Angle-Of-Arrival-Based Detection of Underwater Acoustic OFDM Signals

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***Abstract*—We propose a spatial combining technique for detection of multicarrier underwater acoustic communication signals using a vertical array of receivers. Instead of estimating the channel for each receiver independently and ignoring the angles of arrival of each of the channel paths, the suggested technique is based on spatially filtering each of the channel paths by steering a multi-frequency beamformer to their angles of arrival. As the number of channel paths is usually smaller than the number of the receivers, this process reduces the dimensionality of the model and involves a lower computational load. The signals at the output of the beamformer are then decoded in a coherent or differentially-coherent form. Simulation results show that angle-of-arrival-based detection outperforms the traditional angle-of-arrival-ignorant spatial combining methods for low signal to noise ratio and moderate range.**

***Index Terms*—Multicarrier underwater acoustic communication, multipath diversity, spatial combining, angle of arrival.**

1. Introduction

Orthogonal frequency division multiplexing (OFDM) acoustic communication in shallow water provides the ability to transmit at high data rates with robustness against the frequency-selectivity of the acoustic channel. Further robustness is achieved by the use of vertical arrays and spatial combining [1]–[4]. Processing of OFDM signals has typically been considered in a general form where no assumptions are made about the array structure [5]. As a result, the channel of each receiver is estimated separately. To improve the system performance, the receiver typically capitalizes on the sparsity in the delay domain, attempting to isolate the significant channel taps for each receiver. The data is decoded by combining the signals of all receivers. Differentially coherent combining, which eliminates the need for channel estimation has also been considered [6].

Herein, we adopt a different spatial processing approach for OFDM signals, which is based on multi-frequency beamforming, and is typically studied in the array processing literature [7]. Assuming that the signal is coherent across the array, there is no need to estimate each array element's channel separately. Instead, by combining the outputs of the receivers, we capatilize on the sparsity in the angle-of-arrival (AOA) domain, attempting to isolate P angles $\{\theta_p\}_{p=1}^{P}$ that describe the P propagation paths common to all receivers. These angles are related to the path delays of a single impulse response, which now suffices to describe all the receiving elements.

Reduced-dimensionality signal is obtained by considering the strongest P outputs of the multi-frequency beamforming process. The data is then decoded either via coherent or differentially coherent combining. Numerical simulations demonstrate that the uncoded symbol error rate (SER) of the proposed detectors outperform those of the AOA-ignorant counterpart [5], [6] especially at low signal to noise ratio (SNR), and for short to medium range communication with respect to (w.r.t.) the water depth. However, for longer communication range, where resolution in the AOA-domain is lost, the opposite is true.

2. Signal Model and Assumptions

We consider an OFDM signal with block duration T, and the guard interval T_g. The number of carriers is K, with the kth carrier located at frequency $f_k = f_0 + k\Delta f\,, k = -K/2, \ldots, K/2-1$, where f_0 is the center frequency, and $\Delta f = 1/T$ is the frequency spacing. Denoting by d_k the PSK symbol transmitted on the kth carrier, the OFDM signal corresponding to one block in baseband is

$$x(t) = \sum_{k=-K/2}^{K/2-1} d_k e^{j2\pi k\Delta f t},\, 0 \le t \le T^{'} = T + T_g\,. \quad (1)$$

The passband signal is represented by $\tilde{x}(t) = \Re\{x(t)e^{j2\pi f_0 t}\}$. The signal is transmitted over an underwater channel, and observed by a linear vertical array with M closely-spaced receivers , and d is the spacing between two adjacent receivers (see Fig. 1). The channel between the source and each receiver consists of P time-delayed paths and is assumed to be time-invariant during one block. Although, we assume that the number of paths is a-priori known, it can be determined using model order selection techniques [7]. Usually, P is small, about 5-7 (see Fig. 1 and Table I). Assuming plane wave propagation (justified in the array far-field region), and perfect Doppler compensation [5], the passband signal arriving to the mth receiver where $m = 1, \ldots, M$ is

$$\tilde{z}_m(t) = \sum_{p=1}^{P} h_p \tilde{x}(t - \underbrace{(\tau_p + (m-1)\frac{d}{c}\cos\theta_p)}_{\triangleq \bar{\tau}_{p,m}}) + \tilde{n}_m(t), \quad (2)$$

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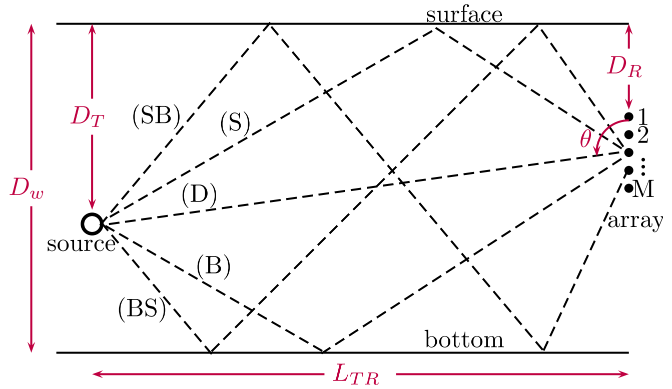

Fig. 1. The source and the sensor array geometry for a flat surface and a flat bottom with several propagation paths, where D, S, B, SB, and BS stands for direct, surface, bottom, surface-bottom and bottom-surface, respectively.

Path	AOA (radians)	Delay (sec)
(D)	$\frac{\pi}{2} - \tan^{-1}(\frac{D_1}{L_{TR}})$	$\sqrt{D_1^2 + L_{TR}^2}/c$
(B)	$\frac{\pi}{2} + \tan^{-1}(\frac{2D_w - D_2}{L_{TR}})$	$\sqrt{(2D_w - D_2)^2 + L_{TR}^2}/c$
(S)	$\frac{\pi}{2} - \tan^{-1}(\frac{D_2}{L_{TR}})$	$\sqrt{D_2^2 + L_{TR}^2}/c$
(SB)	$\frac{\pi}{2} + \tan^{-1}(\frac{2D_w - D_1}{L_{TR}})$	$\sqrt{(2D_w - D_1)^2 + L_{TR}^2}/c$
(BS)	$\frac{\pi}{2} - \tan^{-1}(\frac{2D_w + D_1}{L_{TR}})$	$\sqrt{(2D_w + D_1)^2 + L_{TR}^2}/c$

TABLE I
DELAYS AND AOAS OF THE PATHS IN FIG. 1, WHERE c IS THE PROPAGATION SPEED, AND $D_1 = D_R - D_T$ AND $D_2 = D_R + D_T$.

where $\tilde{n}_m(t)$ is the additive noise, assumed to be independent across the array elements, τ_p denotes the propagation delay of the pth path to the first (reference) receiver (we assume that $\tau_1 = 0$), and θ_p is the path angle of arrival. The parameter h_p represents the real-valued amplitude of the pth path.

The baseband version $z_m(t)$ of the received signal satisfies $\tilde{z}_m(t) = \Re\{z_m(t)e^{j2\pi f_0 t}\}$ and can be written as

$$z_m(t) = \sum_{p=1}^{P} h_p x(t - \bar{\tau}_{p,m}) e^{-j2\pi f_0 \bar{\tau}_{p,m}} + n_m(t), \quad (3)$$

where $n_m(t)$ is the additive noise in baseband. The measurement of the mth received signal at the kth carrier is,

$$\begin{aligned} z_{k,m} &= \frac{1}{T}\int_{t=0}^{T+T_g} z_m(t)e^{-j2\pi k\Delta f t}dt, \quad (4) \\ &\overset{(3)}{=} d_k \underbrace{\sum_{p=1}^{P} h_p e^{-j2\pi\bar{\tau}_{p,m} f_k}}_{\triangleq C_{k,m}} + \underbrace{\frac{1}{T}\int_{t=0}^{T} n_m(t)e^{-j2\pi k\Delta f t}dt}_{\triangleq n_{k,m}}, \end{aligned}$$

where $C_{k,m}$ represents the channel between the transmitter and the mth receiver at the kth frequency.

The problem herein is formulated as follows: Decode the data symbols, d_k, from $\{z_{k,m}\}$, given that the channel coefficients (amplitudes, delays, and AOAs) are unknown.

3. AN AOA-IGNORANT DECODER

In this section we review the basics of multi-channel (array) combining which makes no use of the array geometry, i.e. the AOA structure. Each channel is then estimated independently by capitalizing on the sparseness in the delay domain, i.e., on the fact that there are P significant channel taps paths that represent approximately the entire frequency response of each receiver, where $P \ll J < K$ and J is the maximal channel delay spread (in samples). Usually, sparse estimation techniques [8] utilize this assumption. Therefore, the channel $C_{k,m}$ is translated into a discrete-time equivalent channel parameterized by J complex-valued coefficients $\{g_{q,m}\}_{q=0}^{J-1}$ through

$$H_k(\{g_{q,m}\}_{q=0}^{J-1}) = \sum_{q=0}^{J-1} g_{q,m} e^{-j2\pi qk\Delta f T_s}. \quad (5)$$

The continuous-valued path delays, $\bar{\tau}_{p,m}$, are replaced by discrete values, qT_s, where $T_s = T/K$ is the sampling interval, chosen such that $H_k(\cdot)$ and $g_{q,m}$ are a discrete Fourier pair.

A. Coherent detection

To estimate the channel frequency response, we use L equispaced pilot symbols where $L < K$ and $L \geq J$. The least squares estimate of $g_{q,m}$ is [5, eq. (19)]

$$\hat{g}_{q,m} = \frac{1}{L}\sum_{\ell=1}^{L} d_{k_\ell} z_{k_\ell,m} e^{j2\pi k_\ell q/K}, \quad (6)$$

where k_ℓ is the index of the ℓth pilot symbol d_{k_ℓ}. Given the estimate (6) the channel frequency response $H_k(\cdot)$ is determined for each frequency index, k, using (5). The data symbols are finally estimated as [5, eq. (21)]

$$\hat{d}_k \propto \sum_{m=1}^{M} H_k(\{\hat{g}_{q,m}\}_{q=0}^{J-1})^* z_{k,m}, \quad (7)$$

and the decisions on the transmitted symbols are made by a slicing operation to the nearest hypothesis in the signal space.

B. Differentially coherent detection

Differentially coherent detection is used to eliminate the need for channel estimation. We assume that the data symbols are differentially encoded in the frequency domain such that

$$d_k = d_{k-1}b_k \ , k = -K/2+1, \ldots, K/2-1 \, , \quad (8)$$

where b_k is the PSK data symbol, $b_{-K/2}$ is known, and $d_{-K/2} = 1$. Differentially coherent detection is based on assuming that the channel frequency response is approximately constant between two consecutive carriers, i.e., $C_{k,m} \cong C_{k-1,m}$, such that the detected symbols are given as [6]

$$\hat{b}_k \propto \sum_{m=1}^{M} z_{k-1,m}^* z_{k,m} \, . \quad (9)$$

Final symbol decisions are made by choosing the closest point in the constellation. Despite its simplicity, the differentially coherent detection increases the noise level, and thus results in higher symbol error rate.

4. An AOA-Based Combining

In this approach, we capitalize on the sparsity in the AOA domain and estimate the P directions and their amplitudes and delays that describe the multipath response for all receivers, instead of estimating the channel of each receiver separately. We first re-parameterize the channel $C_{k,m}$ in (4) between the transmitter and the mth receiver at the kth frequency as

$$C_{k,m} = \sum_{p=1}^{P} \underbrace{e^{-j2\pi(m-1)\frac{d}{c}\cos\theta_p f_k}}_{\triangleq \varphi_{k,m}(\theta_p)} (\underbrace{e^{-j2\pi\tau_p \Delta f}}_{\triangleq \lambda_p})^k \underbrace{h_p e^{-j2\pi\tau_p f_0}}_{\triangleq c_p} . \quad (10)$$

Collecting the signals from all receivers at the kth frequency gives a column vector $\mathbf{z}_k = [z_{k,1}, z_{k,2}, \ldots, z_{k,M}]^T$ given as,

$$\mathbf{z}_k = d_k \mathbf{\Phi}_k(\boldsymbol{\theta}) \mathbf{\Lambda}_k \mathbf{c} + \mathbf{n}_k . \quad (11)$$

where $\mathbf{\Lambda}_k = \text{diag}(\lambda_1^k, \lambda_2^k, \ldots, \lambda_P^k)$, $\mathbf{c} = [c_1, c_2, \ldots, c_P]^T$, $\mathbf{n}_k = [n_{k,1}, n_{k,2}, \ldots, n_{k,M}]^T$. Also we define $\mathbf{\Phi}_k(\boldsymbol{\theta}) = [\boldsymbol{\varphi}_k(\theta_1), \boldsymbol{\varphi}_k(\theta_2), \cdots, \boldsymbol{\varphi}_k(\theta_P)]$, where the vector $\boldsymbol{\varphi}_k(\theta) = [\varphi_{k,1}(\theta), \varphi_{k,2}(\theta), \ldots, \varphi_{k,M}(\theta)]^T$, and $\boldsymbol{\theta} \triangleq [\theta_1, \ldots, \theta_P]^T$.

We now want to find a reduced P-dimension vector $\mathbf{r}_k = [r_{k,1}, r_{k,2}, \ldots, r_{k,P}]^T$ such that the vector $\hat{\mathbf{z}}_k = \mathbf{\Phi}_k \mathbf{r}_k$ minimizes the mean squared error (MSE) $\|\mathbf{z}_k - \hat{\mathbf{z}}_k\|^2$. Such minimization requires the path angles to be known. Denoting the estimate of the path direction vector by $\hat{\boldsymbol{\theta}}$, and defining $\hat{\mathbf{\Phi}}_k = \mathbf{\Phi}_k(\hat{\boldsymbol{\theta}})$, $\hat{\mathbf{c}}$ and $\hat{\mathbf{\Lambda}}_k$ as the estimate of $\mathbf{\Phi}_k$, $\mathbf{c}$ and $\mathbf{\Lambda}_k$, respectively (as detailed in the next section), the least squares reduced P-dimension vector is given by,

$$\hat{\mathbf{r}}_k = \hat{\mathbf{\Phi}}_k^\dagger \mathbf{z}_k \stackrel{(11)}{=} \hat{\mathbf{\Lambda}}_k \hat{\mathbf{c}} d_k + \hat{\mathbf{\Phi}}_k^\dagger \mathbf{n}_k , \quad (12)$$

where $\mathbf{\Phi}_k^\dagger = (\mathbf{\Phi}_k^H \mathbf{\Phi}_k)^{-1} \mathbf{\Phi}_k^H$ is the pseudo-inverse of $\mathbf{\Phi}_k$. The data symbols are estimated as according to the minimum mean square error principle [7, Ch. 6]

$$\begin{aligned} \hat{d}_k &= \mathrm{E}\{\mathbf{r}_k d_k^*\}^H E\{\mathbf{r}_k \mathbf{r}_k^H\}^{-1} \mathbf{r}_k , \\ &= \hat{\mathbf{c}}^H \hat{\mathbf{\Lambda}}_k^H (\hat{\mathbf{\Lambda}}_k \hat{\mathbf{c}} \hat{\mathbf{c}}^H \hat{\mathbf{\Lambda}}_k^H + (\hat{\mathbf{\Phi}}_k^H \hat{\mathbf{\Phi}}_k)^{-1} \sigma_n^2)^{-1} \mathbf{r}_k , \quad (13) \end{aligned}$$

and the corresponding decisions are made as before, i.e., by a slicing operation to the nearest hypothesis in the signal space.

5. Channel Estimation

The channel is typically not known and is slowly time varying; thus, the estimation process of the path angles, $\{\theta_p\}_{p=1}^P$ and the channel coefficients $\{c_p, \lambda_p\}_{p=1}^P$ is an integral part of the data detection process.

A. Path AOAs estimation

As mentioned previously, the MSE $\|\mathbf{z}_k - \hat{\mathbf{z}}_k\|^2$ is minimized w.r.t. $\mathbf{r}_k$ when $\hat{\mathbf{r}}_k = \mathbf{\Phi}_k^\dagger \mathbf{z}_k$ and then $\hat{\mathbf{z}}_k = \mathbf{P}_{\mathbf{\Phi}_k} \mathbf{z}_k$, where $\mathbf{P}_{\mathbf{\Phi}_k} = \mathbf{\Phi}_k \mathbf{\Phi}_k^\dagger$ is a projection matrix. The resulting MSE is $\|\mathbf{z}_k\|^2 - \mathbf{z}_k^H \mathbf{P}_{\mathbf{\Phi}_k} \mathbf{z}_k$. This procedure is equivalent to selecting a vector $\hat{\mathbf{z}}_k$ such that the norm of the noise vector, $\mathbf{n}_k$, is minimized. However, $\mathbf{P}_{\mathbf{\Phi}_k}$ depends on the direction vector $\boldsymbol{\theta}$ which is joint to all observations. Hence, the set of path directions are selected such that the norm of the total noise, $\sum_k \|\mathbf{n}_k\|^2$, is minimized. By collecting the measurements at all frequencies we get that the direction vector that minimizes the norm of the total noise is

$$\hat{\boldsymbol{\theta}} = \underset{\tilde{\boldsymbol{\theta}}}{\text{argmax}} \{\sum_k \mathbf{z}_k^H \mathbf{P}_{\mathbf{\Phi}_k} \mathbf{z}_k\} . \quad (14)$$

Assuming that the estimated AOAs are close to their true values, and that the true AOAs are spatially resolvable (as will occur in short to medium range communication links) we have that

$$\hat{\mathbf{\Phi}}_k^H \hat{\mathbf{\Phi}}_k \cong M\mathbf{I} , \quad (15)$$

where $\mathbf{I}$ is the identity matrix. We cannot guarantee that this property will hold in general; however, with a large number of array elements and large signal to noise ratio, it hold at least approximately. We thus obtain that $\mathbf{z}_k^H \mathbf{P}_{\mathbf{\Phi}_k} \mathbf{z}_k \cong 1/M \|\mathbf{\Phi}_k^H \mathbf{z}_k\|^2$ and therefore,

$$\hat{\boldsymbol{\theta}} = \underset{\tilde{\boldsymbol{\theta}}}{\text{argmax}} \{\sum_p \underbrace{\sum_k |\boldsymbol{\varphi}_k^H(\tilde{\theta}_p) \mathbf{z}_k|^2}_{\triangleq Q(\tilde{\theta}_p)}\} . \quad (16)$$

This optimization requires a P-dimensional search in the $\boldsymbol{\theta}$-space. Instead, one can employ a sub-optimal but computationally efficient estimate via a one-dimensional search as

$$\{\hat{\theta}_p\}_{p=1}^P = \arg \max_{\tilde{\theta}} Q(\tilde{\theta}) . \quad (17)$$

We compute the total power (at all frequencies) at each hypothesized $\tilde{\theta}$, assuming that $\tilde{\theta} \in (-\pi/2, \pi/2)$, or restricting it to a narrow span of directions implied by the system geometry. We then select the P directions that are associated with the largest peaks. Since we assume a linear array with equi-spaced elements, the cost function in (16) can be computed using the fast Fourier transform (FFT). Given the reduced-dimensionality vector $\hat{\mathbf{r}}_k$ in (12), the data symbols can be obtained either via coherent or differentially-coherent detection.

B. Coherent decoder

We determine the path coefficients, $\{\lambda_p, c_p\}_{p=1}^P$, using the L equi-spaced pilot carriers. From (12) we have that

$$\hat{\mathbf{r}}_p \cong \mathbf{q}_p c_p , \quad (18)$$

where $\hat{\mathbf{r}}_p = [\hat{r}_{k_1,p}, \ldots, \hat{r}_{k_L,p}]^T$, $\mathbf{q}_p = [d_{k_1} \lambda_p^{k_1}, \ldots, d_{k_L} \lambda_p^{k_L}]^T$, where k_ℓ, $\ell = 1, \ldots, L$ is the index of the ℓth pilot. The estimates of the coefficient c_p and λ_p are determined as

$$(\hat{c}_p , \hat{\lambda}_p) = \underset{c_p, \lambda_p}{\text{argmin}} \|\hat{\mathbf{r}}_p - \mathbf{q}_p c_p\|^2 . \quad (19)$$

Taking the derivative with respect to c_p and setting the result equal to zero yields the estimate of c_p as

$$\begin{aligned} \hat{c}_p &= \mathbf{q}(\lambda_p)^\dagger \hat{\mathbf{r}}_p , \\ &= \frac{1}{\sum_{\ell=1}^L |d_{k_\ell}|^2} \sum_{\ell=1}^L \hat{r}_{k_\ell,p} d_{k_\ell}^* (\lambda_p^*)^{k_\ell} . \quad (20) \end{aligned}$$

Substituting (20) into (19) yields that the estimate of λ_p is

$$\begin{aligned}\hat{\lambda}_p &= \underset{\lambda_p}{\operatorname{argmax}} |\hat{\mathbf{r}}_p^H \mathbf{q}(\lambda_p)|^2 , \\ &= e^{j \cdot \underset{\phi_p}{\operatorname{argmax}} |\sum_{\ell=1}^{L} \hat{r}^*_{k_\ell, p} d_{k_\ell} e^{j\phi_p k_\ell}|^2} , \end{aligned} \quad (21)$$

where we define $\lambda_p = e^{j\phi_p}$. The sum in (21) can be computed efficiently using FFT. The estimate $\hat{c}_p$ is determined by substituting (21) into (20). Coherent detection is then performed by substituting the estimates $\hat{\theta}_p$, $\hat{c}_p$ and $\hat{\lambda}_p$ into (13).

C. Differentially-coherent detection

Instead of estimating the channel parameters using pilot symbols, differential decoding can be employed. The data symbols are differentially encoded in the frequency domain as in (8). Using (12) and assuming that for $p = 1, \ldots, P$, $\lambda_p^{k-1} \cong \lambda_p^k$ we have that,

$$\hat{b}_k \propto \mathbf{r}_{k-1}^H \mathbf{r}_k \, . \quad (22)$$

Given the knowledge of the first data symbol $b_{-K/2}$, the rest can be determined recursively, using the decisions $\hat{b}_{k-1}$.

6. Complexity Load

We compare the complexity load of the AOA-based and AOA-ignorant methods and focus on the coherent detection cases. The dominant complexity of the AOA-ignorant coherent approach involves the computation of $H_k(\cdot)$ in (5) which requires $\mathcal{O}(MK\log_2 J)$ multiplications.

The complexity of the AOA-based coherent approach involves computation of the AOA according to (17) and then channel parameters in (21). The number of search cells in the angle axis is approximately $2\pi/(2/M) \cong 3M$, since the minimal resolution cell is $\Delta\theta = \lambda/(M-1)d \cong 2/M$ (for $d = \lambda/2$). Computing (17) thus requires $\mathcal{O}(KM^2)$ multiplications or even $\mathcal{O}(KM\log_2 M)$ multiplications if this process is performed using FFT. However, in practice, the path AOA is usually slowly time-varying and an initial computation of the path AOA followed by tracking technique is sufficient. The dominant complexity load of computing the channel parameters involves the computation of $\hat{\lambda}_p$ in (21) which requires $\mathcal{O}(PL\log_2 L)$.

Hence, the coherent AOA-based technique involves less computations compared to the coherent AOA-ignorant technique when $PL\log_2 L \ll MK\log_2 J$, i.e., if

$$P \ll \frac{MK}{L} \frac{\log(J)}{\log(L)} \ll M\frac{K}{L} \, , \quad (23)$$

since in practice $L \approx J$. In typical underwater acoustic channels, P is about 5-7, M is between 10-20 and K/L is usually 4-8. We thus obtain that the proposed AOA-based approach involves a significant reduction in complexity compared to that of the AOA-ignorant method.

7. Simulation Examples

We evaluate the performance of the proposed decoding schemes using numerical simulations. We consider an OFDM QPSK signaling with $K = 1024$ carriers operating at a center frequency $f_0 = 20$ KHz with a bandwidth of 12 KHz. The carrier spacing is $\Delta f = 11.72$ Hz and the block duration is $T = 85.3$ msec [5]. The total number of pilots is $L = 256$, equally spaced across the OFDM block. The number of receivers in the vertical array is $M = 25$ with a spacing of $d = \lambda_{\min}/2 = c/(2(f_0 + K/2\Delta f)) = 0.0288$ m. The water depth is $D_w = 80$ m, the transmitter depth is $D_T = 50$ m, and the receiver array depth is $D_R = 40$ m. The delays and the AOAs of the first five paths are computed as detailed in Table I. The performance of the various decoders is expressed using the uncoded SER defined as the ratio between the number of symbols that are incorrectly decoded and the total number of data symbols.

In the first example, the SNR is varied between -4 dB to 8 dB with a step of 1 dB. For each SNR we performed several Monte-Carlo trials, where in each trial only the additive noise is randomly chosen. The distance between the transmitter and the array is $L_{TR} = 600$ m. In Fig. 2 we show the SER as a function of the SNR of all decoders. With this setup, the proposed AOA-based detectors outperform the SER of the AOA-ignorant coherent and differentially-coherent decoders, especially at low SNR.

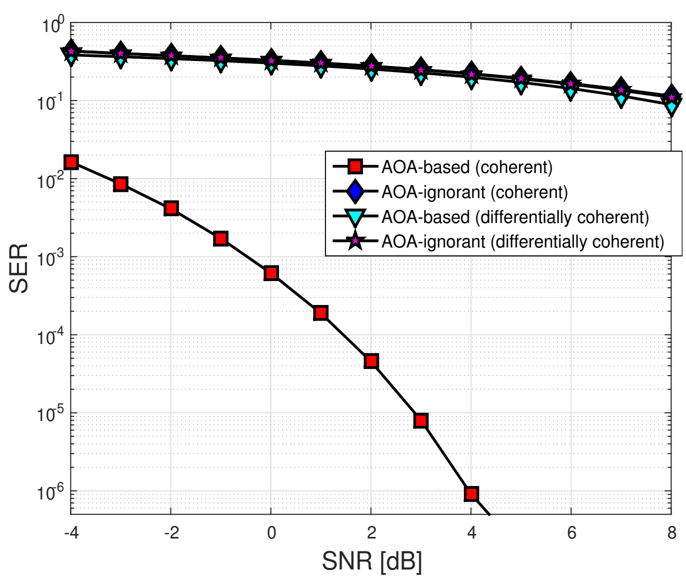

Fig. 2. SER of the decoders vs. SNR.

We further evaluate the SER performance as a function of the distance between the transmitter and the array. We consider the same water depth as with the previous example, where the distance varies from 600 m to 1600 m with a step of 100 m. For distances beyond this set of values, the angles of arrival becomes closely spaced and estimating them require either increasing the array aperture or using high resolution methods with large complexity. While for distances values below this set the delay spread of the channel is very large and may exceed the block duration. This results in effects which are beyond the scope of the current work. In Fig. 3 we show the SER performance versus the distance for SNR of 0 dB and 8 dB, respectively. As can be seen from both figures, for short to medium range communication links the AOA-based coherent and differentially coherent detection outperforms its AOA-ignorant counterparts. However, the opposite occurs for long range communication link. As the range increases, the difference between the AOAs decreases, causing a loss of resolution and making it more difficult to produce accurate AOA estimates.

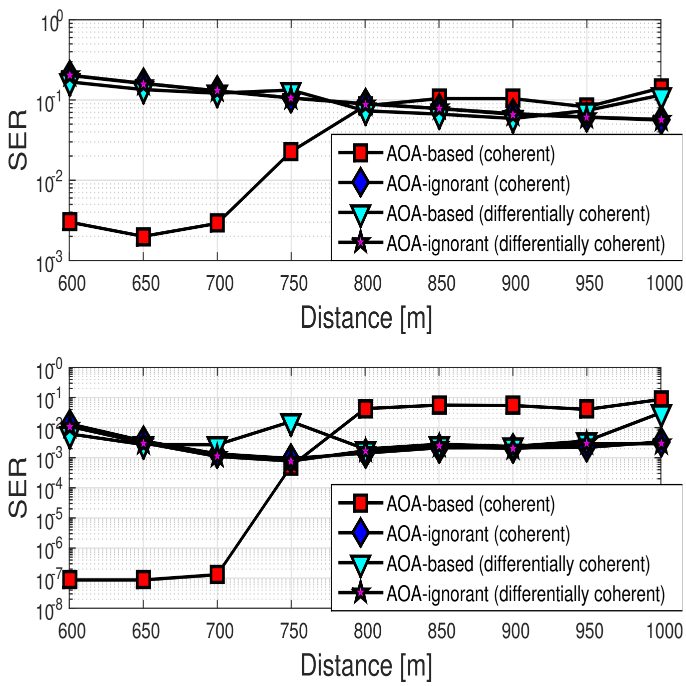

Fig. 3. SER of the decoders vs. transmitter-array distance for SNR=0 dB (upper plot) and SNR=8 dB (lower plot).

8. Conclusion

We suggested an AOA-based detection technique for multicarrier underwater acoustic communication using vertical arrays. The proposed spatial combining is based on a pre-processing stage of a conventional beamformer, which estimates a single multipath profile that suffices to describe the channel responses of all the array elements. Combining all the array elements' signals yields a channel estimate that is more accurate than the one obtained by estimating each element's channel individually. Subsequent array combining for data detection was considered in both coherent and differentially coherent forms, and shown to outperform individual AOA-ignorant channel estimation at low SNR and moderate ranges that preserve resolution in the AOA domain. Future research will focus on extending the proposed approach to time-varying underwater channel, and combining coding, and adaptive channel estimation.

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