ADAPTIVE CHANNEL SHORTENING EQUALIZATION FOR COHERENT OFDM DOUBLY SELECTIVE CHANNELS

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ABSTRACT[#]

In this work we propose a new adaptive channel shortening technique for doubly selective (time-varying frequency-selective) Orthogonal Frequency Division Multiplexing (OFDM) channels. OFDM is considered to be a very bandwidth efficient wireless transmission technique (due to its tightly packed orthogonal sub-carrier structure) for multi-path channels. Transmissions in multipath channels exhibiting long delay spread channels, however, require an excessively long cyclic prefix (CP) to prevent inter-symbol interference (ISI) with OFDM, and thus undesirably reduce the bandwidth and power efficiency of the information transmission. Channel shortening equalization (CSE) is used to shorten the channel delay spread to a certain length (less than the CP length) so that ISI is minimized. Our technique adapts to channel variation and provides a significant bit error rate (BER) improvement in performance over non-adaptive techniques. We show that the adaptive technique improves the demodulated BER by greater than a factor of 10, with diversity order 1, in a mobile long delay spread (frequency-selective) channel, while the non-adaptive method exhibits a BER floor at 10^{-1}.

1. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) is a popular transmission method for highly dispersive wire-line and wireless channels. For OFDM, long delay spread channels require an excessively long cyclic prefix (CP) to prevent inter-symbol interference (ISI), but undesirably reduces the bandwidth and power efficiency of the information transmission (since the CP does not carry additional information content). Using excessive training and blind estimation for long channels can cost extra bandwidth and computational complexity, respectively.

The motivation behind this work is to develop a technique that allows a more efficient CP for time-varying long delay spread channels, and in addition exploits the frequency diversity of coherent OFDM sub-carriers to improve BER performance. There are very few CSE techniques in the literature that do not require excessive channel training (thus consuming excessive overhead), and furthermore there is a strong need for CSE methods that adapt to channel variation for mobile transmitters and/or receivers. In addition, we have not found blind adaptive CSE techniques that can adapt quickly enough for short time-coherence channels.

Some existing CSE techniques require channel knowledge, which reduce bandwidth efficiency by using long training sequences [1]. Blind CSE methods have been introduced to eliminate the bandwidth penalty of training-based methods, such as the subspace method from [2]. However, because the CSE is non-data aided, large recorded sample sets are required, which can result in high computational complexity and long decoding delays, and will be unable to track channels that change fast. Adaptive approaches have been devised to reduce decoding delay, such as the squared autocorrelation algorithm (SAM) [3] and a null-carrier based technique from [4]. Multi-carrier equalization by restoration of redundancy (MERRY) algorithm [5] is devised by exploiting the cyclic redundancy from the CP. The SAM technique in [3] suffers from local minima convergence, while the methods in [4, 5] still need a comparatively larger number of OFDM blocks for satisfactory convergence. SLAM (single lag autocorrelation minimization) [6] is a lower complexity version of the SAM method using the steepest-decent type algorithm. However, this structure also exhibits undesirable local minima in the cost function and requires an effective stop criterion to maximize CSE signal-to-interference ratio.

In this work we develop a low complexity technique that is adaptive to the time-varying nature of each individual channel path. For time-varying channels we introduce a low-overhead periodic initialization block, which in turn supports adaptive update capability of the CSE filter coefficients and the shortened channel time-varying impulse response. To minimize complexity and delay, we propose a segmented polynomial tracking algorithm to provide adaptive updates as the channel changes.

In Section 2 a description of the transmitted signal structure, channel model and CSE estimation method are described, with tracking performance presented in Section 3. Section 4 provides BER performance over fading channels, while Section 5 concludes the paper.

2. COHERENT OFDM SYSTEM

We utilize the same transmission structure described in [7]. In this work, however, we perform coherent modulation and introduce adaptive updates to the CSE coefficients due to the time-varying nature of the channel delay profile. In [7], the channel coherence time was assumed to be long such that the channel was approximately constant between the initialization symbols of the differential block encoder. In this work, the initialization structure from [7] is utilized for training in the receiver's coherent demodulator. In a mobile channel, we can not assume a long channel coherence time as in [7], and thus must provide the capability to adjust the CSE coefficients as the channel changes. In practice, it is often desirable to transmit a signal with a known structure, which in addition to its training purposes, can also be exploited for automatic gain control (AGC) to maximize dynamic range and sensitivity of the receiver. Fortunately, the signaling structure borrowed from [7] allows us to exploit the initialization symbol for channel

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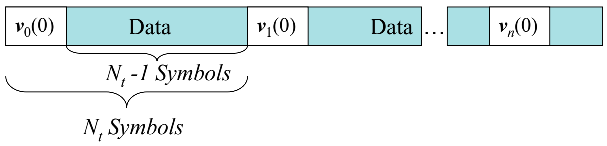

Figure 1: Block Coherent OFDM transmit signal structure.

tracking and AGC purposes.

Let $v(n)$ be the transmitted symbol after IFFT operation and CP (with length L_{cp}) insertion. With perfect synchronization at the receiver, the received symbol $r(n)$ can be expressed as:

$$\begin{aligned} r(n) &= \sum_{l=0}^{L_h} h(l)v(n-l) + w(n) \\ &= \sum_{l=0}^{L_{cp}} h(l)v(n-l) + \sum_{l=L_{cp}+1}^{L_h} h(l)v(n-l) + w(n), \end{aligned} \quad (1)$$

where L_h is the channel order and $w(n)$ is zero-mean, white, complex Gaussian noise with variance N_0. Figure 1 shows a conceptual illustration of the block coherent OFDM transmitted signal structure, and $v_n(0)$ is the initialization symbol.

2.1 Doubly Selective Channel Model

We use the Jakes' model to induce time-varying characteristics on each tap delay of the channel impulse response. The taps are represented by $h(j,m)$, where $h(0,m)$ are generated as zero-mean complex Gaussian variables with variance $\sigma_m^2 = \lambda \cdot \exp(-0.2m)$, where λ is the normalization factor to guarantee $\sum_{m=0}^{L_h} \sigma_m^2 = 1$. Time variation for the jth symbol is introduced by applying Jakes' model independently [8] to each mth channel tap with normalized frequency $f = f_{Dmax}T_s$, where f_{Dmax} is the maximum Doppler frequency, and T_s is the symbol period. Figure 2 shows a sample output from one of the channel taps over time. From Figure 2, we see the periodic transmission of the block differential encoder initialization block, as outlined in [7]. Next we outline the technique used for adaptive CSE estimation. We assume that $h(j,m)$ is approximately constant over the jth OFDM symbol.

2.2 Channel Shortening Equalization (CSE)

This section outlines the basic structure that will be used to estimate the CSE filter coefficients. A novel process to estimate CSE coefficients is detailed in [7], which assumes the channel is constant between initialization blocks. We adopt the same CSE estimation method here, but introduce a novel method to update CSE coefficients as each tap of the channel varies over time. The CSE is defined as a linear filter of length $L_c + 1$ and coefficients, $\boldsymbol{c} = [c(0),\ldots,c(L_c)]^T$. The received symbols filtered through the $(L_c + 1)$-tap CSE are defined as:

$$y(n) = \sum_{l=0}^{L_c} c(l)r(n-l). \quad (2)$$

From (1) and (2), the impulse response of the shortened channel $h_c(m)$ can be written as:

$$h_c(m) = \sum_{l=0}^{L_c} c(l)h(m-l), \forall m \in [0, L_{h_c}], \quad (3)$$

where $L_{h_c} = L_h + L_c$ is the order of the shortened channel. Exploiting the special structure of the transmitted signal defined in [7], and

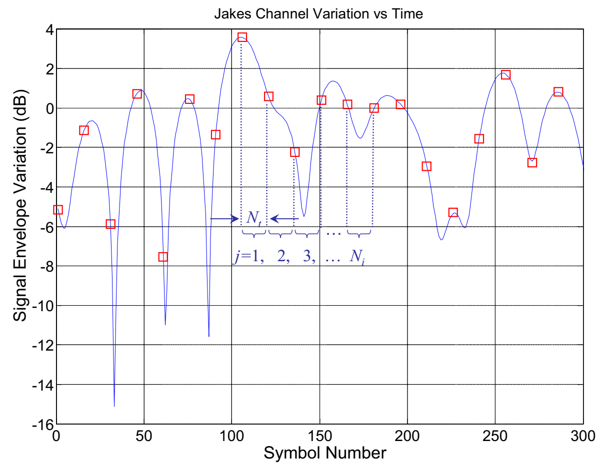

Figure 2: Sample one-tap channel variation using Jakes' model. (solid line–actual channel, square–received initialization block)

assuming that N (number of OFDM sub-carriers) > L_{h_c} the received symbol for the first block is

$$r(n) = \begin{cases} \sqrt{N}h(n-L_{cp}) + w(n), & n \in [L_{cp}, L_h + L_{cp}] \\ w(n), & \text{otherwise.} \end{cases} \quad (4)$$

From the special structure of $\boldsymbol{v}$ at the first block, $r(n)$ is a noisy version of the channel tap $h(n\text{-}L_{cp})$. Since the shortened channel $h_c(n)$ has desirable length L_{cp}, and neglecting the noise term $w(n)$, $y(n) = 0$ when $2L_{cp} \le n \le L_{cp} + L_{h_c}$. It is helpful to rewrite (2) as $y(n) = [r(n), r(n-1), \ldots r(n-L_c)]\boldsymbol{c} := \boldsymbol{r}^T\boldsymbol{c}$, and define a residual matrix as:

$$\boldsymbol{R}_{res} = \begin{bmatrix} \boldsymbol{r}^T(2L_{cp}) \\ \boldsymbol{r}^T(2L_{cp}+1) \\ \vdots \\ \boldsymbol{r}^T(L_{cp}+L_{h_c}) \end{bmatrix},$$

where $\boldsymbol{y}_{res} = \boldsymbol{R}_{res}\boldsymbol{c}$, and $\boldsymbol{y}_{res} := [y(2L_{cp}), \ldots, y(L_{cp}+L_{h_c})]^T$. (5)

When using the residual method from [7], the CSE is designed by minimizing the energy falling into the residual part (5), while keeping the total energy from the CSE coefficients fixed:

$$\begin{cases} \arg\min_c & \boldsymbol{c}^H \boldsymbol{R}_{res}^H \boldsymbol{R}_{res} \boldsymbol{c} \\ \text{subject to} & \boldsymbol{c}^H \boldsymbol{c} = 1. \end{cases} \quad (6)$$

The coefficients are given by $\boldsymbol{c} = \boldsymbol{u}_{min}$, where $\boldsymbol{u}_{min}$ is the eigenvector corresponding to the minimum eigenvalue of $\boldsymbol{R}_{res}^H\boldsymbol{R}_{res}$, and $\|\boldsymbol{u}_{min}\| = 1$. As $h(j,m)$, for $0 \le m \le L_h$, varies from the jth to the $(j+1)$th symbol in time, $\boldsymbol{c}(j) = [c(j,0),\ldots,c(j,L_c)]^T$ must be adaptive to ensure $\boldsymbol{y}_{res} = 0$ for each symbol. The channel model in [7] assumes $h(j,m)$ is approximately constant for N_t symbols. We assume that $h(j,m)$ changes from the jth to the $(j+1)$th symbol according to Jakes' model with Doppler frequency f, and changes independently on each mth tap. In order for block coherent OFDM to perform well, the jth received symbol after channel shortening will need to be equalized using the estimate of the shortened channel, or $\hat{h}_c$, where $\hat{h}_c(j,m) = \sum_{l=0}^{L_c} c(j,l)\hat{h}(j,m-l), \forall m \in [0, L_{h_c}]$. $\hat{h}$ and $\boldsymbol{c}$ are the mobile channel estimate and channel shortening coefficients, respectively, for the jth received symbol. A noisy version of $\hat{h}$ is available during the initialization symbol block and $\boldsymbol{c}$ is computed from (4) in Section 2.2. Tracking $\boldsymbol{c}$ and $\hat{h}_c$, for the time-varying (mobile) channel between initialization symbols, is accomplished using the segmented interpolator described below in Section 2.3.

2.3 Low Delay Adaptive CSE

We have three objectives for the design of an adaptive CSE. First, the number of received symbols for adaptation should be as few as possible. Next the technique should adapt very well to channel variation, but with modest to low computational complexity. Finally, the adaptive CSE coefficient estimation should be independent of the data decisions. In order to be independent of data decisions, we will exploit the initialization symbols to provide adaptive updates of the CSE coefficients, and will use interpolation to minimize the number of blocks required for the adaptation (thus minimizing complexity).

Potentially the most challenging objective is to devise a technique that provides good adaptation performance, but requires minimal delay. Development and evaluation of these tradeoffs is the focus of this section. The evaluation is based on the use of cubic spline interpolation to track channel variation and is capitalized for adaptive CSE estimation. We define $N_t \equiv$ spacing (in symbols) between initialization blocks and $N_i \equiv$ number of spline interpolator segments.

Given a function z, the cubic spline interpolator $s(x)$ can be used to provide an estimate of z, denoted as $\hat{z}$, where $s(x)$ is:

$$s(x)=\begin{cases} s_1(x)=a_{11}+a_{21}x+a_{31}x^2+a_{41}x^3, & x_1 \le x < x_2 \\ s_2(x)=a_{12}+a_{22}x+a_{32}x^2+a_{42}x^3, & x_2 \le x < x_3 \\ \cdots\cdots\cdots & \\ s_l(x)=a_{1l}+a_{2l}x+a_{3l}x^2+a_{4l}x^3, & x_l \le x < x_{l+1} \end{cases}. \quad (7)$$

In our case, we satisfy all conditions typical for a cubic spline interpolator, while using "not-a-knot" end conditions. The value of $l = N_i$, and the spacing N_t must both be made as small as possible to minimize delay. Thus rather than using a large N_i and N_t, we segment the process into many small pieces. Mathematically, a j-segmented spline process $s_j(x)$ is represented by:

$$s_j(x)=\begin{cases} s_{1,1}(x)=a_{11,1}+a_{21,1}x+a_{31,1}x^2+a_{41,1}x^3, & x_{1,1}\le x<x_{2,1} \\ s_{2,1}(x)=a_{21,1}+a_{22,1}x+a_{32,1}x^2+a_{42,1}x^3, & x_{2,1}\le x<x_{3,1} \\ \vdots & \\ s_{N_i,1}(x)=a_{1N_i,1}+a_{2N_i,1}x+a_{3N_i,1}x^2+a_{4N_i,1}x^3, & x_{N_i,1}\le x<x_{N_i+1,1} \\ s_{1,2}(x)=a_{11,2}+a_{21,2}x+a_{31,2}x^2+a_{41,2}x^3, & x_{1,2}\le x<x_{2,2} \\ s_{2,2}(x)=a_{12,2}+a_{22,2}x+a_{32,2}x^2+a_{42,2}x^3, & x_{2,2}\le x<x_{3,2} \\ \vdots & \\ s_{N_i,1}(x)=a_{1N_i,2}+a_{2N_i,2}x+a_{3N_i,2}x^2+a_{4N_i,2}x^3, & x_{N_i,2}\le x<x_{N_i+1,2} \\ \vdots & \\ s_{1,j}(x)=a_{11,j}+a_{21,j}x+a_{31,j}x^2+a_{41,j}x^3, & x_{1,j}\le x<x_{2,j} \\ \vdots & \\ s_{N_i,j}(x)=a_{1N_i,j}+a_{2N_i,j}x+a_{3N_i,j}x^2+a_{4N_i,j}x^3, & x_{N_i,j}\le x<x_{N_i+1,j} \end{cases}, (8)$$

where the spacing between knots, $\left|x_{N_i,j}-x_{N_i+1,j}\right|$ is fixed and equal to N_t. Our objective is to determine the minimum N_tN_i spacing which provides satisfactory performance for adaptive CSE.

In [9], for functions z that have four continuous derivatives, the cubic spline interpolant error is stated to decrease by a factor of 16 for each doubling in N_i. This is contrary to our objective in developing a low-delay adaptive-CSE with minimum estimation error performance. Figure 2 shows a sample channel response and the associated adaptation parameters we seek to optimize. We can see that if $N_t = 10$ and $N_i = 2$, then we have a low delay requirement of $N_tN_i = 20$ symbols. However, [9] may suggest N_i should be much greater than the value we give in this example for the number of spline segments. The appropriate tradeoff in parameters N_tN_i is studied through simulation. The experiment is centered around minimal delay and maximum tracking performance. Thus for our purposes we chose $N_i \ge 2$ and $N_t \le 1/(4f_{Dmax}T_s)$. We compare the segmented spline tracker to a full spline, where full is defined as $N_iNt \ge 1000$. The mean square error (MSE) tracking performance tradeoffs for various values of N_i and N_t are shown in the next section. We define MSE = $E[\left|h-\hat{h}\right|^2]$, where $E[\cdot]$ is the expected value, $\hat{h}$ is the estimate provided by the segmented- and full-spline, $s_j(x)$ and $s(x)$, respectively, and h is the actual channel. CSE coefficients are estimated using (4)-(6) and (7),(8).

3. TRACKING PERFORMANCE

Figure 3 shows the results for segmented- and full-spline trackers at various values of N_i and N_t for the channel described in Section 2, with $L_h = 29$, $f = 0.02$, and SNR = 30 dB. We see that (unexpectedly) the results for the case when $N_i = 2$ actually performs better than the full spline tracker, provided $N_t \le 10$. From $N_t \approx 1/(4f_{Dmax}T_s)$, given in [10], and $f = 0.02$, the MSE performance would be expected to degrade when $N_t \ge 12.5$. This trend is evident for all N_i values, but is much more pronounced when $N_i = 2$. We conjecture then, for our purposes of adaptive CSE estimation, that we can use $N_i \ge 2$ as long as $N_t \le 1/(4f_{Dmax}T_s)$ is satisfied. The evidence that $N_i = 2$ is better than all other choices shown is not clear to us, however. Perhaps it is tied to the nature of the channel variation and the error that is produced subject to the knot- and end-conditions as N_i varies. The trends we see in Figure 3 are consistent at all received SNRs. Figure 4 shows the actual and tracked channel for a single channel tap over time, while Figure 5 shows the frequency response of the shortened channel over time.

4. MOBILE BER PERFORMANCE

When evaluating performance, we use the same channel parameters as defined in Section 3 for $N_t = 10$. Figure 6 shows the performance of the fixed CSE technique introduced in [7] and the adaptive CSE technique we propose in this paper, but for coherent modulation. After CSE, coherent demodulation is performed on the jth symbol using a frequency-domain zero-forcing equalizer or $\hat{Y}[k]=Y[k]./\hat{H}_c[k]$, where $Y[k]$ (k=1:N) is the frequency-domain equivalent to $y(n)$ in (2) after CP removal and $\hat{H}_c[k]$ is the frequency-domain equivalent of $\hat{h}_c(j)$ above (note that we have dropped the index m from $\hat{h}_c(j,m)$). In the fixed CSE (fcse) case, we assume the CSE coefficients are estimated only during the training (initialization) block and not adapted between the training

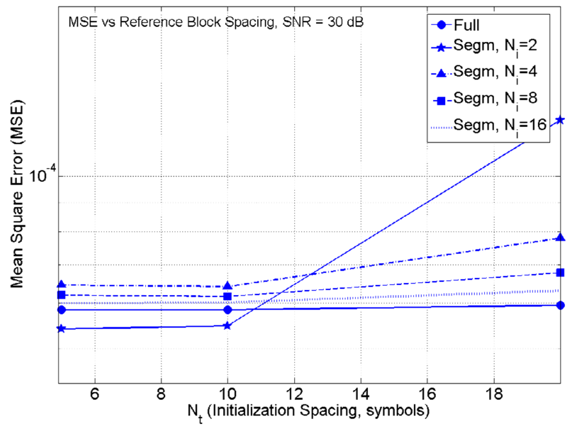

Figure 3: Segmented and full spline performance over N_iN_t.

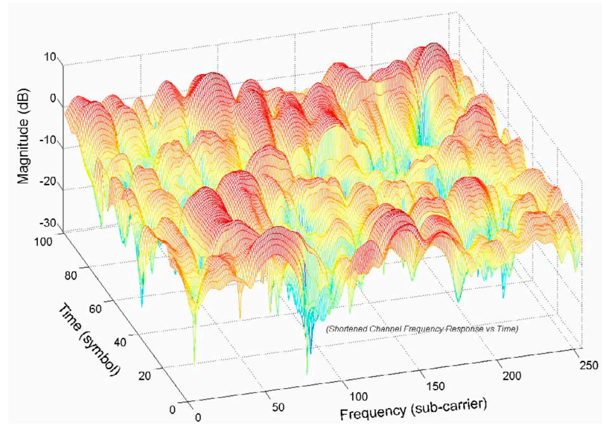

Figure 5: $\hat{H}_c$ versus time when $L_h = 29$ and 30 dB SNR.

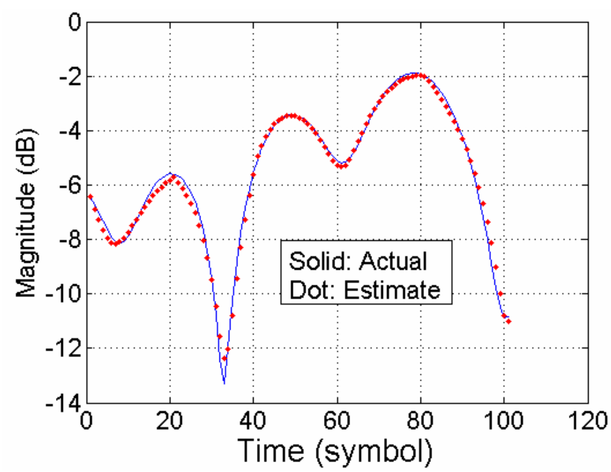

Figure 4: Sample of h and $\hat{h}$ for channel tap-1, 30 dB SNR.

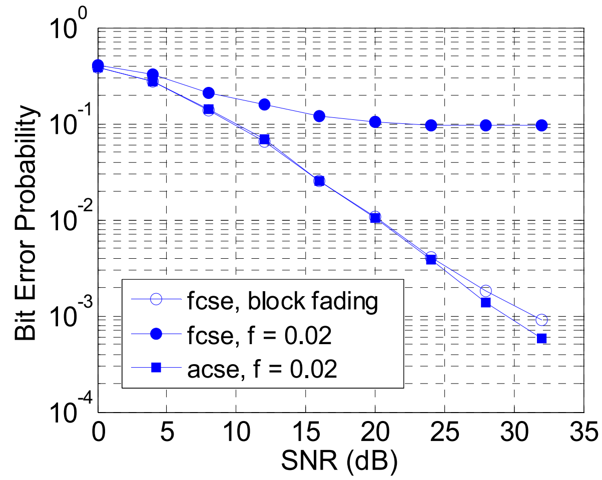

Figure 6: BER for fixed- and adaptive-CSE in doubly selective-fading channel ($f = 0.02$) and fixed-CSE in block fading channel.

intervals. Consequently, the CSE coefficients, $\boldsymbol{c}(j)$, and the shortened impulse response $\hat{h}_c(j)$ both lose track of the actual perturbations as the channel varies. It is clear that this has a debilitating effect on demodulated BER. The adaptive CSE (acse) system tracks both $\boldsymbol{c}(j)$ and $\hat{h}_c(j)$, which shows a significant improvement. The acse performance is equivalent to a fcse coherent system in a block fading channel (where the channel is assumed constant over N_t symbols, but varied after each block of N_t symbols).

5. CONCLUSIONS

In this paper we proposed a low complexity and low delay technique to provide a significant performance improvement in long delay spread mobile OFDM channels. A segmented-spline technique is introduced which is used to provide adaptive channel shortening equalization and block channel estimation for coherent demodulation. A block coding structure is utilized across OFDM sub-carriers which provides maximum frequency diversity in the system. Adaptive channel shortening equalization (CSE) is performed by capitalizing on the block training structure and using the polynomial segmented-spline interpolator between training periods. The signal after adaptive CSE is then corrected from the shortened channel frequency response, which provides a significant improvement in demodulated bit error rate.

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