A Simple Design for Joint Channel Estimation and Data Detection in an Alamouti OFDM System

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***Abstract*— To obtain full benefits of the Alamouti space-time code design on a time-varying acoustic channel, the channel variation is decoupled into two parts: the gain, which is assumed to be constant over an Alamouti pair (two consecutive OFDM blocks), and the phase, which is not restricted in this manner, but allowed to vary from one block to another in a piecewise linear fashion. Under this model, the receiver assumes a simplified structure in which conventional Alamouti data detection and channel estimation are aided by Doppler tracking. System performance is demonstrated using experimental data transmitted over a 1 km shallow water channel in the 8-18 kHz acoustic band, using QPSK modulation and a varying number of carriers (up to 1024). The proposed design shows a gain of up to 2 dB as compared to the reference single-transmitter case.**

I. Introduction

Multi-input multi-output (MIMO) communication is generally considered for one of two purposes: spatial multiplexing or transmit diversity. In the first case, independent data streams are sent simultaneously from multiple transmitters, with the goal of increasing the bit rate over a given channel. This type of MIMO communication is particularly attractive for band-limited acoustic channels, and has been investigated extensively over the past years, notably for multi-carrier systems [1], [2], [3]. In the second case, a single data stream is encoded into multiple (dependent) streams, which are then sent from multiple transmitters. The bit rate now remains the same as if a single transmitter were used (SIMO system) but the performance can be improved through spatial diversity. This type of MIMO communication has also been investigated for underwater acoustic channels, notably for single-carrier systems [9], [10]. Multi-carrier modulation and transmit diversity were considered in Ref. [4], which motivates the present work.

In this paper, we focus on the Alamouti space-time block coding (STBC) coupled with orthogonal frequency division multiplexing (OFDM). OFDM offers a particularly suitable platform for exploiting the MIMO gains on time-dispersive channels, because it relieves the system of the need for complex time-domain equalization. Simple MIMO processing techniques can thus be applied in each narrow subband, while the signal processing efforts are focused on channel estimation.

Adaptive channel estimation for STBC OFDM systems has been extensively studied for terrestrial radio communications, e.g. [5]- [8]. For underwater acoustic channels, Ref. [4] reports on an Alamouti STBC system and the application of a MIMO OFDM channel estimation algorithm enhanced by Doppler shift compensation [1]. In this framework, channel/phase estimation is ignorant of the Alamouti structure, i.e. it treats the transmitted data streams as independent. Such treatment allows the decision-directed estimator to be updated *every* block interval, instead of every *other* block interval, as it would do had it waited for the symbol decisions from the Alamouti decoder. While this approach showed very good performance with experimental data, its computational complexity is higher than that of a conventional Alamouti system where the channel is treated as invariant over *two* consecutive blocks.

Here, we address the possibility of modeling the acoustic channel in a manner that would permit conventional Alamouti detection. This approach will necessarily trade off the performance for implementation simplicity, but the question nonetheless remains as to whether it is possible at all to do so for a time-varying acoustic channel.

To address this question, we propose a receiver design that capitalizes on decoupling the effects of time-variation into those that pertain to the channel gain and those that pertain to the phase. The first type of variation is inherent to the propagation medium, while the second type may be caused by motion. We conjecture that the phase variation is responsible for faster changes, while the inherent channel variation can be assumed fixed over two consecutive OFDM blocks. Further, we assume that the phase variation is such that each transmitter experiences the same Doppler shifting effects. Such an assumption is justified for a system with co-located and closely spaced transmit elements. Under these conditions, we develop a low-complexity receiver algorithm, in which the conventional channel estimator is aided by a phase tracking/prediction algorithm. The latter is based on a model of motion-induced Doppler shifting (non-uniform across the signal bandwidth) and targets estimation of the underlying Doppler scaling factor.

The paper is organized as follows. In Sec.II we outline the system model and discuss the assumptions that lead to the decoupling of the channel gain and phase. The receiver algorithm is developed in Sec.III. In Sec.IV, we present the results of real data processing, using the signals collected during the Surface Processes / Acoustic Communications Experiment (SPACE'08) that took place off the coast of New England in October 2008. We conclude in Sec.V and identify the goals for future work.

II. System Model

We consider a MIMO system with M_T transmitters and M_R receivers. In the Alamouti STBC OFDM scheme, $M_T = 2$,

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and two adjacent OFDM blocks, $2n$ and $2n+1$, are used to transmit the following PSK data symbols on the k-th carrier: $d_k(2n)$, $d_k(2n+1)$ from the first transmitter, and $-d_k^*(2n+1)$, $d_k^*(2n)$ from the second transmitter. We will refer to the blocks $(2n, 2n+1)$ as the n-th pair of blocks.

Let us denote by $A_k^r(2n)$ and $B_k^r(2n)$ the transfer function of the channel observed on the k-th carrier during transmission of block $2n$ from the first and the second transmitter to the r-th receiver, and by $\alpha_k(2n)$ and $\beta_k(2n)$ the corresponding phase distortions caused by the relative motion between the transmitter array and the receiver.[1] The signals received on the k-th carrier during two consecutive OFDM blocks can then be represented as

$$\begin{aligned} y_k^r(2n) &= A_k^r(2n)e^{j\alpha_k(2n)}d_k(2n) - \\ &\quad B_k^r(2n)e^{j\beta_k(2n)}d_k^*(2n+1) + z_k^r(2n) \\ y_k^r(2n+1) &= A_k^r(2n+1)e^{j\alpha_k(2n+1)}d_k(2n+1) + \\ &\quad B_k^r(2n+1)e^{j\beta_k(2n+1)}d_k^*(2n) + \\ &\quad z_k^r(2n+1) \end{aligned} \tag{1}$$

where $z_k^r(2n)$ and $z_k^r(2n+1)$ are the noise components.

To represent the above expressions in a compact form, let us define the vectors

$$\mathbf{y}_k^r[n] = \begin{bmatrix} y_k(2n) \\ y_k(2n+1) \end{bmatrix}, \mathbf{d}_k[n] = \begin{bmatrix} d_k(2n) \\ d_k(2n+1) \end{bmatrix} \tag{2}$$

and their Alamouti arrangements

$$\mathbf{y}_k^{rA}[n] = \begin{bmatrix} y_k^r(2n) \\ -y_k^{r*}(2n+1) \end{bmatrix}, \mathbf{d}_k^A[n] = \begin{bmatrix} d_k(2n) \\ -d_k^*(2n+1) \end{bmatrix} \tag{3}$$

Making similar arrangements with the noise components, we have that

$$\mathbf{y}_k^{rA}[n] = \mathbf{C}_k^r[n]\mathbf{d}_k^A[n] + \mathbf{z}_k^A[n] \tag{4}$$

where $\mathbf{C}_k^r[n] =$

$$\begin{bmatrix} A_k^r(2n)e^{j\alpha_k(2n)} & B_k^r(2n)e^{j\beta_k(2n)} \\ -B_k^{r*}(2n+1)e^{-j\beta_k(2n+1)} & A_k^{r*}(2n+1)e^{-j\alpha_k(2n+1)} \end{bmatrix} \tag{5}$$

We can also collect the signals of all the receivers into a vector

$$\mathbf{y}_k^A[n] = \begin{bmatrix} \mathbf{y}_k^{1A}[n] \\ \vdots \\ \mathbf{y}_k^{M_R A}[n] \end{bmatrix} \tag{6}$$

so that

$$\mathbf{y}_k^A[n] = \mathbf{C}_k[n]\mathbf{d}_k^A[n] + \mathbf{z}_k^A[n] \tag{7}$$

where

$$\mathbf{C}_k[n] = \begin{bmatrix} \mathbf{C}_k^1[n] \\ \vdots \\ \mathbf{C}_k^{M_R}[n] \end{bmatrix} \tag{8}$$

and the noise vector $\mathbf{z}_k^A[n]$ is arranged accordingly.

Expression (7) is useful for data detection, as it directly implies the least squares (LS) estimate

$$\hat{\mathbf{d}}_k^A[n] = [\mathbf{C}_k^H[n]\mathbf{C}_k[n]]^{-1}\mathbf{C}_k^H[n]\mathbf{y}_k^A[n] \tag{9}$$

(H denotes conjugate transpose).

When the channel is time invariant, or slowly time-varying so that we can assume that

$$A_k^r(2n) = A_k^r(2n+1), B_k^r(2n) = B_k^r(2n+1) \tag{10}$$

$$\alpha_k(2n) = \alpha_k(2n+1), \beta_k(2n) = \beta_k(2n+1) \tag{11}$$

we have that

$$\mathbf{C}_k^H[n]\mathbf{C}_k[n] = E_k(n)\mathbf{I}_2 \tag{12}$$

where $E_k(n) = \sum_{r=1}^{M_R} |A_k^r(2n)|^2 + |B_k^r(2n)|^2$, and $\mathbf{I}_2$ is the 2×2 identity matrix. The expression (9) then reduces to

$$\hat{\mathbf{d}}_k^A[n] = \frac{1}{E_k(n)}\mathbf{C}_k^H[n]\mathbf{y}_k^A[n] \tag{13}$$

that is, conventional Alamouti detection is enabled.

In an acoustic channel, however, the conditions (10, 11) may not hold. Below, we discuss these conditions in light of the physics of acoustic propagation, and investigate the possibility to relax them in order to arrive at a data detection and channel estimation principle that is simpler than (9) and exploits the benefits of the Alamouti structure (13).

A. Channel / phase decoupling

In a practical system, the channel is not known and has to be estimated. The frequency at which the estimates need to be updated depends on the channel dynamics, which draw a fine line between the "slow" and "not-so-slow" variations.[2] Within the Alamouti OFDM framework, where *pairs* of blocks represent relevant signal observations, we will make a distinction between those quantities that need to be updated every block, and those that may be updated every other block, i.e. every pair of blocks.

An acoustic channel often exhibits a faster variation in the phase than in the gain. For this reason, we have kept the phases α and β separate from the gains A and B.[3] This fact implies a situation in which the assumption (10) may be valid, but the assumption (11) will not hold. The question then becomes: What happens with the condition (12)?

Substituting the approximation (10) into the matrix $\mathbf{C}_k^r[n]$, but keeping the phases $\alpha_k(2n) \neq \alpha_k(2n+1)$ and $\beta_k(2n) \neq \beta_k(2n+1)$, we find that the condition (12) will still hold if

$$\alpha_k(2n+1) - \alpha_k(2n) = \beta_k(2n+1) - \beta_k(2n) \tag{14}$$

Hence, we want to identify the physical conditions under which the above relationship will hold. We already know that it will hold if the phases are time-invariant, but this is not the assumption that we want to make. Another case in which it

[1]For co-located receive elements, it suffices to associate the same phase distortion with each element, as confirmed by SIMO experiments [1], [3].

[2]Note that a bandwidth-efficient OFDM system design calls for a large number of carriers, implying block durations that may approach the time coherence of the channel.

[3]We call these quantities gains, but note that they are complex-valued channel transfer functions evaluated at different carrier frequencies.

will hold is when the phases are equal, $\alpha_k(2n) = \beta_k(2n)$. This is a much more reasonable assumption, and one that will hold whenever the two transmit elements experience the same relative motion with respect to the receiver array. For example, this will be the case when the transmit elements are co-located and closely spaced on a moving platform (at high acoustic frequencies several wavelengths amount to a spacing on the order of 10 cm).

From our discussion, it is apparent that there are situations in which one can capitalize on the system and channel properties to arrive at a simplified receiver structure. Granted, these conditions may not always hold, in which case one must resort to a full-complexity receiver that estimates both the gains and the phases in every block, and uses the so-obtained estimates in the detector (9). Such a receiver was discussed in Ref. [4]. Here, we proceed to develop a reduced-complexity receiver that exploits the Alamouti structure based on the assumption that

$$\alpha_k(2n) = \beta_k(2n) = \gamma_k(2n) \tag{15}$$

but $\gamma_k(2n+1) \neq \gamma_k(2n)$.

When the conditions (10) and (15) hold, the matrix $\mathbf{C}_k^r[n]$ can be decomposed as

$$\mathbf{C}_k^r[n] = \mathbf{\Gamma}_k^A[n]\mathbf{G}_k^r[n] \tag{16}$$

where

$$\mathbf{\Gamma}_k^A[n] = \begin{bmatrix} e^{j\gamma_k(2n)} & 0 \\ 0 & e^{-j\gamma_k(2n+1)} \end{bmatrix} \tag{17}$$

and

$$\mathbf{G}_k^r[n] = \begin{bmatrix} A_k^r(2n) & B_k^r(2n) \\ -B_k^{r*}(2n) & A_k^{r*}(2n) \end{bmatrix} \tag{18}$$

The matrix $\mathbf{C}_k[n]$ now satisfies the condition (12) and this fact gives rise to the LS data estimate

$$\hat{\mathbf{d}}_k^A[n] = \frac{1}{E_k(n)} \sum_{r=1}^{M_R} \mathbf{G}_k^{rH}[n]\mathbf{\Gamma}_k^{AH}[n]\mathbf{y}_k^{rA}[n] \tag{19}$$

III. Receiver Algorithm

Because the channel and the phases are not known, their estimates are used instead for data detection. The receiver functions thus consist of channel estimation, phase estimation, and data detection.

A. Channel estimation

Channel estimation is based on an alternative compact representation of the received signals (1):

$$\mathbf{y}_k^r[n] = \mathbf{\Gamma}_k[n]\mathbf{D}_k[n]\mathbf{H}_k^r[n] + \mathbf{z}_k[n] \tag{20}$$

where

$$\mathbf{D}_k[n] = \begin{bmatrix} d_k(2n) & d_k(2n+1) \\ -d_k^*(2n+1) & d_k^*(2n) \end{bmatrix} \tag{21}$$

$$\mathbf{\Gamma}_k[n] = \begin{bmatrix} e^{j\gamma_k(2n)} & 0 \\ 0 & e^{j\gamma_k(2n+1)} \end{bmatrix} \tag{22}$$

and

$$\mathbf{H}_k^r[n] = \begin{bmatrix} A_k^r(2n) \\ B_k^r(2n) \end{bmatrix} \tag{23}$$

Note that channel gain / phase decoupling allows us to isolate the channel *vector* $\mathbf{H}_k[n]$ from the phase *matrix* $\mathbf{\Gamma}_k[n]$.[4] This decoupling will enable us to exploit the joint Alamouti channel estimation and data detection in its original form, with the aid of additional phase tracking.

Assuming that the phases are known, and that training data are available, the LS channel estimate can be obtained directly from the expression (20). Because $\mathbf{\Gamma}_k^H[n]\mathbf{\Gamma}_k[n] = \mathbf{I}_2$ and $\mathbf{D}_k^H[n]\mathbf{D}_k[n] = 2\mathbf{I}_2$ (we are assuming unit-amplitude PSK symbols), this reduces to

$$\hat{\mathbf{H}}_k^r[n] = \frac{1}{2}\mathbf{D}_k^H[n]\mathbf{\Gamma}_k^H[n]\mathbf{y}_k[n] \tag{24}$$

Using the phase estimates instead of the unknown values, and symbol decisions or pilots $\tilde{\mathbf{D}}_k[n]$ after training, channel estimates can be formed as

$$\hat{\mathbf{H}}_k^r[n] = \frac{1}{2}\tilde{\mathbf{D}}_k^H[n]\hat{\mathbf{\Gamma}}_k^H[n]\mathbf{y}_k[n] \tag{25}$$

When the channel gains are slowly varying, data detection in each new pair will be accomplished using the last pair's channel estimate. The so-obtained symbol decisions will then be used to estimate the channel. Instead of the single-shot estimate (25), additional smoothing can be performed in time, e.g. as

$$\hat{\mathbf{H}}_k^r[n] = \lambda\hat{\mathbf{H}}_k^r[n-1] + (1-\lambda)\frac{1}{2}\tilde{\mathbf{D}}_k^H[n]\hat{\mathbf{\Gamma}}_k^H[n]\mathbf{y}_k[n] \tag{26}$$

where $\lambda \in (0, 1)$. While this expression suffices to illustrate adaptation in time, it does not exploit the correlation in frequency. To do so, the channel estimation problem can be translated into the impulse response domain by taking the inverse DFT of (26) so that the impulse response coefficients $\hat{\mathbf{h}}_l^r[n] = \text{DFT}^{-1}\hat{\mathbf{H}}_k^r[n]$ are targeted instead of the transfer function coefficients $\hat{\mathbf{H}}_k^r[n]$. This procedure is commonly used for OFDM channel estimation, as it reduces the number of unknowns by exploiting the frequency correlation in an optimal manner. It also enables channel sparsing to reduce the estimation noise [1].

The channel estimates $\hat{\mathbf{H}}_k^r[n]$ are finally used to construct the gain matrix $\hat{\mathbf{G}}_k^r[n]$ according to (18), and this matrix is in turn used to estimate the data symbols according to (19).

B. Phase estimation

Phase tracking is accomplished indirectly, via estimation of the underlying Doppler scaling factor, similarly as in [1]. To this end, the phase distortion is modeled as

$$\gamma_k(2n+1) = \gamma_k(2n) + 2\pi a[n] f_k T' \tag{27}$$

where $a[n]$ is the Doppler factor, $f_k = f_0 + k\Delta f$ is the k-th carrier frequency, $\Delta f = 1/T$ is the carrier spacing, and $T' = T + T_g$ is the block duration that includes the multipath guard interval. Here, we allow the Doppler factor to vary from

[4]The matrix $\mathbf{\Gamma}_k[n]$ captures the block-to-block variations. Had the phase variation been negligible, it could have been eliminated and the factors $\gamma_k(2n) = \gamma_k(2n+1)$ could have been absorbed into the vector $\mathbf{H}_k[n]$, yielding the conventional Alamouti structure.

one block pair to another, but we assume that it is constant within each pair.

At the beginning of each new pair of blocks, the receiver uses phase predictions, $\hat{\hat{\gamma}}_k(2n)$ and $\hat{\hat{\gamma}}_k(2n+1)$,[5] to form tentative symbol estimates as

$$\hat{\hat{\mathbf{d}}}_k^A[n] = \frac{1}{\hat{E}_k[n-1]} \sum_{r=1}^{M_R} \hat{\mathbf{G}}_k^{rH}[n-1] \hat{\hat{\mathbf{\Gamma}}}_k^{AH}[n] \mathbf{y}_k^{rA}[n] \qquad (28)$$

These estimates are used to make tentative symbol decisions $\tilde{\tilde{\mathbf{d}}}_k^A[n]$ (either as closest points in the signal space if no coding is used, in which case scaling by the channel energy can be omitted, or by soft-decision decoding if a channel code is used).

Tentative decisions are used to measure the phase offset that would result had the estimates $\hat{\gamma}_k(2n-2)$ and $\hat{\gamma}_k(2n-1)$ from the last pair of blocks been used. To do so, we first form an "outdated" data estimate

$$\check{\mathbf{d}}_k^A[n] = \sum_{r=1}^{M_R} \hat{\mathbf{G}}_k^{rH}[n-1] \hat{\mathbf{\Gamma}}_k^{AH}[n-1] \mathbf{y}_k^{rA}[n] \qquad (29)$$

(scaling by the channel energy will not influence the phase measurement, so we may omit it). We now measure the phase offset as the angle

$$\psi_k[n] = \angle \tilde{\tilde{\mathbf{d}}}_k^{AH}[n] \check{\mathbf{d}}_k^A[n] \qquad (30)$$

In the absence of noise and estimation errors, this quantity will be equal to $2\pi f_k a[n] \cdot 2T'$. Hence, we make an estimate of the Doppler factor as an average taken over all the carriers (or the pilots):

$$\hat{a}[n] = \frac{1}{K} \sum_k \frac{\psi_k}{4\pi f_k T'} \qquad (31)$$

This estimate is now used to update the phase estimates based on the model (27). Equivalently, we obtain the matrix update

$$\hat{\mathbf{\Gamma}}_k[n] = \hat{\mathbf{\Gamma}}_k[n-1] \cdot e^{j4\pi f_k \hat{a}[n] T'} \qquad (32)$$

The updated phases are then used to obtain the final data estimates

$$\hat{\mathbf{d}}_k^A[n] = \frac{1}{\hat{E}_k[n-1]} \sum_{r=1}^{M_R} \hat{\mathbf{G}}_k^{rH}[n-1] \hat{\mathbf{\Gamma}}_k^{AH}[n] \mathbf{y}_k^{rA}[n] \qquad (33)$$

and the corresponding symbol decisions $\tilde{\mathbf{d}}_k^A[n]$ are made.

These symbol decisions are used to construct the matrix $\tilde{\mathbf{D}}_k[n]$ and to update the channel vector (26). Finally, phase predictions are made for the next pair of blocks:

$$\hat{\hat{\mathbf{\Gamma}}}_k[n+1] = \hat{\mathbf{\Gamma}}_k[n] \cdot e^{j4\pi f_k \hat{a}(n) T'} \qquad (34)$$

[5] We will see later how phase predictions are made.

C. Summary

In each Alamouti block pair n, the following steps are carried out:

- Using the existing channel estimates $\hat{\mathbf{H}}_k^r[n-1]$ and the predicted phases $\hat{\hat{\mathbf{\Gamma}}}_k[n]$, make tentative symbol decisions $\tilde{\tilde{\mathbf{d}}}_k^A[n]$ from (28) for all $k = 0, \ldots K-1$
- Using the existing channel estimates $\hat{\mathbf{H}}_k^r[n-1]$ and the existing phase estimates $\hat{\mathbf{\Gamma}}_k[n-1]$, make "outdated" symbol estimates (29)
- Measure the phase offsets (30)
- Estimate the Doppler factor (31)
- Update the phase matrix $\hat{\mathbf{\Gamma}}_k[n]$ according to (32)
- Make the final symbol decisions $\tilde{\mathbf{d}}_k^A[n]$ from (33)
- Form the updated channel estimates $\hat{\mathbf{H}}_k^r[n]$ using the impulse-response domain version of (26) and sparsing
- Make phase prediction (34) for the next pair of blocks.

IV. Experimental results

Alamouti STBC OFDM signals were transmitted and recorded as part of the Surface Processes / Acoustic Communications Experiment (SPACE'08) conducted by the Woods Hole Oceanographic Institution in October 2008 near Martha's Vineyard island off the shore of New England. The transmitter array (4 elements separated by 50 cm) was deployed at a depth of about 10 m in 15 m of water. The top and the bottom element of the transmitter array (150 cm separation) were used as an Alamouti pair. The receiver array (M_R=12 elements, separated by 12 cm) was deployed 1 km away, at a depth of about 11 m. QPSK modulation and BCH(63,10) coding were used in the 8 kHz-18 kHz acoustic band. Table I lists the signal parameters.

TABLE I
Signal parameters used in the experiment.

Number of subcarriers, K	128, 256, 512, 1024
Subcarrier spacing, Δf [Hz]	78, 39, 19, 10
OFDM block duration, T [ms]	13, 26, 52, 105
Symbols per frame, N_d	16384
Blocks per frame, N	128, 64, 32, 16
Guard interval T_g [ms]	16

Fig. 1 summarizes the results obtained with the signals recorded over the course of three days of the experiment (total of 19 transmissions between Oct. 14 and Oct. 16, 2008.). The system performance is measured by the MSE, taken as the average over time (blocks) and frequency (carriers) at the detector output. The two plots refer to the cases of $K = 512$ and $K = 1024$ carriers. The performance of the proposed detection algorithm, labeled Alamouti 2, is compared to the single-transmitter (SIMO) scheme using the same number of receive elements and the same transmission rate (15 kbps for $K = 512$; 17 kbps for $K = 1024$), as well as to that of Ref. [4], labeled Alamouti 1. Both the SIMO system and the system of Ref. [4], rely on block-by-block adaptation and thus have better channel tracking capabilities than the pair-by-pair adaptive algorithm. For this reason, the Alamouti 1 scheme outperforms the Alamouti 2 scheme. However, this improvement comes at a cost in computational complexity. Whereas the proposed algorithm efficiently performs joint data detection and channel estimation, fully exploiting the benefits of the Alamouti code, Alamouti 1 approach effectively uses two receivers in parallel–one that is ignorant of the Alamouti code and whose sole purpose is to provide channel estimates every block, and another that uses the so-obtained estimates to perform full detection à la (9).

The proposed Alamouti 2 scheme nonetheless achieves a diversity gain over the benchmark SIMO case. The exact amount of gain depends on the current channel conditions, which vary over the course of the experiment. When the channel conditions are harsh such that the SIMO system fails, so do the Alamouti schemes. This is evident around transmission # 17, and is believed to be related to the increased wave activity. In calmer channel conditions, e.g. between transmissions # 6 and # 11, a gain of up to 2 dB is observed. We also note that better overall performance is achieved with 512 carriers than with 1024 carriers, which is explained by the shorter block duration that results when fewer carriers are used within the same total bandwidth.

V. Conclusion

OFDM was considered in conjunction with the Alamouti space-time block code with the goal of providing spatial diversity gain in an underwater acoustic channel. Full benefits of Alamouti coding are contingent upon a slowly varying channel and the receiver's ability to track these variations. An acoustic channel, however, cannot in general be assumed constant over the duration of two consecutive OFDM blocks that comprise one Alamouti pair. To address this problem, we have decomposed the effects of time variation into those that affect the channel gain, and those that are caused by the system motion and affect the phase. Assuming that the gain varies more slowly than the phase, such a decomposition was shown to lead to a simplified receiver structure in which full benefits of Alamouti design can be preserved. The receiver algorithm was applied to experimental data, showing 1-2 dB gain over the reference SIMO structure.

Future work will focus on applying the Alamouti code in the frequency domain (between adjacent carriers) instead of the time domain (between adjacent blocks). By doing so, the system's sensitivity to the time-variation of the channel will be eliminated, while similar design principles as those developed in this paper will apply.

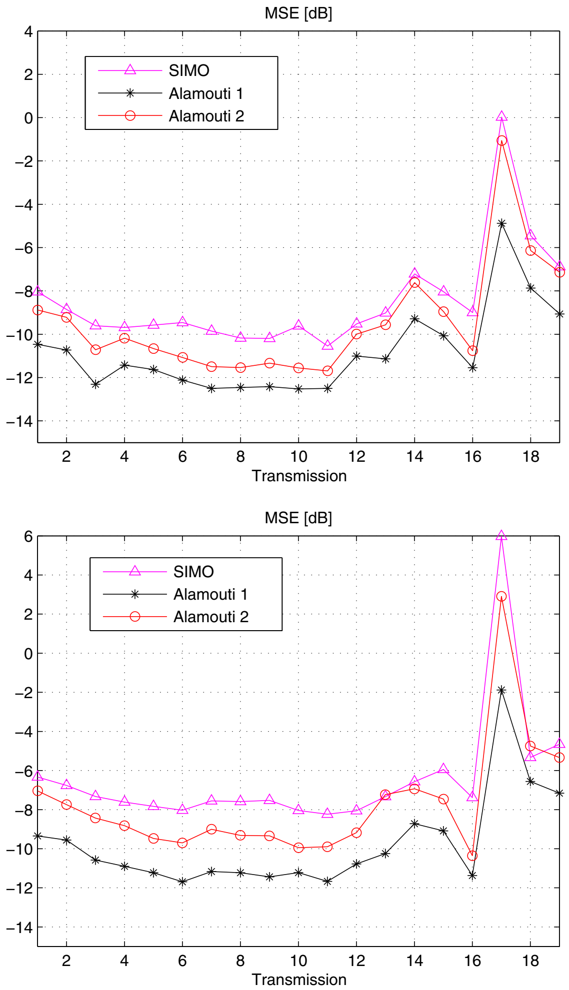

Fig. 1. MSE over three days of experiment; K=512 carriers (top) and K=1024 (bottom). Transmissions are separated by 4 hours. Alamouti 1 denotes the algorithm [4]; Alamouti 2 denotes the presently proposed algorithm.

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