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MLSD Coupled with LMMSE Channel Estimation for OFDM in Time-variant Channels

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***Abstract*—For OFDM systems, high user mobility induces time-variation of channel response over symbol duration, causing orthogonality destruction and introducing intercarrier interference (ICI) among subcarriers. To enhance the performance, sequence detection schemes have been proposed with the availability of corresponding channel state information (CSI), which is generally obtained via channel estimation using some inserted preamble symbols. However, bandwidth efficiency is significantly degraded and long latency is introduced. In this work, without the need of inserting preamble symbols, we propose a linear minimum mean square error (LMMSE) channel estimation method coupled with the Viterbi-algorithm based maximum-likelihood sequence detection (MLSD) scheme to improve bandwidth efficiency and to minimize the detection latency. Hence this proposed scheme can be applied to real-time, delay-sensitive applications. Compared to other pilot-based ICI mitigation methods, our scheme has better performance, as well as better bandwidth efficiency.**

I. Introduction

Orthogonal frequency division multiplexing (OFDM) is a promising technique for wireless mobile communication systems in order to meet the demands for high spectral efficiency and high data rate transmission. In OFDM systems, the cyclic prefix (CP) is inserted between successive OFDM symbols to eliminate the effect of inter-symbol interference (ISI) and to maintain the orthogonality among different subcarriers when the channel response is time-invariant. For high user mobility, however, the channel response may become time-variant over symbol duration and the orthogonality among subcarriers is destroyed, which results in intercarrier interference (ICI) severely degrading receiving performance.

Some works have discussed the effects of ICI in a time-variant channel. In [1], [2], ICI was modeled as an additive white Gaussian random process and a theoretical analysis for the error floor due to ICI on the symbol-error rate was presented. The universal upper bounds on total ICI power were derived in [3], and the exact bounds were derived in [4]. An analysis on the ICI generation mechanism was presented in [5] facilitating further understanding about power spread of each subcarrier due to Doppler frequency shift. The matched-filter bound (MFB), corresponding to the best possible receiving performance that can be achieved over a given channel, for OFDM in time-selective, frequency-flat Rayleigh fading channel was evaluated [6], [7]. Similar works over doubly dispersive channels were derived in [4], [8].

Several methods have been proposed in recent years to eliminate the ICI caused by time-variant channels. An ICI self-cancellation coding and a partial response coding, originally proposed to suppress the ICI caused by carrier frequency offset (CFO) in [9], [10], have been introduced to combat the ICI caused by Doppler frequency shift in [11], [12]. Jeon et al. proposed a method under the assumption that Doppler spread and delay spread are both small [13]. Mostofi et al. introduced a piece-wise linear model to approximate the time variation characteristics for the normalized Doppler frequency less than 0.1 [14]. A minimum mean square error (MMSE) equalization with successive interference cancellation (SIC) and a simplified linear MMSE equalization with parallel interference cancellation (PIC) were proposed in [15] and [4], respectively. In [5], [16], [17], based on the philosophy of maximizing the signal-to-interference-plus-noise ratio (SINR), the methods applying window preprocessing and a bank of linear time-variant (LTV) filters were presented. Hou and Chen designed a modified PIC mechanism with less noise enhancement for a channel with linearly varying response [18]. An ICI mitigation method relying on basis expansion model (BEM) was proposed in [19]. A modified maximum-likelihood sequence detection (MLSD) scheme was proposed in [20], which utilizes adaptive T-algorithm to reduce complexity but sacrifices bandwidth efficiency in acquiring full knowledge of channel response to compute the threshold.

Most of the previous works assume that full channel state information (CSI), i.e. the frequency-domain channel gains of all subcarriers, is available. This assumption is too ideal and cannot be easily achieved. The corresponding channel gains can be possibly obtained by using multiple preamble OFDM symbols accomplished with MMSE estimation. However, this approach has very high computational complexity and may significantly degrade the bandwidth efficiency to 50%. Furthermore, this approach also induces extra latency and thus makes real-time demodulation impossible. In this work, we investigate the ICI elimination performance of Viterbi-algorithm based MLSD scheme for OFDM systems in time-variant channels. Furthermore, we devise a channel estimation scheme, based on the linear minimum mean square error (LMMSE) concept, coupled with MLSD to improve bandwidth efficiency and to realize real-time demodulation.

II. System Model

A. OFDM System Model

The propagation channel is assumed to be the well-known wide sense stationary uncorrelated scattering (WS-

SUS) channel. Due to high user mobility, the channel becomes time-variant in an OFDM symbol interval, and the channel impulse response can be represented as

$$h(t,\tau) = \sum_{l=0}^{L-1} \gamma_l(t)\delta(\tau - \tau_l), \tag{1}$$

where L is the number of nonzero channel taps; τ_l is the time delay of the l-th path; $\gamma_l(t)$ is the corresponding time-variant complex channel gain of the l-th path; and $\delta(\cdot)$ is the Dirac's delta function. The propagation environment is assumed to be isotropic scattering with a Doppler frequency shift f_d. It is reasonable to assume that the Doppler frequency within one symbol duration is constant.

Suppose that, for an OFDM system experiencing a time-variant channel, the number of subcarriers available within the transmission bandwidth is N and the subcarrier separation is Δf. Then we have the sample interval of the OFDM system as $T_c = 1/N\Delta f$. As a result, the time-variant channel can be represented as a discrete impulse response $h[n,l]$, which is modeled by a T_c-spaced transversal filter with L taps. Correspondingly, the normalized Doppler frequency shift is defined as $\bar{f}_d = f_d/\Delta f$.

Let $\mathbf{X} = [X_0\ X_1\ \cdots\ X_{N-1}]^T$ denote the frequency-domain modulated data sequence in an OFDM symbol. By applying N-point inverse discrete Fourier transform (IDFT), we have the transmitted OFDM symbol in the time-domain as

$$\begin{cases} x[n] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k e^{j2\pi nk/N}, & \text{for } 0 \le n \le N-1 \\ x[n] = x[N+n], & \text{for } -N_p \le n \le -1 \end{cases} \tag{2}$$

where N_p is the length of CP. After passing through a time-variant channel with discrete impulse response $h[n,l]$, the received time-domain signal can be represented as

$$y[n] = \sum_{l=0}^{L-1} h[n,l]x[n-l] + w[n], \quad 0 \le n \le N-1, \tag{3}$$

where $w[n]$ is the complex additive white Gaussian noise (AWGN) with zero mean and variance N_0. After DFT, the signal on the k-th subcarrier can be expressed as

$$\begin{aligned} Y_k &= \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} y[n]e^{-j2\pi kn/N},\ 0 \le k \le N-1 \\ &= H_k X_k + I_k + W_k, \end{aligned} \tag{4}$$

where H_k is the frequency-domain channel gain corresponding to X_k on subcarrier k, I_k is the ICI term that subcarrier k suffered due to the time-variant channel, and W_k is the AWGN on subcarrier k. By defining $H_k[n] \triangleq \sum_{l=0}^{L-1} h[n,l]e^{-j2\pi lk/N}$, H_k and I_k can be simply represented as [15]

$$H_k = \frac{1}{N}\sum_{n=0}^{N-1} H_k[n] \tag{5}$$

and

$$I_k = \frac{1}{N}\sum_{m=0,m\ne k}^{N-1}\left(\sum_{n=0}^{N-1} H_m[n]e^{j2\pi n(m-k)/N}\right)X_m. \tag{6}$$

Furthermore, we can also rewrite the received signal in the matrix form as follows. According to (3), the received signal vector in the time-domain is

$$\mathbf{y} = \mathbf{H}_{\mathrm{TV}}\mathbf{x} + \mathbf{w} = \mathbf{H}_{\mathrm{TV}}\mathbf{F}^{\mathrm{H}}\mathbf{X} + \mathbf{w}, \tag{7}$$

where $\mathbf{F}$ is the N-point unitary DFT matrix with $(\cdot)^{\mathrm{H}}$ denoting the Hermitian transpose, and $\mathbf{w}$ is the time-domain noise vector. $\mathbf{H}_{\mathrm{TV}}$ is the time-domain time-variant channel matrix with the elements $(\mathbf{H}_{\mathrm{TV}})_{n+1,l+1} \triangleq h[n, \langle n-l\rangle_N]$, where $h[n,l]=0$ for $l \ge L$ and $\langle\cdot\rangle_N$ denotes the modulo-N operation. In the frequency-domain, we can define the equivalent time-variant channel matrix as $\mathbf{H_{eq}} \triangleq \mathbf{F}\mathbf{H}_{\mathrm{TV}}\mathbf{F}^{\mathrm{H}}$, and the received signal vector can be expressed as

$$\mathbf{Y} = \mathbf{F}\mathbf{y} = \mathbf{H_{eq}}\mathbf{X} + \mathbf{W}, \tag{8}$$

where $\mathbf{W} = \mathbf{F}\mathbf{w}$ is the frequency-domain noise vector.

B. ICI Statistics

Due to Doppler spread, signal power on each subcarrier mainly spreads to a few neighboring subcarriers. However, the destruction of orthogonality among different subcarriers still induces ICI from a subcarrier to other faraway subcarriers. As a result, the ICI coming from a specific subcarrier widely spreads to all other subcarriers, but mainly contributes to only a few neighboring subcarriers. Accordingly, we can rewrite (4), by focusing on the major ICI coming from the neighboring subcarriers around the desired subcarrier, as

$$Y_k = \sum_{m\in\Omega_k} (\mathbf{H_{eq}})_{k,m}X_m + \sum_{m\notin\Omega_k} (\mathbf{H_{eq}})_{k,m}X_m + W_k, \tag{9}$$

where $\Omega_k = \{\langle k-Q\rangle_N, \cdots, k, \cdots, \langle k+Q\rangle_N\}$ is the set of subcarrier indices that contribute significant ICI to subcarrier k, and $2Q+1$ is an odd number denoting the number of elements within Ω_k. The first term contains the desired signal and the major ICI power coming from the neighboring subcarriers. The second term is the residual ICI incurred from all other subcarriers, and would be treated as AWGN for simplicity [1], [2].

III. MLSD-based LMMSE Channel Estimation

A. MLSD ICI Elimination

If the corresponding channel gains of the major ICI, i.e. $(\mathbf{H_{eq}})_{k,m}$ for $m \in \Omega_k$ and $k = 0,\cdots,N-1$, are available, we can apply the MLSD mechanism for data sequence detection. The maximum-likelihood (ML) decision criterion is defined as

$$\hat{\mathbf{X}} = \arg\min_{\tilde{\mathbf{X}}} \sum_{k=0}^{N-1}\left|Y_k - \sum_{m=0}^{N-1}(\mathbf{H_{eq}})_{k,m}\tilde{X}_m\right|^2, \tag{10}$$

where $\hat{\mathbf{X}} = \left[\hat{X}_0\ \hat{X}_1\ \cdots\ \hat{X}_{N-1}\right]^T$ is the decision data sequence, and $\tilde{\mathbf{X}} = \left[\tilde{X}_0\ \ \tilde{X}_1\ \cdots\ \tilde{X}_{N-1}\right]^T$ is a possible data sequence. Then, according to (9) and considering only the major ICI term, the ML criterion can be approximated as

$$\hat{\mathbf{X}} = \arg\min_{\tilde{\mathbf{X}}} \sum_{k=0}^{N-1}\left|Y_k - \sum_{m\in\Omega_k}(\mathbf{H_{eq}})_{k,m}\tilde{X}_m\right|^2. \tag{11}$$

Apparently, this approximation can significantly reduce the computational complexity at the cost of performance degradation. Based on (11) and the MLSD philosophy, we can form a trellis diagram with $N_s = 2^{2MQ}$ states, where the size of signal constellation is 2^M. Define $\mu_{k+Q}^{(\ell)} = \left| Y_k - \sum_{m\in\Omega_k} (\mathbf{H_{eq}})_{k,m}\tilde{X}_m^{(\ell)} \right|^2$ as the branch metric and $\Psi_{k+Q}^{(\ell)} = \sum_j \mu_j^{(\ell)}$ as the path metric corresponding to the surviving sequence ℓ at epoch $k+Q$, where $\tilde{X}_m^{(\ell)}$ is the corresponding data symbol within this surviving sequence and $\mu_j^{(\ell)}$ is the corresponding branch metric at epoch j. Repeating the metric calculation process until epoch N, all the surviving path metrics corresponding to the N_s states can be evaluated. Subsequently, the surviving path with the minimum metric is chosen as the decision data sequence $\hat{\mathbf{X}}$.

B. LMMSE Channel Estimation

To eliminate the ICI caused by a time-variant channel, precise estimation of the channel gain corresponding to the desired signal on each subcarrier is not sufficient; the channel gains corresponding to all ICI signals interfering on a subcarrier must be also correctly determined. In other words, all the elements in the equivalent time-variant channel matrix $\mathbf{H_{eq}}$, or at least the elements $(\mathbf{H_{eq}})_{k,m}$ for $m \in \Omega_k$ and $k = 0, \cdots, N-1$, must be accurately determined.

Based on the Viterbi decision algorithm, we propose an LMMSE scheme estimating the equivalent frequency-domain channel tap gains on each state, by exploiting the frequency-domain autocorrelation property and building a first-order Gauss-Markov autoregressive process for each tap. The proposed LMMSE channel estimation scheme employs no preamble OFDM symbol, but only a fraction of pilot subcarriers in each OFDM symbol. Thus the impact on the overall bandwidth efficiency is minimized. Moreover, channel estimation is processed concurrently with the detection, making it applicable for real-time applications. Since only the channel gains corresponding to the desired signal and the major ICI are incorporated within the Viterbi algorithm, we focus only on the determination of the major channel gains, i.e. the elements $(\mathbf{H_{eq}})_{k,m}$ for $m \in \Omega_k$ and $k = 0, \cdots, N-1$.

Define the frequency-domain correlation coefficient between $(\mathbf{H_{eq}})_{k,m}$ and $(\mathbf{H_{eq}})_{k',m'}$, where $k' = \langle k+v \rangle_N$ and $m' = \langle m+v \rangle_N$, as

$$\begin{aligned} &R_{k,m}(v) \\ &= \frac{E\left[\left((\mathbf{H_{eq}})_{k,m} - E[(\mathbf{H_{eq}})_{k,m}]\right)\left((\mathbf{H_{eq}})_{k',m'} - E\left[(\mathbf{H_{eq}})_{k',m'}\right]\right)^*\right]}{E\left[\left((\mathbf{H_{eq}})_{k,m} - E[(\mathbf{H_{eq}})_{k,m}]\right)\left((\mathbf{H_{eq}})_{k,m} - E[(\mathbf{H_{eq}})_{k,m}]\right)^*\right]}, \end{aligned} \tag{12}$$

which is assumed to be available for the desired receiver. It is noted that $R_{k,m}(-v) = R_{k,m}(v)^*$ for all v. Then the autoregressive model for $(\mathbf{H_{eq}})_{k',m'}$ is given as

$$(\mathbf{H_{eq}})_{k',m'} = R_{k,m}(v)\,(\mathbf{H_{eq}})_{k,m} + \eta_{k',m'}, \tag{13}$$

where $\eta_{k',m'}$ is the innovation term assumed to be a Gaussian random variable with zero mean and variance

$$\begin{aligned} &\sigma_{k',m'}^2 = \\ &E\left[\left((\mathbf{H_{eq}})_{k',m'} - E\left[(\mathbf{H_{eq}})_{k',m'}\right]\right)\left((\mathbf{H_{eq}})_{k',m'} - E\left[(\mathbf{H_{eq}})_{k',m'}\right]\right)^*\right] \\ &- E\left[\left((\mathbf{H_{eq}})_{k,m} - E[(\mathbf{H_{eq}})_{k,m}]\right)\left((\mathbf{H_{eq}})_{k',m'} - E\left[(\mathbf{H_{eq}})_{k',m'}\right]\right)^*\right]. \end{aligned} \tag{14}$$

According to (9) and (13) the frequency-domain received signal can be rewritten as

$$\begin{aligned} Y_k = &\sum_{m\in\Omega_k} \left[R_{k'',m''}(v)\,(\mathbf{H_{eq}})_{k'',m''} + \eta_{k,m}\right] X_m \\ &+ \sum_{m\notin\Omega_k} (\mathbf{H_{eq}})_{k,m} X_m + W_k, \end{aligned} \tag{15}$$

where $k'' = k - v$ and $m'' = m - v$.

To estimate the channel gains at epoch i, we take the received signals of $2Q+1$ successive subcarriers (i.e., $Y_{\langle i-2Q \rangle_N}, \cdots, Y_i$ taken from subcarrier $\langle i-2Q \rangle_N$ to subcarrier i) into consideration, since all the data symbols related to these received signals are currently available for the correct surviving path in MLSD detector (i.e., $X_{\langle i-3Q \rangle_N}, \cdots, X_{\langle i+Q \rangle_N}$). Consequently, we have $2Q+1$ linear equations corresponding to the correct surviving path expressed as

$$\begin{aligned} \mathbf{Y_i} &= \left[Y_{\langle i-2Q \rangle_N} \quad Y_{\langle i-2Q+1 \rangle_N} \quad \cdots \quad Y_{\langle i-1 \rangle_N} \quad Y_i\right]^T \\ &\triangleq \mathbf{A_i H_i} + \mathbf{\Theta_i} + \mathbf{\Gamma_i} + \mathbf{W_i} \\ &= \mathbf{A_i H_i} + \tilde{\mathbf{\Gamma}}_\mathbf{i} + \mathbf{W_i}, \end{aligned} \tag{16}$$

where $(\mathbf{A_i})_{p,q} = X_{\langle i-3Q-2+p+q \rangle_N} R_{i,\langle i-Q-1+q \rangle_N}(p - 2Q - 1)$ for $1 \le p \le 2Q+1$ and $1 \le q \le 2Q+1$, $\mathbf{H_i} = \left[(\mathbf{H_{eq}})_{i,\langle i-Q \rangle_N} \ \cdots \ (\mathbf{H_{eq}})_{i,i} \ \cdots \ (\mathbf{H_{eq}})_{i,\langle i+Q \rangle_N}\right]^T$ is the estimated channel gain vector corresponding to the desired signal and the major ICI on subcarrier i, $\mathbf{\Theta_i}$ is an error vector due to the approximation of the autoregressive model, $\mathbf{\Gamma_i}$ is the residual ICI vector, $\mathbf{W_i}$ is the frequency-domain AWGN vector, and $\tilde{\mathbf{\Gamma}}_\mathbf{i} = \mathbf{\Theta_i} + \mathbf{\Gamma_i}$. By applying the LMMSE estimator, the estimated frequency-domain channel gain vector can be obtained as

$$\hat{\mathbf{H}}_\mathbf{i} = \mathbf{C_{H_iY_i}} \mathbf{C_{Y_iY_i}^{-1}} \mathbf{Y_i}, \tag{17}$$

where $\mathbf{C_{H_iY_i}} = E\left[\left(\mathbf{H_i} - E[\mathbf{H_i}]\right)\left(\mathbf{A_i}\left(\mathbf{H_i} - E[\mathbf{H_i}]\right)\right)^H\right]$ is the covariance matrix between $\mathbf{H_i}$ and $\mathbf{Y_i}$, $\mathbf{C_{Y_iY_i}} = \mathbf{A_i} E\left[\left(\mathbf{H_i} - E[\mathbf{H_i}]\right)\left(\mathbf{H_i} - E[\mathbf{H_i}]\right)^H\right]\mathbf{A_i}^H + \mathbf{C_{\tilde{\Gamma}_i\tilde{\Gamma}_i}} + N_0\mathbf{I}$ is the auto-covariance matrix of $\mathbf{Y_i}$ with $\mathbf{I}$ denoting the $(2Q+1) \times (2Q+1)$ identity matrix, and $\mathbf{C_{\tilde{\Gamma}_i\tilde{\Gamma}_i}} = \operatorname{diag}\left(\sigma_{i-2Q}^2, \cdots, \sigma_i^2\right)$ with σ_j^2 denoting the average residual signal power on subcarrier j, including the residual ICI and the innovation error. Since the total received signal power is fixed for the data sequence $X_{\langle i-3Q \rangle_N}, \cdots, X_{\langle i+Q \rangle_N}$, we have the following constraint [4]

$$\mathbf{C_{\tilde{\Gamma}_i\tilde{\Gamma}_i}} + \operatorname{diag}\left(\mathbf{A_i} E\left[\mathbf{H_i H_i}^H\right]\mathbf{A_i}^H\right) = \left(\sum_{l=0}^{L-1} \sigma_{h_l}^2\right) \cdot \mathbf{I}, \tag{18}$$

where $\sigma_{h_l}^2$ is the average signal power corresponding to each tap of delay spread. Accordingly, $\mathbf{C_{\tilde{\Gamma}_i\tilde{\Gamma}_i}}$ can be obtained via (18) based on the channel statistics, which are assumed to be available based on the channel types.

To combine with a Viterbi decoder, for each surviving path the LMMSE channel estimator uses $4Q+1$ successive data symbols to form matrix $\mathbf{A_i}$, including $4Q$ latest data symbols stored within each surviving sequence and the one for current state. The $(2Q+1)$-tap channel gains obtained from LMMSE estimator in each state are then applied to compute the branch metric of each state at current epoch. The same operation is processed for all states and repeated at each epoch until epoch $N-1$.

C. Pilot Subcarriers Placement

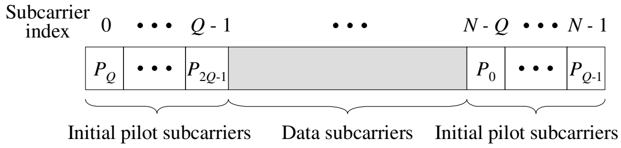
For Viterbi algorithm, a determinate initial condition, as well as a determinate final state, is essential for correction decision. To set up the initial state and the final state for the MLSD ICI elimination scheme, a proper placement of pilot subcarriers is necessary. Assuming that all the corresponding channel gains are available, the placement of pilot subcarriers in an OFDM symbol for a Viterbi decoder with 2^{2MQ} states can be as that shown in Fig. 1(a). Due to the cyclic-shifting property of OFDM systems, we require at least $2Q$ pilot subcarriers for the determination of both initial and final conditions. Specifically, the first Q subcarriers, i.e. subcarrier i for $i = 0, \cdots, Q-1$, and the last Q subcarriers, i.e. subcarrier i for $i = N-Q, \cdots, N-1$, are all assigned as pilot subcarriers. In practice, guard subcarriers carrying with no signal are generally assigned at the band edges to increase spectral compactness and can be used as substitute for these $2Q$ pilot subcarriers. Thus, the degradation in bandwidth efficiency is minimized.

On the other hand, if the corresponding channel gains are unavailable and the propose LMMSE channel estimator is applied for channel estimation, more pilot subcarriers are required. Specifically, the first $3Q$ subcarriers and the last Q subcarriers are assigned as pilot subcarriers. Similarly, some pilot subcarriers can be assigned as guard subcarriers carrying with no signal. Notice that there are $4Q$ pilot symbols available for initial channel estimation by the cyclic-shifting property, and thus the proposed LMMSE channel estimator can be processed. Furthermore, it must be noted that both the channel gains and data symbols are simultaneously estimated/detected in the MLSD receiver. Since there is ambiguity in the determination of the signs (phases) of the channel gains, as well as the data symbols, for the MLSD receiver, it is highly possible that a wrong decision could be made on the data sequence which has inverse phase relative to the desired one. This type of error occurrence is referred to as the phase inversion error.

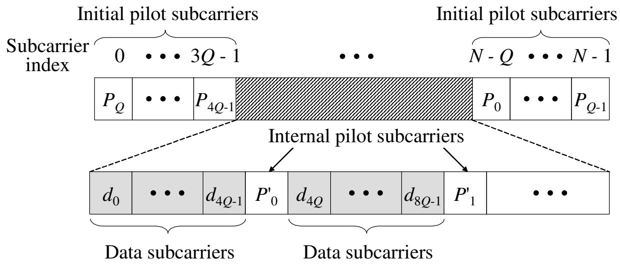
To prevent the phase inversion error, suitable amount of pilot subcarriers must be equally spread out among all data subcarriers. In other words, one out of every $Z \leq 4Q+1$ subcarriers is assigned as a pilot subcarrier, achieving a bandwidth efficiency approximated to $\alpha = 1 - 1/Z$. Specifically, as shown in Fig. 1(b), the first $3Q$ subcarriers are assigned as pilot/guard subcarriers, and then the subsequent $4Q$ subcarriers are assigned to be data subcarriers, followed by an internal pilot subcarrier to prevent phase inversion ambiguity. Subsequently, $4Q$ consecutive data subcarriers followed by one internal pilot subcarrier are successive assigned. Finally, the last Q subcarriers are assigned as guard/pilot subcarriers to complete the setting of both the initial and final states. It should be noted that the number of consecutive data subcarriers must be no more than $4Q$ to prevent the phase inversion error.

IV. Performance Evaluation and Comparison

In this section, we examine the bit error rate (BER) performance of the MLSD scheme with LMMSE channel estimation via simulation. Specifically, the number of subcarriers is $N = 128$, the length of CP is $N_p = N/4$ samples, and BPSK modulation is applied. The time-variant channel, generated by using Jake's method, is



(a) When perfect CSI is available.



(b) When LMMSE channel estimation is applied.

Fig. 1. Pilot subcarrier placement for MLSD ICI suppression scheme.

assumed to have an exponential decay power profile with delay spread length $L = 3$. The normalized Doppler frequency is assumed to be $\bar{f}_d$ = 0.05 or 0.3. The applied bandwidth efficiency is $\alpha = 3/4$ or 8/9. Moreover, perfect carrier and symbol synchronization are assumed. Two other schemes with pilot-based channel estimation are evaluated for performance comparison, including the MMSE scheme based on BEM developed in [19], with the corresponding bandwidth efficiency $\alpha = 5/8$, and the ICI mitigation scheme based on piece-wise linear approximation designed in [14], with the corresponding bandwidth efficiency $\alpha = 3/4$. Furthermore, two schemes with perfect CSI are investigated, including the traditional MMSE [4] and Viterbi-based MLSD.

Fig. 2 shows the BER performance comparison for normalized Doppler frequency $\bar{f}_d$ = 0.05. For $\bar{f}_d$ = 0.05, corresponding to low user mobility, the BER performance of different schemes is very close in low signal-to-noise ratio (SNR) region, but diverge in high SNR region. The performance of MLSD scheme is slightly better than MMSE scheme. The performance of MLSD-LMMSE is worse than that of the schemes with perfect CSI, but better than that of the other two schemes with pilot-based channel estimation. However, it must be noted that the bandwidth efficiency of the proposed scheme is $\alpha = 3/4$ or 8/9, which is far better than 50%. While, in order to obtain perfect CSI, the bandwidth efficiency of the two schemes with perfect CSI is only about 50%.

Fig. 3 shows the BER performance comparison for normalized Doppler frequency $\bar{f}_d$ = 0.3, corresponding to high user mobility. The performance of all schemes is severely degraded due to the strong ICI caused by high Doppler spread. The proposed scheme significantly outperforms the other two schemes with pilot-based channel estimation. To further improve BER performance as well as bandwidth efficiency, we can improve the accuracy of channel estimation and data detection by increasing the computational complexity, i.e. by increasing the parameter Q. If $Q = 3$ is applied, the performance of the proposed scheme can be further improved, approaching to that of the MMSE scheme with perfect CSI. In other words, BER performance can be significantly improved at the cost of

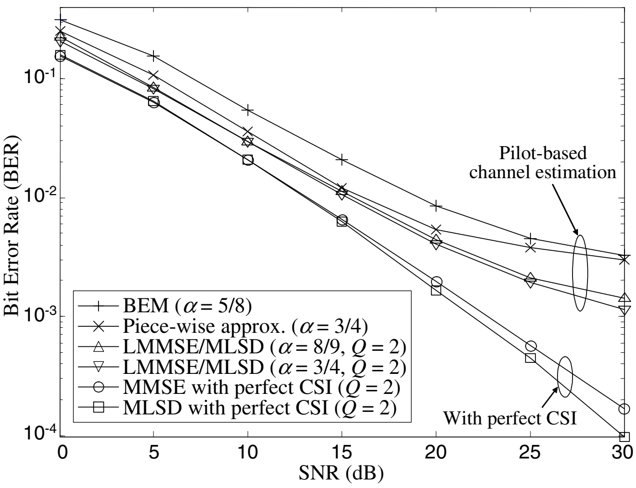


Fig. 2. BER performance comparison for $\bar{f}_d = 0.05$ in a multipath Rayleigh fading channel with $L = 3$.

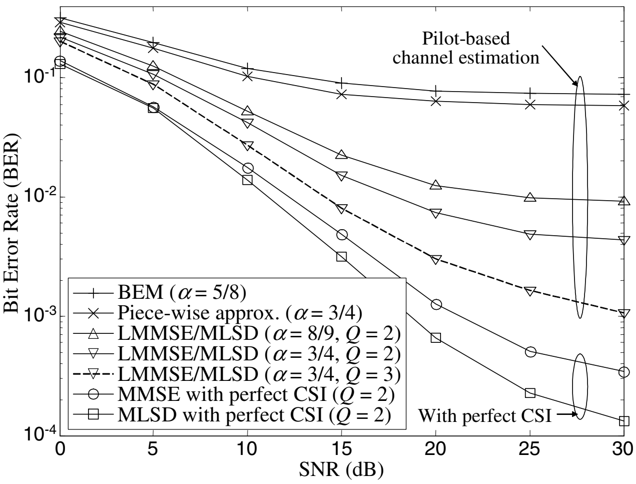


Fig. 3. BER performance comparison for $\bar{f}_d = 0.3$ in a multipath Rayleigh fading channel with $L = 3$.

increase in computational complexity. A tradeoff between bandwidth efficiency and computational complexity can be made under a target BER performance.

V. Conclusion

In this work, we have proposed an LMMSE channel estimation method coupled with the Viterbi-algorithm based MLSD scheme for OFDM systems in a time-variant channel. Without the need of inserting any preamble OFDM symbols, the proposed scheme employs only a fraction of pilot subcarriers in each OFDM symbol, significantly improving the overall bandwidth efficiency. Moreover, channel estimation is processed concurrently with the data detection, making it feasible for real-time, delay-sensitive applications. Compared to other pilot-based ICI mitigation methods, our scheme has better receiving performance, as well as better bandwidth efficiency. Furthermore, by incorporating more ICI subcarriers into consideration, the performance of the proposed scheme can be significantly improved at the cost of increasing computational complexity.

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