

Progressive Inter-carrier Interference Equalization for OFDM Transmission over Time-varying Underwater Acoustic Channels

Jianzhong Huang*, Shengli Zhou*, Jie Huang*, Christian R. Berger[†], and Peter Willett*

*Department of Electrical and Computer Engineering, University of Connecticut, Storrs, CT 06269, USA

[†]Dept. of Electrical and Computer Engineering, Carnegie Mellon University, Pittsburgh, PA 15213, USA

Abstract—In this paper we propose a progressive receiver for orthogonal-frequency-division-multiplexing (OFDM) transmission over time-varying underwater acoustic (UWA) channels. The progressive receiver is in nature an iterative receiver. However, it distinguishes itself from existing iterative receivers in that the system model for channel estimation and data detection is itself continually updated during the iterations. When the decoding in the current iteration is not successful, the receiver increases the span of the inter-carrier interference (ICI) in the model and utilizes the currently available soft information from the decoder to assist the next iteration which deals with a channel with larger Doppler spread. Numerical simulation and experimental data collected from the SPACE08 experiment show that the proposed receiver can self adapt to channel variations, enjoying low complexity in calm channel conditions while maintaining excellent performance in tough channel conditions.

I. INTRODUCTION

Recently, multicarrier modulation in the form of orthogonal-frequency-division-multiplexing (OFDM) has been actively pursued for underwater acoustic communications; see performance results based on data recorded from various field experiments in [1]–[12]. Different receivers are built based on different assumptions on the underlying channel models. Specifically, the receivers in [1]–[8] assume that inter-carrier interference (ICI) can be neglected after proper resampling and Doppler shift compensation, while the receivers in [10]–[12] explicitly deal with ICI. Ignoring the ICI, the receivers in [1]–[7] have low complexity on channel estimation and data detection, however, their performance degrades quickly in channels with large Doppler spreading. Explicitly accounting for the ICI, the receiver in [12] achieves robust performance in tough channel conditions. However, it requires a significant pilot overhead for channel estimation, and the receiver complexity is considerably higher than its ICI-ignorant counterparts.

In practice, underwater acoustic (UWA) channels are fast time-varying due to environmental variations such as wind speed, wave height, and the motion of the transceiver platforms. So far, there is no commonly-agreed model for UWA channels. One important area of research is to investigate how environmental factors affect the performance of a particular receiver structure, see e.g., [13], [14].

This work was supported by the ONR grants N00014-07-1-0805 (YIP), N00014-09-1-0704 (PECASE) and the NSF grant CNS-0721834.

In this paper, we propose an OFDM receiver that can adapt to different channel conditions in an automatic fashion. In nature, the proposed progressive receiver is an iterative receiver. However, it distinguishes itself from existing iterative receivers, e.g., [15], [16], in that the system model itself for channel estimation and data detection keeps being updated during the iterations. When the decoding in the current iteration is not successful, the receiver increases the span of the ICI in the system model and in the next iteration utilizes the available soft information from the channel decoder to deal with a channel with a larger Doppler spread. For channel estimation, we use compressive sensing techniques to exploit the sparse nature of UWA channels [12], incorporating the soft information from the decoder as well. The ICI mitigation problem in the frequency-domain is equivalent to that of inter-symbol interference (ISI) equalization in the time-domain, but with time-varying ISI coefficients. In this paper, we adopt the minimum-mean-squared-error (MMSE) equalizer from [15], [17] and the Markov Chain Monte Carlo (MCMC) equalizer from [18]–[20], both of which can effectively utilize *a priori* information for data detection.

We conduct extensive tests on the progressive receiver using both simulations and experimental data from the SPACE08 experiment, conducted off the coast of Martha's Vineyard, MA, from Oct. 14 to Nov. 1, 2008. The reported experimental results include eight consecutive days, with two storm cycles. The progressive receiver achieves consistent performance despite large channel variations. In calm channel conditions, it has as low complexity as the ICI-ignorant receiver in [1]. In tough channel conditions, it maintains excellent performance, but avoids the drawback of the ICI-aware receiver in [12] of large pilot overhead. Compared with an iterative receiver that uses a fixed ICI model, the proposed progressive receiver has comparable performance but much lower complexity.

The rest of this paper is organized as follows. Section II presents the system model and Section III describes the proposed progressive receiver in details. Section IV presents the simulation results and Section V contains the experimental results. Section VI contains the conclusions.

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. $\mathbf{I}_N$ stands for an identity matrix with size N .

II. SYSTEM MODEL

Zero-padded (ZP) OFDM is used as in [1], [12]. Let T denote the OFDM symbol duration and T_g the guard interval. The total OFDM block duration is $T' = T + T_g$ and the subcarrier spacing is $1/T$. The k th subcarrier is at frequency

$$f_k = f_c + k/T, \quad k = -K/2, \dots, K/2 - 1, \quad (1)$$

where f_c is the carrier frequency and K subcarriers are used so that the bandwidth is $B = K/T$. Let $s[k]$ denote the information symbol to be transmitted on the k th subcarrier. The non-overlapping sets of active subcarriers $\mathcal{S}_A$ and null subcarriers $\mathcal{S}_N$ satisfy $\mathcal{S}_A \cup \mathcal{S}_N = \{-K/2, \dots, K/2 - 1\}$. The transmitted passband signal is

$$\tilde{x}(t) = \text{Re} \left\{ \left[\sum_{k \in \mathcal{S}_A} s[k] e^{j2\pi \frac{k}{T} t} q(t) \right] e^{j2\pi f_c t} \right\}, \quad t \in [0, T'] \quad (2)$$

where $q(t)$ is the pulse shaping filter. In this paper we use a rectangular pulse shaping filter

$$q(t) = \begin{cases} 1, & t \in [0, T], \\ 0, & \text{otherwise.} \end{cases} \quad (3)$$

Assume that the UWA multipath channel consists of N_p discrete paths [12]

$$h(\tau, t) = \sum_{p=1}^{N_p} A_p(t) \delta(\tau - \tau_p(t)) \quad (4)$$

where $A_p(t)$ and $\tau_p(t)$ are the amplitude and delay of the p th path. Within the block duration T' , we assume that i) the path amplitudes do not change $A_p(t) \approx A_p$, and ii) the path delays can be approximated by

$$\tau_p(t) = \tau_p - a_p t, \quad (5)$$

where τ_p is the delay at the start of the OFDM block and a_p is the Doppler rate corresponding to the p th path. As such, the channel model simplifies to

$$h(\tau, t) = \sum_{p=1}^{N_p} A_p \delta(\tau - (\tau_p - a_p t)), \quad (6)$$

and the passband signal at the receiver is

$$\tilde{y}(t) = \sum_{p=1}^{N_p} A_p \tilde{x}((1 + a_p)t - \tau_p) + \tilde{v}(t), \quad (7)$$

where $\tilde{v}(t)$ is the additive noise.

As in [1], [12], the receiver applies a resampling operation to remove the dominant Doppler effect, which leads to a resampled passband signal $\tilde{z}(t) = \tilde{y}(t/(1 + \hat{a}))$, where $1 + \hat{a}$ is the resampling factor. Let $z(t)$ denote the baseband signal corresponding to $\tilde{z}(t)$. After the Doppler shift compensation $e^{-j2\pi\epsilon t}$ (see [1], [12] for details on how to estimate ϵ), the FFT output on the m th subcarrier is

$$z[m] = \frac{1}{T} \int_0^{T+T_g} z(t) e^{-j2\pi\epsilon t} e^{-j2\pi \frac{m}{T} t} dt. \quad (8)$$

Plugging in $z(t)$ and carrying out the integration, we obtain:

$$z[m] = \sum_{p=1}^{N_p} \left(A'_p e^{-j2\pi(f_m + \epsilon)\tau'_p} \left(\sum_{k \in \mathcal{S}_A} \varrho_{m,k}^{(p)} s[k] \right) \right) + v[m], \quad (9)$$

where $v[m]$ is the additive noise and

$$b_p = \frac{a_p - \hat{a}}{1 + \hat{a}}, \quad A'_p = \frac{A_p}{1 + b_p}, \quad \tau'_p = \frac{\tau_p}{1 + b_p}, \quad (10)$$

$$\beta_{m,k}^{(p)} = (k - m) + \frac{b_p f_m - \epsilon}{1 + b_p} T, \quad (11)$$

$$\varrho_{m,k}^{(p)} = \frac{\sin(\pi \beta_{m,k}^{(p)})}{\pi \beta_{m,k}^{(p)}} e^{j\pi \beta_{m,k}^{(p)}}. \quad (12)$$

With the definition of

$$H_{m,k} = \sum_{p=1}^{N_p} A'_p e^{-j2\pi(f_m + \epsilon)\tau'_p} \varrho_{m,k}^{(p)} \quad (13)$$

we can rewrite the input-output relationship as

$$z[m] = \sum_{k \in \mathcal{S}_A} H_{m,k} s[k] + v[m]. \quad (14)$$

Clearly $H_{m,k}$ is the ICI coefficient that determines how the symbol transmitted on the k th subcarrier contributes to the output on the m th subcarrier. Using a matrix-vector notation, we can rewrite (14) as

$$\underbrace{\begin{pmatrix} z[-\frac{K}{2}] \\ \vdots \\ z[\frac{K}{2} - 1] \end{pmatrix}}_{:=\mathbf{z}} = \underbrace{\begin{pmatrix} v[-\frac{K}{2}] \\ \vdots \\ v[\frac{K}{2} - 1] \end{pmatrix}}_{:=\mathbf{v}} + \underbrace{\begin{pmatrix} H_{-\frac{K}{2}, -\frac{K}{2}} & \cdots & H_{-\frac{K}{2}, \frac{K}{2}-1} \\ \vdots & \ddots & \vdots \\ H_{\frac{K}{2}-1, -\frac{K}{2}} & \cdots & H_{\frac{K}{2}-1, \frac{K}{2}-1} \end{pmatrix}}_{:=\mathbf{H}} \underbrace{\begin{pmatrix} s[-\frac{K}{2}] \\ \vdots \\ s[\frac{K}{2} - 1] \end{pmatrix}}_{:=\mathbf{s}} \quad (15)$$

where $\mathbf{z}$, $\mathbf{s}$, and $\mathbf{v}$ collect the FFT outputs, the transmitted data symbols, and the noise elements across all subcarriers $m = -K/2, -K/2 + 1, \dots, K/2 - 1$, and $\mathbf{H}$ is the channel mixing matrix.

III. THE PROGRESSIVE RECEIVER

Section III-A presents the overview of the proposed progressive receiver, while the key components are specified in Sections III-B – III-E.

A. The Progressive Receiver Structure

The channel models used in the paper are parameterized by a parameter D that specifies the ICI span

$$H_{m,k} \simeq 0, \quad |m - k| > D. \quad (16)$$

In other words, each symbol only affects its D direct neighbors on each side, a reasonable assumption used in many existing works, e.g., [10], [21], [22]. Let $\mathbf{H}_D$ denote the matrix carved

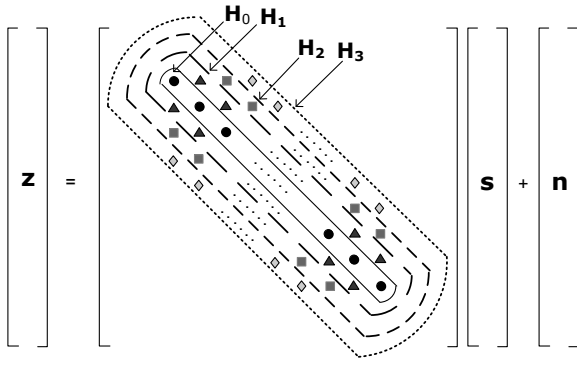


Fig. 1. Equivalent system model with channel matrix $\mathbf{H}_D$.

from $\mathbf{H}$ keeping only the main diagonal and $2D$ off-diagonals, as shown in Fig. 1. The effective system model used for channel estimation and data demodulation is:

$$\begin{aligned} \mathbf{z} &= \mathbf{H}_D \mathbf{s} + (\mathbf{H} - \mathbf{H}_D) \mathbf{s} + \mathbf{v} \\ &= \mathbf{H}_D \mathbf{s} + \mathbf{n}, \end{aligned} \quad (17)$$

where $\mathbf{n} = (\mathbf{H} - \mathbf{H}_D) \mathbf{s} + \mathbf{v}$ is the effective noise. In the proposed progressive receiver, the parameter D increases during each iteration, and hence more severe ICI can be addressed as the receiver processing proceeds to deal with channels with large Doppler spread.

Fig. 2 depicts the progressive receiver structure, which consists of the following steps.

- *Step 1: Pre-processing.* For each received OFDM block, the receiver applies the pre-processing operation to remove the dominant Doppler effect [1]. Set $D = 0$.
- *Step 2: Channel estimation.* Estimate the equivalent channel mixing matrix $\mathbf{H}_D$ based on the assumed channel model given in (17).
- *Step 3: Noise variance estimation.* After channel estimation, the variance of the effective noise $\mathbf{n}$ is computed. This quantity is needed for ICI equalization.
- *Step 4: ICI equalization.* By using the estimated channel matrix $\mathbf{H}_D$, the equivalent noise variance $\hat{N}_0$, and *a priori* information from the nonbinary LDPC decoder in the previous iteration, the ICI equalizer generates soft output on the reliability of the data symbols.
- *Step 5: Nonbinary LDPC decoding.* The nonbinary LDPC decoder yields the decoded information symbols and the soft information that can be used for channel estimation and ICI equalization [4]. During the decoding process, the decoder will declare success if all the parity check conditions are satisfied.
- *Step 6: Iteration among steps 2 to 5.* Increase D in the system model, and increase the Doppler spread of the channel to be estimated. Feed back the soft information to the channel estimator and the ICI equalizer. Iteration stops when the decoder declares a success, or when D reaches a pre-specified number $D_{\max}$.

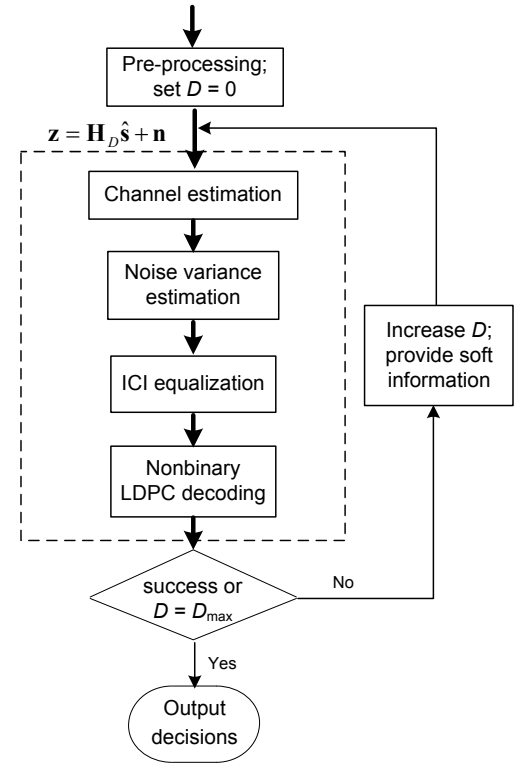


Fig. 2. Progressive receiver structure.

B. Sparse Channel Estimation

The inputs to the channel estimator are the observations in $\mathbf{z}$, the pilot symbols, and the *a posteriori* probabilities (APP) on the information symbols from the LDPC decoder. Different strategies on how to use APP can be found in [8]. Here, we use the soft feedback strategy, which produces an MMSE estimate on the information symbols as

$$\bar{s}[k] = \sum_{i=1}^M P_{\text{app}}(s[k] = \alpha_i) \alpha_i, \quad \forall k \in \mathcal{S}_D \quad (18)$$

where M is the constellation size, α_i is the i th constellation point, $P_{\text{app}}(\cdot)$ is the APP from LDPC decoder, and $\mathcal{S}_D$ is the set of data subcarriers in $\mathcal{S}_A$. Hence, an estimate of $\mathbf{s}$ can be formed from

$$\hat{s}[k] = \begin{cases} s[k], & k \in \mathcal{S}_P, \\ 0, & k \in \mathcal{S}_N, \\ \bar{s}[k], & k \in \mathcal{S}_D \end{cases} \quad (19)$$

where $\mathcal{S}_P$ is the set of pilot subcarriers from $\mathcal{S}_A$.

Despite having $K(2D + 1)$ non-zero entries, the matrix $\mathbf{H}_D$ is determined by N_p triplets of (A'_p, b_p, τ'_p) as described in (13). Since UWA channels are sparse in nature, it is possible that these N_p paths can be identified based on only a limited number of measurements via advanced methods, such as compressed sensing. We will use the compressive sensing approach in [12] as the building block.

The compressed sensing based channel estimator [12] tries to identify the discrete paths from an overcomplete dictionary.

On the delay and Doppler plane, a set of uniformly spaced points can be constructed from

$$\tau' \in \left\{ 0, \frac{T}{\lambda K}, \frac{2T}{\lambda K}, \dots, \frac{(N_\tau - 1)T}{\lambda K} \right\}, \quad (20)$$

$$b^D \in \{-b_{\max}^D, -b_{\max}^D + \Delta b, \dots, b_{\max}^D\}, \quad (21)$$

where the time resolution is chosen as a fraction, λ , of the baseband sampling time T/K , leading to $N_\tau = \lambda K T_g / T$ tentative delays. For the Doppler rates, we assume that they are spread around zero after pre-processing, and Δb is the Doppler resolution. $b_{\max}^D$ can be chosen based on the assumed Doppler spread for the current system model. There are

$$2b_{\max}^D / (\Delta b) + 1 = N_b^D, \quad (22)$$

tentative Doppler rates, leading to a delay-Doppler dictionary of size $N_\tau N_b^D$.

During the iterations, the progressive receiver keeps the delay dictionary unchanged, but enlarges the Doppler dictionary size by increasing $b_{\max}^D$ as D increases. This way, channels with large Doppler spread can be handled.

C. Noise Variance Estimation

The variance of the equivalent noise is estimated by measuring the energy on the null subcarriers after each channel estimation iteration. In particular, it is estimated as

$$\hat{N}_0 = E \left[\left| z_m - \sum_{k=-D}^{k=D} \hat{H}_{m,m-k} \hat{s}[m-k] \right|^2 \right], \quad (23)$$

where the expectation is carried out on the null subcarriers in $\mathcal{S}_N$ that are within the signal band. For $D = 0$, all the ICI terms are treated as the additive noise, and $\hat{N}_0$ in (23) measures the energy on the null subcarriers [1]. A simple noise-whitening approach as in [9] will be used when $D = 0$, as ICI is the dominant source for the noise. As the iteration goes on with $D > 0$, the split over from the neighboring subcarriers to the null subcarriers is extracted as in (23), and no whitening is applied. As less ICI is viewed as additive noise, the effective signal-to-noise ratio (SNR) is expected to increase.

D. ICI Equalization

Data detection follows channel estimation. Based on (17), the observation on the m th subcarrier can be written as

$$z[m] = \sum_{k=-D}^{k=D} \hat{H}_{m,m-k} s[m-k] + n[m], \quad (24)$$

which is in the convolution form but with coefficients that are changing from symbol to symbol. Hence, the ICI mitigation problem in the frequency domain is equivalent to an inter-symbol-interference (ISI) equalization problem in the time domain with time-varying channel taps. Existing methods for ISI equalization can be used for ICI equalization. Here we test the MMSE equalizer from [17] and the Markov Chain Monte Carlo (MCMC) equalizer from [18]–[20], both of which can effectively incorporate soft information from the channel decoder.

1) *MMSE Equalizer*: An MMSE equalizer with *a priori* information from [17] is used here for ICI mitigation. The channel decoder feeds back the extrinsic information, denoted as λ^e , based on which the mean $\mu[k] := E(s[k])$ and the variance $\sigma^2[k] := \text{Cov}(s[k], s[k])$ can be calculated. The *a priori* information is assumed to be independent.

Based on (24), we collect the FFT outputs that are directly related to $s[m]$ as in (25) on the top of the next page. Now, let $\hat{\mathbf{h}}_m$ be the $(D+1)$ th column of $\hat{\mathbf{H}}_m$:

$$\hat{\mathbf{h}}_m := [\hat{H}_{m-D,m}, \dots, \hat{H}_{m+D,m}]^T, \quad (26)$$

which is of length $2D+1$. Define $\hat{\mathbf{H}}_m^-$ to be $\hat{\mathbf{H}}_m$ with $\hat{\mathbf{h}}_m$ removed, and $\mathbf{s}_m^-$ to be $\mathbf{s}_m$ with $s[m]$ removed. As such, we rewrite (25) as

$$\mathbf{z}_m = \hat{\mathbf{h}}_m s[m] + \hat{\mathbf{H}}_m^- \mathbf{s}_m^- + \mathbf{n}_m. \quad (27)$$

Based on the extrinsic information, define the mean and covariance of $\mathbf{s}_m^-$ as:

$$\boldsymbol{\mu}_m = E(\mathbf{s}_m^-), \quad \boldsymbol{\Sigma}_m = \text{Cov}(\mathbf{s}_m^-, \mathbf{s}_m^-). \quad (28)$$

Note that no *a priori* information of $s[m]$ is included for the MMSE estimation of it, hence, $E(s[m]) = 0$ and $\sigma^2[m] = E_s$, where E_s is the average symbol energy. Further, the equivalent noise $\mathbf{n}$ is assumed Gaussian with a covariance matrix $\hat{N}_0 \mathbf{I}$. The MMSE estimator for $s[m]$ based on (27) is then

$$\begin{aligned} \hat{s}[m] = & E_s \hat{\mathbf{h}}_m^H \left[E_s \hat{\mathbf{h}}_m \hat{\mathbf{h}}_m^H + \hat{N}_0 \mathbf{I} + \hat{\mathbf{H}}_m^- \boldsymbol{\Sigma}_m (\hat{\mathbf{H}}_m^-)^H \right]^{-1} \\ & \times (\mathbf{z}_m - \hat{\mathbf{H}}_m^- \boldsymbol{\mu}_m) \end{aligned} \quad (29)$$

Assuming that $\hat{s}[m]$ is Gaussian distributed, the probabilities $P(\hat{s}[m] | s[m] = \alpha_i)$, $i = 0, \dots, M-1$, can be computed [17]. These probabilities are passed to the nonbinary LDPC decoder.

Since matrix inversion is involved in (29), the complexity of the MMSE equalizer is at least quadratic with D . However, D is usually small, and the complexity is not sensitive to the constellation size M .

2) *MCMC Equalizer*: The MCMC method has been successfully applied for MIMO detection in [18], [19] and ISI equalization in [20]. Note that both bit-wise and group-wise MCMC detectors have been proposed [20]. In our case with nonbinary LDPC codes, the size of the Galois field is matched to the constellation size, with each entry in the finite field directly mapped to a constellation point [4]. Since there is no symbol-to-bit and bit-to-symbol conversion, we now present the MCMC detector in a symbol-wise notation.

The procedure of the MCMC sampling can be found in [18]–[20]. Assume that Ω importance samples are obtained after removing the redundant samples as in [18]. The log-likelihood-ratio vector (LLRV) for $s[m]$ is defined as $\boldsymbol{\lambda}[m] = [\lambda_0[m], \lambda_1[m], \dots, \lambda_{M-1}[m]]^T$, where

$$\lambda_i[m] = \ln \frac{P(s[m] = \alpha_i | \mathbf{z}, \boldsymbol{\lambda}^e)}{P(s[m] = 0 | \mathbf{z}, \boldsymbol{\lambda}^e)}. \quad (30)$$

$$\underbrace{\begin{pmatrix} z[m-D] \\ \vdots \\ z[m+D] \end{pmatrix}}_{:=\mathbf{z}_m} = \underbrace{\begin{pmatrix} \hat{H}_{m-D,m-2D} & \cdots & \hat{H}_{m-D,m} & & \\ & \ddots & \vdots & \ddots & \\ & & \hat{H}_{m+D,m} & \cdots & \hat{H}_{m+D,m+2D} \end{pmatrix}}_{:=\hat{\mathbf{H}}_m} \underbrace{\begin{pmatrix} s[m-2D] \\ \vdots \\ s[m+2D] \end{pmatrix}}_{:=\mathbf{s}_m} + \underbrace{\begin{pmatrix} n[m-D] \\ \vdots \\ n[m+D] \end{pmatrix}}_{:=\mathbf{n}_m} \quad (25)$$

With the importance samples, the LLRV is computed as

$$\lambda_i[m] = \ln \frac{\sum_{n=1}^{\Omega} P^{(n)}(\mathbf{z}_m | \mathbf{s}_m^-, s[m] = \alpha_i) P(\boldsymbol{\lambda}^e)}{\sum_{n=1}^{\Omega} P^{(n)}(\mathbf{z}_m | \mathbf{s}_m^-, s[m] = 0) P(\boldsymbol{\lambda}^e)}, \quad (31)$$

where $P^{(n)}(\mathbf{z}_m | \mathbf{s}_m^-, s[m] = \alpha_i)$ is calculated based on the n th sample in Ω , and can be simplified as

$$P^{(n)}(\mathbf{z}_m | \mathbf{s}_m^-, s[m] = \alpha_i) \sim \exp \left\{ -\frac{1}{\hat{N}_0} \left\| \mathbf{z}_m - \hat{\mathbf{h}}_m \alpha_i - \hat{\mathbf{H}}_m^- \mathbf{s}_m^- \right\|^2 \right\}. \quad (32)$$

The computational complexity of the MCMC equalizer is determined by the sample size Ω and the constellation size M . Processing the probabilities in the log-domain, the complexity of calculating (32) is $6D + 6$ complex multiplications and $6D + 1$ complex additions. Hence, the complexity of drawing samples for each symbol is roughly $\Omega M(6D + 6)$ multiplications and $\Omega M(6D + 1)$ additions. Computing the output LLRVs as in (31) needs $\Omega M(6D + 6)$ multiplications and $\Omega M(6D + 1)$ additions. Therefore, the total complexity for the MCMC equalizer is $2\Omega M K(6D + 6)$ complex multiplications and $2\Omega M K(6D + 1)$ complex additions. The complexity is linear with D , but is sensitive to the constellation size M .

E. Nonbinary LDPC Decoding

Nonbinary LDPC decoding as in [4] is performed based on the equalizer outputs. The decoder outputs the decoded information symbols and the updated a posteriori and extrinsic probabilities, which are used for the next round of channel estimation and equalization, respectively. During the decoding process, if all the parity check conditions are satisfied, the decoder declares success.

IV. SIMULATION RESULTS

The system parameters are the same as used in the SPACE08 experiment, with bandwidth $B = 9.77$ kHz, symbol duration $T = 104.86$ ms, and guard time $T_g = 24.6$ ms. Out of $K = 1024$ subcarriers, there are 256 pilot subcarriers, 96 null subcarriers, and 672 data subcarriers. With 16-QAM constellation and rate 1/2 nonbinary LDPC coding, the data rate is $R = \frac{1}{2} \cdot 672 \cdot \log_2 16 / (T + T_g) = 10.4$ kb/s.

We generate sparse channels with $N_p = 15$ discrete paths, where the inter-arrival times are distributed exponentially with inter-arrival mean of 1 ms, leading to a total average delay spread of 15 ms. The amplitudes are Rayleigh distributed with the average power decreasing exponentially with delay.

Each path has a separate Doppler rate, which is drawn from a uniform distribution with standard deviation of σ_v m/s. We choose a zero-mean Doppler distribution, as a non-zero mean could be removed through the resampling operation.

Fig. 3 plots the block error rate (BLER) of the progressive receiver, using one receive-phone. The results are averaged over at least 1000 channel realizations or when 50 block errors are detected. The related parameters are $D_{\max} = 3$, the time resolution $\lambda = 2$, the Doppler resolution $\Delta b = 4.5 \cdot 10^{-3}$. During the iteration process, $N_b^D = 7, 11$ and 15 for $D = 1, 2$ and 3, respectively.

Fig. 3 shows that the performance difference between MMSE and MCMC equalization is negligible in the progressive receiver setup, while both of them achieve significant performance improvement relative to the ICI-ignorant receiver. When the channel conditions are calm, see Fig. 3(a), more than 90% OFDM blocks can be decoded in the first round, i.e., using the ICI-ignorant receiver, when the operating SNR is above 11 dB. The block success rate increases to 97% when the progressive receiver reaches $D = 1$. When the channel conditions deteriorate, as in Fig. 3(b), the ICI-ignorant receiver has very poor performance, while the progressive receiver has only small performance loss relative to that in the calm channels. Note that only fewer than 10% of the OFDM blocks need $D = 3$ for SNR larger than 11 dB, which indicates that the progressive receiver enjoys low complexity by gradually increasing D during the iterations.

V. EXPERIMENTAL RESULTS

We use data recorded during the *surface processes and acoustic communications experiment* (SPACE) experiment, conducted off the coast of Martha's Vineyard, MA, from Oct. 14 to Nov. 1, 2008. The water depth was about 15 meters. A transmission occurred every two hours, resulting in 12 recorded files each day. For each transmission, there are 20 OFDM blocks with the parameters specified in the simulation setting. Hence, the data rate is 10.4 kb/s, with 16-QAM and rate-1/2 nonbinary LDPC coding over a bandwidth of 9.77 kHz.

We report performance results for Julian dates 295 – 302 (Oct. 21 - 28) and consider three receivers, labeled as S1, S3, and S5, which were 60 m, 200 m, and 1,000 m from the transmitter, respectively. The Doppler resolution and the dictionary size are the same as used in the simulation.

A. Performance overview

Tables I and II report the number of decoded blocks across the eight days considered using the MMSE and MCMC equalizers, respectively, with different number of phones combined.

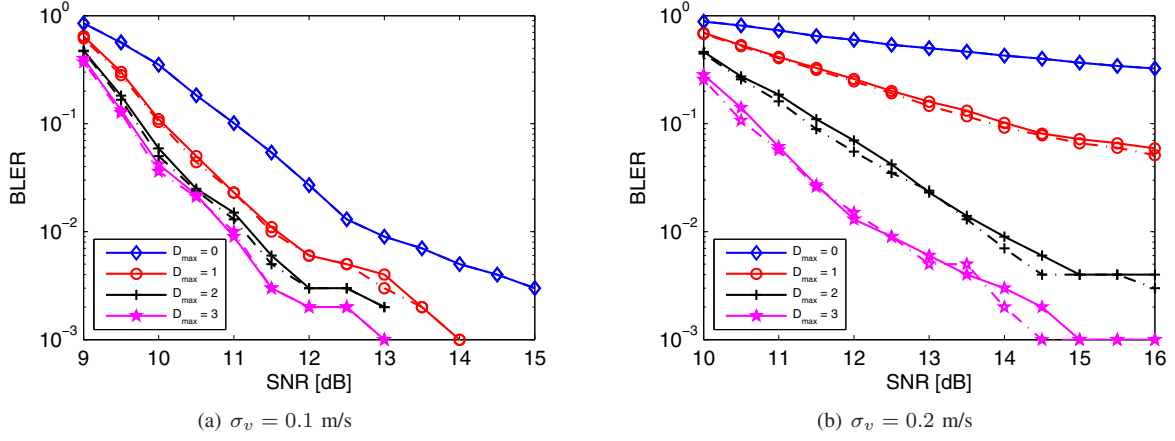


Fig. 3. Simulated performance for the progressive receiver with different $D_{\max}$. Dashed lines: MCMC equalizer; solid lines: MMSE equalizer.

TABLE I
THE NUMBER OF OFDM BLOCKS DECODED FOR JULIAN DATES 295-302; 16-QAM, RATE 1/2 CODING, MMSE BASED ICI EQUALIZATION

System	S1 (60 m)				S3 (200 m)				S5 (1000 m)			
# of Phones	$D = 0$	$D = 1$	$D = 2$	$D = 3$	$D = 0$	$D = 1$	$D = 2$	$D = 3$	$D = 0$	$D = 1$	$D = 2$	$D = 3$
1	382	130	97	74	411	88	51	44	791	51	15	11
2	829	166	134	133	865	96	49	47	1205	58	41	19
3	1210	135	92	50	1170	102	69	40	1421	43	17	15
4	1408	75	39	19	1427	87	46	23	1491	23	10	10
5	1490	38	19	8	1552	41	23	9	1518	17	14	14
6	1524	17	11	4	1596	27	8	1	1532	22	16	7
7	1536	12	6	4	1624	7	5	4	1547	19	13	3
8	1541	9	6	3	1629	8	3	0	1555	23	8	2
9	1544	9	4	3	1636	2	2	0	1563	18	5	1
10	1547	7	4	2	1638	2	0	0	1574	10	2	1
11	1549	6	3	2	1640	0	0	0	1575	8	6	0
12	1551	4	3	2	1640	0	0	0	1577	11	2	1

TABLE II
THE NUMBER OF OFDM BLOCKS DECODED FOR JULIAN DATES 295-302; 16-QAM, RATE 1/2 CODING, MCMC BASED ICI EQUALIZATION

System	S1 (60 m)				S3 (200 m)				S5 (1000 m)			
# of Phones	$D = 0$	$D = 1$	$D = 2$	$D = 3$	$D = 0$	$D = 1$	$D = 2$	$D = 3$	$D = 0$	$D = 1$	$D = 2$	$D = 3$
1	382	137	108	78	411	93	58	44	791	51	17	14
2	829	175	159	116	865	102	54	56	1205	62	42	26
3	1210	148	90	52	1170	105	77	46	1421	43	22	11
4	1408	77	42	17	1427	93	48	24	1491	24	14	11
5	1490	39	20	7	1552	46	21	9	1518	18	22	12
6	1524	17	13	3	1596	28	7	4	1532	25	19	1
7	1536	13	6	3	1624	7	6	2	1547	23	12	1
8	1541	10	5	4	1629	8	3	0	1555	25	8	2
9	1544	9	5	2	1636	3	1	0	1563	20	4	1
10	1547	6	5	2	1638	2	0	0	1574	10	4	0
11	1549	6	3	2	1640	0	0	0	1575	9	5	0
12	1551	4	3	2	1640	0	0	0	1577	12	1	0

Since some recorded files are corrupted, there are a total of 1560, 1640 and 1600 blocks processed for S1, S3 and S5, respectively. Comparing Tables I and II, we see that the MCMC equalizer performs slightly better when only a small number of hydrophones are combined, and the gap closes when more hydrophones are available. Combining 12 hydrophones, all blocks in S1 and S3 are decoded correctly using the progressive receiver when it reaches $D = 3$. There are 9 (with MMSE) or 10 (with MCMC) blocks that cannot be decoded in S5. Since the performance difference between MMSE and MCMC is small, in the following we use the

MMSE results for illustration.

Fig. 4 shows the block success rate averaged over the eight consecutive days using the proposed progressive receiver with the MMSE equalizer. At short (S1) to medium (S3) ranges, we expect rich multipath and significant Doppler variation due to the geometry. When the number of hydrophones is small, the performance of the ICI-ignorant receiver ($D = 0$) is limited, and many more OFDM symbols can be decoded by applying the progressive procedure, with a larger D . When the number of hydrophones is large, the ICI-ignorant receiver already achieves excellent results for all the blocks. Checking

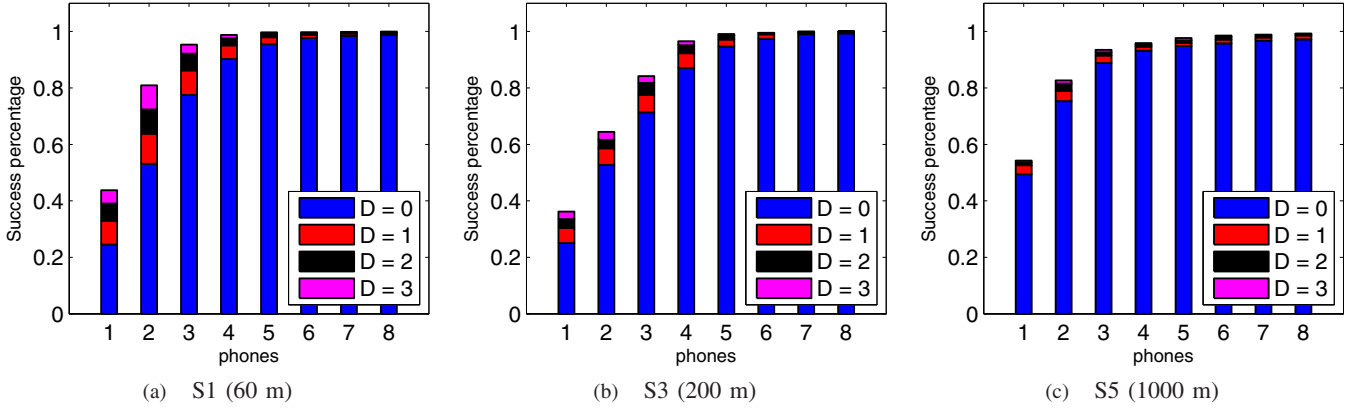


Fig. 4. The block success percentage averaged over Julian dates 295-302, SPACE08, MMSE based ICI equalization.

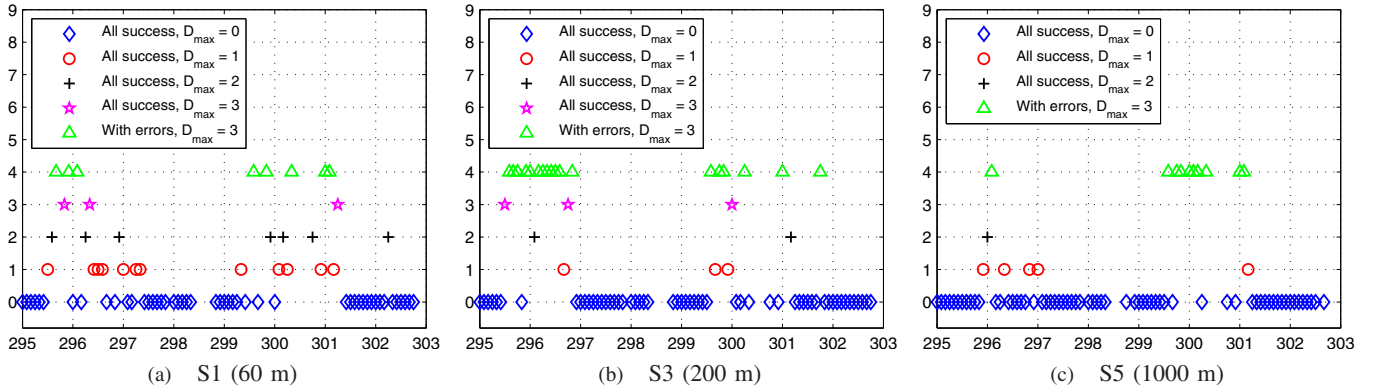


Fig. 5. Success level for each transmission of 20 OFDM blocks; four hydrophones. Markers are placed at convenient heights for illustration purposes only.

the results using four hydrophones, about 90% OFDM blocks can be decoded at the $D = 0$ stage, and the success rate increases to 95% when $D = 1$, and up to 98.8% when $D = 3$.

For S5, we see similar trends as S1 and S3, but the gap between the ICI-ignorant and progressive receivers gets small. When four hydrophones are combined, over 93% blocks can be decoded by ignoring the ICI, and the success rate increases to 96% when the progressive receiver reaches $D = 3$.

B. Environmental impact

Using four hydrophones for combining, Table I shows that there are 19 out of 1560 blocks with decoding errors in S1, 57 out of 1640 blocks with decoding errors in S3, and 66 out of 1600 blocks with decoding errors in S5, for the progressive receiver with $D_{\max} = 3$. Fig. 5 illustrates the success level of each transmission of 20 OFDM blocks across the 8-day period. Each day, we have about 12 files recorded (a few files are corrupted). “All success” means that all 20 blocks in that file can be decoded, while “With errors” means that some blocks cannot be decoded out of 20 blocks in the file.

The average wind speed and wave height are shown in Fig. 6. We can observe some correlations between Fig. 5 and Fig. 6. There are two periods that the progressive receiver with $D > 0$ are used: Julian dates 296-297 and Julian dates 300-301, during which the wind speed and the wave height

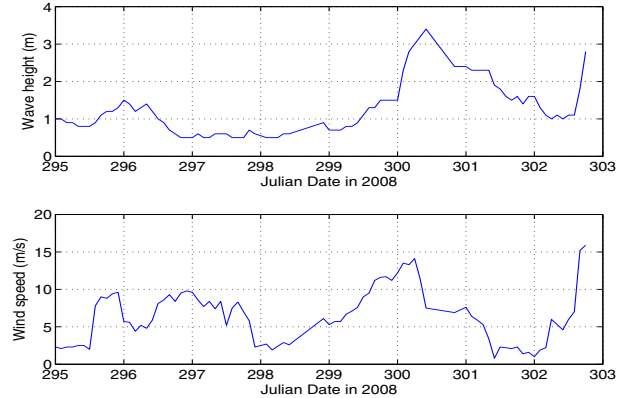


Fig. 6. Average wave height and wind speed for selected days; SPACE08

are high. For the rest of days, the ICI-ignorant receivers can decode all the blocks. Fig. 5 confirms that the progressive receiver can self adapt to channel conditions, maintaining both good performance and low complexity.

C. Progressive versus Iterative ICI-aware receivers

In Fig. 7, we compare the performance between the proposed progressive receiver and an iterative ICI-aware receiver

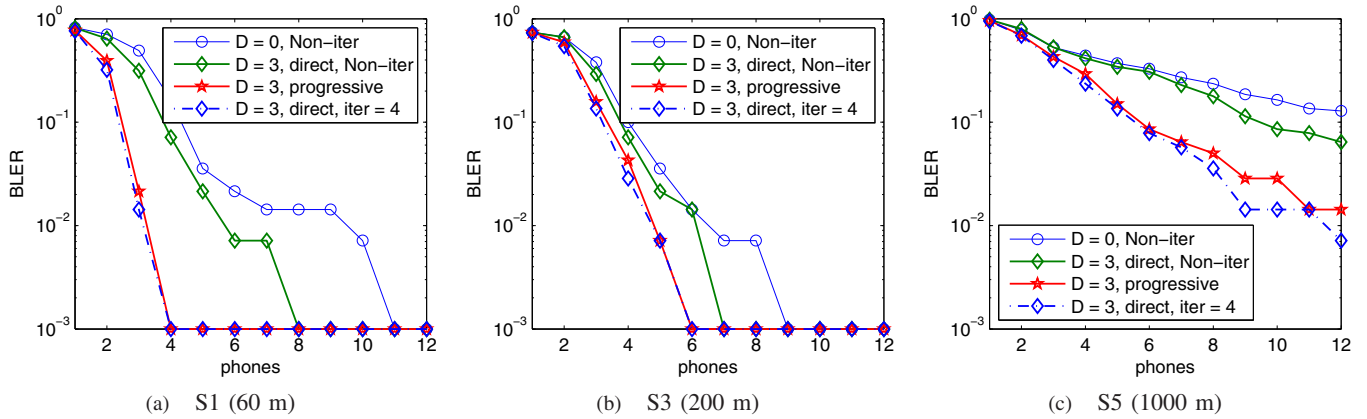


Fig. 7. Performance comparisons between progressive and iterative ICI-aware receivers; Julian date 300.

that fixes the channel model at $D = 3$ but iterates several times, over channels with large Doppler spread (Julian Date 300). Obviously, the latter receiver has much higher complexity. After the first iteration, we see that the ICI-aware receiver outperforms the ICI-ignorant receiver. As the iterations continue, the progressive receiver catches up the iterative ICI-aware receiver, and the performance difference is negligible. Hence, the progressive receiver collects the performance benefits as the iterative ICI-aware receiver, but enjoys much lower complexity in various channel conditions.

VI. CONCLUSIONS

In this paper we developed a progressive receiver for OFDM transmission over time-varying underwater acoustic channels. During the iterations, it updates the “system model” to account for channels with large Doppler spreads, while fully utilizing soft information from the previous iteration for enhanced sparse channel estimation and inter-carrier interference equalization. Extensive tests based on experimental data showed that the proposed receiver enjoys low complexity in calm channel conditions while maintaining excellent performance even when the channel deteriorates. Adapting to channel variations without any *a priori* information, the proposed receiver is very promising for practical underwater systems.

ACKNOWLEDGEMENT

We thank Dr. James Preisig and his team for conducting the SPACE08 experiment.

REFERENCES

- [1] B. Li, S. Zhou, M. Stojanovic, L. Freitag, and P. Willett, “Multicarrier communication over underwater acoustic channels with nonuniform Doppler shifts,” *IEEE Journal of Oceanic Engineering*, Apr. 2008.
- [2] M. Stojanovic, “Low complexity OFDM detector for underwater channels,” in *Proc. of OCEANS Conf.*, Boston, MA, Sept. 18-21, 2006.
- [3] —, “OFDM for underwater acoustic communications: Adaptive synchronization and sparse channel estimation,” in *Proc. of Intl. Conf. on ASSP*, Las Vegas, NV, Mar. 30 – Apr. 4, 2008.
- [4] J. Huang, S. Zhou, and P. Willett, “Nonbinary LDPC coding for multicarrier underwater acoustic communication,” *IEEE J. Select. Areas Commun.*, vol. 26, no. 9, pp. 1684–1696, Dec. 2008.
- [5] T. Kang and R. A. Iltis, “Iterative carrier frequency offset and channel estimation for underwater acoustic OFDM systems,” *IEEE J. Select. Areas Commun.*, vol. 26, no. 9, pp. 1650–1661, Dec. 2008.
- [6] Y. Emre, V. Kandasamy, T. M. Duman, P. Hursky, and S. Roy, “Multi-input multi-output OFDM for shallow-water UWA communications,” in *Acoustics’08 Conference*, Paris, France, July 2008.
- [7] B. Li, J. Huang, S. Zhou, K. Ball, M. Stojanovic, L. Freitag, and P. Willett, “MIMO-OFDM for high rate underwater acoustic communications,” *IEEE Journal of Oceanic Engineering*, vol. 34, pp. 634–644, Oct. 2009.
- [8] J. Huang, J.-Z. Huang, C. R. Berger, S. Zhou, and P. Willett, “Iterative sparse channel estimation and decoding for underwater MIMO-OFDM,” in *Proc. of OCEANS Conference*, Oct. 26-29, 2009.
- [9] C. R. Berger, S. Zhou, W. Chen, and J. Huang, “A simple and effective noise whitening method for underwater acoustic OFDM,” in *Proc. of OCEANS Conference*, Biloxi, Mississippi, USA, Oct. 26-29, 2009.
- [10] G. Leus and P. van Walree, “Multiband OFDM for covert acoustic communications,” *IEEE Journal on Selected Areas in Communications*, vol. 26, no. 9, pp. 1662–1673, Dec. 2008.
- [11] K. Tu, D. Fertonani, T. Duman, and P. Hursky, “Mitigation of intercarrier interference in OFDM systems over underwater acoustic channels,” in *Acoustics’08 Conference*, Paris, France, July 2008.
- [12] C. R. Berger, S. Zhou, J. Preisig, and P. Willett, “Sparse channel estimation for multicarrier underwater acoustic communication: From subspace methods to compressed sensing,” *IEEE Trans. Signal Processing*, vol. 58, no. 3, pp. 1708 – 1721, 2010.
- [13] A. Song, M. Badiey, H.-C. Song, W. S. Hodgkiss, M. B. Porter, and the KauaiEx Group, “Impact of ocean variability on coherent underwater acoustic communications during the Kauai experiment (KauaiEx),” *The Journal of the Acoustical Society of America*, pp. 856–865, 2008.
- [14] L. Freitag and S. Singh, “Performance of micro-modem PSK signalling under variable conditions during the 2008 RACE and SPACE experiments,” in *Proc. of OCEANS Conf.*, Biloxi, Mississippi, USA, Oct. 2009.
- [15] R. Koetter, A. C. Singer, and M. Tuchler, “Turbo equalization,” *IEEE Signal Processing Magazine*, vol. 21, no. 1, pp. 67–80, Jan. 2004.
- [16] D. N. Liu and M. P. Fitz, “Iterative MAP equalization and decoding in wireless mobile coded OFDM,” *IEEE Trans. Commun.*, July 2009.
- [17] M. Tüchler, A. C. Singer, and R. Koetter, “Minimum mean squared error equalization using a priori information,” *IEEE Trans. Signal Processing*, vol. 50, no. 3, pp. 673–683, 2002.
- [18] B. Farhang-Boroujeny, H. Zhu, and Z. Shi, “Markov chain monte carlo algorithms for CDMA and MIMO communication systems,” *IEEE Trans. Signal Processing*, vol. 54, no. 5, pp. 1896–1909, May 2006.
- [19] R.-R. Chen, R. Peng, B. Farhang-Boroujeny, and A. Ashikhmin, “Approaching MIMO capacity using bit-wise markov chain monte carlo detection,” *IEEE Trans. Commun.*, vol. 58, pp. 423–428, Feb. 2010.
- [20] R. Peng, R.-R. Chen, and B. Farhang-Boroujeny, “Markov Chain Monte Carlo detectors for channels with intersymbol interference,” *IEEE Trans. Signal Processing*, vol. 58, no. 4, pp. 2206–2217, Apr. 2010.
- [21] P. Schniter, “Low-complexity equalization of OFDM in doubly selective channels,” *IEEE Trans. Signal Processing*, pp. 1002–1011, 2004.
- [22] L. Rugini, P. Banelli, and G. Leus, “Simple equalization of time-varying channels for OFDM,” *IEEE Commun. Lett.*, pp. 619–621, July 2005.