Partial FFT Demodulation for MIMO-OFDM Over Time-Varying Underwater Acoustic Channels

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***Abstract*—Partial FFT demodulation is a newly-emerging technique to mitigate the inter-carrier interference (ICI) of orthogonal frequency division multiplexing (OFDM) systems over time-varying underwater acoustic channels. In this letter, we extend the partial FFT demodulation method for a single-input single-output (SISO) configuration to the multiple-input multiple-output (MIMO) case. By assuming no channel knowledge, we design an adaptive algorithm which performs sliding-window channel estimation, partial FFT combining and data detection across subcarriers iteratively. Furthermore, a new parameter "residual ICI span" is introduced to counteract the post-combining ICI and provide a better system performance.**

***Index Terms*—MIMO-OFDM, partial FFT, time-varying channels, underwater acoustic communications.**

I. Introduction

UNDERWATER Acoustic (UWA) channels are recognized as one of the most challenging communication media, characterized by long multipath propagation and severe time variation [1]. To combat the delay spread and achieve high-rate transmission over UWA channels, orthogonal frequency division multiplexing (OFDM) has been widely investigated since it avoids intersymbol interference (ISI) using a guard interval and mitigates the frequency selectivity of a multipath channel with a simple frequency-domain one-tap equalizer [2]. However, due to the low velocity of acoustic waves (nominally 1500 m/s), the Doppler effect measured by the normalized carrier frequency offset is often on the order of 10^{-4} in UWA channels with mobile transceivers, which is several orders of magnitude greater than in wireless radio channels. The resulting time variation within one OFDM block corrupts the orthogonality among the subcarriers and generates a power leakage, known as inter-carrier interference (ICI), which may cause significant degradation in system performance.

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How to cope with ICI in OFDM systems depends heavily on the assumptions of the time variation in the underlying channel models. For UWA channels, the receiver in [3] assumes a common Doppler scale among all paths and neglects ICI after proper resampling and carrier frequency offset compensation. Other receivers use a basis expansion model (BEM) [4], [5] or a path-specific Doppler model [6], [7] to approximate doubly selective UWA channels and take ICI mitigation into account explicitly. Despite the benefit in performance, these systems usually require a significant pilot overhead and extra ICI equalization. As such, recently, a novel partial FFT demodulation based on a subblock-wise fading model is proposed for single-input single-output (SISO) OFDM systems over UWA channels [8]. Mathematically, it is equivalent to imposing a different window for each subcarrier, and hence it has the capability to alleviate ICI effectively. Similar strategies have been adopted for hybrid carrier modulation [9], [10], multichannel and differential OFDM systems [11], [12].

In this letter, assuming channel state information is *a priori* unknown, we design an adaptive algorithm to extend the partial FFT demodulation to multiple-input multiple-output (MIMO) OFDM systems, which has the following features:

- The algorithm jointly performs channel estimation, weight updating and data detection. By this means, partial FFT combining is applied across subcarriers iteratively to accommodate for the variation of ICI.
- To address the problem of the increased number of model parameters in channel estimation, a sliding window of subcarriers is established by exploiting the frequency correlation among adjacent subcarriers.
- A new parameter "residual ICI span" is also introduced to counteract the post-combining ICI and enhance the accuracy of channel estimation. By this way, the data detection becomes more robust against the time variation of UWA channels. Numerical results verify its benefit on the bit error rate (BER) performance.

Notation: $(\cdot)^*$ stands for conjugate, $(\cdot)^T$ for transpose, $(\cdot)^H$ for Hermitian transpose, $(\cdot)^\dagger$ for Moore-Penrose pseudo-inverse, and $|\cdot|$ for absolute value. $\mathrm{diag}\{\mathbf{x}\}$ is a diagonal matrix with the vector $\mathbf{x}$ on its diagonal. $\mathbf{1}_N$ and $\mathbf{0}_N$ denote the $N \times 1$ all-one and all-zero vectors respectively.

II. System Model

We consider a MIMO-OFDM system with N transmitters, M receivers, and K subcarriers. The OFDM symbol duration is T with the corresponding subcarrier spacing $\Delta f = 1/T$, and a cyclic prefix (CP) of duration T_g is added. For each transmitter

n, let d_k^n denote the complex symbol to be modulated on the kth subcarrier at the frequency $f_k = f_0 + (k-1)\Delta f$. Then the transmitted signal in passband can be written as

$$s^n(t) = \sqrt{\frac{1}{T}}\,\mathrm{Re}\left\{\sum_{k=1}^{K} d_k^n e^{j2\pi f_k t}\right\}, \quad t \in [-T_g, T]. \tag{1}$$

The UWA channel between transmitter n and receiver m which consists of P discrete paths is modeled as

$$h^{m,n}(\tau, t) = \sum_{p=1}^{P} h_p^{m,n}(t)\delta\left(\tau - \tau_p^{m,n}(t)\right), \tag{2}$$

where $h_p^{m,n}(t)$ and $\tau_p^{m,n}(t)$ are the time-varying amplitude and delay of the pth path, respectively. At receiver m, after removing the CP and frequency f_0, the complex baseband received signal can be expressed as

$$y^m(t) = \sqrt{\frac{1}{T}}\sum_{n=1}^{N}\sum_{k=1}^{K}\tilde{H}_k^{m,n}(t)d_k^n e^{j2\pi(k-1)\Delta f t} + z^m(t), \quad t \in [0, T], \tag{3}$$

where $z^m(t)$ is the additive white Gaussian noise, and

$$\tilde{H}_k^{m,n}(t) = \sum_{p=1}^{P} h_p^{m,n}(t) e^{-j2\pi f_k \tau_p^{m,n}(t)} \tag{4}$$

represents the channel frequency response.

To cope with the time variation of UWA channels, we extend the partial FFT demodulation first proposed for SISO-OFDM in [8] to a MIMO scenario. Specifically, partial FFT demodulation is implemented by dividing the OFDM symbol duration $[0, T]$ into Q non-overlapping intervals and performing a Fourier transform on each windowed segment. The output for the qth segment on the kth subcarrier at receiver m is then

$$\begin{aligned} y_k^m(q) &= \sqrt{\frac{1}{T}}\int_{\frac{(q-1)T}{Q}}^{\frac{qT}{Q}} y^m(t)e^{-j2\pi(k-1)\Delta f t}dt \\ &= \frac{1}{T}\sum_{n=1}^{N}\sum_{l=1}^{K} d_l^n \int_{\frac{(q-1)T}{Q}}^{\frac{qT}{Q}} \tilde{H}_l^{m,n}(t)e^{j2\pi(l-k)\Delta f t}dt \\ &\quad + \bar{z}_k^m(q), \quad q = 1, 2, \ldots, Q. \end{aligned} \tag{5}$$

where $\bar{z}_k^m(q)$ is the noise term.

For the subblock-wise fading channel model, we assume that, although the UWA channel may not be deemed as quasi-static within an entire OFDM symbol, the channel parameters do not change significantly over each segment. As such, the time-varying frequency response in the qth interval $[(q-1)T/Q, qT/Q]$ can be approximated by its midpoint value $\bar{H}_l^{m,n}(q) = \tilde{H}_l^{m,n}(\frac{2q-1}{2Q}T)$, and thus the output in (5) is simplified as

$$y_k^m(q) \approx \sum_{n=1}^{N}\sum_{l=1}^{K} d_l^n \bar{H}_l^{m,n}(q) v_{l-k}(q) + \bar{z}_k^m(q). \tag{6}$$

Here, $v_i(q)$ represents the effect of partial integration over the qth interval, i.e., [8]

$$\begin{aligned} v_i(q) &= \frac{1}{T}\int_{\frac{(q-1)T}{Q}}^{\frac{qT}{Q}} e^{j2\pi i \Delta f t}dt \\ &= \frac{1}{Q}e^{j\frac{\pi i(2q-1)}{Q}}\,\mathrm{sinc}\left(\frac{\pi i}{Q}\right), \\ i &= -K+1, \ldots, K-1, \end{aligned} \tag{7}$$

where $\mathrm{sinc}(x) = \frac{\sin x}{x}$. Furthermore, for brevity, we can rewrite the matrix-vector counterpart of (6) as

$$\mathbf{y}_k^m = \sum_{n=1}^{N}\sum_{l=1}^{K} d_l^n \bar{\mathbf{H}}_l^{m,n}\mathbf{v}_{l-k} + \bar{\mathbf{z}}_k^m, \tag{8}$$

where we have $\mathbf{y}_k^m = [y_k^m(1), \ldots, y_k^m(Q)]^T$, $\mathbf{v}_{l-k} = [v_{l-k}(1), \ldots, v_{l-k}(Q)]^T$, $\bar{\mathbf{z}}_k^m = [\bar{z}_k^m(1), \ldots, \bar{z}_k^m(Q)]^T$, and define $\bar{\mathbf{H}}_l^{m,n} = \mathrm{diag}\{[\bar{H}_l^{m,n}(1), \ldots, \bar{H}_l^{m,n}(Q)]\}$ as the pre-combining channel matrix.

III. Partial FFT Algorithm

In this section, we will introduce the proposed partial FFT algorithm for MIMO-OFDM systems. We will start from the iterative processing of channel estimation, weight updating and data detection at each subcarrier, and proceed to the initialization and termination operations.

A. Iterative Processing

Partial FFT demodulation compensates the Doppler distortion and reduces the ICI by weighted combining of the partial FFT outputs in (8) judiciously. With full channel knowledge, zero-forcing (ZF) or minimum mean square error (MMSE) criteria can be utilized to estimate the partial FFT weights. However, for practical time-varying UWA channels, since the channel frequency response parameters at each segment's midpoint are *a priori* unknown and nontrivial to estimate, we develop a four-step iterative approach, and its operations on the kth subcarrier are as follows.

1) Partial FFT Combining: We assume all transmitters are colocated and experience approximately the same time-varying effect. In this case, on the kth subcarrier, let us define $\mathbf{d}_k = [d_k^1, \ldots, d_k^N]^T$ as the transmitted symbol vector, and $\mathbf{w}_k^m = [w_k^m(1), \ldots, w_k^m(Q)]^T$ for each receiver m to combat the ICI of OFDM signals from all transmitters. The combining of partial FFT outputs yields

$$x_k^m = \mathbf{w}_k^{mH}\mathbf{y}_k^m = \sum_i \mathbf{H}_{k,i}^{mT}\mathbf{d}_{k-i} + z_k^m, \tag{9}$$

where $z_k^m = \mathbf{w}_k^{mH}\bar{\mathbf{z}}_k^m$ is the noise term, and $\mathbf{H}_{k,i}^m$ is the $N \times 1$ post-combining channel vector defined as

$$\mathbf{H}_{k,i}^m = \left[\bar{\mathbf{H}}_{k-i}^{m,1}\mathbf{v}_{-i}, \ldots, \bar{\mathbf{H}}_{k-i}^{m,N}\mathbf{v}_{-i}\right]^T \mathbf{w}_k^{m*}. \tag{10}$$

Under an ideal assumption that the weighted combining of the partial FFT outputs eliminates most of the ICI, i.e., $\mathbf{H}_{k,i}^m \approx \mathbf{0}_N$, $i \neq 0$, we can further write (9) as

$$x_k^m = \mathbf{H}_{k,0}^{mT}\mathbf{d}_k + \underbrace{\sum_{i\neq 0}\mathbf{H}_{k,i}^{mT}\mathbf{d}_{k-i} + z_k^m}_{\xi_k^m}, \tag{11}$$

where ξ_k^m contains the additive noise and all residual ICI. This is also the strategy adopted by [8] for SISO-OFDM. However, in practice, the residual ICI usually can not be ignored completely, especially for the MIMO context considered herein. It degrades the accuracy of channel estimation in the following step, and eventually incurs a performance loss of partial FFT demodulation. Therefore, inspired by [13], [14], we introduce a new parameter I referred to as "residual ICI span" in this letter, by which (9) is reformulated as

$$x_k^m = \sum_{i=-I}^{I} \mathbf{H}_{k,i}^{mT} \mathbf{d}_{k-i} + \underbrace{\sum_{|i|>I} \mathbf{H}_{k,i}^{mT} \mathbf{d}_{k-i} + z_k^m}_{\eta_k^m}. \tag{12}$$

2) Post-Combining Channel Estimation: Since the variation of the frequency responses is much faster than that of the partial FFT weights across subcarriers in OFDM systems, a coupled post-combining channel estimator is necessary to prevent potential convergence or instability problems.

It can be seen in (12) that there are $N(2I+1)$ channel coefficients to be estimated, while only one observation is available. Therefore, to guarantee the problem not to be underdetermined, we design a sliding-window method, in which $K_w \geq N(2I+1)$ adjacent OFDM subcarriers are grouped in a window and shifted by one at each iteration. Moreover, we leverage the frequency correlation of UWA channels to assume that the frequency response over each window is flat and represented by its middle subcarrier, i.e.,

$$\mathbf{H}_{k+l,i}^{m,n} = \mathbf{H}_{k,i}^{m,n}, \quad \mathbf{w}_{k+l}^m = \mathbf{w}_k^m, \tag{13}$$

for $l = -K_w/2, \ldots, K_w/2-1$ and $i = -I, \ldots, I$. Here, K_w is chosen as an even number. We then obtain (14), shown at the bottom of the page. In the matrix $\mathbf{D}_k$, $\{\mathbf{d}_{k+l} | l = -K_w/2-I, \ldots, K_w/2-2+I\}$ are estimated from previous iterations, and only $\mathbf{d}_{k+K_w/2-1+I}$ is unknown. However, the channel estimates for the corresponding subcarrier are not available at this point, hence we resort to the frequency correlation assumption again for a tentative decision. To this end, an ICI-ignored scheme is employed, i.e.,

$$\hat{\mathbf{d}}_{k+K_w/2-1+I} = \mathrm{dec}\left\{\hat{\mathbf{H}}_{k-1,0}^{\dagger} \mathbf{x}_{k+K_w/2-1+I}\right\}. \tag{15}$$

Here, $\mathrm{dec}\{\cdot\}$ is the slicer operation producing the tentative decision $\hat{\mathbf{d}}_{k+K_w/2-1+I}$, $\hat{\mathbf{H}}_{k-1,0} = [\hat{\mathbf{H}}_{k-1,0}^1, \ldots, \hat{\mathbf{H}}_{k-1,0}^M]^T$ denotes the estimated post-combining channel matrix with $\hat{\mathbf{H}}_{k-1,0}^m$ as the estimate of $\mathbf{H}_{k-1,0}^m$, and $\mathbf{x}_{k+K_w/2-1+I} = [\mathbf{w}_k^{1H}\mathbf{y}_{k+K_w/2-1+I}^1, \ldots, \mathbf{w}_k^{MH}\mathbf{y}_{k+K_w/2-1+I}^M]^T$.

Based on (14), the estimate of the stacked post-combining channel vector $\mathbf{H}_k^m$ is updated as

$$\hat{\mathbf{H}}_k^m = \alpha \hat{\mathbf{H}}_{k-1}^m + (1-\alpha)\,\hat{\mathbf{D}}_k^{\dagger} \bar{\mathbf{x}}_k^m \tag{16}$$

where $\alpha \in [0, 1)$ is the tracking parameter, and $\hat{\mathbf{D}}_k$ is a decision matrix with similar structure as $\mathbf{D}_k$.

3) Symbol Decisions Refining: Given the frequency response $\hat{\mathbf{H}}_k^m$ in (16), the refined decisions $\hat{\mathbf{d}}_k$ at the middle subcarrier of the sliding window can be obtained based on a similar ICI-ignored scheme as in (15). These refined symbols are then inserted into $\hat{\mathbf{D}}_k$ and substituted for the tentative ones.

4) Partial FFT Weights Updating: Unlike the typical windowing methods for ICI cancellation which utilize a common window for all subcarriers [13], partial FFT demodulation imposes a frequency-dependent window for each subcarrier [8], thus a better performance can be expected. Bearing this in mind, we update the partial FFT weights using the recursive least squares (RLS) algorithm

$$\mathbf{w}_{k+1}^m = \mathbf{w}_k^m + \mathrm{RLS}\left\{\mathbf{y}_k^m, e_k^m\right\}, \tag{17}$$

where the error on the kth subcarrier is defined to maximize the ICI cancellation, i.e.,

$$e_k^m = \hat{\mathbf{H}}_{k,0}^{mT} \hat{\mathbf{d}}_k - x_k^m. \tag{18}$$

Remark 1: When $I = 0$, the partial FFT demodulation proposed here reduces to a straightforward MIMO extension of the recursive weight (RW) estimation algorithm in [8]. However, in this case, the channel estimation in (16) may suffer from lower accuracy due to the indiscrimination between interference and noise in (11), and may finally lead to a convergence problem in (17).

Remark 2: The channel frequency correlation assumption is repeatedly utilized for the subcarrier grouping in (13) and the tentative decisions in (15). Quantitatively, it implies the following constraint on the channel coherence bandwidth B_c:

$$B_c \gg \frac{1}{T}\max\left\{K_w, \frac{K_w}{2} + I\right\} = \frac{K_w}{T}. \tag{19}$$

Equivalently, for a given bandwidth B_c, this inequality leads to an upper bound on the residual ICI span I.

Remark 3: The tentative and refined decisions in steps 2 and 3 can also use a partial ICI cancellation scheme which removes the ICI contributed by the previously detected I subcarriers. Generally, it can be expected to improve performance at the expense of additional complexity. However, to highlight the ICI reduction performance of the partial FFT demodulation, we focus here on the ICI-ignored scheme only.

$$\underbrace{\begin{bmatrix} \mathbf{w}_k^{mH}\mathbf{y}_{k-K_w/2}^m \\ \vdots \\ \mathbf{w}_k^{mH}\mathbf{y}_k^m \\ \vdots \\ \mathbf{w}_k^{mH}\mathbf{y}_{k+K_w/2-1}^m \end{bmatrix}}_{\bar{\mathbf{x}}_k^m} = \underbrace{\begin{bmatrix} \mathbf{d}_{k-K_w/2-I}^T & \cdots & \mathbf{d}_{k-K_w/2}^T & \cdots & \mathbf{d}_{k-K_w/2+I}^T \\ \vdots & & \vdots & & \vdots \\ \mathbf{d}_{k-I}^T & \cdots & \mathbf{d}_k^T & \cdots & \mathbf{d}_{k+I}^T \\ \vdots & & \vdots & & \vdots \\ \mathbf{d}_{k+K_w/2-1-I}^T & \cdots & \mathbf{d}_{k+K_w/2-1}^T & \cdots & \mathbf{d}_{k+K_w/2-1+I}^T \end{bmatrix}}_{\mathbf{D}_k} \underbrace{\begin{bmatrix} \mathbf{H}_{k,I}^m \\ \vdots \\ \mathbf{H}_{k,0}^m \\ \vdots \\ \mathbf{H}_{k,-I}^m \end{bmatrix}}_{\mathbf{H}_k^m} + \underbrace{\begin{bmatrix} \eta_{k-K_w/2}^m \\ \vdots \\ \eta_k^m \\ \vdots \\ \eta_{k+K_w/2-1}^m \end{bmatrix}}_{\boldsymbol{\eta}_k^m} \tag{14}$$

B. Initialization and Termination

1) Initialization: The algorithm starts at the $(K_w/2 + I + 1)$th subcarrier. To initialize the iteration among steps 1-4, we set $\mathbf{w}^m_{K_w/2+I+1} = \mathbf{1}_Q$ and

$$\hat{\mathbf{H}}^m_{K_w/2+I,i} = \begin{cases} \mathbf{1}_N, & i = 0, \\ \mathbf{0}_N, & 0 < |i| \leq I. \end{cases} \tag{20}$$

Moreover, to guarantee the convergence of the algorithm, we use a similar pilot arrangement as in [8], which includes the pilots on the first $K_a = K_w + 2I + 2Q$ subcarriers to train initially, and some extra K_b pilots equally spaced within the remaining $K - K_a$ subcarriers to alleviate the error propagation induced by deep fades in the channel frequency response.

2) Termination: The length-K_w sliding window stops moving at the $(K - K_w/2 - I + 1)$th subcarrier, and we set

$$\hat{\mathbf{H}}^m_{k,i} = \hat{\mathbf{H}}^m_{K-K_w/2-I+1,i}, \tag{21}$$

for $k = K - K_w/2 - I + 2, \ldots, K$ and $|i| \leq I$. Also, after the iterations end, a similar recomputation for the channel frequency responses is needed to smooth the estimates [8].

Furthermore, the iteration among steps 1-4 can be carried out multiple times. In this case, soft decisions and full ICI cancellation can be employed for tentative and refined symbol estimation to improve performance. However, including these aspects here would be out of the scope of this letter.

IV. Simulation Results

In this section, numerical simulation results are provided to illustrate the BER performance of our proposed partial FFT demodulation algorithm. In the following simulations, each OFDM modulator utilizes $K = 1024$ subcarriers with a total bandwidth of 4.096 kHz at a center frequency of $f_c = 10$ kHz. The subcarrier spacing is $\Delta f = 4$ Hz and the corresponding OFDM symbol period is $T = 1/\Delta f = 0.25$ s. We assume that the $N \times M$ UWA channels are independent, and each channel uses a simple two-ray uniform power delay profile with a maximum delay spread of $\tau_{\max} = 2$ ms. Therefore, the channel coherence bandwidth can be coarsely calculated as $B_c \approx 1/\tau_{\max} = 125/T$, which justifies the frequency correlation assumption in (19). Moreover, a cyclic prefix of duration $T_g = 5$ ms is chosen to eliminate ISI.

Fig. 1 compares the BER performance of the proposed algorithm to that of the RW algorithm in [8] for a $M = 1$, $N = 1$ scenario. We simulate different time-varying UWA channels by changing the normalized Doppler scale, namely, $f_d = af_cT =$ 0.25, 0.375 and 0.50, which correspond to Doppler scaling factors $a = 1 \times 10^{-4}$, 1.5×10^{-4} and 2×10^{-4}, respectively. We set the tracking parameter to $\alpha = 0.2$ and use a RLS forgetting factor of 0.99. First, it can be seen that the RW algorithm suffers a severe performance degradation as f_d increases. This observation supports our analysis in Remark 1 of Section III. The superiority of our algorithm is attributed to the nonzero "residual ICI span" I. However, it is interesting to note that our algorithm is surpassed slightly by the RW algorithm at $f_d = 0.25$ and it performs better with $I = 1$ than with $I = 2$. This suggests that the benefit of introducing a larger I may be offset by estimation errors due to the increased number of model parameters, and a judicious tradeoff should be taken accordingly.

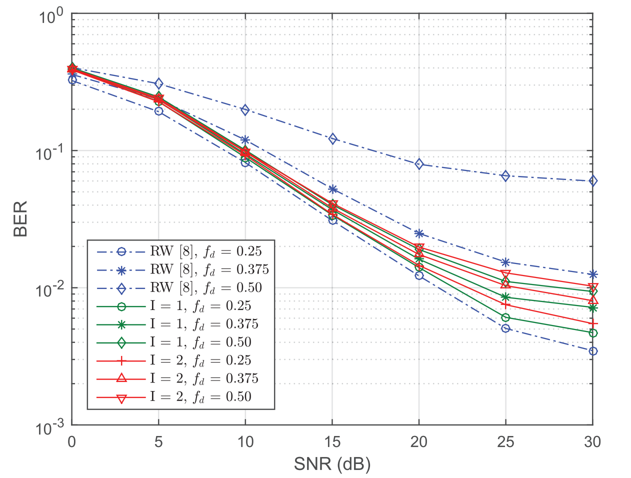

Fig. 1. BER performances of the partial FFT algorithms for a 1×1 scenario with various normalized Doppler scales.

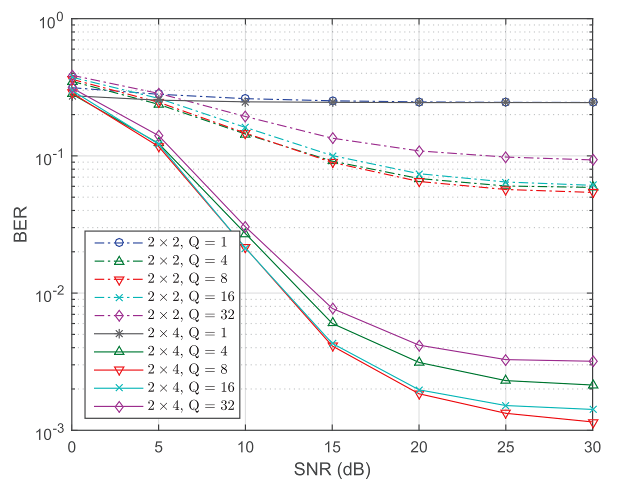

Fig. 2. BER performances of the partial FFT algorithms for 2×2 and 2×4 scenarios with a normalized Doppler scale of 0.25, $I = 1$ and various Q.

Fig. 2 depicts the BER performance of the proposed algorithm for MIMO scenarios with $N \times M = 2 \times 2$, 2×4. We use $I = 1$, $f_d = 0.25$ and various Q. Here, $Q = 1$ represents the conventional full FFT demodulation, which fails to keep up with the channel variations. Clearly, the proposed algorithm outperforms its full FFT counterpart and collects more spatial diversity gains in the 2×4 system. Meanwhile, the performance saturation mentioned in [8] is also witnessed, i.e., for the specific f_d, the proposed algorithm with $Q = 8$ yields the lowest BER. Larger values of Q may cause slower convergence of (17) and inferior ICI compensation.

V. Conclusion

An adaptive partial FFT demodulation algorithm is proposed for MIMO-OFDM systems. Simulations show its performance merits over time-varying UWA channels and its performance tradeoffs in choosing the algorithm design parameters.

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