Performance Results of Two Iterative Receivers for Distributed MIMO OFDM with Large Doppler Deviations

Jianzhong Huang, Student Member, IEEE, Shengli Zhou, Senior Member, IEEE, Zhaohui Wang, Student Member, IEEE

Abstract—This paper studies a distributed OFDM system with multiple quasi-synchronous users, where different users may transmit different numbers of parallel data streams. The distinction from most existing work is that the multipath channels for different users have significantly different Doppler scales. Such a setting with two single-transmitter users was first studied in a recent publication by Tu et al. This paper presents two iterative receivers, termed as multiuser detection (MUD)-based and single-user detection (SUD)-based receivers, respectively. The MUD-based receiver adopts a frequency-domain-oversampling front end on each receive element, then performs joint channel estimation and multiuser data detection iteratively. The SUD-based receiver adopts conventional single-user processing modules, but adds a critical step of multiuser interference (MUI) cancellation, where the MUI reconstruction explicitly considers different resampling factors used by different users. Experimental data sets from MACE10 and SPACE08 are used to emulate a distributed OFDM system with different numbers of users and different numbers of data streams per user. Performance results in different settings validate the effectiveness of the proposed iterative receivers.

Index Terms—Iterative receiver, Doppler scale, multiuser interference cancellation, underwater acoustic communications, distributed MIMO OFDM

I. INTRODUCTION

In addition to various approaches recently developed for single-transmitter systems [2]–[22], multi-input multi-output (MIMO) techniques have been actively pursued for underwater acoustic communications, which enable high spectral efficiency through spatial modulation for both single-carrier [23]–[30] and multicarrier transmissions [31]–[35].

• For single-carrier MIMO transmissions, various receivers have been developed with, e.g., time-domain equalization [24], [25], [27], [28], [30], or frequency-domain equalization [29]. A multi-channel decision feedback equalizer (DFE) has been presented in [25], while a single-channel DFE following a time reversal preprocessing module has been used in [24], [30]. In [27], iterative block decision-feedback equalizer (BDFE) was proposed with successive interference cancellation (SIC) in each iteration. In [29], a frequency-domain turbo equalizer combined with phase rotation and soft successive interference cancellation was proposed.

• For multicarrier transmissions in the form of orthogonal-frequency-division-multiplexing (OFDM), various MIMO receivers have been developed with, e.g., block-by-block processing [31], [32], [34], [35], or adaptive block-to-block processing [33]. The block-by-block receivers rely on embedded pilot symbols in each OFDM block for channel estimation [32], which can be further refined by using soft symbol estimates from the channel decoder [34], [35]. The adaptive receiver uses the channels estimated from the previous block for data detection of the current block, after proper phase compensation [33].

All these works on MIMO transmissions focus on the co-located case as shown in Fig. 1(a), where the parallel data streams are transmitted by co-located transducers and experience similar mobilities.

A recent publication by Tu et al [1] has considered a distributed two-user system, where the moving speeds of these users are significantly different. For example, the users can move on different directions, leading to Doppler scales with opposite signs. The major challenge for such a distributed system is that multiple dominant Doppler factors are present and it is hard for the receiver to compensate them through resampling. Reducing the Doppler scale for one user may make the Doppler scales for other users get larger. In [1], a front end with multiple resampling operations on each receive element was proposed.

The contributions of this paper are the following.

• We consider a system with multiple distributed users as depicted in Fig. 1(b), where different users could have different numbers of transmitters. We present two iterative receivers, termed as multiuser detection (MUD) based and single-user detection (SUD) based receivers, respectively. The MUD-based receiver adopts a frequency-domain-oversampling front end on each hydrophone, then performs joint channel estimation and multiuser data detection in an iterative fashion. The SUD-based receiver adopts conventional single-user processing modules, but
adds a critical step of multiuser interference (MUI) cancellation. Since different users have different resampling operations, judicious amplitude, delay and Doppler scale corrections need to be applied to the estimated multipath channels to reconstruct the MUI to other users.

- We present extensive performance results in various emulated settings, using experimental data sets from the Surface Processes and Acoustic Communications Experiment 2008 (SPACE08) and the Mobile Acoustic Communications Experiment 2010 (MACE10), where the transducer array in SPACE08 was stationary and that in MACE10 was slowly moving. We compare the MUD- and SUD-based receivers in a system with two single-transmitter users moving with different directions. We then present the performance of the SUD-based receiver in a two-user system with one slowly moving single-transmitter user and another stationary user having two or three data streams, as well as in a three-user system with two mobile single-transmitter users and one stationary user having two transmitters.

The problem considered in this paper is motivated by [1]. Compared to [1] and the substantial results presented in [36] and [37] using both simulation and emulated data sets, the work in this paper has the following distinctions: 1) the system setting is expanded where different users could have different numbers of data streams, 2) the receiver processing and the signalling format are different, and 3) the performance results include different emulated settings with up to four parallel data streams.

The rest of the paper is organized as follows. Section II describes the system model. Sections III and IV present the proposed MUD-based and SUD-based receivers, respectively. Section V contains performance results for a two-user system using MACE10 data sets, while Section VI contains performance results in various settings by using the MACE10 and SPACE08 data sets. Conclusions are contained in Section VII.

Notation: Bold upper case and lower case letters denote matrices and column vectors, respectively: $(\cdot)^T$, $(\cdot)^*$, and $(\cdot)^H$ denote transpose, conjugate, and Hermitian transpose, respectively. $I_N$ stands for an identity matrix with size $N$.

II. SYSTEM MODEL

We consider an underwater system with $U$ users, where the $i$th user has $N_i$ transmitters. The basic signalling format is zero-padded (ZP) OFDM, with $N_i$ parallel data streams transmitted from $N_i$ transmitters [3], [32]. Hence, the total number of data streams is $N_t = \sum_{i=1}^{U} N_i$. Let $T$ denote the OFDM symbol duration and $T_g$ the guard interval. The duration of the overall OFDM block is $T_{bl} = T + T_g$ and the subcarrier spacing is $1/T$. The $k$th subcarrier is at frequency

$$f_k = f_c + \frac{k}{T}, \quad k = -\frac{K}{2}, \ldots, \frac{K}{2} - 1,$$

where $f_c$ is the carrier frequency and $K$ subcarriers are used so that the bandwidth is $B = K/T$.

Within each OFDM block for the $i$th user, $N_i$ independent data streams are separately encoded with a nonbinary low-density parity-check (LDPC) code [6]. Let $s^{(i)}_{\mu}[k]$ denote the encoded information symbol, e.g., quadratic phase-shift-keying (QPSK) or quadratic amplitude modulation (QAM), to be transmitted on the $k$th subcarrier by the $\mu$th transducer of user $i$, where $i = 1, 2, \ldots, U$ and $\mu = 1, 2, \ldots, N_i$. Let $S_D$, $S_P$, and $S_N$ denote the nonoverlapping sets of data, pilot, and null subcarriers. The set of active subcarriers is then $S_A = S_P \cup S_D$ and the set of all subcarriers is denoted as $S_{all} = S_A \cup S_N = \{-K/2, \ldots, K/2-1\}$. All the data streams from different users have the same subcarrier allocation but have different pilot and data symbols. Such a signal design has been adopted in the SPACE08 and MACE10 experiments [34], [35].

The passband signal transmitted by the $\mu$th transducer of user $i$ is

$$\tilde{x}^{(i)}_{\mu}(t) = 2\text{Re} \left\{ \sum_{k \in S_A} s^{(i)}_{\mu}[k] e^{j2\pi f_k t} g(t) \right\} e^{j2\pi f_{\mu} t}, \quad t \in [0, T_{bl}]$$

where $g(t)$ is the pulse shaping filter whose Fourier transform is denoted by $G(f)$. In this paper, we use a rectangular pulse.
shaper where
\[
G(f) = \frac{\sin(\pi f T)}{\pi f T} e^{-j\pi f T}.
\] (3)

Consider an underwater acoustic (UWA) multipath channel which consists of \( P_{\nu,\mu}^{(i)} \) discrete paths between the \( \mu \)th transmitter of user \( i \) and the \( \nu \)th hydrophone [11]. The channel model can be approximated as
\[
h_{\nu,\mu}^{(i)}(\tau, t) = \sum_{p=1}^{P_{\nu,\mu}^{(i)}} A_{\nu,\mu,p}^{(i)} \delta \left( \tau - [\tau_{\nu,\mu,p}^{(i)} - a_{\nu,\mu,p}^{(i)} t] \right),
\] (4)
in which two assumptions are made [3], [11]: (i) the path amplitudes \( A_{\nu,\mu,p}^{(i)} \) are nearly constant within one OFDM block, and (ii) the path delays are approximated by a first-order polynomial with \( \tau_{\nu,\mu,p}^{(i)} \) and \( a_{\nu,\mu,p}^{(i)} \) representing the delay at the start of the OFDM block and the Doppler scale for the \( p \)th path, respectively.

Assume that the users are quasi-synchronous via some coordinations\(^1\), and the guard interval is larger than the maximum channel delay spread plus the slight asynchronism among users. As such, there is no interblock interference (IBI), and block-by-block processing as in [3], [11] can be applied. For one OFDM block, the passband signal at the \( \nu \)th hydrophone is
\[
\tilde{y}_\nu(t) = \sum_{i=1}^{U} \sum_{\mu=1}^{N_i} \sum_{p=1}^{P_{\nu,\mu}^{(i)}} A_{\nu,\mu,p}^{(i)} \hat{y}_{\nu,\mu,p}^{(i)} \left( 1 + a_{\nu,\mu,p}^{(i)} t - \tau_{\nu,\mu,p}^{(i)} \right) + \tilde{n}_\nu(t),
\] (5)
where \( \tilde{n}_\nu(t) \) is the additive noise.

III. MULTIUSER DETECTION (MUD) BASED ITERATIVE RECEIVER

Since different users have significantly different Doppler scales, the receiver designs for co-located MIMO OFDM as
\(^1\)Quasi-synchronism among distributed transmitters can be achieved via delay control if the distances between the transmitters and the receiver can be estimated. A preliminary result for multiuser OFDM reception without the quasi-synchronism assumption is available in [38].
in [32], [34], [35] can no longer work well. In this section, we develop an iterative receiver based on multiuser channel estimation and data detection, as shown in Fig. 2.

First, we adopt the frequency-domain oversampling approach in [18] as the receiver front end, which converts the received continuous-time signal into discrete samples in the frequency domain. With an integer frequency-domain oversampling factor \( \alpha \), a total of \( \alpha K \) frequency-domain samples on each hydrophone are obtained for one OFDM block. Define
\[
\bar{f}_{\bar{m}} = f_c + \frac{\bar{m}}{\alpha T}, \quad \bar{m} = \frac{\alpha K}{2}, \ldots, \frac{\alpha K}{2} - 1,
\] (6)
where \( \bar{m} \) is the index of the oversampled measurements. The measurement \( y_{\nu,\bar{m}} \) on the frequency \( \bar{f}_{\bar{m}} \) is related to \( \tilde{y}_\nu(t) \) as
\[
y_{\nu,\bar{m}} = \int_{0}^{T+T_g} \tilde{y}_\nu(t) e^{-j2\pi \bar{f}_{\bar{m}} t} dt,
\] (7)
which can be implemented by an \( \alpha K \)-point FFT operation after padding zeros to the sampled baseband signal.

Due to intercarrier interference (ICI), the measurement on the \( \bar{m} \)th frequency is potentially affected by all transmitted symbols \( s_{\mu,k}^{(i)} \), as,
\[
y_{\nu,\bar{m}} = \sum_{i=1}^{U} \sum_{\mu=1}^{N_i} \sum_{k \in S_{all}} \bar{H}_{\nu,\mu}^{(i)}[\bar{m}, k] s_{\mu,k}^{(i)} + n_{\nu,\bar{m}},
\] (8)
where \( \bar{H}_{\nu,\mu}^{(i)}[\bar{m}, k] \) is the coefficient that specifies how the symbol transmitted on the \( k \)th subcarrier of the \( \mu \)th transmitter from user \( i \) contributes to the output on the \( \bar{m} \)th subcarrier at the \( \nu \)th hydrophone. Following the derivation in [18], \( \bar{H}_{\nu,\mu}^{(i)}[\bar{m}, k] \) can be related to the channel parameters in (5) as
\[
\bar{H}_{\nu,\mu}^{(i)}[\bar{m}, k] = \sum_{p=1}^{P_{\nu,\mu}^{(i)}} \frac{A_{\nu,\mu,p}^{(i)}}{1 + a_{\nu,\mu,p}^{(i)}} e^{-j2\pi f_{\bar{m}}\bar{m}/\alpha T + a_{\nu,\mu,p}^{(i)} T_g} G \left( \frac{\bar{f}_{\bar{m}}}{1 + a_{\nu,\mu,p}^{(i)}} - f_k \right).
\] (9)

For each hydrophone, collect the \( \alpha K \) frequency-domain samples into a vector \( y_{\nu, \nu = 1, \ldots, N_r} \). For each transmitter
of user $i$, collect the $K$ transmitted symbols into a vector $s^{(i)}_\mu$. Now define
\[
y = \left[ y_1, y_2, \ldots, y_{N_r} \right]^T, \quad s = \left[ s^{(1)}_1, \ldots, s^{(1)}_{N_r}, \ldots, s^{(U)}_1, \ldots, s^{(U)}_{N_r} \right]^T.
\] (10)

Define the channel mixing matrix
\[
\hat{H} = \begin{bmatrix}
\hat{H}^{(1)}_{1,1} & \cdots & \hat{H}^{(1)}_{1,N_i} & \cdots & \hat{H}^{(U)}_{1,1} & \cdots & \hat{H}^{(U)}_{1,N_u} \\
\vdots & \ddots & \vdots & \ddots & \vdots & \ddots & \vdots \\
\hat{H}^{(1)}_{N_r,1} & \cdots & \hat{H}^{(1)}_{N_r,N_i} & \cdots & \hat{H}^{(U)}_{N_r,1} & \cdots & \hat{H}^{(U)}_{N_r,N_u}
\end{bmatrix}
\] (11)

where the submatrix $\hat{H}^{(i)}_{\nu,\mu}$ of size $\alpha K \times K$ has an $(\hat{m}, k)$th entry as shown in (9). The matrix-vector channel input-output relationship is then
\[
y = \hat{H}s + n
\] (12)

where $n$ is the additive noise similarly defined as $y$.

Two remarks are in order.

- When $\alpha = 1$, overlap-add operation is performed following the FFT operation to obtain the frequency-domain samples, which incurs information loss, as only $K$ frequency-domain samples are retained on each receive element while there are $K = T_{s,\nu}K/T > K$ time-domain samples available [18]. The information loss incurred by the overlap-add operation can be avoided with $\alpha > 1$ [18].

- The matrix-vector formulation in (12) looks similar as the one for the co-located MIMO OFDM [35]. The key difference is that in co-located MIMO OFDM, the submatrix $\hat{H}^{(i)}_{\nu,\mu}$ can be assumed to be banded after a resampling operation, which limits the ICI to affect only near neighbors. This is not true in the considered system because different users have significantly different Doppler scales. There is no resampling operation performed in the MUD-based receiver, and the submatrices $\hat{H}^{(i)}_{\nu,\mu}$ have different ICI patterns for different users.

\section{Multiuser channel estimation}

We adopt the sparse channel estimator as described in [35] for multiuser channel estimation, with the following modifications.

- The Doppler search ranges for different data streams are different. For each user, the dominant Doppler scale due to platform motion can be estimated, e.g., by preamble preceding data transmission or training sequences embedded in the transmission.

- On the first iteration of the iterative receiver, the data symbols are unknown, and the measurements on pilot subcarriers are severely contaminated by ICI. For the $\mu$th data stream of user $i$, we generate the frequency-domain observation template as
\[
\hat{\phi}^{(i)}_{\nu,\mu}[\hat{m}] = \sum_{k \in S_{\nu}} G \left( \frac{\hat{f}_{\hat{m}}}{1 + \hat{a}^{(i)}_{\nu,\mu}} - f_k \right) s^{(i)}_\mu[k],
\] (13)

where $\hat{a}^{(i)}_{\nu,\mu}$ is the estimated mean Doppler scale for the $\mu$th data stream of user $i$ (Note that the Doppler scales can be quite different for different data streams of the same user, which is not applicable to the SUD-based receiver to be developed in Section IV). Only those measurements with $\hat{\phi}^{(i)}_{\nu,\mu}[\hat{m}]$ larger than a given threshold for some user $i$ are used for channel estimation, and other measurements are excluded. See more discussions on how to choose different measurements in the presence of ICI due to unknown data symbols in [39]. In later iterations, the observations on all subcarriers can be utilized for channel estimation as tentative decisions on all information symbols are available.

\section{Multiuser data detection and LDPC decoding}

After obtaining the estimated channel mixing matrix $\hat{H}$, joint MIMO detection with a priori information fed back from the nonbinary LDPC decoding [6] can be applied. Based on the channel input-output model in (12), we use here the linear minimum mean square error (MMSE) equalizer from [40], where the extrinsic information from the channel decoders is used as the a priori information to the equalizer. Note that the size of the channel mixing matrix $H$ is $(\alpha K N_r \times K N_r)$, so inverting a $(K N_r \times K N_r)$ matrix is involved. The outputs of the MMSE detector are fed into $N_i$ separate LDPC decoders [6].

\section{Single User Detection (SUD) Based Iterative Receiver}

In this section, we propose a single-user-detection (SUD) based iterative receiver which has much lower complexity than the MUD-based receiver. Here we assume that the data streams from the same user will experience the same dominant Doppler scale, and will adopt all the processing modules developed for co-located MIMO OFDM [34], [35]. A MUI cancellation module is added to deal with the co-channel interference among multiple users. The receiver diagram is depicted in Fig. 3.

Let $\hat{a}^{(i)}_{\nu}$ denote the dominant Doppler scale for all the received $N_i$ data streams for user $i$. The receiver for user $i$ will apply the resampling operation on the $\nu$th hydrophone and then perform OFDM demodulation to generate frequency-domain measurements
\[
z^{(i)}_{\nu}[m] = \int_0^{T + T_0} y_{\nu}(t) \frac{t}{1 + \hat{a}^{(i)}_{\nu}} e^{-j2\pi f_m t} dt, \quad m = \frac{K}{2}, \ldots, \frac{K}{2} - 1
\] (14)

where the superscript $(i)$ stresses that the output is associated with user $i$. We can represent the measurement on the $m$th subcarrier as
\[
z^{(i)}_{\nu}[m] = \sum_{\mu=1}^{N_i} \sum_{k \in S_{\nu}} H^{(i)}_{\nu,\mu}[m,k] s^{(i)}_\mu[k] + \sum_{l=1, l \neq i} U \chi^{(\nu \rightarrow i)}_{\nu}[m] + n^{(i)}_{\nu}[m]
\] (15)

where $\chi^{(\nu \rightarrow i)}_{\nu}[m]$ is the interference from user $l$ to user $i$ on the $m$th subcarrier, and the channel coefficient can be expressed
Suppose that an estimate of $\chi_{\nu,\mu}$ in which estimates from the previous iteration are used for MUI reconstruction and cancellation, and the soft information from the channel decoder is used to improve UN

Fig. 3. SUD-based iterative receiver with MUI cancellation. There are $UN_r$ not $N_rN_v$, resampling operations. In each iteration, the channel and symbol estimates from the previous iteration are used for MUI reconstruction and cancellation, and the soft information from the channel decoder is used to improve channel estimation and data detection.

as

$$H_{\nu,\mu}[m,k] = \sum_{p=1}^{P_{\nu,\mu}} z_{\nu,\mu,p} e^{-j2\pi \frac{\nu,\mu}{2} G \left( \frac{f_m}{1 + \tilde{b}_{\nu,\mu,p}} - f_k \right)}$$

(16)

$$b_{\nu,\mu,p} = \frac{\tilde{a}_{\nu,\mu,p} - \tilde{a}_{\nu}}{1 + \tilde{a}_{\nu}}$$

(17)

where $b_{\nu,\mu,p}$ is the baseband complex gain for $p$th path.

Suppose that an estimate of $\chi_{\nu}^{(l\rightarrow i)}[m]$ is available, the receiver will obtain

$$\tilde{z}_{\nu}^{(l)}[m] = z_{\nu}^{(l)}[m] - \sum_{l=1,\nu\neq l}^{U} \tilde{\chi}_{\nu}^{(l\rightarrow i)}[m]$$

(19)

$$\tilde{z}_{\nu}^{(l)}[m] = \sum_{\nu=1}^{N_v} \sum_{\mu=1}^{N_r} H_{\nu,\mu}[m,k] z_{\nu,\mu}^{(l)}[k] + w_{\nu}^{(l)}[m]$$

(20)

where $w_{\nu}^{(l)}[m]$ is the equivalent noise containing both the ambient noise and the residual interference.

Since the front end for user $i$ removes the dominant Doppler effect on the multipath channels for user $i$, the "limited leakage" assumption holds that

$$H_{\nu,\mu}[m,k] = 0, \quad |m - k| > D, \quad \forall \nu, \mu$$

(21)

where the parameter $D$ specifies the ICI depth. Hence, single user channel estimation and data detection for user $i$ as in [34], [35] can be directly applied based on $\{\tilde{z}_{\nu}^{(l)}[m]\}_{m=-k/2}$ from all $N_r$ hydrophones.

Remarks:

- The residual Doppler compensation as described in [3] is not included. The MUI cancellation is directly applied on the frequency-domain samples.
- If only a single resampling operation is adopted, the SUD-based iterative receiver would not work well with a small ICI depth $D$. In this case, the MUD-based receiver with either $\alpha = 1$ or $\alpha > 1$ is applicable, which however has to use a large $D$, leading to high complexity.

A. MUI construction

The key issue is how to reconstruct the MUI knowing that different users have carried out channel estimation and data detection based on the measurements from different front ends. Let $\tilde{a}_{\nu}^{(l)}$ and $\tilde{b}_{\nu}^{(l)}$ denote the resampling factors used in the front ends of user $i$ and $l$, respectively.

For the $l$th data stream from user $l$, assume that $\tilde{z}_{\nu,\mu}^{(l)}$, $\tilde{\chi}_{\nu,\mu}^{(l)}$, $\hat{\tau}_{\nu,\mu}^{(l)}$, $\tilde{\gamma}_{\nu,\mu}^{(l)}$, $\tilde{\gamma}_{\nu,\mu}^{(l)}$, and $\hat{\gamma}_{\nu,\mu}^{(l)}$ have been estimated in the previous iteration. Then, we can construct a virtual signal as

$$\tilde{z}_{\nu,\mu}^{(l)}(t) = \sum_{p} (1 + \tilde{b}_{\nu,\mu,p}^{(l)}) \tilde{\gamma}_{\nu,\mu,p}^{(l)} e^{j2\pi f_c \hat{\gamma}_{\nu,\mu,p}^{(l)}}$$

(22)

whose Fourier transform satisfies

$$\tilde{Z}_{\nu,\mu}^{(l)}(f) = \sum_{p} \tilde{\gamma}_{\nu,\mu,p}^{(l)} e^{-j2\pi \hat{\gamma}_{\nu,\mu,p}^{(l)}}$$

$$\times \sum_{k} G \left( \frac{f_c}{1 + \tilde{b}_{\nu,\mu,p}^{(l)}} - f_k \right) \tilde{\gamma}_{\nu,\mu,p}^{(l)}$$

(23)

which is compatible with the channel and symbol estimates. With $N_l$ data streams, we construct:

$$\tilde{z}_{\nu,\mu}^{(l)}(t) = \sum_{\mu=1}^{N_l} \tilde{z}_{\nu,\mu}^{(l)}(t).$$

(24)
Fig. 4. The estimated moving speed in tow 1 from MACE10.

One can view \( \hat{\chi}_\nu^{(l-i)}(t) \) as the reconstructed signal after the \( l \)th user’s front-end processing, and hence the corresponding version before the resampling operation is \( \hat{y}_\nu^{(l-i)}(t) = \hat{\chi}_\nu^{(l-i)}((1 + a_\nu^{(l-i)})t) \). Letting \( \hat{y}_\nu^{(l-i)}(t) \) pass through the \( i \)th user’s front end, the MUI from the user \( l \) to user \( i \) can be expressed as:

\[
\hat{\chi}_\nu^{(l-i)}[m] = \int_0^{T^+_\nu} \hat{z}_\nu^{(l-i)} \left( \frac{1 + \hat{a}_\nu^{(l-i)}}{1 + a_\nu^{(l-i)}} \right) e^{-j2\pi f_m t} dt
\]

(25)

After straightforward manipulation, we obtain

\[
\hat{\chi}_\nu^{(l-i)}[m] = \sum_{p=1}^{N_1} \sum_{k \in S_{\text{all}}} \hat{H}_{\nu,\mu}^{(l-i)}[m, k] \hat{s}_\mu^{(l-i)}[k]
\]

(26)

where \( \hat{H}_{\nu,\mu}^{(l-i)}[m, k] \) can be computed as

\[
\hat{H}_{\nu,\mu}^{(l-i)}[m, k] = \sum_{p=1}^{P_{\mu,\nu}} \hat{\xi}_{\nu,\mu,\nu,\mu}^{(l-i)}(t) e^{-j2\pi \frac{m}{m_{\nu,\mu,\nu,\mu}} G} \left( \frac{f_m}{1 + b_{\nu,\mu,\nu,\mu}} - f_k \right)
\]

(27)

with

\[
1 + \hat{b}_{\nu,\mu,\nu,\mu}^{(l-i)}(t) = \frac{1 + \hat{a}_\nu^{(l-i)}}{1 + a_\nu^{(l-i)}}(1 + \hat{a}_\nu^{(l-i)}) \hat{s}_{\nu,\mu}^{(l-i)}(t), \quad \hat{\xi}_{\nu,\mu,\nu,\mu}^{(l-i)}(t) = \frac{1 + \hat{a}_\nu^{(l-i)}}{1 + a_\nu^{(l-i)}} \hat{r}_{\nu,\mu,\nu,\mu}^{(l-i)}
\]

(28)

\[
\hat{\xi}_{\nu,\mu,\nu,\mu}^{(l-i)} = \frac{1 + \hat{a}_\nu^{(l-i)}}{1 + a_\nu^{(l-i)}} \hat{\xi}_{\nu,\mu,\nu,\mu}^{(l-i)} e^{-j2\pi f_c (\hat{\chi}_{\nu,\mu}^{(l-i)} - \hat{\chi}_{\nu,\mu}^{(l-i)})}
\]

(29)

Hence, the amplitudes, delays, and Doppler scales need to be properly modified when reconstructing the MUI from one user to another user. This procedure is unique to the distributed MIMO-OFDM system considered in this paper and has not been discussed in the literature.

V. AN EMULATED TWO-USER SYSTEM USING MACE10

DATA

The MACE10 experiment was conducted off the coast of Martha’s Vineyard, MA, July 23, 2010. The bandwidth was \( B = 4.88 \) kHz. With \( K = 1024 \) subcarriers, the OFDM time duration is \( T = K / B = 209.8 \) ms. The carrier frequency was \( f_c = 13 \) kHz and the guard interval was \( T_g = 40.3 \) ms. Out of 1024 subcarriers, there were 256 pilot subcarriers, 96 null subcarriers, and 672 data subcarriers. Nonbinary LDPC codes with rate 1/2 were used. For each data stream, the spectral efficiency is \( \frac{1}{2} \cdot \log_2 M \cdot \frac{672}{1024 \cdot T} - T_g \) m/s, where \( M \) is the constellation size.

The signal was transmitted from a depth of about 80 meters and received by a 12-element array, which was moored. The transmitter was towed from the minimum range (about 500 m) out to the maximum range (about 4500 m) and then back to the minimum range. The relative speed between the transmitter and the receiver was estimated as \( \hat{v} = \frac{\hat{a}}{c} \), using a nominal sound speed of \( c = 1500 \) m/s, where \( \hat{a} \) is the Doppler scale estimated by minimizing the average energy on the null subcarriers with two single-transmitter users from MACE10. The relative speed was about 1.2 m/s when the transmitter was moving away from the receiver array, and about 1 m/s when it was towed back. We use the single-transmitter signals in MACE10 to emulate a distributed system with two users. Note however that the Doppler scales are pre-estimated as in the single user case, which is reasonable with e.g., time-division multiplexed preambles. See [41] for Doppler scale estimation in a distributed MIMO-OFDM system.

Tow 1 has \( n = 31 \) transmissions, with 20 OFDM blocks in each transmission. To emulate a distributed MIMO-OFDM system with different mobilities, we add the received passband signals of the 1st and the \( n \)th transmissions, the 2nd and \( (n-1) \)th transmissions together, and so on, where the hydrophones are numbered as 1, 2, . . . , 12 from the top to the bottom of the array and the received signals from the same hydrophone index are added together. A total of 15 emulated data sets for a two-user system are obtained while the 16th transmission is excluded. Note that the OFDM blocks in one transmission are reversed in order so that the overlapping blocks have different
pilot and data symbols. From Fig. 4, the absolute value of projected velocity was around 1 m/s, which will lead to a frequency shift as about $\pm 7 \sim \pm 10$ Hz while the subcarrier spacing in MACE10 was 4.8 Hz. We plot a measured channel response in Fig. 5. We see that the channel energy for two users are well separated in the Doppler plane. The Doppler spreads are around 0.1 m/s and the delay spread is around 10 ms.

We use the block-error rate (BLER) as the performance metric, which is the ratio of the number of OFDM blocks decoded in error to the total number of OFDM blocks transmitted from all the users.

A. MUD-based receiver with and without frequency-domain oversampling

Fig. 6 shows the performance results by using the MUD-based receiver for both the conventional sampling ($\alpha = 1$) and the frequency-domain oversampling with $\alpha = 2$. The ICI is not limited to only near neighbours, and we assume $H_{\nu,\mu}(m, \nu, \mu) \neq 0, \forall |\bar{f}_m - f_k| \leq \frac{\Delta f}{f_T}$ for all the channel matrices $H_{\nu,\mu}$ in (11). The frequency-domain oversampling method outperforms the conventional sampling uniformly at early iterations. As the number of iterations gets large, the performance gap decreases to a negligible level. In the following, we use the conventional sampling with $\alpha = 1$ for the MUD-based receiver.

B. SUD-based versus MUD-based receivers

Fig. 7 shows the performance for distributed MIMO-OFDM systems with two users, where 8 iterations are used. In the SUD-based receiver, we use $D = 0$ in (21) to achieve low-complexity processing, which ignores the ICI from the same user. When constructing the MUI $\tilde{x}_{l\rightarrow i}(m)$ in (26), the contributions from all the transmitted symbols of user $l$ on each measurement are considered. We can see that both the SUD-based and MUD-based receivers work very well for QPSK, 8-QAM and 16-QAM. The following observations are in order$^2$.

- At the first several iterations, the SUD-based receiver is much worse than the MUD-based receiver. This is due to the severe residual MUI at early iterations.
- With continuing iterations, the performance of the SUD-based receiver catches up that of the MUD-based receiver as more and more MUI is cancelled out. For the 8-QAM in this data set, the SUD-based receiver even slightly outperforms the MUD-based counterpart, while on the other hand the MUD-based receiver work slightly better than the SUD-based counterparts for QPSK and 16-QAM.

$^2$Note that in Fig. 7(a), two BLER curves increase when more phones are combined. This strange behavior happens when working with real data sets. See e.g., [39], for an investigation on this issue, where some phones were found to have larger local noise than others.
Fig. 8 shows an example of the estimated channel impulse responses (CIRs) for the distributed two-user system. It is obvious from Fig. 8 that as the iteration goes on, the CIRs look like MUI-free ones, which verifies the effectiveness of the MUI cancellation in the SUD-based receiver.

The constellation scattering plots in Fig. 9 show the soft-decision symbols at the output of the MMSE detector for users 1 and 2. The improvement over iterations is clearly observed by comparing subfigures in each row in Fig. 9. Also, from Figs. 8 and 9, we can see that during the start up stage, the CIRs and scatter plots for both users look very noisy due to the severe MUI, as expected.

VI. EMULATED MIMO OFDM WITH MACE10 AND SPACE08 DATA

In this section, we use MACE10 and SPACE08 data to emulate various distributed multiuser settings. In particular, we use some SPACE08 data sets that have the same basic ZP-OFDM parameter setup as MACE10, except two differences: the carrier frequency was $f_c = 11$ kHz and the guard interval was $T_g = 25$ ms. The signals were transmitted by a mounted array which was approximately 4 meters from the sea floor, and the top of the receive arrays were about 3.25 meters above the sea floor. There were three receivers labeled as S1, S3, and S5, which were 60 m, 200 m, and 1,000 m from the transmitter. In this paper, we use the signals from the receiver S3, Julian date 296 - 297, to emulate the distributed multiuser
settings. There were also 12 hydrophones on the receivers in SPACE08, numbered as 1, 2, …, 12 from the top to the bottom of the array, and the signals on the same hydrophone index are added together when mixing the SPACE08 and MACE10 data. Note that the user from SPACE08 data has multiple co-located transducers, as specified later. Also, as reported in, e.g., [16], [22], the channels in SPACE08 are more challenging than those in MACE10.

A. One mobile single-transmitter user plus one stationary two-transmitter user

In this setting, one user is from MACE10 with transmissions 1 to 15 (a negative Doppler scale), as shown in Fig. 4. The second user owns the data from SPACE08, with two transmitters. This distributed MIMO OFDM hence has two users and three data streams. As the carrier frequencies and the guard intervals used in MACE10 and SPACE08 were different, we add the baseband signals together on the OFDM block level. Fig. 10 shows the decoding results for this setting with the SUD-based receiver, using QPSK constellation. Due to the high computation complexity, the MUD-based receiver performance is not reported.

From Fig. 10, we can see that the performance at the first iteration is pretty bad due to the severe MUI. The performance improvement from the first iteration to second iteration is impressive, which is similar with the setting in Fig. 7. Good performance is achieved after four to five iterations.

B. One mobile single-transmitter user plus one stationary three-transmitter user

In this setting, one user is from MACE10 with transmissions 1 to 15, and the other user is using SPACE08 data with three transmitters. The resulting distributed MIMO OFDM has two users and four data streams. We can see that the system still work well with the SUD-based receiver. Due to the larger number of data streams, the improvement from first iteration to the second iteration is shrunk compared with the settings in Fig. 7 and Fig. 10. However, satisfactory performance can still be achieved and the system performance saturates after five iterations.

C. Two mobile single-transmitter users plus one stationary two-transmitter user

In this setting, two users are from MACE10 as described in Section V. A third user has data from SPACE08 with two transmitters. Hence, the distributed MIMO OFDM has three users and four data streams. Fig. 12 depicts the overall coded BLER with the SUD-based receiver. With 12 hydrophones, the BLER after the first iteration is around 0.8, but reduces to below $10^{-2}$ after 6 iterations.

Fig. 10. Distributed multiuser setting with one mobile single-transmitter user from MACE10 (0.55 bits/s/Hz) plus one stationary two-transmitter user from SPACE08 (1.10 bits/s/Hz). SUD-based receiver, QPSK.

Fig. 11. Distributed MIMO OFDM with one mobile single-transmitter user from MACE10 (0.55 bits/s/Hz) plus one stationary three-transmitter user from SPACE08 (1.65 bits/s/Hz). SUD-based receiver, QPSK.

Fig. 12. Distributed MIMO OFDM with two mobile single-transmitter users (0.55 bits/s/Hz per user) from MACE10 plus one stationary two-transmitter user from SPACE08 (1.10 bits/s/Hz). SUD-based receiver, QPSK.
VII. CONCLUSIONS

In this paper, we reported performance results of two iterative receivers for distributed MIMO-OFDM systems in various settings, where different users have significantly different Doppler scales and different numbers of data streams. The MUD-based receiver outperforms the SUD-based counterparts at the first several iterations, but the performance gap decreases quickly as the iteration goes on, hence the SUD-based receiver has a better performance-complexity tradeoff. It has been shown that the SUD-based receiver can handle various distributed MIMO-OFDM settings with satisfactory performance.

REFERENCES


Jianzhong Huang (S’09) received his B.S. degree in 2003 and his M.Sc. degree in 2006, from Xidian University, Xian, Shaanxi China, both in communication engineering. He received his Ph.D. degree in electrical engineering from the University of Connecticut, Storrs, in 2012. His research interests lie in the areas of communications and signal processing, currently focusing on multicarrier modulation algorithms and channel coding for underwater acoustic communications.

Shengli Zhou (SM’11) received the B.S. degree in 1995 and the M.Sc. degree in 1998, from the University of Science and Technology of China (USTC), Hefei, both in electrical engineering and information science. He received his Ph.D. degree in electrical engineering from the University of Minnesota (UMN), Minneapolis, in 2002. He has been an assistant professor with the Department of Electrical and Computer Engineering at the University of Connecticut (UCONN), Storrs, 2003-2009, and now is an associate professor. He holds a United Technologies Corporation (UTC) Professorship in Engineering Innovation, 2008-2011. His general research interests lie in the areas of wireless communications and signal processing. His recent focus is on underwater acoustic communications and networking.


Zhaohui Wang (S’10) received the B.S. degree in 2006, from the Beijing University of Chemical Technology (BUCT), and the M.Sc. degree in 2009, from the Institute of Acoustics, Chinese Academy of Sciences (IACAS), Beijing, China, both in electrical engineering. She is currently working toward the Ph.D degree in the Department of Electrical and Computer Engineering at the University of Connecticut (UCONN), Storrs, USA. Her research interests lie in the areas of communications, signal processing and detection, with recent focus on multicarrier modulation algorithms and signal processing for underwater acoustic communications.