Receivers Design for OFDM Signals under both High Mobility and Carrier Frequency Offset

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***Abstract*—In highly mobile scenarios, mobility-induced channel time variation (CTV) will destroy the orthogonality among subcarriers in orthogonal frequency division multiplexing (OFDM) systems, which causes intercarrier interference (ICI) and thus degrades the system performance. On the other hand, carrier frequency offset (CFO) between the oscillators of the transmitter and the receiver also causes ICI. As a result, when the CFO and the channel time variation coexist, ICI caused by the two factors become more severe and thus complicate the receiver design. In order to cope with the ICI caused by these two factors, this paper investigates two receiver structures for OFDM systems. One separately addresses the effects of the CFO and the channel time variation, and the other makes an integrated compensation of the two factors. Performances of the two receivers are studied and compared through numerical simulations. Results obtained from simulations show that the separate receiver takes advantages over the integrated one, and is an appropriate option for OFDM receiver under both high mobility and carrier frequency offset.**

***Index Terms*—orthogonal frequency division multiplexing, carrier frequency offset, time varying channel, equalizer**

I. Introduction

OFDM is a frequency efficient transmission technique owing to its capability of maintaining the subcarriers orthogonality. Moreover, it is computationally efficient as the modulation and demodulation can be realized by using fast discrete Fourier transform (FFT) [1]. Till now, OFDM has become an important transmission technique in various wireless communication standards including IEEE 802.11a, 802.16e, 802.20 and 802.22 etc [2]. The aforementioned merit of such a transmission technique is obtained assuming no channel variation for the duration of an OFDM symbol [3]. In mobile application environments, however, there exist many challenges for this technique to be addressed. For one thing, the channel time variation (CTV) in mobile environments will destroy the orthogonality among subcarriers and result in intercarrier interference (ICI) which thereby degrades the system performance. For the other, carrier frequency offset [4], the frequency discrepancies between the oscillators of the transmitter and the receiver, will also destroy the subcarrier orthogonality and cause ICI. As a result, the ICI will be severe under both the CFO and the channel time variation, so in order to reduce their effects, associated mitigation method is indispensable.

For mitigating ICI caused by the channel time variation, Stamoulis *et al.* proposed a time-domain equalization approach by designing a series of block ICI-mitigating filters [5]. Compared with the time-domain approach, the method in the frequency domain is more advantageous in the sense of the compromise between system performance and complexity. So rather than still working in the time domain, Choi *et al.* proposed a minimum mean squared error based frequency-domain equalization method [6]. A remarkable feature of this method is its ability of exploiting the inherent time diversity provided by the channel time variation, achieved by using an ordered serial interference cancellation (OSIC) during the MMSE equalization of OFDM symbols. Unfortunately, however, this method has an unaffordable computational load as it involves tedious subcarriers sorting and complex matrix inversion. With the aim of reducing computational complexity of Ref. [6] and based on the observation of ICI diminishing rapidly with subcarrier intervals increasing, Cai and Giannakis proposed a reduced-complexity MMSE equalizer with SIC by neglecting the ICI from subcarriers with far spacing [7]. Compared with Ref. [6], the complexity of Ref. [7] was reduced from $O(N^4)$ to $O(N^2)$, where N is the symbol length in terms of subcarriers, unfortunately however, such a computational load is still high for practical applications. For further reducing the computational complexity, [8]-[10] independently proposed several ICI equalization methods of linear complexity with the symbol length. Fang *et al.* proposed a block turbo MMSE equalizer by exploiting the band structure of the channel frequency domain (CFR) matrix [8]. In Ref. [8], the intensive computations of matrix inversion is lessened by means of a band LDL factorization. In Ref. [9], Schniter proposed a two-stage turbo equalizer also by exploiting the band structure of the CFR matrix. In this algorithm, the first stage used an optimal linear processing filter for windowing the received signal in order to restrict the ICI support. Using the windowed signal, the second stage adopted a serial turbo equalizer to recover the transmit symbols. In addition to the turbo equalizers [8-9], Li *et al.* proposed a one-tap frequency-domain equalizer with parallel ICI cancellation [10]. In the parallel interference cancellation of Ref. [10], only parts of the neighboring ICI are considered in

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order to reduce the computational complexity which thereby sacrificed the equalization performance. In [11], a CFO compensation technique was proposed to improve the channel estimation reliability using piece-wise models, but its focus is on reducing the ICI impact on the channel estimation.

As stated previously, when the channel time variation and the CFO coexist, the ICI becomes more severe. Existing OFDM receiver designs generally consider only one or neither of the two factors of the CFO and the channel time variation [12-13]. This doe not conform to the actual conditions of mobile applications. By now, few literatures addressed the receiver designs for OFDM system under such cases. In this paper, we consider the receiver design for OFDM operating over time varying channels and in the presence of CFO. Two types of receivers, the hybrid time-frequency domain receiver and the pure frequency domain receiver, are investigated and compared under the considered cases. The hybrid time-frequency domain receiver works in two stages. In the first stage, the effect of CFO is compensated from the received signals in the time domain. As the CFO has been compensated in the first stage, the second stage only needs to cope with the channel time variation. Specifically, the received signals after CFO compensation are firstly converted into the frequency domain, and then feeds into an ICI equalizer to recover the transmit symbols. The frequency domain receiver, as its name indicates, only works in the frequency domain. After converting the overall received signal from the time domain to the frequency domain, it jointly equalizes the channel time variation and the CFO. In order to evaluate the two receivers, comparisons in terms of system performance and complexity are made through numerical simulations.

The rest of the paper is organized as follows. Section Ⅱ presents the system model for OFDM under the effect of both time varying channels and CFO. Section Ⅲ addresses the receiver design for OFDM affected by time varying channels and CFO. Section Ⅳ presents some simulations to evaluate the receiver performance. Finally, section Ⅴ concludes with a summary of some key results.

II. SYSTEM MODEL WITH BOTH CFO AND CTV

In the transmitter, the bit streams are mapped into complex symbols $X(k)$ according to the adopted modulation scheme such as QPSK. After a serial to parallel conversion, the QPSK symbols are converted into the time domain by using an N-point inverse DFT. The resulting OFDM modulated signal is given by

$$x(n)=\frac{1}{\sqrt{N}}\sum_{k=0}^{N-1}X(k)\exp(\mathrm{j}2\pi kn/N)\quad n=-G,\cdots,N-1,\quad (1)$$

where N and $X(k)$ are the OFDM symbol length in terms of subcarriers number and the data modulated on the kth subcarrier, respectively. G is the cyclic prefix (CP) length bigger than the maximal channel delay. The OFDM signals $\{x(n),\ n=-G,\ \ldots,\ N-1\}$ are transmitted over doubly selective channels. After removing CP samples at the receiver, we get the received signal in the time domain [7]

$$r(n)=\mathrm{e}^{\mathrm{j}2\pi\frac{\varepsilon n}{N}}\sum_{l=0}^{L-1}h(n,l)x(n-l)+w(n),\qquad (2)$$

where ε is the CFO between the oscillators of the transmitter and the receiver; $h(n, l)$ is the channel impulse response (CIR) corresponding to the lth transmission path at time n; $w(n)$ is the additive white Gaussian noise (AWGN) with zeros mean and variance of σ^2. After discarding the CP and performing an N-point DFT to the received signal, we arrive at [7]

$$\begin{aligned}R(k)&=\frac{1}{\sqrt{N}}\sum_{n=0}^{N-1}r(n)\exp(-\mathrm{j}2\pi kn/N)\\&=H(k,k)X(k)+\sum_{u=0,k\neq u}^{N-1}H(k,u)X(u)+W(k),\end{aligned}\qquad (3)$$

where $R(k)$ and $H(k, k)$ are the received signal in the frequency domain and the associated channel complex gains on the kth subcarrier, respectively; $W(k)$ is the additive noise in the frequency domain. The second term in the right hand side (RHS) of (3) is the ICI superimposed on the kth subcarrier and $H(k, u)$ denotes the weighted ICI coefficient from the uth subcarrier. The formulas of $H(k, k)$ and $H(k, u)$ are given by

$$H(k,k)=\sum_{l=0}^{L-1}\bar{h}(l)\exp(-\mathrm{j}2\pi kl/N)\qquad (4)$$

and

$$H(k,u)=\sum_{l=0}^{L-1}\left(\frac{1}{N}\sum_{n=0}^{N-1}h(n,l)\mathrm{e}^{-\mathrm{j}\frac{2\pi}{N}(k-u-\varepsilon)n}\right)\mathrm{e}^{-\mathrm{j}\frac{2\pi}{N}ul},\qquad (5)$$

respectively. $\bar{h}(l)$ in (4) is defined as the time-averaged product of the CIR and the CFO over a symbol:

$$\bar{h}(l)=\frac{1}{N}\sum_{n=0}^{N-1}h(n,l)\exp(\mathrm{j}2\pi\varepsilon n/N).\qquad (6)$$

When the channel is static and the CFO is zero, $H(k, u)$=0 for any $k\neq u$, i.e., ICI no longer exist under time invariant channels and perfect carrier synchronization. If the channel is static, $H(k, u)$ becomes [4,14-15]

$$H(k,u)=\frac{\sin(\pi\varepsilon)}{N\sin\left(\frac{\pi(\varepsilon+u-k)}{N}\right)}\exp\left(\frac{\mathrm{j}\pi(N-1)\varepsilon-u+k}{N}\right).\qquad (7)$$

For the convenience of derivations, (3) can be further written in a vector form [7]:

$$\boldsymbol{R} = \boldsymbol{H}\boldsymbol{X} + \boldsymbol{W}, \tag{8}$$

where $\boldsymbol{R} = [R(0), R(1), \ldots, R(N\text{-}1)]^{\mathrm{T}}$ is defined as he received signal vector; $\boldsymbol{X} = [X(0), X(1), \ldots, X(N\text{-}1)]^{\mathrm{T}}$ is the transmitted symbol vector and $\boldsymbol{W} = [W(0), W(1), \ldots, W(N\text{-}1)]^{\mathrm{T}}$ is the superimposed noise vector. $\boldsymbol{H}$ is an $N\times N$ channel frequency response matrix:

$$\boldsymbol{H} = \begin{bmatrix} H(1,1) & H(1,2) & \cdots & H(1,N) \\ H(2,1) & H(2,2) & \cdots & H(2,N) \\ \vdots & \vdots & \ddots & \vdots \\ H(N,1) & H(N,2) & \cdots & H(N,N) \end{bmatrix}. \tag{9}$$

As can be seen from (9), non-zero elements on the non-diagonal matrix $\boldsymbol{H}$ incur ICI that need be addressed in the receiver designing.

III. OFDM RECEIVERS DESIGN UNDER BOTH CFO AND CTV

Based on the signal models derived in Section 2, this section investigates two types of receivers, the hybrid time-frequency domain receiver and the pure frequency domain receiver, in presence of both the CFO and the CTV. In the hybrid receiver, effects of the CFO and the channel time variation are compensated separately. Specifically, the CFO is firstly removed from the received time-domain signals. The resultant received signals are then converted from the time domain into the frequency domain. Finally the ICI only caused by the channel time variation are mitigated in the frequency domain. As compared with the hybrid/separate receiver, the pure frequency receiver works in a different manner. It jointly addresses the effects of the two factors in the frequency domain. For the convenience of discussion, two receivers of the separate and the joint/integrated are denoted by the type-I receiver and the type-II receiver respectively.

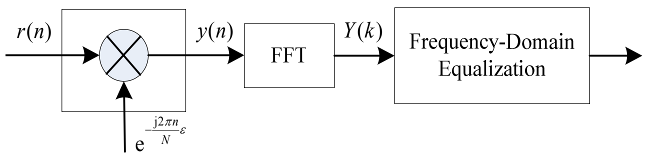

Fig. 1. Hybrid time-frequency domain receiver.

A. Hybrid time-frequency domain receiver

The hybrid time-frequency domain receiver, as shown in Fig 1, is based on the received signal model (2). In this structure, the CFO is firstly removed by multiplying the received signal (2) with sequences $\exp(-\mathrm{j}2\pi\varepsilon n/N), n = 0, \cdots, N-1$. The resulting received signal after CFO compensation is given by

$$y(n) = \sum_{l=0}^{L-1} h(n,l)x(n-l) + v(n), \tag{10}$$

where $v(n)=w(n)\exp(-\mathrm{j}2\pi\varepsilon n/N)$. As $w(n)$ is identically and independently distributed (*i.i.d.*), $v(n)$ is also *i.i.d.* and has the identical distributions as $w(n)$. Performing an N-point DFT to (10), we arrive at

$$\begin{aligned} Y(k) &= \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} y(n) \exp(-\mathrm{j}2\pi kn/N) \\ &= G(k,k)X(k) + \sum_{u=0,k\neq u}^{N-1} G(k,u)X(u) + V(k), \end{aligned} \tag{11}$$

where $G(k, k)$ and $G(k, u)$ are the special cases of $H(k, k)$ and $H(k, u)$, respectively, with the exception of $\varepsilon = 0$.

Similar as (8), for the derivation convenience, (11) can be written in a vector form by (12)

$$\boldsymbol{Y} = \boldsymbol{G}\boldsymbol{X} + \boldsymbol{V}, \tag{12}$$

where $\boldsymbol{G}$ is the ICI coupling matrix defined as

$$\boldsymbol{G} = \begin{bmatrix} G(1,1) & G(1,2) & \cdots & G(1,N) \\ G(2,1) & G(2,2) & \cdots & G(2,N) \\ \vdots & \vdots & \ddots & \vdots \\ G(N,1) & G(N,2) & \cdots & G(N,N) \end{bmatrix}. \tag{13}$$

Comparing (12) with (8), we find they are in a similar form with exception that the ICI coefficients in the ICI coupling matrix $\boldsymbol{G}$ are caused only by the CTV while in $\boldsymbol{H}$ they are caused by both the CFO and the CTV. As the received signal is only affected by the CTV, the following task of the receiver is to mitigate the CTV induced ICI in the frequency domain. From (12), the equalized symbol $\boldsymbol{X}$ based on the MMSE criteria is given by

$$\boldsymbol{X} = (\boldsymbol{G}^{\mathrm{H}}\boldsymbol{G} + \sigma^2\boldsymbol{I})^{-1}\boldsymbol{G}^{\mathrm{H}}\boldsymbol{Y}, \tag{14}$$

where $\boldsymbol{I}$ is an $N\times N$ diagonal identity matrix. As can be seen from (14), direct MMSE equalization in the presence of ICI involves $N\times N$ matrix inversion which is a heavily computational task for practical realization. Fortunately, previous study has shown that the ICI caused by only CTV diminishes rapidly with the increasing of subcarrier intervals. Based on this observation, the frequency domain equalization can be realized in a low complexity, by considering ICI only from neighboring subcarriers, without obvious performance degradation.

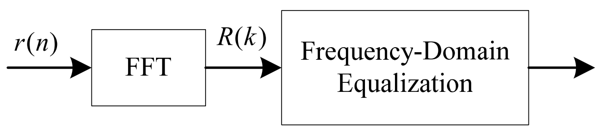

Fig. 2. Pure frequency domain receiver.

B. Pure frequency-domain receiver

Different from the hybrid receiver, the pure frequency-domain receiver, as shown in Fig 2, is derived on the basis of signal model (8). It jointly compensates the ICI induced by both the CFO and the channel time variation. The transmitted symbol can also be recovered by using MMSE equalization:

$$\boldsymbol{X} = (\boldsymbol{H}^{\mathrm{H}}\boldsymbol{H} + \sigma^2\boldsymbol{I})^{-1}\boldsymbol{H}^{\mathrm{H}}\boldsymbol{R}. \tag{15}$$

As (15) takes on the form of (14), so the type-II receiver can also be realized in a low complexity so long as the ICI induced by both the CFO and the CTV decrease rapidly with the increasing of subcarrier intervals.

In order to reveal the ICI distributions in the ICI coupling matrix, we illustrate in Fig 3 the normalized amplitude of ICI coefficients $|H(k,u)|$ versus the subcarrier interval for a typical OFDM system with N=64. Transmission channels consist of 6 independent paths with exponential power delay profile generated by Jakes model [17]. The CIR of each path varies with time and the channel time variation speed is measured by the normalized (by subcarrier spacing) Doppler frequency f_{d}. Fig 3 considered three cases: (1) Only CFO exists (denoted by "only CFO"); (2) Only channel time variation exists (denoted as "only CTV"); (3) Both CFO and channel time variation exist (denoted as "both CTV and CFO"). We see from Fig 3 that in all cases, the ICI power decreases sharply with the increasing subcarrier spacing, and moreover, when the CFO and channel time variation coexist, the ICI becomes more severe than the case of only CFO or channel time variation presents.

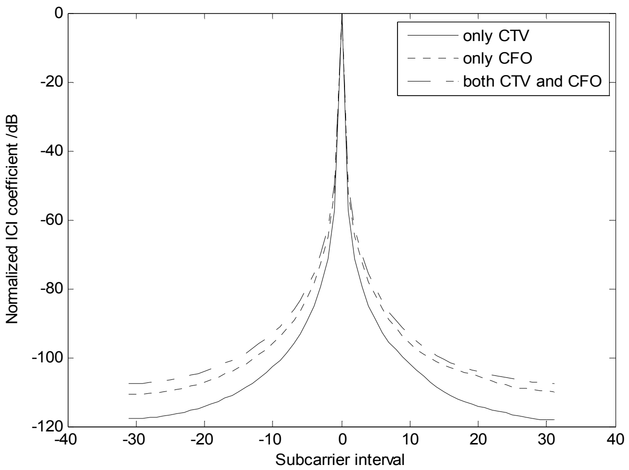

Fig. 3a. Illustration of the ICI distributions (f_{d}=0.1, $\varepsilon = 0.1$)

Although both the receivers can be realized with low complexity, the type-II may involves more computations than the type-I, as the ICI induced by two factors is more severe than by only one, and a bigger bandwidth of banded matrix may be assumed in type-II receiver to maintain the same performance as that of the type-I receiver. In the next section, we will evaluate the system performance of these two receivers through numerical simulations.

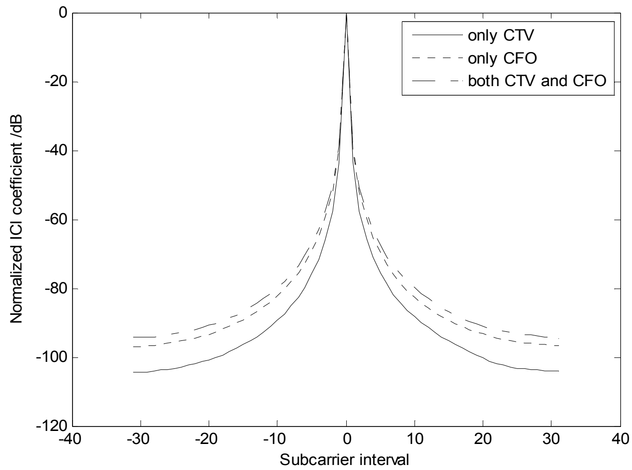

Fig. 3b. Illustration of the ICI distributions (f_{d}=0.2, $\varepsilon = 0.2$)

C. Complexity-Reduced Equalization

Owing to the approximate band structure of CFR matrix, there exists many low-complexity equalization methods such as the complexity-reduced MMSE equalizer with ordered SIC (MMSE-OSIC) [16]. Next we take the MMSE-OSIC [16] as an example to illustrate how to use the band structure of CFR matrix for reducing computational complexity. For the convenience of formulation, we define the vectors: $\boldsymbol{Y}_k = [Y(k\text{-}D),\ldots, Y(k\text{+}D)]^{\mathrm{T}}$, $\boldsymbol{X}_k = [X(k\text{-}2D),\ldots, X(k\text{+}2D)]^{\mathrm{T}}$, $\boldsymbol{W}_k = [W(k\text{-}D),\ldots, W(k\text{+}D)]^{\mathrm{T}}$, and the matrix:

$$\boldsymbol{H}_k = \begin{bmatrix} H(k-D,k-2D) & \cdots & H(k-D,k+2D) \\ \vdots & \ddots & \vdots \\ H(k+D,k-2D) & \cdots & H(k+D,k+2D) \end{bmatrix}, \tag{16}$$

where D is the number of considered ICI terms. It is worth noting that the indexes in above expressions of this subsection are all taken by modulo-N operation. For properly selected parameter D, $\boldsymbol{Y}_k$ can be approximated as

$$\boldsymbol{Y}_k = \boldsymbol{H}_k\boldsymbol{X}_k + \boldsymbol{W}_k. \tag{17}$$

From (17), the MMSE estimate of $X(k)$ can be obtained as

$$\hat{X}(k) = \boldsymbol{H}_k^{\mathrm{H}}(:,k)(\boldsymbol{H}_k\boldsymbol{H}_k^{\mathrm{H}} + \sigma^2\boldsymbol{I}_k)^{-1}\boldsymbol{Y}_k, \tag{18}$$

where $\boldsymbol{H}_k(:,k)$ is the $(2D$+1)th column of $\boldsymbol{H}_k$ and $\boldsymbol{I}_k$ denotes a $(2D+1)\times(2D+1)$ identity matrix. In order to improve detection performance, the ICI caused by $X(k)$ is subsequently subtracted from the received signal. This operation is the so called SIC. Although the SIC can improve detection performance to some extend, the performance improvement is restricted by error propagation inherent in the SIC. In response, the authors in [16] proposed a simple yet effective ordering method based on Frobenius norm of the column vector of matrix $\boldsymbol{H}_k$ to relief the error propagation of SIC. The overall complexity-reduced MMSE-OSIC detection procedure proposed in [16] is summarized as follows:

Step 1. Obtain the subcarrier index k for which the subcarrier is detected by (19)

$$k = \arg\max_{m} \left\| \boldsymbol{H}_m(:,m) \right\|_{\mathrm{F}}, \tag{19}$$

where $\|\bullet\|_{\mathrm{F}}$ denotes the Frobenius norm operator.

Step 2. Obtain the MMSE estimate of $X(k)$ using (18) and then make a decision according to the modulation constellation adopted.

Step 3. Perform SIC: $\boldsymbol{Y}$=$\boldsymbol{Y}$-$\boldsymbol{H}(:,k)X(k)$, where $\boldsymbol{H}(:,k)$ is the kth column of the CFR matrix $\boldsymbol{H}$.

Step 4. Set $\boldsymbol{H}(:,k)=\mathbf{0}$.

Step 5. Go to step 1 until all subcarriers have been detected.

According to the discussion of [16], the complexity-reduced MMSE-OSIC equalizer is advantageous over the MMSE equalizer with full CFR matrix in terms of both performance and complexity.

IV. Numerical simulations

We consider an OFDM system with 64 subcarriers per symbol. Each symbol appends a CP of 16 samples to avoid intersymbol interference. Transmission channels consist of 6 independently paths with exponential power delay profile, and the CIR of each path are Rayleigh distributed samples generated according to the classic Jakes model [17]. In the simulations the normalized Doppler frequency f_{d} is set to be 0.1 and 0.2, and the CFO are also chosen to be 0.1 and 0.2. Other simulation parameters are the same as in Section 3. In order to reduce the receiver complexity, we use the complexity-reduced MMSE-OSIC method [16] to perform ICI equalization in the frequency domain. The parameter D of the MMSE-OSIC is chosen to be 4. Moreover, the MMSE equalizer with full CFR (i.e., all ICI are considered) is used to benchmark the equalization performance. We also assumed perfect knowledge of CFO and channel state information at the receiver.

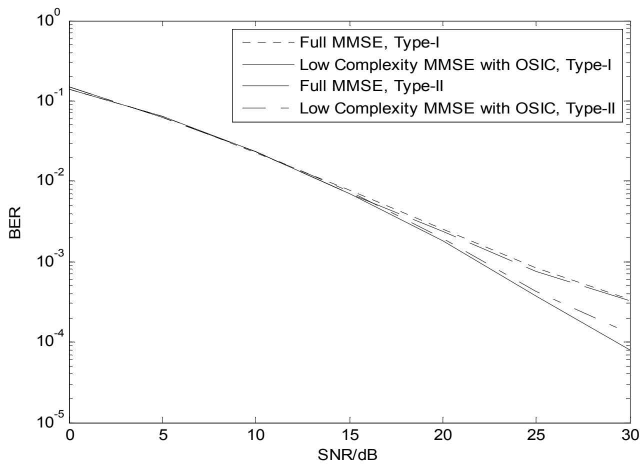

Fig. 4a. BER Performance under both CFO and CTV (f_{d}=0.1, $\varepsilon = 0.1$).

Fig 4 shows the bit error rate (BER) versus signal to noise ratio (SNR) using the two type receivers. In these plots, "Full MMSE" denotes the MMSE equalizer with full CFR and "Low complexity MMSE with OSIC" the low-complexity MMSE equalizer with ordered serial interference cancellation [14]. We see from these plots that under various conditions and for both receivers the MMSE equalizer with OSIC is superior to the full MMSE equalizer. This may contribute to the diversity induced by the channel time variation. Due to the channel time variation, the power of each subcarrier is distributed over all subcarriers of a symbol, which produce potential diversity gains that thus be exploited through collecting the scattered subcarrier energies by the MMSE equalizer with OSIC. On the other hand, under severe ICI, the MMSE equalizer cannot fully exploit this diversity and can only mitigate the ICI to some extent.

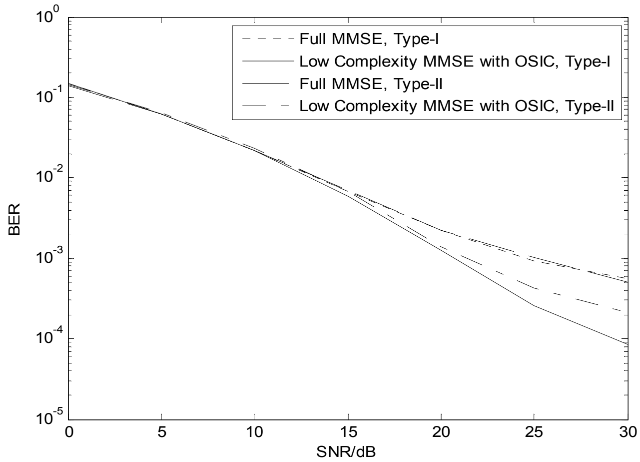

Fig. 4b. BER Performance under both CFO and CTV (f_{d}=0.2, $\varepsilon = 0.2$).

Further more, we see from Fig 4 that under various conditions and under the MMSE equalization with OSIC, the type-I outperforms the type-II, especially in the case of large CFO. For instance, at a high SNR of 30 dB and under CTV and CFO values of f_{d}=0.1 and $\varepsilon = 0.1$, the BER of the type-II is 1.6 times the BER of the type-I, while at the same SNR and under large CTV and CFO values of f_{d}=0.2 and $\varepsilon = 0.2$, the BER of the type-II is 2.5 times that of the type-I.

In order to discover the reason why the type-I outperforms the type-II in the case of, especially large, CFO, we need resort to insight analysis of the effects of the CFO on the OFDM signals. Although in the presence of CFO the power of each subcarrier is also distributed over all subcarriers which also means a possible diversity, unfortunately however, the ICI distributions in this case make the CFR matrix highly correlated as (7) indicates. Actually, given a k, the CFR $H(k,u)$ for any value of u are the same when the channels is flat and static. As is well known, matrix with correlated elements means an ill conditioned matrix which thereby leading poor performance of the type-II receiver.

To corroborate this claim, Fig 5 shows the simulation results for the type II receiver where there is no CTV and only CFO exists. It is seen from the simulations that the performance of MMSE equalization with OSIC is even inferior to that of the MMSE equalization with full CFR. This is due to the ill

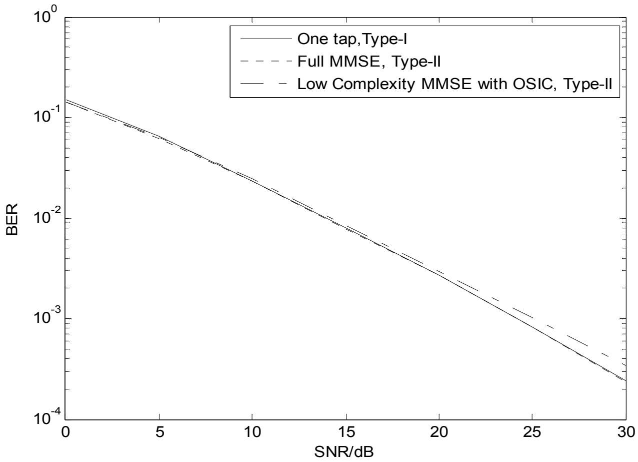

Fig. 5a. BER performance with only CFO ($\varepsilon = 0.1$)

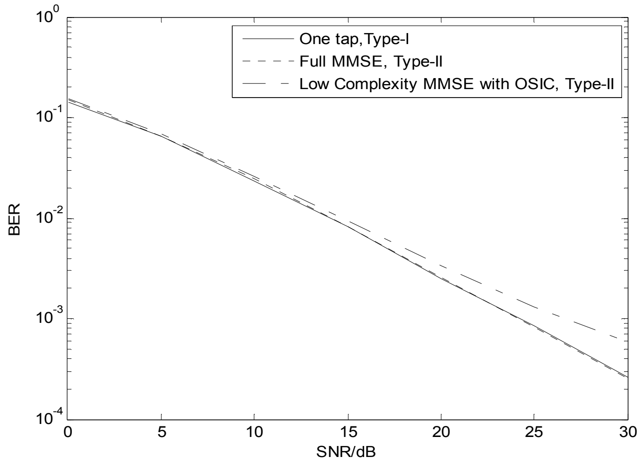

Fig. 5b. BER performance with only CFO ($\varepsilon = 0.2$)

conditioned CFR matrix $\boldsymbol{H}$ caused by the CFO and the resulting diversity suppression.

On the other hand, for the type-I receiver, the effect of CFO is compensated before the MMSE equalization with OSIC, so ill conditioned CFR matrix no longer exists in this case, and as a result the diversity gain can be exploited by the MMSE equalizer with the OSIC. The analysis above makes us naturally conclude that it is inappropriate for OFDM system to compensate the CFO in the frequency domain.

V. Conclusion

This paper investigates the receiver design for OFDM system operating over both the CFO and the channel time variation. The ICI under such cases becomes severe and the receiver confronts challenges of intensive computations and performance degradation. For solving this problem, this paper investigates two receive schemes, denoted by the separate scheme and the integrated scheme respectively, for OFDM system operating in such conditions. Performance of these two receivers are studied and compared through numerical simulations. Results obtained from the simulations show that the separate scheme outperforms the integrated one in terms of the BER performance, and is a more appropriate option for OFDM system operating over both the CFO and the channel time variation.

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