A SIMPLE ICI MITIGATION METHOD FOR A SPACE-FREQUENCY CODED COOPERATIVE COMMUNICATION SYSTEM WITH MULTIPLE CFOS

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ABSTRACT

A cooperative system may have both timing errors and carrier frequency offsets (CFOs) from relay nodes. To combat timing errors, space-frequency (SF) coded OFDM system has been recently proposed for cooperative communications to achieve both full cooperative and multipath diversities without the time synchronization requirement among relay nodes. In this paper, we consider the signal detection problem in an SF coded OFDM system for cooperative communications where multiple CFOs may occur from relay nodes. By exploiting the structure of the SF codes that are rotational based in this paper, we propose a novel inter-carrier interference (ICI) mitigation method called the multiple fast Fourier transform (M-FFT) method. Simulation results illustrate that for the same symbol error rate (SER) level, the M-FFT based detection methods require less computations than other classic approaches.

Index Terms— CFO, cooperative communication, SF code, ICI.

1. INTRODUCTION

Due to the fact that OFDM systems are robust to timing errors, the SF coded cooperative system has been proposed to achieve full asynchronous cooperative diversity [1]. However, it is well-known that OFDM is sensitive to CFO which often leads to inter-carrier interference (ICI). In a SF coded cooperative communication system, because SF codes are transmitted from various distributed nodes, there may exist multiple CFOs at the destination node, which makes it difficult for the receiver to synchronize the signals from multiple relays at the same time. Therefore, ICI exists and ICI mitigation is needed to improve the system performance.

The ICI mitigation problem due to multiple CFOs in cooperative communication systems has been studied recently in [2], [3], [4]. In [2], a subcarrier-wise Alamouti coded OFDM system is considered and a simplified zero-forcing (ZF) equalizer is applied to suppress the ICI. In [3], delay diversity is considered and a minimum mean squared error (MMSE) decision feedback equalizer (DFE) is employed at the receiver. In [4], rotational based SF codes [5], [1] are considered. Compared with codes used in [2] and [3], these SF codes are powerful in the sense that they can achieve both the full cooperative diversity and the full multipath diversities, and their rate is always equal to one regardless of the number of transmit antennas. However, in [4] the SF code structure is not utilized.

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In this paper, we consider ICI mitigation for an SF coded OFDM system with multiple CFOs. By fully exploiting the structure of the SF codes in [5] and [1], we propose a novel ICI mitigation method called the multiple FFT (M-FFT) method which can be realized by simply rearranging the SF code with little loss in system performance. Both analysis and simulation results show that for the same SER level, the M-FFT based methods require less computations than other classic approaches, e.g., the well-known Q taps (Q-T) MMSE filtering method [6]. This paper is organized as follows. In Section 2 the structure of the SF codes described in [5] is reviewed. The system and the received signal models are introduced in Sections 3 and 4, respectively. The proposed ICI mitigation method is described in Section 5. Finally, simulation results are shown in Section 6. Throughout this paper, full channel knowledge including CFOs at the receiver is assumed.

Notation: We use $\mathrm{A}(l,k)$ to denote the (l,k)th entry of A, and $\mathrm{x}(k)$ to denote the kth entry of vector x. Superscripts $^{\mathcal{T}}$, *, and $^{\mathcal{H}}$ stand for transpose, conjugate, and Hermitian, respectively. $\mathrm{E}[x]$ represents the expectation of variable x. The notation $\|\mathrm{C}\|_F$ stands for the Frobenius norm of C. Integer ceiling and floor are denoted by $\lceil\cdot\rceil$ and $\lfloor\cdot\rfloor$, respectively. I_N represents the identity matrix of size $N\times N$. $\mathrm{diag}(d_0 \ \cdots \ d_{N-1})$ denotes an $N\times N$ diagonal matrix with diagonal scalar entries $d_0,\cdots,d_{N-1}$, and an $NM\times NM$ block diagonal matrix with diagonal $M\times M$ matrix entries $D_0,\cdots,D_{N-1}$ is denoted by $\mathrm{diag}(D_0 \ \cdots \ D_{N-1})$. F_N is the $N\times N$ normalized FFT matrix. The Kronecker product is denoted by $\otimes$. The notation $()_N$ means modular operation.

2. STRUCTURE OF SPACE-FREQUENCY CODES

In this section we briefly review the structure and properties of the SF codes in [5], [1], which we will utilize. Here we adopt the code structure proposed in [5]. For the coding strategy proposed in [5], each SF codeword C is a concatenation of some matrices G_P

$$C = \begin{bmatrix} G_1^{\mathcal{T}} & G_2^{\mathcal{T}} & \cdots & G_P^{\mathcal{T}} & 0^{\mathcal{T}}_{(N-N')\times M_t} \end{bmatrix}^{\mathcal{T}}, \tag{1}$$

where N is equal to the number of subcarriers in one OFDM symbol, M_t is equal to the number of transmit antennas, $P = \lfloor N/(\Gamma M_t)\rfloor$, and $N' = P\Gamma M_t$. The zero padding matrix $0_{(N-N')\times M_t}$ is an $(N-N')\times M_t$ all zero matrix which is used if the number of subcarriers N is not an integer multiple of ΓM_t. In the remainder of this paper, unless otherwise specified, we always assume that N is an integer multiple of ΓM_t. Each matrix G_p, $1\le p\le P$, has the same structure given by

$$G_p = \sqrt{M_t}\mathrm{diag}(X_1^p, X_2^p, \cdots, X_{M_t}^p), \tag{2}$$

where $X_m^p = [x_{(m-1)\Gamma+1}^p x_{(m-1)\Gamma+2}^p \cdots x_{m\Gamma}^p]^{\mathcal{T}}$, $m = 1, 2, \cdots, M_t$, and all x_k^p, $k = 1, 2, \cdots, \Gamma M_t$, are complex symbols and are mapped from an information subvector $S^p = [s_1^p, s_2^p, \cdots, s_{\Gamma M_t}^p]^{\mathcal{T}}$ by a rotation operation

$$[X_1^{p\mathcal{T}}, X_2^{p\mathcal{T}}, \cdots, X_{M_t}^{p\ \mathcal{T}}]^{\mathcal{T}} = \Theta S^p, \tag{3}$$

where s_l^p is $((p-1)\Gamma M_t + l - 1)$th entry of S for $1 \le l \le \Gamma M_t$ and satisfies the energy constraint $\mathrm{E}\left[|s_l^p|^2\right] = 1$. Θ is an $M_t\Gamma \times M_t\Gamma$ rotation matrix.

One property of the SF code is that each subcarrier is only used by one transmit antenna. From (2), the mth column of C, denoted as C_m, can be written as

$$C_m = \sqrt{M_t}\mathrm{P}_m\left(\mathrm{I}_P \otimes \Theta\right) S = \sqrt{M_t}\mathrm{P}_m\mathrm{X}, \tag{4}$$

where $\mathrm{X} = (\mathrm{I}_P \otimes \Theta) S$ and P_m is a diagonal matrix corresponding to the mth column of C. For the SF code in [5], P_m has the following form

$$\mathrm{P}_m(l, l') = \begin{cases} 1, & \text{if } l = l' = (t-1)\Gamma M_t + (m-1)\Gamma + i \\ 0, & \text{else} \end{cases}, \tag{5}$$

where $0 \le i \le \Gamma - 1$ and $1 \le t \le P$.

Furthermore, if the signal constellations are QAM and PAM, it can be shown that the optimum rotation matrix Θ is a unitary matrix [5]. Based on this property, it is not difficult to verify that the elements of X are uncorrelated.

At the receiver, each SF submatrix G_p is decoded by ML or sphere decoding independently and the maximum diversity order $\Gamma M_t M_r$ can be achieved. Here M_r is the number of receive antennas.

3. COOPERATIVE PROTOCOL

The cooperative communication system we use in this work includes one source node, one destination node, and a number of relay nodes. Here, the *decode-and-forward* cooperative protocol is adopted. In the first phase, the source node broadcasts the information while the relays receive the same information. In the second phase, the M_t relays, which have detected the received information symbols correctly, will help the source to transmit. The detected symbols are parsed into blocks of size N and the bth block, $b = 0, 1, \cdots$, is encoded to an $N \times M_t$ SF codeword matrix C^b in a distributed fashion [1]. Here N is equal to the number of subcarriers in one OFDM symbol. Finally, the mth relay transmits C_m^b, the mth column of C^b, by the standard OFDM technology. At the destination node, the received OFDM symbols are used to decode. Although in a cooperative communication system, each column of an SF code corresponds to one relay, to be consistent with the notation used in the SF coding literature we will use the term *antenna* in place of *relay*.

4. RECEIVE SIGNAL MODEL

At the destination node, after standard steps, the bth received OFDM symbol Z^b in the frequency domain is given by

$$\mathrm{Z}^b = \sqrt{\frac{\rho}{M_t}} \sum_{m=1}^{M_t} e^{j\theta_{\varepsilon_m}^b} \mathrm{U}_{\varepsilon_m}\mathrm{H}_m C_m^b + \mathrm{W}^b, \tag{6}$$

where H_m is an $N \times N$ diagonal matrix whose diagonal elements are the channel frequency response from the mth relay to destination, W^b is an $N \times 1$ vector with each entry being a zero mean unit variance complex Gaussian random variable, and ρ stands for the signal-to-noise ratio (SNR) at the destination node. $\varepsilon_m = \triangle f_m T$ is the normalized CFO where $\triangle f_m$ is the CFO between the mth relay and the destination node, and T is the OFDM symbol duration. In (6), $\theta_{\varepsilon_m}^b = 2\pi\varepsilon_m(bN + bL_{cp} + L_{cp})/N + \theta_{0,m}$ where L_{cp} is the length of cyclic prefix, and $\theta_{0,m}$ is the phase rotation between the phase of the destination node local oscillator and the carrier phase of the mth relay at the start of the received signal. $\mathrm{U}_{\varepsilon_m}$ is the ICI matrix induced by ε_m and is given by

$$\mathrm{U}_{\varepsilon_m} = \mathrm{F}_N \Omega_{\varepsilon_m} \mathrm{F}_N^{\mathcal{H}}, \tag{7}$$

where $\Omega_{\varepsilon_m} = \mathrm{diag}[1, e^{j2\pi\varepsilon_m/N}, \cdots, e^{j2\pi\varepsilon_m(N-1)/N}]$. From the definition of $\mathrm{U}_{\varepsilon_m}$, clearly $\mathrm{U}_{\varepsilon_m}$ is a unitary matrix and the element at its lth row and kth column is

$$\mathrm{U}_{\varepsilon_m}(l, k) = u_{\varepsilon_m,(k-l)_N} = \frac{\sin[\pi((k-l)_N + \varepsilon_m)] e^{j\pi((k-l)_N + \varepsilon_m)(N-1)/N}}{N\sin[\pi((k-l)_N + \varepsilon_m)/N]}, \tag{8}$$

where $0 \le l, k \le N - 1$.

From (7) and (8) we can see that if $\varepsilon_m \ne 0$, the matrix $\mathrm{U}_{\varepsilon_m}$ is no longer diagonal and that (8) shows the effect of ICI. Furthermore, if more than one CFO exists, which is possible in a cooperative communication system, it is difficult for the receiver to compensate all the CFOs at the same time.

5. ICI MITIGATION AND SIGNAL DETECTION IN THE PRESENCE OF MULTIPLE CFOS

It is shown in [7] that CFO may not decrease the diversity order of SF codes and the performance degradation can be regarded as that due to the loss of equivalent SNR, i.e., signal-to-interference-plus-noise ratio (SINR). Based on this observation, we propose to deal with ICI mitigation due to multiple CFOs by first increasing the SINR , and then decoding the SF code by ML or sphere decoding.

Substituting (4) into (6), we get the following signal model:

$$\mathrm{Z}^b = \sqrt{\rho}\mathrm{D}^b\mathrm{X}^b + \mathrm{W}^b, \tag{9}$$

where $\mathrm{D}^b = \sum_{m=1}^{M_t} e^{j\theta_{\varepsilon_m}^b}\mathrm{D}_m\mathrm{P}_m$ for $\mathrm{D}_m = \mathrm{U}_{\varepsilon_m}\mathrm{H}_m$ and the elements of X^b are uncorrelated from one another. For (9), the linear MMSE filter can maximize the output SINR [8]. Thus we can first maximize the SINR by a linear filtering operation, and then decode the SF code by ML or sphere decoding. We call this detection method the MMSE-F method. Note that when we decode the SF codes, we use two approximations. The first is that the filtered Gaussian noise is treated as if it were white. The second is that the residual ICI terms are regarded as additional Gaussian noise.

The major drawback of the MMSE-F method is its computational complexity since calculating MMSE filter coefficient is of the order of $O(N^3)$. It could be a burden when N is large. The Q taps (Q-T) method is a common way to reduce computational complexity for the OFDM system, e.g., [6], where for the concerned subcarrier only its L circular left and right subcarriers are considered. That means we use Q (Q $= 2L + 1$) taps filter instead of the full N-taps filter. Roughly speaking, if $N\mathrm{Q}^3 \le N^3$, the computational complexity can be reduced at the expense of a certain performance loss. To partially tackle the above problem, in the remainder of this section, we propose a simple and efficient ICI mitigation method, the M-FFT method.

First, at the transmitter, we re-arrange the SF code C in (1)-(2) by some row permutations and denote this row-wisely permuted SF code by C' which has the following structure:

$$C' = \mathrm{diag}\left([X_1^{1\mathcal{T}}, X_1^{2\mathcal{T}}, \cdots, X_1^{P\mathcal{T}}]^{\mathcal{T}}, \cdots, [X_{M_t}^{1\ \mathcal{T}}, X_{M_t}^{2\ \mathcal{T}}, \cdots, X_{M_t}^{P\ \mathcal{T}}]^{\mathcal{T}}\right), \tag{10}$$

where X_m^p is defined in (2). From (10) we can see that after row permutations, the non-zero entries of the mth column of C are grouped. Note that the subcarriers from $P\Gamma(m-1)$ to $P\Gamma m - 1$ are used by the mth transmit antenna. We call these subcarriers the mth group. Assume that after permutations the $((m-1)\Gamma + i)$th row, $0 \leq i \leq \Gamma - 1$, of G_p , is located at the $n^p_{(m-1)\Gamma+i}$th row, $0 \leq n^p_{(m-1)\Gamma+i} \leq N-1$, of C', i.e., $x^p_{(m-1)\Gamma+i+1}$ will be transmitted at the $n^p_{(m-1)\Gamma+i}$th subcarrier by the mth transmit antenna. Then, for (10) we have $n^p_{(m-1)\Gamma+i} = (m-1)\Gamma P + (p-1)\Gamma + i$.

In [5], it has been shown that the diversity order and coding advantage of this kind of SF codes depends only on the relative positions of the permuted rows corresponding to the entries of X_m^p with respect to the position $n^p_{(m-1)\Gamma}$. Since we do not change the relative positions for the entries of X_m^p, i.e., $n^p_{(m-1)\Gamma+i} - n^p_{(m-1)\Gamma} = i$ for each pair of m and p, C' should have the same diversity order and coding advantage as that of C. When we use C', we can get the following frequency domain SISO-OFDM model from (9) as

$$\mathrm{Z}'^b = \sqrt{\rho}\mathrm{D}'^b\mathrm{X}'^b + \mathrm{W}^b, \tag{11}$$

where $\mathrm{X}'^b = \mathrm{T}\mathrm{X}^b$, T is a permutation matrix whose element at its $\left(n^p_{(m-1)\Gamma+i}\right)$th row and $((p-1)M_t\Gamma + (m-1)\Gamma + i)$th column is equal to 1, $\mathrm{D}'^b = \sum_{m=1}^{M_t} e^{j\theta^b_{\varepsilon_m}}\mathrm{D}_m\mathrm{P}'_m$, and P'_m is defined as

$$\mathrm{P}'_m(l,l') = \begin{cases} 1, & \text{if } l = l' \text{ and } (m-1)P\Gamma \leq l \leq mP\Gamma - 1 \\ 0, & \text{else} \end{cases} . \tag{12}$$

At the receiver in the time domain we apply frequency shift (FS) and FFT operations M_t times. Each time CFO ε_l, $1 \leq l \leq M_t$, is completely compensated by the FS. Then, after the FFT operation we can get signal vector Z_l^b in the frequency domain as

$$\mathrm{Z}'^b_l = \sqrt{\rho}\mathrm{D}'^b_l\mathrm{X}'^b + \mathrm{W}^b_l, \tag{13}$$

where $\mathrm{D}'^b_l = \sum_{m=1}^{M_t} e^{j(\theta^b_{\varepsilon_m} - \theta^b_{\varepsilon_l})}\mathrm{D}_{m,l}\mathrm{P}'_m$, $\mathrm{D}_{m,l} = \mathrm{U}_{\varepsilon_{m,l}}\mathrm{H}_m$ and $\varepsilon_{m,l} = \varepsilon_m - \varepsilon_l$.

Since ε_l is compensated, we have $\mathrm{D}'^b_l\mathrm{P}'_l = \mathrm{H}_l\mathrm{P}'_l$. This means that the subcarriers of the lth group do not leak their power to all other subcarriers whether they are in the same group or not. Unfortunately, due to the effect of multiple CFOs, the ICI power arriving from other groups may be increased by this FS operation. However, it is also highly probable that the SINR of the subcarriers in the lth group can be significantly increased.

From (7)-(8) for the ICI coefficients, one can see that for the desired kth subcarrier in the lth group, most of the ICI power comes from a few of its neighbors and most of which are in the same group. Considering another subcarrier, say the ith subcarrier, as the distance $|k-i|$ increases, the ICI power coming from the ith subcarrier drastically decreases. Therefore, although the ICI power arriving from other groups may be increased by this FS, its effect is insignificant compared with the canceled ICI power of the same group. Furthermore, the subcarriers in the middle of the lth group should suffer from smaller ICI power than those subcarriers which are at the boundaries of lth group do. Because both sides of the middle subcarrier are separated from other groups and as the position of subcarrier shifts from the middle to left or right boundary, the ICI power coming from $((l-2)_{M_t}+1)$th or $((l)_{M_t}+1)$th group will rapidly increase.

Therefore, after the lth FS and FFT operation, we may only want to preserve the signals of the lth group denoted by z'^b_l which is a $P\Gamma\times 1$ vector. Finally, we collect all of z'^b_l for $1 \leq l \leq M_t$ and construct an equivalent received signal vector $\bar{\mathrm{Z}}'^b = [\mathrm{z}'^{b\mathcal{T}}_1, \mathrm{z}'^{b\mathcal{T}}_2, \cdots, \mathrm{z}'^{b\mathcal{T}}_{M_t}]^{\mathcal{T}}$. It is easy to show that shifting the frequency of the received signal in the time domain by $-\varepsilon_l$ is equivalent to multiplying Z'^b by $(e^{j\theta^b_{\varepsilon_l}}\mathrm{U}_{\varepsilon_l})^{\mathcal{H}}$. Let $\mathrm{U}^b = \sum_{l=1}^{M_t} e^{j\theta^b_{\varepsilon_l}}\mathrm{U}_{\varepsilon_l}\mathrm{P}'_l$. Then, $\bar{\mathrm{Z}}'^b$ can be expressed as

Method	Number of multiplications
MMSE-F	$O(14N^3)$
Q-T	$O(13NQ^3)$
M-FFT	$M_tN + (M_t-1)\frac{N}{2}\log_2 N$

Table 1. Comparison of computational complexity.

$$\begin{aligned} \bar{\mathrm{Z}}'^b &= \sqrt{\rho}\mathrm{U}^{b\mathcal{H}}\mathrm{D}'^b\mathrm{X}'^b + \mathrm{U}^{b\mathcal{H}}\mathrm{W}^b \\ &= \sqrt{\rho}\mathrm{U}^{b\mathcal{H}}\left(\sum_{l=1}^{M_t} e^{j\theta^b_{\varepsilon_l}}\mathrm{U}_{\varepsilon_l}\mathrm{P}'_l\right)\left(\sum_{l=1}^{M_t}\mathrm{P}'_l\mathrm{H}_l\right)\mathrm{X}'^b + \mathrm{U}^{b\mathcal{H}}\mathrm{W}^b, \end{aligned} \tag{14}$$

where the second equality follows from the fact that P'_m has the same property as that of P_m described in (5). Because $\left(\sum_{l=1}^{M_t}\mathrm{P}'_l\mathrm{H}_l\right)$ is a diagonal matrix, $\bar{\mathrm{Z}}'^b$ is actually the matched filter output of Z'^b. Then, based on (14) we can directly decode X'^b by ML or sphere decoding by using the same approximations as we do for the MMSE-F or the Q-T methods. Furthermore, for this M-FFT method, we have the following observations:

- Referring to [9], the computational complexities of different methods are evaluated in Table 1. We can see that the computation complexity of the M-FFT method is much less than that of other two methods for reasonable value of N and M_t.
- If we combine M-FFT with the successive interference cancellation (SIC), the decoding order of G_p should depend on the subcarriers where G_p are located. For example, considering the structure of C', $G_{\lfloor P/2 \rfloor}$ or $G_{\lceil P/2 \rceil}$ should first be decoded. Although this is not the optimal decoding order, we do not need additional operations to decide the decoding order.
- Since the boundary subcarriers of each group have the worst quality/SINR, we may not use them to improve the SER performance at the cost of even small efficiency loss. On the other hand, if zero padding exists, we can use them to separate groups. We call this scheme the M-FFT-Zn method, where n is the number of unused subcarriers.
- M-FFT can also be used for the preprocessing of the Q-T method to improve their performance. We call this two-stage scheme the M-FFT-Q-T method.

6. SIMULATION RESULTS

An $M_t = 2$ system with 64 OFDM tones is simulated. The bandwidth is 20 MHz and the cyclic prefix is fixed to 16. The channels from relays to destination are all frequency-selective fading with two equal power rays $[\tau_m(0), \tau_m(1)] = [0, 0.5]\mu s$ for $1 \leq m \leq M_t$. We also assume that the receiver only has one receive antenna. For each channel realization, each ε_m, is uniformly selected from $[-\varepsilon_{Max}, \varepsilon_{Max}]$. The SF code proposed in [5] is applied.

As shown in Fig.1, as SNR increases, the OFDM system will quickly suffer from an error floor if we directly decode (referred to as DD). On the other hand, in the simulated SNR range, the MMSE-F method works well. Compared with the MMSE-F method, the performance loss of the Q-T method is obvious when Q is small. The M-FFT method outperforms the Q-T method with $Q = 5$. Finally, to demonstrate that the grouping operation of subcarriers plays an important role in the M-FFT method, we also simulate a scheme that we directly apply multiple FS and FFT operations without the grouping operation and it is denoted by M-FFT-WG . From Fig.1 we can see that the performance of the M-FFT-WG method is worse than that of the directly decoding. This is because without the grouping step, in the FS operation, the increased ICI power from other antennas can even exceed the canceled ICI power.

Fig.2 shows some extensions of the M-FFT method. For M-FFT-Z4, the 0th to 29th subcarriers are used by the first transmit antenna, the 32nd to 61th subcarriers are used by the second transmit antenna and all the other subcarriers are not used. We can see that the performance of the M-FFT method is greatly increased and is comparable with that of the Q-T method with $Q = 15$. The cost is that the frequency efficiency is decreased by 6.25%. The effect of zero padding can be explained as follows. Since no signals are transmitted at these four subcarriers, no power leaks from them. More importantly, for each group, a large part of the ICI power from the other groups falls on these 4 unused subcarriers. Note that for the Q-T method, zero padding can also increase SER performance. However, the improvement is small because only the subcarriers near these zero pads can significantly benefit from them. For the M-FFT-Q-T method, when $Q = 4$, it outperforms the M-FFT-Z4 method. Finally, we combine the M-FFT-Z4 method with SIC. For our simulated system, the decoding order is $G_8, G_9, G_7, \cdots, G_{15}, G_1$ according to the average SINR of subcarriers. We also plot the curve of the M-FFT-Q-T method with $Q = 7$. For a fair comparison, four zero paddings are used at the same subcarriers as that of the M-FFT-Z4 method. Therefore, it is denoted by M-FFT-Q-T-Z4. We can see that both these two schemes have a comparable performance with the MMSE-F method. Note that for the Q-T method, 15 is the maximum value of Q when $N = 64$ since if $Q > 15$, the Q-T methods would lose their computational efficiency compared with the MMSE-F method.

7. CONCLUSION

In this paper, we have proposed a novel ICI mitigation method, M-FFT, for the multiple CFOs problem in an SF coded cooperative communication system. Our proposed M-FFT method takes the SF code into account and has a low computational complexity. Simulation results have illustrated the effectiveness of the method.

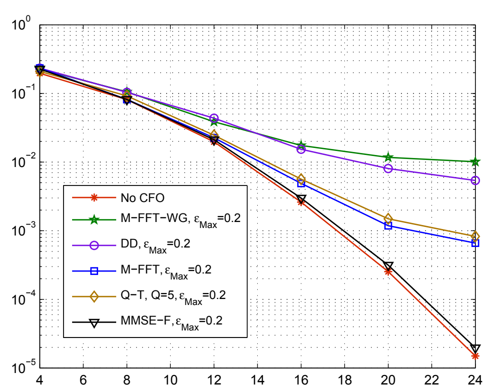

Fig. 1. SER performance of the MMSE-F, Q-T, and M-FFT methods.

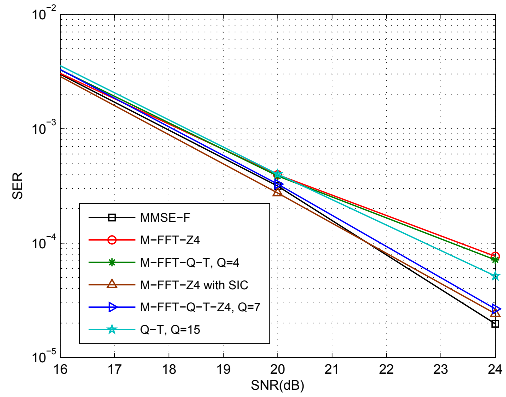

Fig. 2. Extensions of the M-FFT method when $\varepsilon_{Max} = 0.2$.

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