \hat{\mathbf{X}}'\mathbf{U}'_\eta\mathbf{H}'_\eta\mathbf{H}_\eta\mathbf{U}_\eta\hat{\mathbf{X}} \tag{9}$$

and $\mathbf{R}'\mathbf{R}$ is independent of $\hat{\mathbf{X}}$, the ML detection can be carried out by searching for $\hat{\mathbf{X}}$ to maximize the quantity

$$\Omega(\hat{\mathbf{X}}) = 2\,\mathrm{Re}\left\{\mathbf{R}'\mathbf{H}_\eta\mathbf{U}_\eta\hat{\mathbf{X}}\right\} - \hat{\mathbf{X}}'\mathbf{U}'_\eta\mathbf{H}'_\eta\mathbf{H}_\eta\mathbf{U}_\eta\hat{\mathbf{X}}\,. \tag{10}$$

Let the transmitted data symbol vector be $\mathbf{X}$. Expressing the estimate $\hat{\mathbf{X}}$ as $\hat{\mathbf{X}} = \mathbf{X} + \mathbf{e}$, where $\mathbf{e}$ is an error vector, and substituting (7) into (10) yield

$$\Omega(\mathbf{X}+\mathbf{e}) - \Omega(\mathbf{X}) = 2\,\mathrm{Re}\left\{\mathbf{e}'\mathbf{U}'_\eta\mathbf{H}'_\eta\mathbf{V}\right\} - \mathbf{e}'\mathbf{U}'_\eta\mathbf{H}'_\eta\mathbf{H}_\eta\mathbf{U}_\eta\mathbf{e}\,. \tag{11}$$

We see that, given an error vector $\mathbf{e}$, $\Omega(\mathbf{X}+\mathbf{e}) - \Omega(\mathbf{X})$ is a Gaussian distributed variable with mean $-\delta$ and variance $2\sigma_V^2\delta$, where $\delta = \mathbf{e}'\mathbf{U}'_\eta\mathbf{H}'_\eta\mathbf{H}_\eta\mathbf{U}_\eta\mathbf{e}$ is referred to as the distance between $\mathbf{X}+\mathbf{e}$ and $\mathbf{X}$ after precoding and multipath channel, and σ_V^2 is the variance of the noise $V[k]$. Suppose that we use the Gray-coded QPSK (i.e., there is only one bit difference between two adjacent constellation points) for the bit-to-symbol mapping before precoding. Thus, any one bit error occurring at the k th data symbol in $\hat{\mathbf{X}}$ will result in an error vector

$$\mathbf{e}_k = (0,\cdots,0,\underbrace{\pm\sqrt{2}\sigma_X}_{k\text{th element}},0,\cdots,0)^T \text{ or } \mathbf{e}_k = (0,\cdots,0,\underbrace{\pm j\sqrt{2}\sigma_X}_{k\text{th element}},0,\cdots,0)^T \tag{12}$$

where $(\cdot)^T$ denotes matrix transposition, and σ_X^2 is the average power of the data symbol. The mean and variance of $\Omega(\mathbf{X}+\mathbf{e}_k) - \Omega(\mathbf{X})$ are therefore found to be $-\delta_k$ and $2\sigma_V^2\delta_k$ respectively, where $\delta_k = 2\sigma_X^2\sum_{l=0}^{N-1}\left|H_\eta[l]u_\eta(l,k)\right|^2$. According to the ML detection principle, if $\Omega(\mathbf{X}+\mathbf{e}_k) > \Omega(\mathbf{X})$, then $\mathbf{X}+\mathbf{e}_k$ will be declared as the detected data vector and hence one bit error occurs. Therefore, the probability for one bit error is evaluated as the probability with which $\Omega(\mathbf{X}+\mathbf{e}_k)-\Omega(\mathbf{X})>0$, i.e.

$$P(\Omega(\mathbf{X}+\mathbf{e}_k)-\Omega(\mathbf{X})>0)$$
$$=Q\left(\sqrt{\frac{\delta_k}{2\sigma_V^2}}\right)=Q\left(\sqrt{\gamma_{in}\sum_{l=0}^{N-1}\left|H_\eta[l]u_\eta(l,k)\right|^2}\right) \quad (13)$$

where $\gamma_{in}=\sigma_X^2/\sigma_V^2$ is the input signal-to-noise ratio (SNR) and $Q(\cdot)$ is the Q-function.

Note that the BER of the ML detection is determined by the minimum distance $\delta_{\min}$, which requires an exhaust search of the error vector(s) leading to the minimum distance. However, since the distance given by the error vector $\mathbf{e}_k$ is always greater than or equal to the minimum distance, the one bit error probability (13) can serve as a lower bound of the BER.

Also note that the above BER lower bound relies on a realization of the channel frequency response $H_\eta[l]$, $0\le l\le N-1$, or equivalently, the channel impulse response $h[n]$. Thus, for a frequency-selective channel, the average BER lower bound will be

$$P_e^{ML}=E_h\left\{\frac{1}{N}\sum_{k=0}^{N-1}Q\left(\sqrt{\gamma_{in}\sum_{l=0}^{N-1}\left|H_\eta[l]u_\eta(l,k)\right|^2}\right)\right\} \quad (14)$$

where $E_h\{\}$ denotes the ensemble averaging over all $h[n]$.

B. BER of MMSE Equalization

For linear equalization, when the MMSE criterion is used, i.e., designing $\mathbf{C}$ so that the mean squared error (MSE) between $\mathbf{d}$ and $\mathbf{X}$

$$\varepsilon^2=E\left\{(\mathbf{d}-\mathbf{X})'(\mathbf{d}-\mathbf{X})\right\} \quad (15)$$

is minimized, the diagonal element in $\mathbf{C}$ is found to be

$$C[k]=\frac{H_\eta^*[k]}{\left|H_\eta[k]\right|^2+\dfrac{1}{\gamma_{in}}} \quad (16)$$

and the output SNR in the decision variable $\mathbf{d}$ for data symbol $X[k]$ can be expressed as [8]

$$\gamma_\eta(k)=\frac{\displaystyle\sum_{l=0}^{N-1}\frac{\left|H_\eta[l]u_\eta(l,k)\right|^2}{\left|H_\eta[l]\right|^2+\frac{1}{\gamma_{in}}}}{1-\displaystyle\sum_{l=0}^{N-1}\frac{\left|H_\eta[l]u_\eta(l,k)\right|^2}{\left|H_\eta[l]\right|^2+\frac{1}{\gamma_{in}}}}. \quad (17)$$

We see that the output SNR is also determined by $H_\eta[l]$, or equivalently, $h[n]$. Assuming QPSK modulation for data symbols and making a Gaussian distribution approximation for ISI, the average BER of the equalizer for a given realization of $h[n]$ can be evaluated as $\frac{1}{N}\sum_{k=0}^{N-1}Q\left(\sqrt{\gamma_\eta(k)}\right)$, and consequently the average BER for a frequency-selective fading channel is evaluated as

$$P_e^{MMSE}=E_h\left\{\frac{1}{N}\sum_{k=0}^{N-1}Q\left(\sqrt{\gamma_\eta(k)}\right)\right\}. \quad (18)$$

V. MONTE CARLO SIMULATION RESULTS

The performance P_e^{ML} and P_e^{MMSE} analyzed in the above section are evaluated using the Monte Carlo simulation assuming that the channel has a full multipath diversity of order L. That is, all channel coefficients $h[n]$, $n=0,1,\cdots,L-1$, are independent and identically distributed (i.i.d.) complex Gaussian random variables with zero-mean and variance $1/L$. To evaluate the ensemble average over all $h[n]$, we generate sufficient realizations of these independent Gaussian variables, calculate the BER for each realization, and then take an average.

To show the performance potential of OFDM with CFO, we first assume that the channel provides the maximum diversity order, i.e., $L=N$, and evaluate the BER lower bounds of the ML detection and the BERs of the MMSE equalization under different CFOs. The input SNR γ_{in} is expressed as $2\frac{E_b}{N_0}$ for QPSK, where E_b is the signal energy per bit and N_0 is the noise power spectral density. The results are shown in Fig. 2 with N =256. Different numbers of N, such as 16, 32, 64, and 128, are also tested, and the results are all the same. From Fig. 2 we see that when η =±0.5 the best performance is achieved. When η =0, the performance is the worst, which is the same as the one in flat fading (diversity order one). For the ML detection the best performance achieves a diversity order of four (see [9]) and for the MMSE equalization the best performance shows a more than 4 dB improvement at 15 dB normalized SNR and 5 dB at 20 dB normalized SNR.

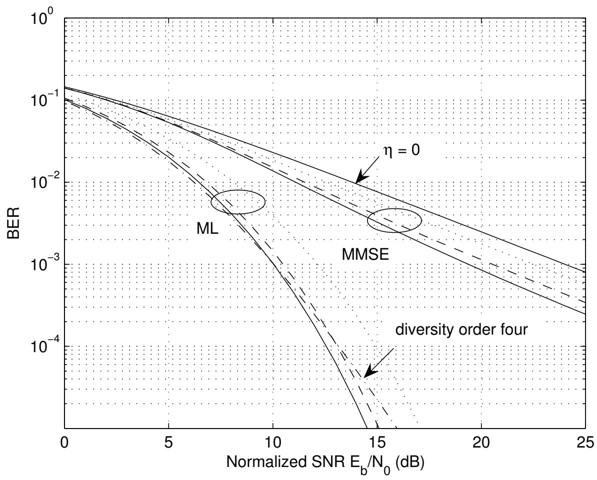

Fig. 2. Performance potential of OFDM with CFO (solid lines for η =±0.5; dashed lines for η =±0.375 or ±0.625; dotted lines for η =±0.25 or ±0.75).

Fig. 3 shows the performance for a more practical system setting with $L = N/4$ and N =128. As the diversity order provided by the channel decreases, the performance is degraded accordingly. For comparison purpose, the performance curves with $L = N$ are also displayed. We see that the ML lower bound has about 3dB loss but is still much better than the performance of diversity order two [9]. The MMSE equalization incurs about 1 dB loss but still provide much better performance than that of the conventional OFDM without CFO or with CFO compensation.

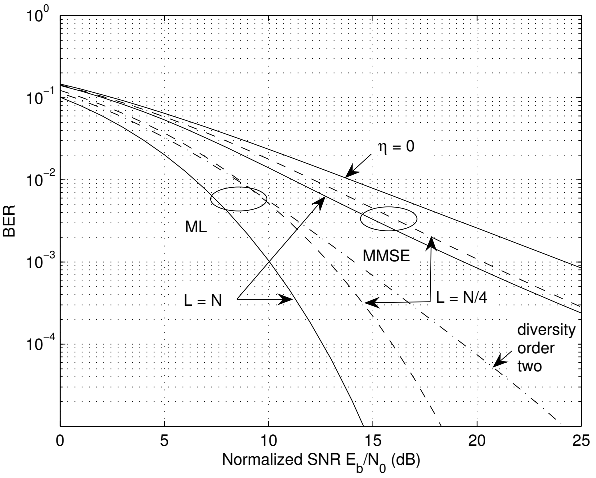

Fig. 3. Performance of a practical OFDM with CFO η =±0.5 and N =128.

The above performance results suggest that when the right equalization and detection techniques are used the CFO in an OFDM system should be set to half of the subcarrier spacing in order to achieve the best diversity performance. This observation leads to the new receiver architecture to conduct frequency synchronization in an OFDM system, i.e., after the CFO estimation, the CFO should be adjusted to half of the subcarrier spacing rather than be compensated.

As a final remark on the receiver complexity, we point out that the complexity would be the same as that of a precoded OFDM with the same precoding matrix size. However, by exploiting the characteristics of the interpolation function, the complexity can be greatly reduced, i.e., the frequency-domain interpolation can be simplified by considering only several adjacent subcarriers. This can be easily seen from the interpolation function amplitude shown in Fig. 4. Because the significant values of the interpolation function are located around k =0, a subcarrier can be simply interpolated using only several adjacent subcarriers. Furthermore, if proper pre and post processing is performed before and after the frequency-domain interpolation respectively, a real valued interpolation function can be used, and hence further overall complexity reduction is possible. Since our purpose here is to demonstrate the performance potential by exploiting the carrier frequency offset, the complexity reduction and how this reduction impacts on system performance are beyond the scope of this paper.

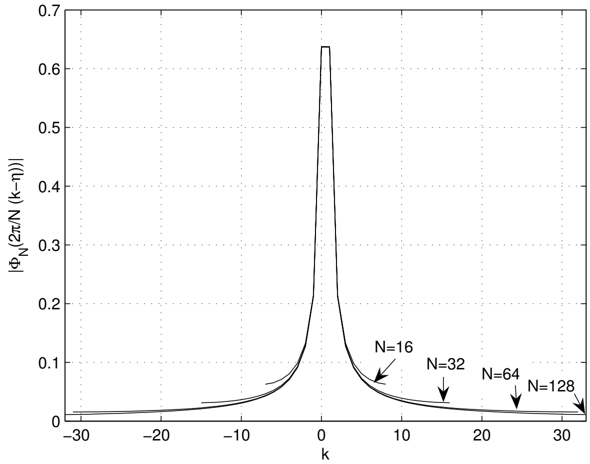

Fig. 4. Amplitude of the interpolation function $\Phi_N(\omega)$ sampled at $\omega = \frac{2\pi}{N}(k - \eta)$ with η =0.5 for different N .

VI. Conclusions

We have shown that the CFO in an OFDM system introduces correlation among modulated subcarriers. It achieves the same effect as linear precoding but without explicit precoding operation at the transmitter. In terms of combating frequency-selective multipath fading, the CFO is actually beneficial rather than destructive. Instead of being compensated as impairment, the CFO should be set to half of the subcarrier spacing in order to make full use of this diversity advantage.

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