Low-Complexity CFO Correction of Frequency-Spreading SMT in Uplink of Multicarrier Multiple Access Networks

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***Abstract*—This paper presents an alternative method to compensate carrier frequency offset (CFO) for frequency-spreading structure of staggered modulated multitone (FS-SMT) system in the uplink of multiple access scenario. The proposed method reduces computational complexity and latency of the receiver without losing the performance. Moreover, a possibility for merging compensation of timing offset (TO) and channel effect with CFO correction is presented. Simulation results show that the proposed method offers a reasonable performance with mentioned improvements.**

I. Introduction

Multicarrier modulation techniques are particularly appropriate for realization of broadband communication systems. Nowadays, orthogonal frequency division multiplexing (OFDM), as the most famous example, has been adapted in various communication standards, i.e., ADSL, IEEE802.11a, 802.16 and LTE [1]. Nonetheless, spectral efficiency of OFDM is reduced due to the need for a cyclic prefix (CP) in order to cope with a multipath channel. Furthermore, it uses a pulse shaping filter with sinc-shaped frequency response which dose not have proper frequency localization and leads to more interference [2]. Moreover, OFDM is not robust enough against carrier frequency offset (CFO) which causes loss of orthogonality and results in performance loss [3]. As a result of these shortcomings of OFDM, filter bank multicarrier (FBMC) has been suggested as an alternative technique [2],[4]. Several systems of FBMC have been introduced such as cosine-modulated multitone (CMT), filtered multitone (FMT) and the most spectrally efficient one, staggered modulated multitone (SMT) [3]. SMT is also known as FBMC/OQAM or OFDM/OQAM that are different acronyms for the same modulation technique. There are two structures to implement SMT, polyphase structure and frequency-spreading (FS) [5]. It is shown that although FS is more complex than polyphase structure, it offers some advantages in terms of synchronization and equalization issues [6].

Synchronization is a crucial issue when multicarrier modulation is applied for a multiple access network. Synchronization is performed in two steps, estimation and correction. In this paper, we focus on the second step. The TO correction is straightforward in OFDM system due to presence of CP. The CFO correction for OFDM has been extensively researched and documented [7]. The impact of TO and CFO on the SMT modulation has been also considered and compared with OFDM in [8]. The correction is more important in the uplink mode where each user has its own CFO and TO. In this case, corrections must be performed individually for each user. In uplink scenario, CFO can be compensated before or after discrete Fourier transform (DFT). We tend to perform CFO correction after DFT which offers less computational complexity in uplink mode. There are some post-DFT correction methods for orthogonal frequency division multiple access (OFDMA) [7]. Similarly, in FBMC systems, CFO correction can be performed after DFT for polyphase structure of SMT [9]. Also, authors in [10] have suggested a post-DFT correction method for FS structure that we call Dore-Berg-Cassiau-Ktenas (DBCK). But it imposes some extra computational complexity on the receiver side. In this paper, an alternative approach to post-DFT CFO correction for FS structure has been proposed. The computational complexity and performance have been investigated and compared with other methods. Simulation results show the proposed method presents reasonable performance in comparison with DBCK with less delay and lower computational complexity.

The paper is organized as follows. In Section II we start with system model. In Section III the SMT structures, specially FS as a desired structure, are described. Then, in Section IV the CFO correction of SMT multiple access (SMTMA) system is considered. In this section, after the case of the pre-DFT, prior work on the post-DFT correction is presented and the modified method is proposed. The complexity comparison of all CFO correction methods is illustrated in Section IV-C and simulation results are reported in Section V. Finally, conclusion are drawn in Section VI.

II. System model

We consider a multicarrier multiuser network which is based on SMTMA. There are P users communicating with a base station (BS). We assume that the block subcarrier

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allocation has been used. So if we leave one carrier unused between users as a gaurd, $(N-P)/P$ of all N subcarriers are allocated to each user. Then, we focus on the uplink scenario of this system where each user sends its signal to BS. If the channel response between BS and the pth user is represented by $c_p[i]$, the received signal can be written as

$$y[i] = \sum_{p=1}^{P} (r_p[i] * c_p[i]) + z[i], \qquad (1)$$

where $*$ denotes linear convolution and $z[i]$ is an additive white Gaussian noise (AWGN) vector. In the presence of CFO $r_p[i]$ is equal to

$$r_p[i] = x_p[i].e^{j\frac{2\pi}{N}\delta_p i}, \qquad p = 1, 2, \cdots, P \qquad (2)$$

where $x_p[i]$ is the transmitted signal from the pth user and δ_p is the normalized CFO with respect to the carrier spacing. In this model δ_p is in the range of [-0.5,0.5].

III. Structures of SMT

SMT, like any other multicarrier communication system, transmits signals across several subcarriers. This system applies specific pulse shaping filters which overlap in the frequency domain and satisfy Nyquist criteria to make necessary orthogonality between subcarriers. Two structures for implementation of SMT have been introduced as follow.

A. Polyphase Structure of SMT

In the polyphase structure, prototype filter is designed in time domain and all pulse shaping filters are driven from different phases of it [3]. However, pulse shaping filter of FS is designed and applied in the frequency domain [11]. It is shown that although the FS is more complex than polyphase structure, it offers a crucial advantages in terms of synchronization and equalization issues [6]. In order to take advantage of FS with less complexity, FS may be applied only in the receiver side and transmitter can be implemented by polyphase structure.

B. Frequency Spreading Structure of SMT

In the FS structure of SMT, the frequency sampling technique is applied to design a prototype filter [12]. Hence, the filter with definite non-zero samples in the frequency domain may be achieved. Therefore, where K is the overlapping factor, i.e., the number of multicarrier symbols which overlap in the time domain, the filter length $L = KN$ with $2K-1$ non-zero samples is required in the frequency domain. In [13], some optimization criteria are considered and the coefficients of this filter are calculated for different values of K.

In the transmitter, as OQAM modulation is used, real part and imaginary part of each symbol are modulated separately and added together with $N/2$ delays difference. Thereupon, real-valued symbols, $a_{n,k}$ related to the kth subcarrier of the nth multicarrier symbol, are applied to pulse shaping filters. The transmitter structure is illustrated in Fig.1, where $H_{k'}$ is the k'th coefficient of the filter that are chosen to be real-valued and symmetric. Since filters in the frequency domain are only overlapped with their two neighbors, the interference

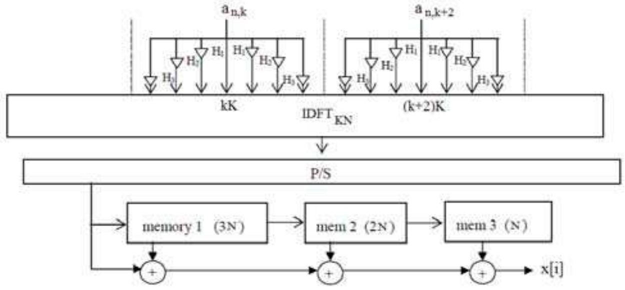

Fig. 1. The in-phase part of a FS-transmitter [11]

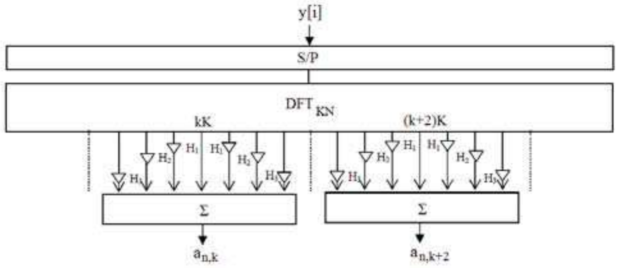

Fig. 2. The in-phase part of a FS-receiver [11]

can be avoided by quadrature, i.e., make a $\pi/2$ phase difference between neighbors [11]. Then, they are put together and fed to the KN-point inverse discrete Fourier Transform (IDFT). After that, the IDFT output converts from parallel to serial. Finally, K IDFT output blocks overlap in the time domain to abide by the rate of the multicarrier symbols which is supposed to be $1/N$.

In the receiver, KN samples of received signal are fed to KN-point DFT. Then, matched filters extract the data symbols on their relative frequencies. After the summation, as depicted in Fig.2, real-valued symbols are extracted. But it must be noted that the next KN samples have to be picked up from just N samples later and it can be the counterpart of the overlap and sum operations of the transmitter [11]. This procedure has to be performed with $N/2$ delays to extract real-valued symbols related to in-phase part of the data symbols.

IV. CFO compensation methods for SMTMA

We assume that the receiver knows the CFO values by perfect estimation. Each subcarrier receives multiple access interference (MAI) only from neighbour users, hence there is no need to use interference cancellation (IC) method, thanks to the very frequency selective filters in SMT [9].

A. Pre-DFT CFO Correction

For the first time, the single user detection (SUD) method was introduced in OFDMA system [7]. An equivalent method can be used in SMTMA system. The received time domain signal is the sum of all users' signal which have their own CFOs. Any correction for desired user can lead to more distortion for other users. For this reason, users with different

CFO values should be compensated separately. So the CFO compensated signal is formed as

$$r_p^{'}[i] = x_p[i].e^{-j\frac{2\pi}{N}\delta_p i}. \tag{3}$$

This procedure is convenient for downlink scenario where each user aims to extract just its own signal and do not mind other signals. But in uplink scenario where a BS is looking for every user's signal, it needs to have a separate receiver for each user. Hence, the computational complexity goes up as the number of users increases (see more details in Section IV-C). It is worth noting that the CFO correction after DFT reduces computational complexity.

B. Post-DFT CFO Correction

CFO correction can be applied after DFT. In the case of polyphase structure, the receiver structure will change [9]. In contrast, post-DFT correction is quite straightforward in FS structure. In this section, we first go through prior work on this area [10], then we propose an alternative scheme in order to improve the correction procedure.

1) DBCK method: Authors in [10] suggested a CFO correction scheme for FS structure. The vector of the received time domain signal to extract the mth multicarrier symbol can be written as

$$y_m = \sum_{p=1}^{P} (d_p x_p^{(m)} e^{j2\pi\delta_p m}) + z_m, \tag{4}$$

where z_m is the $KN \times 1$ column noise vector and d_p is a KN diagonal matrix defined as

$$d_p[k,k] = e^{j2\pi\frac{\delta_p k K}{KN}}. \tag{5}$$

Since the CFO correction is the same for all users after DFT, without loss of generality subscript index p can be eliminated and the received frequency domain signal can be written as

$$Y_m = CX_m.e^{j2\pi\delta m} + Z_m, \tag{6}$$

where $X_m = DFT[x^{(m)}]$, Z_m is AWGN vector and C is a $KN \times KN$ Toeplitz matrix defined by

$$C = FdF^H, \tag{7}$$

where F stands for the $KN \times KN$ DFT matrix as [10]

$$F = \frac{1}{\sqrt{KN}} \begin{bmatrix} \mathbf{1} & 1 & \cdots & 1 \\ \mathbf{1} & W_{KN} & \cdots & W_{KN}^{KN-1} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{1} & W_{KN}^{KN-1} & \cdots & W_{KN}^{(KN-1)(KN-1)} \end{bmatrix}, \tag{8}$$

where $w_{KN} = e^{j\frac{2\pi}{KN}}$. Also the coefficients of C can be written as [10]

$$C[i,k] = \frac{sinc(K\delta - (i-k))}{sinc(\frac{K\delta-(i-k)}{N})} e^{-j\pi\frac{(N-1)(K\delta-(i-k))}{N}}. \tag{9}$$

As shown in FS structure, the data on each subcarrier is spread over the subcarriers of the respective subchannel. So this normalized CFO over N subcarriers is devided into the integer part and the fractional part over the new KN subcarriers. The parameter δ may be decomposed as

$$\delta = \frac{q}{K} + \epsilon, \tag{10}$$

where $\epsilon \in \mathbb{R}$ with $\epsilon \in [-\frac{1}{2K}, \frac{1}{2K}]$ is the fractional part and $q \in \mathbb{Z}$ is the integer part which can be compensated with a simple shift. The phase term should be corrected by a phase correction factor. But the reduction of interference due to fractional part needs a filter to mitigate the effect of C. In [10] zero-forcing (ZF) criterion is considered to derive the coefficients of this filter.The $KN \times KN$ matrix W should be chosen so that satisfy

$$WC = I. \tag{11}$$

With substitution of (7) in (11), we have

$$W = Fd^H F^H. \tag{12}$$

Since W is a Toeplitz matrix, only KN complex coefficients are sufficient to obtain all other coefficients. The largest values for the W power coefficients are in the vicinity of the diagonal elements. Therefore it can be approximated with a matrix that has $2Q+1$ non-zero coefficients on each column and centered on the diagon. Q should be chosen to make a trade-off between the complexity and the performance. In other words higher Q leads to better performance with more computational complexity. However calculation of new matrix is still inconvenient even with this simplification.

As proposed in [10], a practical implementation is utilization of combination of some pre-computed matrixes. $F = 4$ matrixes are computed to be used in cascade scheme. Each matrixes can be applied or bypassed by a multiplexer (MUX). In case of negative CFO, the inverse command may be performed and appropriate matrixes are derived by a permutation and complex conjugation of the prior matrixes. For instance, W_i can be calculated in order to compensate the CFO equal to δ_i which is defined as

$$\delta_i = 2^i \times 1\%, \qquad i = 0, 1, \cdots, F-1 \tag{13}$$

Thereby correction range of $[-15\%, 15\%]$ with 1% precision is possible. As an example, for correcting a CFO equal to 11.7%, the MUX commands with numbers 0 , 1 and 3 should be applied and MUX command with number 2 should be bypassed. Therefore, the effective CFO corrected is $1\% + 2\% + 8\% = 11\%$.

It should be noted that this scheme almost needs more than one matrix multiplication and even in some cases all pre-computed matrix may be inserted. In the following, an alternative procedure is proposed to correct CFO in FS structure with lower computational complexity.

2) Proposed CFO correction method: Since the total received signal is defined as

$$y[i] = \sum_{p=1}^{P} r_p[i], \tag{14}$$

where

$$r_p[i] = x_p[i].e^{\frac{j2\pi\delta_p i}{N}}, \tag{15}$$

the pth user's CFO compensated signal before DFT can be obtained by

$$r_p^{'}[i] = r_p[i].e^{\frac{-j2\pi\delta_p i}{N}}. \quad (16)$$

We consider only a user and eliminate the subscript index p. As it was explained in Section III-B, the receiver uses KN samples to extract N data symbols on each multicarrier symbol. Therefore, the column vector needed for achieving the mth SMT symbol can be defined as r_m

$$r_m = \begin{bmatrix} r[mN-(KN-1)] \\ r[mN-(KN-2)] \\ \vdots \\ r[mN] \end{bmatrix}. \quad (17)$$

By substituting (16) in (17), we have

$$r_m^{'} = \begin{bmatrix} r[mN-(KN-1)].e^{\frac{-j2\pi\delta_p K(mN-(KN-1))}{KN}} \\ r[mN-(KN-2)].e^{\frac{-j2\pi\delta_p K(mN-(KN-2))}{KN}} \\ \vdots \\ r[mN].e^{\frac{-j2\pi\delta_p KmN}{KN}} \end{bmatrix}. \quad (18)$$

By some manipulation, it can be rewritten as

$$r_m^{'} = \begin{bmatrix} r[mN-(KN-1)] \\ r[mN-(KN-2)] \\ \vdots \\ r[mN] \end{bmatrix} \odot \begin{bmatrix} e^{\frac{-j2\pi\delta_p K(mN-KN+1)}{KN}} \\ e^{\frac{-j2\pi\delta_p K(mN-KN+2)}{KN}} \\ \vdots \\ e^{\frac{-j2\pi\delta_p KmN}{KN}} \end{bmatrix}$$

$$= r_m \odot \begin{bmatrix} e^{\frac{+j2\pi\delta_p K(KN-1)}{KN}} \\ e^{\frac{+j2\pi\delta_p K(KN-2)}{KN}} \\ \cdots \\ 1 \end{bmatrix} .e^{-j2\pi\delta_p m}, \quad (19)$$

where $\odot$ denotes point-wise multiplication of the vectors. After that

$$r_m^{'} = r_m \odot c(\delta_p).e^{-j2\pi\delta_p m}, \quad (20)$$

where

$$c(\delta_p) = \begin{bmatrix} e^{\frac{+j2\pi\delta_p K(KN-1)}{KN}} \\ e^{\frac{+j2\pi\delta_p K(KN-2)}{KN}} \\ \cdots \\ 1 \end{bmatrix}. \quad (21)$$

By applying DFT block,

$$R_m^{'} = R_m \otimes DFT[c(\delta_p)].e^{-j2\pi\delta_p m}, \quad (22)$$

where $\otimes$ denotes circular convolution of vectors. The most power of the vector $DFT(c(\delta_p))$ is concentrated at the coefficients in upper and lower sides of vector. In order to a fair comparison between the proposed scheme and DBCK, we retain $2Q+1$ components (the first $Q+1$ components and Q components at the end) and force other coefficients to zero. The components of this vector can also be computed off-line and just loaded whenever they are needed.

Consequently, in addition to phase correction, a circular convolution between the vector $\mathrm{DFT}(c(\delta_p))$ and the received signal in the frequency domain is sufficient for compensation of the CFO and there is no need for several matrix multiplications. Hence, not only the complexcity of receiver is reduced, but also the latency of correction can be reduced.

3) Proposed joint CFO, TO, and channel correction method: The other advantage of the proposed scheme is possibility to compensate the timing offset (TO), channel effect and CFO correction simultaneously. According to circular convolution, (22) can be rewritten for $k = 1, 2, \cdots, KN$ as

$$R_m^{'}(k) = e^{-j2\pi\delta_p m} \sum_{i=1}^{KN} C_{app}(mod_{KN}(i-k+1))R_m(i), \quad (23)$$

where C_{app} is equal to $C = DFT[c(\delta_p)]$,where $KN-2Q+1$ of its components are zero.

Since the TO (n_τ) appears as a phase shift in frequency domain, it can be compensated with multiplying $e^{j\frac{2\pi}{KN}kn_\tau^{(p)}}$ by the value on the kth subcarrier.

In the presence of a channel $H_c(f)$, according to a zero-forcing criterion, the single-tap equalizer may be introduced as

$$H_{eq}(k) = H_c^{-1}(k/KN). \quad (24)$$

Thus, the corrected received signal of each user in the frequency domain is given by

$$R_{corr}(k) = H_{eq}(k)e^{j\frac{2\pi}{KN}kn_\tau}R_m^{'}(k). \quad (25)$$

If we define

$$C(k,i) \triangleq C_{app}(mod_{KN}(i-k+1)), \quad (26)$$

using (23) we have

$$R_{corr}(k) = \sum_{i=1}^{KN} H_{eq}(k)e^{j\frac{2\pi}{KN}(kn_\tau - \delta mKN)}C(k,i)R(i), \quad (27)$$

thereupon, the corrected signal can be achieved by

$$R_{cor} = A.R, \quad (28)$$

where A is $NK \times KN$ matrix as

$$A(i,j) \triangleq H_{eq}(i)C(i,j)e^{j\frac{2\pi}{KN}(in_\tau - \delta mKN)}. \quad (29)$$

It can be computed only for necessary components required for each user according to its subcarriers (i.e. for i equals to the number of subcarriers assign to each user and $2Q+1$ components on each row). Therefore, only with $(N/P-1)(2Q+1)$ multiplication, all CFO, TO and channel effect can be compensated and there is no need for extra multiplication in order to equalize the channel. Notice that the less delay, the better in the uplink scenario even at the cost of additional memory or parallel computation.

The most important advantage of this correction method is the matrix A remains unchanged as well as the estimated channel frequency response is valid. In other word, A can be used for some iterations in case of slow fading channel and updated only for a user which its channel changes. So matrix

A is computed parallel to data stream and once for a period of time.

C. Complexity Comparison

In the present subsection we assess the computational complexity of each CFO correction method. Here, the complexity expressions refer to the number of complex multiplications (CMs) [14]. It is assumed that there are P users and by considering one subcarrier as a guard, $\frac{N}{P}-1$ subcarriers from total of N are assigned to each user.

In the case of pre-DFT compensation, P separate receivers are needed. Each receiver after a simple shift for TO correction, performs KN multiplications to compensate CFO effect. Then P DFTs, each of size KN, have to be performed. Assuming that N is a power of 2 and FFT technique can be used, the CM for each block of FFT, by using split radix technique [15], is equal to $(KN\log_2 KN-3KN+4)/2$. At the end, each receiver uses $(N-P)/P$ real-valued matched filters related to its own subcarriers, where this subtracted P is due to filters related to the unused subcarriers between users as a guard band. Since each filter has $2K-1$ values, $(\frac{2K-1}{2})(\frac{N-P}{P})$ complex multiplications are needed for filtering in the frequency domain. Finally, KN/P multiplications per user are performed to compensate the channel effect on subcarriers of each user. So the total CMs for all P users by considering both in-phase and quadrature parts leads to

$$\begin{aligned} CM^{pre} &= 2P\{KN+\frac{1}{2}(KN\log_2 KN-3KN+4) \\ &+ (\frac{2K-1}{2})(\frac{N-P}{P})+\frac{KN}{P}\}. \end{aligned} \tag{30}$$

If CFO correction is performed after DFT by DBCK method, first of all, a KN points DFT has to be performed. After that, depending on the amount of the CFO, from 1 to 4 $KN\times KN$ matrixes with $2Q+1$ non-zero components on each column are required for each user. Thus, if we perform multiplication only for relative subcarriers, CMs for multiplication of each matrix to DFT output will be $\frac{KN}{P}\times(2Q+1)$ and by considering 4 matrixes the total CMs may be reached to $4\frac{KN}{P}\times(2Q+1)$. Then, there is $\frac{N-P}{P}$ following filters. Therefore, by considering the equalization of the channel, the total CMs can be written as

$$\begin{aligned} CM_1^{post} &= KN\log_2 KN-3KN+4+2P\{G(2Q+1)\frac{KN}{P} \\ &+ (\frac{2K-1}{2})(\frac{N-P}{P})+\frac{KN}{P}\}, \end{aligned} \tag{31}$$

where G varies from 1 to 4 according to the amount of CFO.

In the proposed method one circular convolution has to be performed in order to compensate the CFO effect. So CMs is $\frac{KN}{P}(2Q+1)$. In this case, $\frac{KN}{P}(2Q+1)$ multiplications are needed to prepare the matrix A for each user; But it is performed once. Thus the total CMs for the receiver is formed as

$$\begin{aligned} CM_2^{post} &= KN\log_2 KN-3KN+4+2P\{(2Q+1)\frac{KN}{P} \\ &+ (\frac{2K-1}{2})(\frac{N-P}{P})\}. \end{aligned} \tag{32}$$

Numerical evaluation of the computational complexities of FS for different CFO compensation techniques are listed in table I. For fair comparison between correction methods, we ignore equalization part in computing complexity. We consider the worst case and let G=4. As seen in the table, CM^{pre} is less than CM_1^{post} when P is small and becomes higher as P increases. For small amount of P, CM_2^{post} is very near to CM^{pre}. However, for all cases, CM_1^{post} is about 3 times more than CM_2^{post}.

TABLE I. COMPUTATIONAL COMPLEXITY OF FS-SMT FOR DIFFERENT CFO COMPENSATION TECHNIQUES

N	Type	$P=2$	$P=4$	$P=16$
64	FS pre-DFT	4,026	7,604	29,072
	post-DFT [10]	16,054	16,040	15,956
	Proposed post-DFT	5,302	5,288	5,204
256	FS pre-DFT	20,218	38,644	149,200
	post-DFT [10]	66,294	66,280	66,196
	Proposed post-DFT	23,286	23,272	23,188
1024	FS pre-DFT	97,274	187,380	728,016
	post-DFT [10]	273,398	273,384	273,300
	Proposed post-DFT	101,366	101,352	101,268

V. Simulation Results

Computer simulations are performed to compare the performance of the two post-DFT CFO correction schemes. We let $N=64$ and assume that there are $P=4$ users in the network and block subcarrier allocation is applied. The CFO has been chosen randomly from the uniform distribution in interval $-0.5<\delta<0.5$ and is independent of other users. In both CFO correction method $Q=3$. The bit error rate (BER) versus E_b/N_0 has been shown in Fig. 3 in the case of random CFO in the presence of the multipath channel ITU-PedA recommended by WiMAX Forum. For more details of this channel see [16]. The channel coding is a rate $1/2$ convolutional code with a constraint length of 5 and data symbols are from a 4-QAM constellation.

As seen, two methods have the reasonable performance and follow the curve obtained when there is no CFO. However, the proposed method has less computational complexity and reduces latency.

VI. Conclusion

In this paper, a CFO compensation technique has been proposed for FS structure of SMT. This approach lessens computational complexity and latency of CFO correction without losing the performance. The possibility of merging the compensation of TO and channel effect with CFO correction by this method, was clarified. Moreover, prior work on post-DFT CFO compensation has been clarified and compared with the proposed method. Simulation results demonstrated the proposed method had the same performance as the prior CFO correction technique with less complexity.

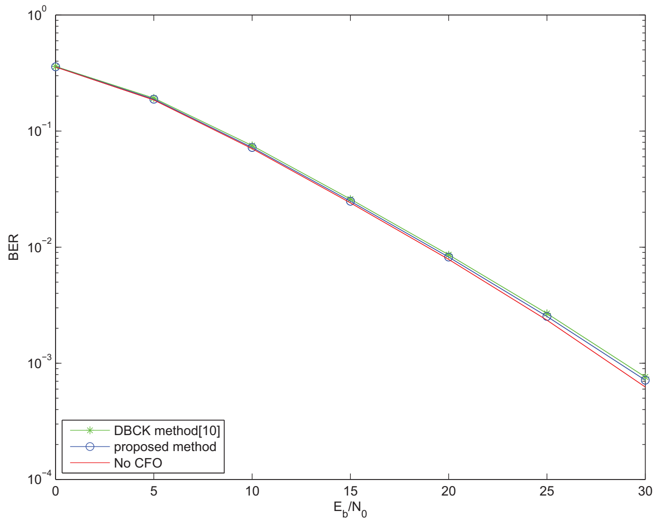

Fig. 3. BER performance for two CFO correction methods

REFERENCES

[1] A. Larmo, M. Lindstrom, M. Meyer, G. Pelletier, J. Torsner, and H. Wiemann, "The LTE link-layer design," *IEEE Communications Magazine,* vol. 47, no. 4, pp. 52-59, April 2009.

[2] B. Farhang-Boroujeny, "OFDM versus filter bank multicarrier," *IEEE Signal Processing Magazine*, vol. 28, no. 3, pp. 92–112, May 2011.

[3] B. Farhang-Boroujeny, *Signal processing techniques for software radios*. Lulu publishing house, 2008.

[4] J. Louveaux *et al.* .(2008) PHYDYAS document *1: Equalization and demodulation in the receiver (single antenna)*, [Online]. Available: http://www. ict-phydyas. org/delivrables/PHYDYAS-D3. 1. pdf/view

[5] M. Bellanger. (2010) PHYDYAS team *FBMC physical layer: a primer*, [Online]. Available: www. ict-phydyas. org

[6] M. Bellanger. "FS-FBMC: A flexible robust scheme for efficient multicarrier broadband wireless access," in *Globecom Workshops (GC Wkshps) 2012 IEEE*, Anaheim, CA, 2012, pp. 192–196.

[7] M. Morelli, C. Kuo, and M.-O. Pun, "Synchronization techniques for orthogonal frequency division multiple access (ofdma): A tutorial review," *Proc. IEEE* , vol. 95, no. 7, p. 1394, July 2007.

[8] H. Lin, M. Gharba, and P. Siohan, "Impact of time and carrier frequency offsets on the fbmc/oqam modulation scheme," *Signal Processing*, vol. 102, pp. 151–162, Sep. 2014.

[9] H. Saeedi-Sourck, Y. Wu, J. W. Bergmans, S. Sadri, and B. Farhang-Boroujeny, "Complexity and performance comparison of filter bank multicarrier and ofdm in uplink of multicarrier multiple access networks," *IEEE Transaction on Signal Processing*, vol. 59, no. 4, pp. 1907–1912, Jan. 2011.

[10] J.-B. Doré, V. Berg, N. Cassiau, and D. Kténas, "FBMC receiver for multi-user asynchronous transmission on fragmented spectrum," *EURASIP Journal on Advances in Signal Processing*, vol. 2014, no. 1, p. 41, Mar. 2014.

[11] M. Bellanger, "FS-FBMC : an alternative scheme for filter bank based multicarrier transmission," in *Processing of Communications Control and Signal, ISCCSP*, May 2012, vol. 24, no. 9, pp. 1–4.

[12] M. G. Bellanger, "Specification and design of a prototype filter for filter bank based multicarrier transmission," in *Proceedings of IEEE International Conference on Acoustics, Speech, and Signal Processing, ICASSP'01*, May 2001, vol. 4., pp. 2417–2420.

[13] A. Viholainen, M. Bellanger, and M. Huchard,(2009) PHYDYAS document "Prototype filter and structure optimization,"[Online]. Available: www. ict-phydyas. org: Document D5. 1 deliverable

[14] S. Manohar, D. Sreedhar, V. Tikiya, and A. Chockalingam, "Cancellation of multiuser interference due to carrier frequency offsets in uplink ofdma," *IEEE Transactions on Wireless Communications*, vol. 6, no. 7, pp. 2560–2571, Jul. 2007.

[15] H. V. Sorensen, M. Heideman, and C. S. Burrus, "On computing the split-radix fft," *IEEE Transactions on Acoustics, Speech and Signal Processing*, vol. 34, no. 1, pp. 152–156, Feb. 1986.

[16] R. Jain,(2007) *Channel models a tutorial1*, [Online]. Available: http://www1.cse.wustl.edu/ jain/wimax/ftp/channel-model-tutorial.pdf