Pilot-tone based ZP-OFDM Demodulation for an Underwater Acoustic Channel

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***Abstract*— Existing coherent underwater acoustic communication systems rely on single carrier transmission and adaptive decision feedback equalization to deal with time-varying and highly dispersive underwater acoustic (UWA) channels. Equalization complexity prevents any substantial rate improvement with the existing single-carrier approach, as the channel frequency selectivity increases considerably when the symbol rate increases. Multicarrier modulation in the form of orthogonal frequency division multiplexing (OFDM), on the other hand, converts a frequency selective channel into a set of parallel frequency-flat subchannels, thus greatly simplifying receiver equalization. Motivated by the success of OFDM in radio channels, we investigate its use for underwater acoustic channels. In this paper, we develop a pilot-tone based receiver design for zero-padded OFDM transmissions, and test it in a real underwater acoustic channel. Our proposed receiver performs carrier frequency offset compensation, channel estimation, and data demodulation on the basis of individual OFDM block. This approach is appealing to applications with short data bursts, or fast varying channels, as it does not rely on channel dependence across OFDM blocks.**

I. Introduction

Although land-based wireless sensor networks have proliferated in many applications, the use of underwater sensor networks has been limited. Nonetheless, there has been a growing interest in building distributed and scalable underwater wireless sensor networks (UWSN) that will bring significant advantages and benefits in a wide spectrum of underwater applications, such as ocean observation for scientific exploration, commercial exploitation, coastline protection and target detection in military events [1], [5]. Improving underwater acoustic communications among distributed sensor nodes is one of the major design issues.

Due to the reverberation effect where the receiver observes multipath signals bounced from the surface and the bottom, underwater acoustic channels usually have large delay spread, leading to strong frequency selectivity. On the other hand, UWA channels exhibit high time-variation temporally and spatially. Being both frequency- and time-selective, UWA channels pose great challenges for high performance and high rate communications.

Existing coherent underwater communication uses single carrier transmission and relies on linear or non-linear equalization techniques to suppress inter-symbol interference (ISI) [9]. The canonical receiver in [13], which demonstrated the feasibility of phase coherent modulation, relies on an adaptive decision feedback equalizer coupled with delay and phase tracking. As the data rate increases, the symbol duration decreases, and thus a channel with the same delay spread contains more channel taps when converted to the baseband discrete-time model. This imposes great challenges for the channel equalizer, whose complexity will prevent substantial rate improvement with the existing single-carrier approach.

Multicarrier modulation in the form of orthogonal frequency division multiplexing (OFDM) has been quite successful in broadband wireless communication over radio channels, e.g., wireless local area networks (IEEE 802.11a/g/n) [6], and wireless metropolitan area networks (IEEE 802.16) [8]. OFDM divides the available bandwidth into a large number of overlapping subbands, so that the symbol duration is long compared to the multipath spread of the channel. Consequently, ISI may be neglected in each subband, that greatly simplifies the receiver complexity on channel equalization.

The question is then: *can the success of OFDM be replicated in underwater acoustic channels?* Motivated by this question, researchers have long attempted to apply OFDM in underwater acoustic channels. However, the existing literature focuses mostly on conceptual system analysis and simulation based studies [2], [10], [11], [15], while the experimental results are scarce [4], [3].

In this paper, we develop a pilot-tone based receiver design for zero-padded OFDM transmissions. Our focus is on a receiver design that operates on each OFDM block separately. Such a design is appealing to applications with short data bursts, or fast varying channels, as it does not rely on channel dependence across OFDM blocks. Based on pilot tones, we perform carrier frequency offset and channel estimation for each block, followed by data demodulation. We test our receiver design in a real underwater acoustic channel. We obtain solid system performance, when signals from multiple receive-elements are properly combined.

The rest of this paper is organized as follows. We describe the receiver design for ZP-OFDM in Section II, and present the experimental setting and numerical results for a real UWA channel in Section III. We draw conclusions in Section IV.

Notation: Bold upper and lower letters denote matrices and column vectors, respectively; $(\cdot)^T$, $(\cdot)^*$, and $(\cdot)^{\mathcal{H}}$ denote transpose, conjugate, and Hermitian transpose, respectively; $\|\cdot\|$ denotes the two-norm of a vector; $\mathbf{I}_N$ is the $N \times N$ identity

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matrix; $\mathbf{0}_{M\times N}$ denotes an all-zero matrix of size $M\times N$; $\mathbf{F}_N$ denotes a unitary $N\times N$ FFT matrix with the $(p+1,q+1)^{\rm st}$ entry as $\frac{1}{\sqrt{N}}e^{-j\frac{2\pi}{N}pq}$.

II. Receiver Design for ZP-OFDM

A. ZP-OFDM Basics

OFDM is a multicarrier modulation with efficient implementation based on fast-Fourier-transform (FFT). Consider an OFDM transmission over a frequency selective channel, that is described by its discrete-time baseband impulse response vector $\mathbf{h} := [h(0),\ldots,h(L)]^T$, where L stands for the channel order. The channel impulse response includes the effects of transmit-receive filters and physical multipath. Inverse FFT operation and cyclic prefix (CP) insertion at the transmitter together with CP removal and FFT processing at the receiver diagonalize the associated channel matrix [12]. As such, OFDM converts an ISI channel into parallel ISI-free subchannels with gains equal to the channel's frequency response values on the FFT grid. This means low equalization complexity regardless of the dispersive channel.

On the other hand, CP insertion can be replaced by zero-padding (ZP), leading to the so called ZP-OFDM [12]. In this work, we focus on ZP-OFDM rather than CP-OFDM, due to the following reasons: i) the UWA channel has large delay spread, thus the CP portion would consume a considerable fraction of the transmission power; and ii) ZP-OFDM is robust against channel nulls, leading to better performance than CP-OFDM, when appropriate receivers are used [12].

The ZP-OFDM transmitter partitions the information symbols into blocks each with length K. Let $\mathbf{s}$ denote one information block: $\mathbf{s} := [s(0),\ldots,s(K-1)]^T$. IFFT on $\mathbf{s}$ leads to $\mathbf{F}_K^{\mathcal{H}}\mathbf{s}$. Padding $L_{\rm zp}$ zeros after $\mathbf{F}_K^{\mathcal{H}}\mathbf{s}$ can be described by a matrix-vector multiplication as $\mathbf{T}_{\rm zp}\mathbf{F}_K^{\mathcal{H}}\mathbf{s}$, where we define $\mathbf{T}_{\rm zp} := [\mathbf{I}_K, \mathbf{0}_{K\times L_{\rm zp}}]^T$. After parallel to serial conversion, the block $\mathbf{T}_{\rm zp}\mathbf{F}_K^{\mathcal{H}}\mathbf{s}$ of length $P = K + L_{\rm zp}$ is transmitted through the channel $\mathbf{h}$. Assuming $L_{\rm zp} \geq L$ to avoid inter-block interference, the received symbol block of length P is

$$\mathbf{y} = \mathbf{H}\mathbf{T}_{\rm zp}\mathbf{F}_K^{\mathcal{H}}\mathbf{s} + \mathbf{n}, \tag{1}$$

where $\mathbf{H}$ is a $P\times P$ Toeplitz matrix with first column $[h_0,\ldots,h_L,0,\ldots,0]^T$ and first row $[h_0,0,\ldots,0]$, and $\mathbf{n} = [n(0),\ldots,n(P-1)]^T$ stands for additive white Gaussian noise.

The information symbols can be recovered using a linear zero-forcing (ZF) receiver as:

$$\hat{\mathbf{s}} = \mathbf{F}_K^{\mathcal{H}}(\mathbf{H}_0^{\mathcal{H}}\mathbf{H}_0)^{-1}\mathbf{H}_0^{\mathcal{H}}\mathbf{y}, \tag{2}$$

where $\mathbf{H}_0 := \mathbf{H}\mathbf{T}_{\rm zp}$ is defined for notational brevity. More advanced receivers, such as linear minimum-mean-square-error (MMSE), non-linear decision-feedback-equalizer (DFE) and sphere-decoders can also be used. Note that $\mathbf{H}_0$ has full column rank irrespective of the channel nulls, thus $(\mathbf{H}_0^{\mathcal{H}}\mathbf{H}_0)^{-1}$ is guaranteed invertible. Such a channel-irrespective invertibility is not available for CP-OFDM [12].

Though attractive performance-wise, ZF receiver entails matrix inversion, that may place a computational burden on the implementation. In this paper, we adopt the low-complexity overlap-add (OLA) based demodulation for OFDM [12]. The idea of OLA is to convert a linear convolution into a circular convolution, and then relies on FFT based signal processing.

Define a matrix $\mathbf{R}_{\rm ola} := [\mathbf{I}_K, \mathbf{I}_{\rm zp}]$, where $\mathbf{I}_{\rm zp}$ is the first $L_{\rm zp}$ columns of $\mathbf{I}_K$. The OLA operation can be described as

$$\tilde{\mathbf{y}} = \mathbf{R}_{\rm ola}\mathbf{y} = \underbrace{\mathbf{R}_{\rm ola}\mathbf{H}\mathbf{T}_{\rm zp}}_{:=\tilde{\mathbf{H}}}\mathbf{F}_K^{\mathcal{H}}\mathbf{s} + \mathbf{R}_{\rm ola}\mathbf{n}, \tag{3}$$

which amounts to retaining the first K entries of $\mathbf{y}$ while adding the last $L_{\rm zp}$ entries of $\mathbf{y}$ to the first $L_{\rm zp}$ entries. The resulting channel matrix $\tilde{\mathbf{H}}$ is circulant with the (i,j)th entry as $h((i-j) \bmod K)$. As a circulant matrix can be diagonalized by (I)FFT matrices, we obtain:

$$\begin{aligned}\mathbf{z} = \mathbf{F}_K\tilde{\mathbf{y}} &= \mathbf{F}_K\tilde{\mathbf{H}}\mathbf{F}_K^{\mathcal{H}}\mathbf{s} + \mathbf{F}_K\mathbf{R}_{\rm ola}\mathbf{n}\\ &= \mathbf{D}_H\mathbf{s} + \mathbf{v},\end{aligned} \tag{4}$$

where $\mathbf{v} = \mathbf{F}_K\mathbf{R}_{\rm ola}\mathbf{n}$ is the processed noise, $\mathbf{D}_H := \mathrm{diag}(H(0),\ldots,H(K-1))$, and $H(k)$ is the channel's frequency responses on the kth subcarrier:

$$H(k) = \sum_{l=0}^{L} h(l)e^{-j\frac{2\pi}{K}kl}, \quad k = 0,\ldots,K-1. \tag{5}$$

The scalar version of (4) is

$$\begin{aligned}\begin{bmatrix} z(0)\\ \vdots\\ z(K{-}1)\end{bmatrix} &= \begin{bmatrix} H(0) & & \\ & \ddots & \\ & & H(K{-}1)\end{bmatrix}\begin{bmatrix} s(0)\\ \vdots\\ s(K{-}1)\end{bmatrix}\\ &\quad + \begin{bmatrix} v(0)\\ \vdots\\ v(K{-}1)\end{bmatrix}.\end{aligned} \tag{6}$$

Therefore, channel equalization can be achieved by scalar inversion on each subcarrier:

$$\hat{s}(k) = \frac{z(k)}{H(k)}, \quad k = 0,\ldots,K-1. \tag{7}$$

B. Pilot Tone based Channel Estimation

The receiver needs to estimate the channel $\mathbf{h}$ (or $\mathbf{D}_H$) before equalization. We here adopt a low-complexity pilot tone based channel estimator. Out of K symbols in $\mathbf{s}$, N_p of them are known pilot data. The pilot tones are designed according to the following two guidelines:

d1) The N_p pilot symbols are equally spaced at subcarriers

$$0, M, 2M, \ldots, (N_p-1)M, \tag{8}$$

where $M = K/N_p$ is an integer by design.

d2) The pilot symbols are PSK signals with unit amplitude.

For notational convenience, we define a selection matrix $\mathbf{T}_{\rm sc}$ that are N_p rows of $\mathbf{I}_K$ with the row indices specified in (8). From (4), we isolate the channel outputs on those N_p subcarriers as

$$\mathbf{T}_{\rm sc}\mathbf{z} = (\mathbf{T}_{\rm sc}\mathbf{D}_H\mathbf{T}_{\rm sc}^T)(\mathbf{T}_{\rm sc}\mathbf{s}) + \mathbf{T}_{\rm sc}\mathbf{v}. \tag{9}$$

Define $\mathbf{D}_s = \text{diag}(\mathbf{T}_{\text{sc}}\mathbf{s})$, and an $N_p \times L$ Vandermonde matrix $\mathbf{V}$ as

$$\mathbf{V} = \begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{-j\frac{2\pi}{K}M} & \cdots & e^{-j\frac{2\pi}{K}ML} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & e^{-j\frac{2\pi}{K}(N_p-1)M} & \cdots & e^{-j\frac{2\pi}{K}(N_p-1)ML} \end{bmatrix}. \quad (10)$$

The matrix $\mathbf{V}$ is nothing but the first L columns of a scaled FFT matrix $\sqrt{N_p}\mathbf{F}_{N_p}$. Based on (5) and (10), we can re-express (9) as

$$\mathbf{T}_{\text{sc}}\mathbf{z} = \mathbf{D}_s\mathbf{V}\mathbf{h} + \mathbf{T}_{\text{sc}}\mathbf{v}. \quad (11)$$

Treating $\mathbf{h}$ as a deterministic unknown vector, the least square estimate of $\mathbf{h}$ is

$$\begin{aligned} \hat{\mathbf{h}}_{\text{LS}} &= \arg\min_{\mathbf{h}} \|\mathbf{T}_{\text{sc}}\mathbf{z} - \mathbf{D}_s\mathbf{V}\mathbf{h}\|^2 \\ &= (\mathbf{V}^{\mathcal{H}}\mathbf{D}_s^{\mathcal{H}}\mathbf{D}_s\mathbf{V})^{-1}\mathbf{V}^{\mathcal{H}}\mathbf{D}_s^{\mathcal{H}}\mathbf{T}_{\text{sc}}\mathbf{z}. \end{aligned} \quad (12)$$

Thanks to the equi-spaced pilots in d1), we have $\mathbf{V}^{\mathcal{H}}\mathbf{V} = N_p\mathbf{I}_L$. Also we have $\mathbf{D}_s^{\mathcal{H}}\mathbf{D}_s = \mathbf{I}_{N_p}$ due to d2). Therefore, the LS solution in (12) simplifies to

$$\hat{\mathbf{h}}_{\text{LS}} = \frac{1}{N_p}\mathbf{V}^{\mathcal{H}}\mathbf{D}_s^{\mathcal{H}}\mathbf{T}_{\text{sc}}\mathbf{z}, \quad (13)$$

which does not involve matrix inversion, and can be implemented by N_p-point IFFT. With the time-domain channel estimate $\hat{\mathbf{h}}_{\text{LS}}$, we use (5) to obtain $\hat{H}(k)$.

The LS fitting error corresponding to (12) is

$$\begin{aligned} \mathcal{E}_{\text{LS}} &= \|\mathbf{T}_{\text{sc}}\mathbf{z} - \mathbf{D}_s\mathbf{V}\hat{\mathbf{h}}_{\text{LS}}\|^2 \\ &= \|\mathbf{T}_{\text{sc}}\mathbf{z}\|^2 - N_p^{-1}\|\mathbf{V}^{\mathcal{H}}\mathbf{D}_s^{\mathcal{H}}\mathbf{T}_{\text{sc}}\mathbf{z}\|^2. \end{aligned} \quad (14)$$

C. CFO Estimation

One main challenge for OFDM in a UWA channel is that fast variation of the channel within each OFDM block will destroy the orthogonality among subcarriers, giving rise to inter-subcarrier-interference (ICI). Explicit modeling and compensation of channel variation is thus a must. Advanced channel variation models will be pursued in our future work. In this paper, we assume that the channel variation is solely due to a carrier frequency offset (CFO), denoted by ϵ. The CFO itself could change from block to block.

Define $\mathbf{\Gamma}(\epsilon) = \text{diag}(1, e^{j2\pi T_s\epsilon}, \cdots, e^{j2\pi T_s\epsilon\cdot(P-1)})$, where T_s is the sampling interval. In the presence of CFO, the received vector in (1) becomes

$$\mathbf{y} = \mathbf{\Gamma}(\epsilon)\mathbf{H}\mathbf{T}_{\text{zp}}\mathbf{F}_K^{\mathcal{H}}\mathbf{s} + \mathbf{n}. \quad (15)$$

The CFO has to be estimated and compensated. Otherwise, the pilot tone based channel estimation and OLA based data demodulation will be severely affected by ICI.

We perform a one-dimensional search to estimate ϵ. For each tentative ϵ, we compensate CFO to obtain $\mathbf{\Gamma}(\epsilon)^{-1}\mathbf{y}$, and perform the pilot-tone based channel estimation based on

$$\mathbf{z}(\epsilon) = \mathbf{F}_K\mathbf{R}_{\text{ola}}\mathbf{\Gamma}(\epsilon)^{-1}\mathbf{y}. \quad (16)$$

The LS fitting error in (14) is used as the performance indicator to find the best fit for ϵ. In short, our CFO estimator is:

$$\hat{\epsilon} = \arg\min_{\epsilon} \left\{\|\mathbf{T}_{\text{sc}}\mathbf{z}(\epsilon)\|^2 - N_p^{-1}\|\mathbf{V}^{\mathcal{H}}\mathbf{D}_s^{\mathcal{H}}\mathbf{T}_{\text{sc}}\mathbf{z}(\epsilon)\|^2\right\}. \quad (17)$$

D. Multi-channel Combining

Multi-channel reception greatly improves the system performance utilizing receive-diversity; see e.g. [14] on multi-channel combining for single-carrier transmissions over UWA channels.

Multi-channel combining can be easily done on each OFDM subcarrier. Suppose that we have N_r receive elements, and let $z_r(k)$, $H_r(k)$, and $v_r(k)$ denote the channel output, the channel frequency response, and the additive noise at the kth OFDM subcarrier of the rth element. We thus have:

$$\underbrace{\begin{bmatrix} z_1(k) \\ \vdots \\ z_{N_r}(k) \end{bmatrix}}_{:=\bar{\mathbf{z}}(k)} = \underbrace{\begin{bmatrix} H_1(k) \\ \vdots \\ H_{N_r}(k) \end{bmatrix}}_{:=\bar{\mathbf{h}}(k)} s(k) + \underbrace{\begin{bmatrix} v_1(k) \\ \vdots \\ v_{N_r}(k) \end{bmatrix}}_{:=\bar{\mathbf{v}}(k)}. \quad (18)$$

Assuming that $\bar{\mathbf{v}}(k)$ has independent and identically distributed entries, the optimal maximum-ratio combining (MRC) yields

$$\hat{s}(k) = [\bar{\mathbf{h}}^{\mathcal{H}}(k)\bar{\mathbf{h}}(k)]^{-1}\bar{\mathbf{h}}^{\mathcal{H}}(k)\bar{\mathbf{z}}(k). \quad (19)$$

In summary, our ZP-OFDM receiver is as follows:

1) first perform CFO estimation as in (17) on each receive element.
2) after CFO compensation, perform channel estimation as in (13) on each receive element.
3) finally perform MRC combining as in (19) for data demodulation.

The CFO/channel estimation and data demodulation are carried out on each OFDM block.

III. Offline ZP-OFDM Demodulation based on Experimental Data

A. Experiment Setting

ZP-OFDM signals have been transmitted and collected in several experiments carried out by the WHOI acoustic communication group. We will present some results regarding the experiment performed off the coast of Buzzard Bay in Sept. 2005. The transmission range was 2.5 km, with 12 receive-elements. The transmitter and receivers were anchored about 12 m in deep water.

The ZP-OFDM signal occupies the band of $22 - 46$ kHz with bandwidth $B = 24$ kHz. The sampling rate f_s is 96 kHz. A total of $N_d = 2^{15}$ QPSK symbols are transmitted as one data burst (termed as one packet here) for each setting with different number of subcarriers. With K subcarriers per OFDM symbol, we thus have N_d/K OFDM blocks for one data burst, as shown in Fig. 1. Each OFDM block is of duration $T = K/B$, followed by a guard time $T_g = 25$ ms. The packet is preceded by a probe signal of duration $T_{sw} = 100$ ms and a 50 ms pause. (The probe signal in this experiment is a PN sequence of length 127, quadrature modulated at 24ksps using the center frequency of 34kHz.) Another pause of 50 ms ends the packet.

The experiment includes OFDM transmissions of $K = 128, 256, 512, 1024, 2048$. In this paper, we focus on the $K =$

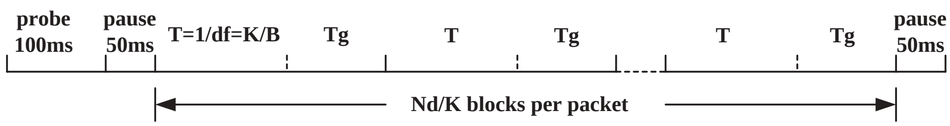

Fig. 1. The partitioning of one data packet with N_d symbols

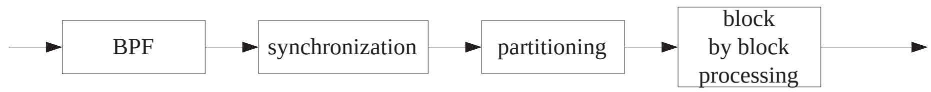

Fig. 2. Pre-processing to partition the received data into blocks.

Fig. 3. Block by block processing

1024 case. Excluding the probing and pause signal portion, the raw bit rate is:

$$R_{\mathrm{raw}} = 2B \cdot \frac{T}{T+Tg} = 2 \cdot 24 \cdot \frac{K/B}{K/B+T_g} = 30.2 \text{ kbps}. \quad (20)$$

We will use $N_p = K/4$ pilot tones for CFO and channel estimation, that leads to the achieved rate of

$$R = R_{\mathrm{raw}} \frac{K - K/4}{K} = 22.7 \text{ kbps}. \quad (21)$$

B. Receiver Description

Figs. 2 and 3 depict the receiver diagram. Synchronization is done via correlating the received samples with the known probing sequence. Note that in OFDM systems, we only need very coarse synchronization, because timing offsets can be included as zero taps in the overall multipath channel. With coarse synchronization, the receiver partitions the received data into OFDM blocks, based on the transmission structure shown in Fig. 1.

We focus on block by block OFDM demodulation without exploiting any channel dependence across OFDM blocks. After downshifting the real passband signals to complex baseband, the receiver carries out joint CFO and channel estimation on each receive antenna, and then applies MRC combining for data demodulation.

C. Performance results

Fig. 4 depicts the estimated channel taps on one receive-element. It can be seen that the channel has non-zero support within:

$$L \times T/K = 6.25\mathrm{ms}, \quad (22)$$

where the channel order is set $L = 150$ and T stands for the OFDM symbol period.

Fig. 5 shows the estimated CFO for $N_d/K = 32$ blocks on one receive-element. We observe that CFO is around several Hertz in this experiment. With $f_c = 33$ kHz, a CFO of 3 Hz induced by Doppler shift translates to a moving speed of 0.13 m/s (or 0.26 knots).

If we demodulate the OFDM signal from one receive-element, we observe bit error rate (BER) varying between 10^{-3} and 10^{-2}. However, if we combine signals from four or more elements, we observe no bit errors (BER=0). The scatter diagrams after combining four elements and all twelve elements are shown in Figs. 6 and 7, respectively.

IV. Conclusions and Future Work

In this paper, we developed a pilot-tone based ZP-OFDM receiver, where CFO compensation, channel estimation, and data demodulation are carried out on the basis of each OFDM block. Since it does not rely on any channel dependence across OFDM blocks, this receiver design is appealing for application with short data bursts, and/or fast varying channels. The experimental results are encouraging, with no bit errors when MRC is applied on four or more receive-elements.

Future work includes

- advanced receiver design, where channel estimation and data demodulation are *iteratively* coupled. This can further improve the BER performance and also reduce the number of pilots needed.
- better modeling of channel variation. Now we have modeled the channel variation within each OFDM block by one CFO variable. We could use multiple CFOs for different portions of the channel taps, or, employ the general basis expansion model (BEM) for doubly-selective channels [7].
- more experiments with different UWA channels. The current CFO estimate roughly corresponds to a moving platform of 0.26 knots. Much faster channel variation could occur when the platform moves on the order of several knots.

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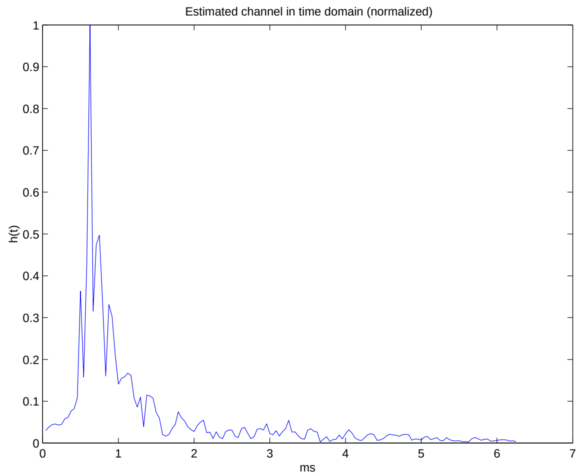

Fig. 4. The estimated channel (non-zero support roughly 6.25 ms)

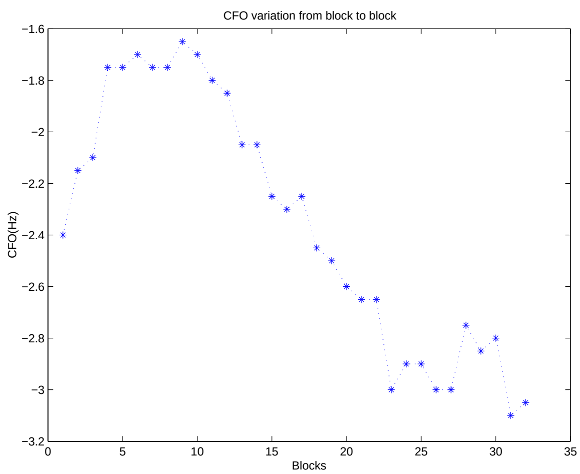

Fig. 5. CFO variation from block to block on element 6.

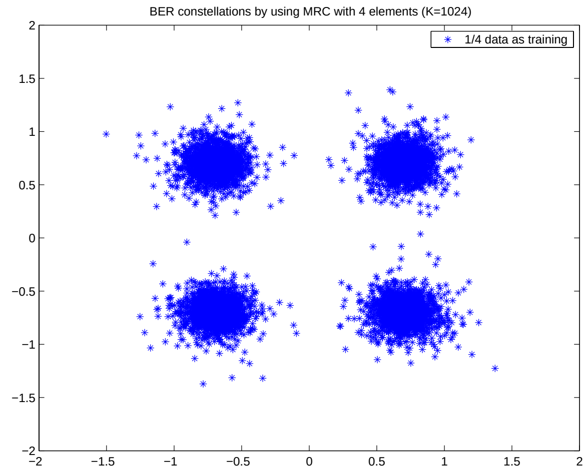

Fig. 6. MRC with elements 9,10,11,12 (K=1024)

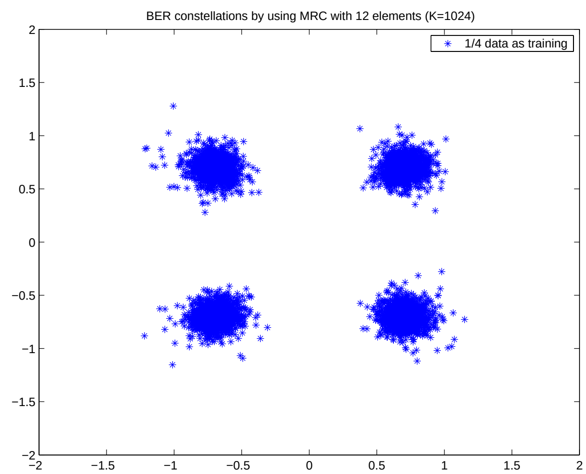

Fig. 7. MRC with 12 elements (K=1024)

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