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Carrier frequency offset estimation for OFDM systems with time-varying DC Offset

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Abstract

Orthogonal frequency division multiplexing (OFDM) systems with direct-conversion architecture suffer from both carrier frequency offset (CFO) and dc offset (DCO). In this paper, we study CFO estimation problem for OFDM systems with time-varying DCO (TV-DCO) caused by gain mode switch of low noise amplifier (LNA). Based on linear approximation of TV-DCO, a blind algorithm is proposed for CFO estimation by means of DCO compensation and power leakage minimization. Performance of the proposed algorithm is demonstrated by simulations.

Introduction

Orthogonal frequency division multiplexing (OFDM) [1] is a promising technology for wireless communications to achieve efficient spectrum utilization, robustness to multi-path fading and easy implementation based on fast Fourier transform (FFT) and inverse FFT (IFFT), and has been widely adopted by emerging wireless applications such as digital audio broadcasting (DAB) [2], digital video broadcasting (DVB) [3], wireless local area network (WLAN) [4] and 3GPP long term evolution (LTE) [5] etc.

Despite of the attractive advantages, OFDM is vulnerable to various disturbances in practice. Carrier frequency offset (CFO) is one of most well-known disturbances for OFDM. It generates inter-carrier interference (ICI) and degrades OFDM performance [1]. In order to mitigate the negative influence, CFO is usually estimated and compensated accordingly during OFDM reception. CFO estimation for OFDM systems had been excessively studied and various algorithms had been proposed in literatures such as [6-8]. In [6], maximum likelihood (ML) CFO estimation for OFDM systems in additive white Gaussian noise (AWGN) channel was presented, while its performance degrades in multi-path dispersive channel. Liu et al. proposed in [7] a MUSIC-like blind CFO estimator which was proved in [9] to be equivalent to ML estimator in fading channel.

In addition to CFO, OFDM may also suffer from other disturbances such as direct current offset (DCO). For the sake of cost and power efficiency, mobile receiver architecture is under evolution from superheterodyne to direct conversion [10-13] in recent years. DCO is one of the most common disturbances of direct conversion receiver (DCR). It comes from self-mixing of local oscillator (LO) signal or radio frequency (RF) signal due to the finite isolation between input ports of mixer [10-13], as shown in Figure 1. In OFDM systems, DCO not only degrades demodulation performance but also violates CFO estimation [14-19]. CFO estimation for OFDM systems with static DCO had been well addressed in literatures [14-19]. Impacts of static DCO on CFO estimation can either be eliminated by analog high pass filter (HPF) [20] or be compensated in digital domain with data-aided [14-16] or blind approaches [17-19]. Besides static DCO, DCR may also introduce time-varying DCO (TV-DCO). In order to cover the high dynamic range of faded OFDM signals, low noise amplifier (LNA) with multiple gain modes [20-27] is usually employed by DCR in OFDM systems. During the gain mode switch stage shown in Figure 2, a sudden change of DCO level may occur [20,28-36] and high frequency components of the sudden change may pass through the HPF succeeding LNA, which results in TV-DCO [28-36]. Only a few works were reported to address CFO estimation for OFDM systems with TV-DCO. Inamori et al. proposed in [29] to suppress the influence of TV-DCO on CFO estimation with differential filter (DFE). Yunus et al. presented a least square estimation (LSE) algorithm in [34,35], which

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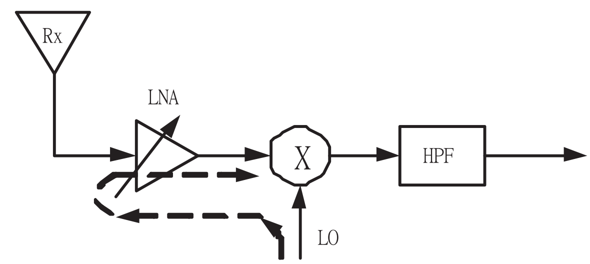

Figure 1 Generation of DCO in DCR.

achieves better performance than DFE at costs of higher computation efforts. The DFE and LSE algorithms were respectively extended in [28,30] and [36] to address CFO estimation for OFDM systems with both TV-DCO and in-phase/quadrature (I/Q) imbalance.

The established CFO estimation algorithms for OFDM systems with TV-DCO, e.g. DFE and LSE, are all data-aided algorithms. They depend on transmission of training sequences and/or pilots, which reduces effective data transmission rate. Different from the established works, we in this paper propose a blind CFO estimation algorithm for OFDM systems with TV-DCO. Based on linear approximation of TV-DCO, the proposed algorithm estimates CFO by means of DCO compensation and power leakage minimization. As it is a generalization of our recent work, the eigen-decomposition based estimator (EDE) [37], the proposed algorithm is named LVD-EDE, i.e. EDE in the presence of linear varying DCO. Performance of LVD-EDE is demonstrated by simulations in comparison with established algorithms including the maximum likelihood CFO estimation algorithm for OFDM systems with no DCO [7], with static DCO [17-19] and blind version of the DFE [29] and LSE algorithms [34,35].

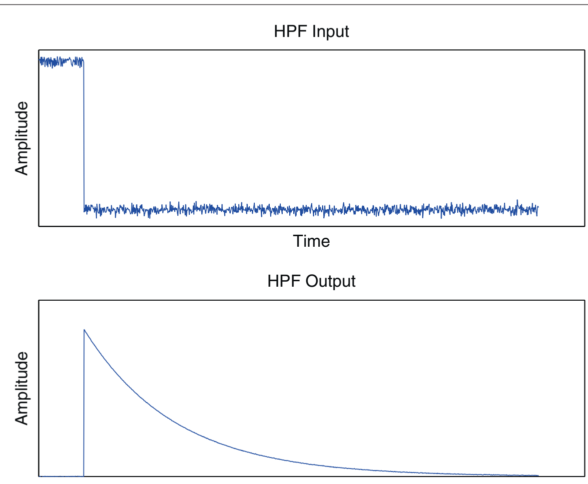

Figure 2 Generation of TV-DCO during LNA gain mode switch stage.

The reminder of this paper is organized as follows. Model of OFDM system with TV-DCO is established in the second section. The proposed LVD-EDE algorithm is developed in detail in the third section. In the fourth section, simulation results and corresponding analysis are provided to demonstrate the performance of LVD-EDE. Finally, conclusions are drawn in the last section.

Model of OFDM system with CFO and TV-DCO

Consider an OFDM system with totally N sub-carriers, among which the K sub-carriers occupied by data transmission are referred to as real sub-carriers, and the other $N-K$ unoccupied ones are referred to as virtual sub-carriers. Let $T_s \triangleq \frac{T}{N}$ denote the sample spacing in digital signal processing (DSP) stage, where T is the duration of OFDM block without cyclic prefix (CP). After CP removal, the received samples that belong to the m-th OFDM block can be expressed as

$$r(n,m) = e^{j\phi(m)} \sum_{k\in C_r} H(k,m)S(k,m)e^{j\frac{2\pi}{N}(k+\varepsilon)n} + d(n,m) + w(n,m) \quad (1)$$

for $n = 0,\ldots,N-1$. $C_r \triangleq \{k_0,\ldots,k_{K-1}\}$ denotes the indices set of all the K real sub-carriers. $S(k,m)$ is the modulated symbol mapped onto the k-th sub-carrier of the m-th OFDM block and $H(k,m)$ is the corresponding frequency domain channel response. Both $S(k,m)$ and $H(k,m)$ are assumed to be zero-mean and independent to each other. $\phi(m) \triangleq 2\pi\varepsilon(m(N_{CP}+N)+N_{CP})/N$ denotes a cumulative phase offset, where ε and N_{CP} refer to the CFO normalized to sub-carrier spacing and length of CP in samples, respectively. TV-DCO at output of HPF and AWGN are denoted by $d(n,m)$ and $w(n,m)$, respectively. As $d(n,m)$ is excited by a sudden level change at input of HPF, exact expression of $d(n,m)$ depends on step response of the HPF. For a first order HPF, its step response can be expressed as [20]

$$d(t) = \beta e^{-\frac{t}{\tau}}, \quad (2)$$

where both β and τ are positive real constants. τ is referred to as time constant, which is inverse to cut-off frequency of the HPF. In order to keep low degree of inter-symbol interference, a HPF with low cut-off frequency is usually used [20,28-36]. Thus, (2) can be approximated by a linear model

$$d(t) \approx \beta - \frac{\beta}{\tau}t. \quad (3)$$

Based on (3), we can formulate the TV-DCO in (1) with two parameters $a(m)$ and $b(m)$ as

$$r(n,m) \approx e^{j\phi(m)} \sum_{k\in C_r} H(k,m)S(k,m)e^{j\frac{2\pi}{N}(k+\varepsilon)n} + a(m) + b(m)*n + w(n,m), \tag{4}$$

where $a(m)$ and $b(m)*n$ represent respectively the static and linear varying parts of TV-DCO within the m-th OFDM block. The linear model of TV-DCO in (4) was also widely adopted in literatures [28-36] as a valid approximation of TV-DCO at output of HPF. For notation simplicity, we replace '≈' with '=' and then rewrite (4) in matrix form as

$$\boldsymbol{r}_N(m) = \boldsymbol{P}_N(\varepsilon)\boldsymbol{U}_N\boldsymbol{x}_K(m) + \boldsymbol{l}_N a(m) + \boldsymbol{c}_N b(m) + \boldsymbol{w}_N(m), \tag{5}$$

where the vectors of received samples, symbols on real sub-carriers and noise samples are denoted respectively by $\boldsymbol{r}_N(m) \triangleq [r(0,m),\ldots,r(N-1,m)]^T$, $\boldsymbol{x}_K(m) \triangleq [H(k_0,m)S(k_0,m),\ldots,H(k_{K-1},m)S(k_{K-1},m)]^T e^{j\phi(m)}$ and $\boldsymbol{w}_N(m) \triangleq [w(0,m),\ldots,w(N-1,m)]^T$ with the superscript ^{T} representing transpose of matrix. $\boldsymbol{l}_N \triangleq [1,1,\ldots,1]^T$ and $\boldsymbol{c}_N \triangleq [0,1,\ldots,N-1]^T$ represent respectively all ones vector and linear vector of length N. The inverse discrete Fourier transform (IDFT) on real sub-carriers are denoted by an $N\times K$ matrix $\boldsymbol{U}_N$, whose (n,l)-th entry is $e^{j\frac{2\pi}{N}u_l n}$ with $u_l \in C_r$. The diagonal matrix $\boldsymbol{P}_N(\varepsilon) \triangleq diag\{1, e^{j\frac{2\pi}{N}\varepsilon},\ldots,e^{j\frac{2\pi(N-1)}{N}\varepsilon}\}$ represents the incremental phase offset caused by CFO.

Proposed CFO estimation algorithm LVD-EDE

Basic ideas of the proposed LVD-EDE algorithm originate from two facts. First, a DCO-free signal $\boldsymbol{y}_N(m)$ can simply be obtained through weighted linear combination (WLC)

$$\boldsymbol{y}_N(m) \triangleq \boldsymbol{Y}_N(m)\boldsymbol{g}_o, \tag{6}$$

with $\boldsymbol{Y}_N(m) \triangleq [\boldsymbol{r}_N(m), \boldsymbol{l}_N, \boldsymbol{c}_N]$ and weighting vector $\boldsymbol{g}_o \triangleq [g_o(0), g_o(1), g_o(2)]^T$ satisfying

$$b(m)g_o(0) + g_o(2) = 0 \tag{7}$$

and

$$a(m)g_o(0) + g_o(1) = 0. \tag{8}$$

Second, with perfect CFO compensation there will be no power leakages on the virtual sub-carriers, i.e.

$$\|\boldsymbol{V}_N^H \boldsymbol{P}_N(-\varepsilon)\boldsymbol{y}_N(m)\|_2^2 = 0, \tag{9}$$

where $\|.\|_2$ denotes 2-norm and $\boldsymbol{V}_N$ is a $N\times(N-K)$ matrix, whose (n,l)-th entry is $e^{j\frac{2\pi}{N}v_l n}$ with $v_l \in C_v \triangleq \{k_K,\ldots,k_{N-1}\}$ being the indices set of all the $N-K$ virtual sub-carriers. Note that noise is ignored in (9) and hereafter for clarity of illustration, while the noise effects are included in simulations for evaluation of LVD-EDE's performance.

Base on the two facts, we construct a cost function defined as

$$f(\nu,\boldsymbol{g}) \triangleq \sum_m \|\boldsymbol{V}_N^H \boldsymbol{P}_N(-\nu)\boldsymbol{Y}_N(m)\boldsymbol{g}\|_2^2, \tag{10}$$

by first linearly combining $\boldsymbol{r}_N(m)$, $\boldsymbol{l}_N$ and $\boldsymbol{c}_N$ with a trail weighting vector $\boldsymbol{g}$ to suppress DCO, compensating CFO with a trail value ν and then projecting the combined signal to virtual sub-carriers. It is easy to verify that $f(\varepsilon,\boldsymbol{g}_o) = 0$ if noise is ignored which means that $(\varepsilon,\boldsymbol{g}_o)$ achieves minimum of the non-negative cost function. Based on this observation, we may think about estimating CFO through the following search

$$\begin{aligned}(\hat{\varepsilon},\hat{\mathbf{g}}_o) &= \arg\min_{\nu,\mathbf{g}} f(\nu,\mathbf{g}) \\ &= \arg\min_{\nu,\mathbf{g}} \mathbf{g}^H \Omega_3(\nu)\mathbf{g},\end{aligned} \tag{11}$$

where $\boldsymbol{\Omega}_3(\nu) \triangleq \sum_m \boldsymbol{Y}_N^H(m)\boldsymbol{P}_N(\nu)\boldsymbol{V}_N\boldsymbol{V}_N^H\boldsymbol{P}_N(-\nu)\boldsymbol{Y}_N(m)$. However, $(\varepsilon,\boldsymbol{g}_o)$ is not the only solution to (11). Substituting (5) into $\boldsymbol{Y}_N(m)$ and then into (10) yields that

$$\begin{aligned}f(\nu,\boldsymbol{g}) &= \sum_m \|\boldsymbol{V}_N^H\boldsymbol{P}_N(-\nu)\boldsymbol{r}_N(m)g(0) + \boldsymbol{V}_N^H\boldsymbol{P}_N(-\nu)\boldsymbol{l}_N g(1) \\ &\quad + \boldsymbol{V}_N^H\boldsymbol{P}_N(-\nu)\boldsymbol{c}_N g(2)\|^2 \\ &= \sum_m \|\boldsymbol{V}_N^H\boldsymbol{P}_N(\varepsilon-\nu)\boldsymbol{U}_N\boldsymbol{x}_K(m)g(0) \\ &\quad + \boldsymbol{V}_N^H\boldsymbol{P}_N(-\nu)\boldsymbol{l}_N[g(1)+a(m)g(0)] \\ &\quad + \boldsymbol{V}_N^H\boldsymbol{P}_N(-\nu)\boldsymbol{c}_N[g(2)+b(m)g(0)]\|_2^2.\end{aligned} \tag{12}$$

Forcing the right hand side (RHS) of (12) to zero is equivalent to

$$\begin{aligned}&\boldsymbol{V}_N^H\boldsymbol{P}_N(\varepsilon-\nu)\boldsymbol{U}_N\boldsymbol{x}_K(m)g(0) + \boldsymbol{V}_N^H\boldsymbol{P}_N(-\nu)\boldsymbol{l}_N[g(1) \\ &\quad + a(m)g(0)] + \boldsymbol{V}_N^H\boldsymbol{P}_N(-\nu)\boldsymbol{c}_N[g(2) \\ &\quad + b(m)g(0)] = \boldsymbol{0}_{N-K},\end{aligned} \tag{13}$$

which has multiple solutions. Besides the desired solution $(\varepsilon,\boldsymbol{g}_o)$, there are a couple of undesired solutions. One is the homogeneous solution $\boldsymbol{g} = \boldsymbol{0}_3$ with $\boldsymbol{0}_3$ being the all zeros vector of 3×1. The other solution is that $\boldsymbol{V}_N^H\boldsymbol{P}_N(\varepsilon-\nu)\boldsymbol{U}_N\boldsymbol{x}_K(m)$, $\boldsymbol{V}_N^H\boldsymbol{P}_N(-\nu)\boldsymbol{l}_N$ and $\boldsymbol{V}_N^H\boldsymbol{P}_N(-\nu)\boldsymbol{c}_N$ are linearly dependent. Existence of the undesired solutions prevents us from estimating CFO directly through (11), but fortunately it can be solved in a certain way. The homogeneous solution can be avoided by imposing a constraint $\|\boldsymbol{g}\|_2^2 = 1$ to the minimization. Although the solution based on linear dependency cannot be avoided, it exists with little possibility because the realization of

$\boldsymbol{x}_K(m)$ is independent of the TV-DCO. Therefore, its influence can be ignored in practice without damaging CFO estimation. Finally, in LVD-EDE, CFO estimation is achieved by solving (11) subject to $\|\boldsymbol{g}\|_2^2 = 1$, which leads to [38]

$$\hat{\varepsilon} = \arg\min_{\nu} \lambda_{min}\{\Omega\}_3(\nu)\}, \tag{14}$$

and

$$\hat{\mathbf{g}}_o = V_{min}\{\Omega_3(\hat{\varepsilon})\}, \tag{15}$$

where $\lambda_{min}\{\boldsymbol{Z}\}$ and $V_{min}\{\boldsymbol{Z}\}$ denote the smallest eigenvalue and corresponding eigenvector of matrix $\boldsymbol{Z}$ respectively.

Computation efforts of LVD-EDE is mainly determined by the eigen-decomposition of $\Omega_3(\nu)$ in the 1-D search operation for CFO estimation, as it will execute in each searching step while other operations execute only once. Regarding to the eigen-decomposition, since $\Omega_3(\nu)$ is a small Hermitian matrix of dimensions 3×3, there exist efficient methods for calculating its smallest eigenvalue and corresponding eigenvector [38].

Simulation results

In this section, performance of LVD-EDE is demonstrated by simulations in comparison with four reference algorithms, MUE [7], NBE [17-19], CP-DFE and CP-LSE. MUE and NBE are ML CFO estimation algorithms for OFDM systems with respectively no DCO and only static DCO. CP-DFE and CP-LSE are blind versions of DFE [28] and LSE [34,35], respectively. Up to our awareness, DFE and LSE are the only algorithms reported for CFO estimation under TV-DCO. As data-aided algorithms, they rely on transmission of identical training sequences [28,34,35], which prevents us to directly compare them with the blind algorithm LVD-EDE. In order to have a fair comparison, we made the two blind versions, CP-DFE and CP-LSE. These two blind ones follow the basic ideas of DFE and LSE respectively, and remove the original dependency on training sequences by making use of ending part of OFDM symbol and its identical copy in CP instead.

Conditions for the simulations are summarized in Table 1. The OFDM system is a WLAN [4] like system. At beginning of reception, the TV-DCO is 10dB higher than signal and decays according to the cut-off frequency of HPF. The estimation performance is evaluated by normalized mean square error (NMSE) defined as $E\{|\hat{\varepsilon} - \varepsilon|^2\}$, where $E\{.\}$ denotes expectation operation. To track the variation of CFO as quickly as possible, only one OFDM block is used in the estimation.

In addition to simulations, Cramér-Rao lower bound (CRLB) for CFO estimation under desired case is also given for comparison. In accordance with the simulations, CRLB is derived for the case that CFO is estimated from only one OFDM block, so that the index m of OFDM block is dropped in derivation. First, define the vector of parameters to be estimated $\mathbf{z} \triangleq [\mathbf{x}_R, \mathbf{x}_I, a_R, a_I, b_R, b_I, \varepsilon]^T$ and mean observation vector $\mathbf{q} \triangleq [(\mathbf{P}(\varepsilon)\mathbf{U}\mathbf{x})_R^T + a_R\mathbf{l}^T + b_R\mathbf{c}^T, (\mathbf{P}(\varepsilon)\mathbf{U}\mathbf{x})_I^T + a_I\mathbf{l}^T + b_I\mathbf{c}^T]^T$, where the subscript $_R$ and $_I$ represent the real and imaginary parts of a complex number, respectively. Since in (5) $\mathbf{w}_N$ is assumed to be AWGN, the calculation of Fisher Information Matrix $\mathbf{F}$ can be simplified [39] to

Table 1 Simulation conditions

Simulation conditions	
Number of total sub-carriers	64
Number of real sub-carriers	48
Real sub-carrier allocation	1–24, 40–63
Sub-carrier modulation	QPSK
Channel fading	Rayleigh
Channel power delay profile	$e^{-p/5}, p = 0, \ldots, 9$
Initial signal to DCO ratio	-10 dB
HPF	first order Butterworth
HPF cut-off frequency	100 kHz

$$\boldsymbol{F} = [\,\partial\boldsymbol{q}/\partial\boldsymbol{z}]^T\,[\,\partial\boldsymbol{q}/\partial\boldsymbol{z}]\,, \tag{16}$$

where

$$\frac{\partial\boldsymbol{q}}{\partial\boldsymbol{z}} = \begin{bmatrix} \boldsymbol{A}_R & -\boldsymbol{A}_I & \boldsymbol{l}_N & \mathbf{0} & \boldsymbol{c}_N & \mathbf{0}_N & \boldsymbol{B}_R \\ \boldsymbol{A}_I & \boldsymbol{A}_R & \mathbf{0} & \boldsymbol{l}_N & \mathbf{0}_N & \boldsymbol{c}_N & \boldsymbol{B}_I \end{bmatrix}, \tag{17}$$

with $\boldsymbol{A} \triangleq \boldsymbol{P}(\varepsilon)\boldsymbol{U}$, $\boldsymbol{B} \triangleq \boldsymbol{QP}(\varepsilon)\boldsymbol{Ux}$, $\boldsymbol{Q} \triangleq diag\left\{0, j\frac{2\pi}{N}, \ldots, j\frac{2\pi(N-1)}{N}\right\}$ and $\mathbf{0}_N$ denoting the $N \times 1$ all zeros vector. CRLB of CFO can be calculated from the last element on diagonal of $\mathbf{F}^{-1}$.

Performance of CFO estimation within SNR range of interest by different algorithms is shown in Figure 3. In

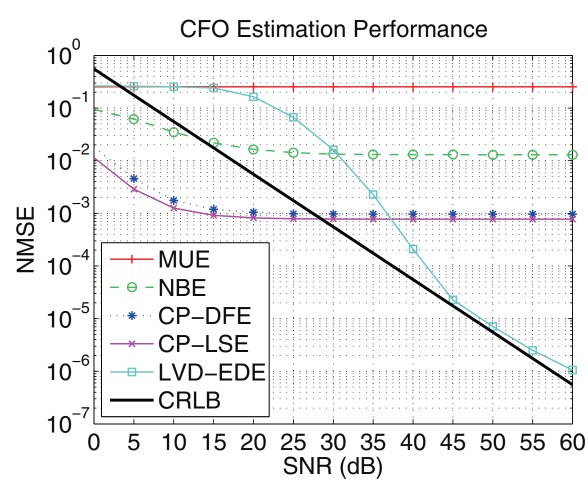

Figure 3 CFO estimation performance within SNR range of interests (ε = 0.1).

medium to high SNR region, all the algorithms except for LVD-EDE exhibit error floor. For MUE and NBE, it is due to the missing of TV-DCO in their signal models. For CP-DFE and CP-LSE, it is because the inter-symbol interference (ISI) caused by multi-path dispersive channel makes CP no longer be identical to the ending part of its associated OFDM symbol. With the linear modeling of TV-DCO and removal of CP to avoid ISI, LVD-EDE exhibits no error floor and approaches CRLB asymptotically with increasing SNR. In low to medium SNR region, LVD-EDE suffers from performance degradation due to the threshold effects [40,41] of eigen-decomposition (or equivalently singular value decomposition). This is one of the major cons of LVD-EDE. Figure 4 demonstrates that LVD-EDE is valid for the whole CFO range of $|\varepsilon| < 0.5$ and outperforms the others.

The purpose of CFO estimation in OFDM systems is to compensate CFO accordingly to achieve acceptable demodulation performance. Figure 5 shows demodulation performance after compensation of CFO and TV-DCO in terms of symbol error rate (SER), where knowledge of the fading channel is assumed to be perfectly known at receiver, CFO is compensated according to the estimates, and TV-DCO is compensated according to its real value. In accordance with the CFO estimation performance, only LVD-EDE achieves consistent SER improvement with increasing SNR. In low to medium SNR region, NBE, CP-DFE and CP-LSE achieves better performance than LVD-EDE, however none of them leads to an acceptable SER.

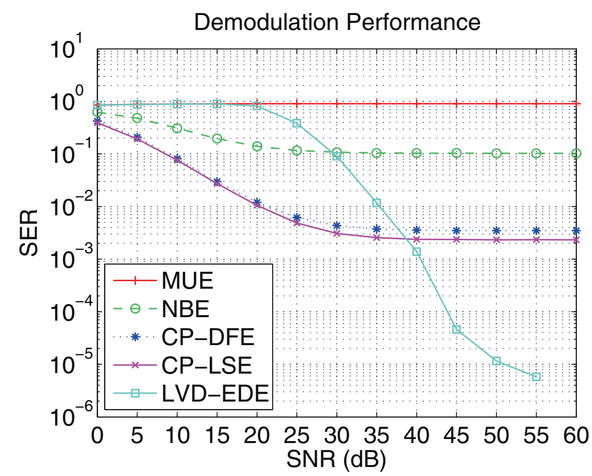

Figure 5 SER performance after compensation of CFO and DCO.

Conclusions

OFDM systems with DCR suffer from both CFO and DCO. In this paper, we propose a blind CFO estimation algorithm for OFDM systems with TV-DCO. The proposed LVD-EDE algorithm is an extension of our previous work EDE based on linear approximation of TV-DCO. Different from the established algorithms, LVD-EDE does not depend on specific preamble or training sequence. Performance of LVD-EDE is demonstrated by simulations.

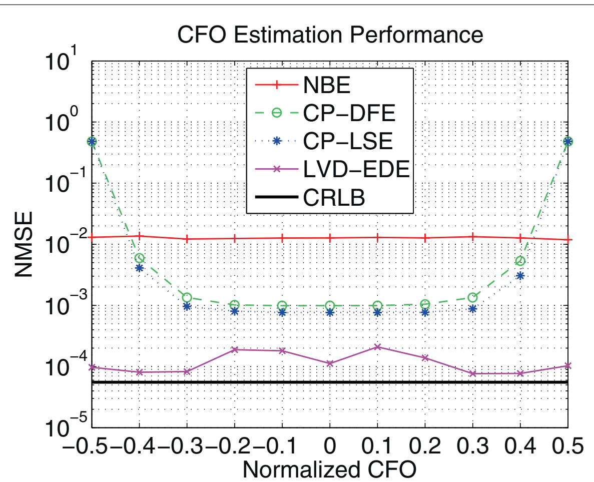

Figure 4 CFO estimation performance within CFO range of interests ($SNR = 40$ dB).

Competing interests

Both authors declare that they have no competing interests.

Author's contributions

Our contributions in this paper is the proposed blind CFO estimation algorithm LVD-EDE for OFDM systems with time-varying DCO.

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