CHANNEL ESTIMATION FOR OFDM SYSTEMS BASED ON COMB-TYPE PILOT ARRANGEMENT IN FREQUENCY SELECTIVE FADING CHANNELS

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ABSTRACT

In this paper, the channel estimation methods for OFDM systems based on comb-type pilot sub-carrier arrangement are investigated. The channel estimation algorithm based on comb-type pilots is divided into pilot signal estimation and channel interpolation. The pilot signal estimation based on LS or MMSE criteria, together with channel interpolation based on piecewise-linear interpolation or piecewise second-order polynomial interpolation is studied. Owing to the MMSE estimate of pilot signals, the inter-carrier interference and additive white Gaussian noise are reduced considerably. The computational complexity of pilot signal estimation based on MMSE criterion can be reduced by using a simplified LMMSE estimator with low-rank approximation using singular value decomposition. Phase compensators before and after interpolation are also presented to combat the phase changes of subchannel symbols arising from the frame synchronization errors. Comparing to the transform-domain processing based channel estimation algorithm, the MMSE estimate of pilot signals together with phase compensated linear interpolation algorithm provides better performance and requires less computations.

I. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) technique has received a lot of interest in mobile communication research lately. For wideband mobile communication systems, the radio channel is usually frequency selective and time variant. Hereafter, in OFDM systems, the channel transfer function of radio channel appears unequal in both frequency and time domains. Therefore, a dynamic estimation of the channel is necessary for the demodulation of OFDM signals.

In wideband mobile channels, the pilot-based signal correction scheme has been proven a feasible method for OFDM systems [1]. Most channel estimation methods for OFDM transmission systems have been developed under the assumption of a slow fading channel, where the channel transfer function is assumed stationary within one OFDM data block. In addition, the channel transfer function for the previous OFDM data block is used as the transfer function for the present data block. In practice, the channel transfer function of a wideband radio channel may have significant changes even within one OFDM data block. Therefore, it is preferable to estimate channel characteristic based on the pilot signals in each individual OFDM data block.

Recently, an elegant channel estimation method for OFDM mobile communication systems has been proposed by Zhao and Huang [3]. In this method, the additive white Gaussian noise (AWGN) and the inter-carrier interference (ICI) in the pilot subcarriers are reduced by low-pass filtering in a transform domain, and the channel transfer function for all the subcarriers is obtained by the high-resolution interpolation realized by zero padding and DFT/IDFT. Comparing to the conventional linear interpolation method, this method provides about 1 dB and 3 dB improvement in E_b/N_0 for the same bit error rate values in slow- and fast-fading noisy radio channel, respectively.

In this paper, we present a different approach for channel estimation, which is based on a minimum mean-squared error (MMSE) estimate of pilot signals. Because of the robustness of the MMSE estimator, the AWGN and the ICI components are reduced significantly in fast- or slow-fading noisy radio channel environments. The computational complexity of the MMSE estimator can be reduced by using a simplified linear minimum mean-squared error (LMMSE) estimator with low-rank approximation by singular value decomposition (SVD) [6]. In addition, because of the in-sensitivity to parameter mismatch, a generic low-rank estimator can be used for the same kind of channels. The channel transfer function of data subcarriers is then interpolated based on the MMSE estimate of pilot signals. Two interpolation methods, the piecewise-linear and the piecewise second-order polynomial interpolations, are studied. Furthermore, by taking the frame position error into account, we propose a phase pre-compensator and post-compensator to mitigate the model mismatch error of the interpolator and the MMSE estimator. Based on our simulations over fast- and slow-fading mobile channels, we see that a significant improvement in E_b/N_0 for the same BER is provided.

The paper is organized as follows. In Section II, the pilot-based OFDM system is described, and two type of pilot arrangements, the block-type arrangement and the comb-type arrangement, are discussed. Section III discusses the estimation of pilot signals, and a low-complexity estimator is studied. Interpolation methods and a way to mitigate the model mismatch problem are deliberated in Section IV. Section V presents the simulation results, which indicate the BER improvements. The computational complexity of the proposed method is evaluated in Section VI. Section VII concludes the paper.

II. SYSTEM DESCRIPTION

Fig. 1 shows a typical block diagram of OFDM system with pilot signal assisted. The binary information data are grouped and mapped into multi-amplitude-multi-phase signals. In this paper, we consider the 16-QAM modulation. After pilot insertion, the modulated data $\{X(k)\}$ are sent to an IDFT and are

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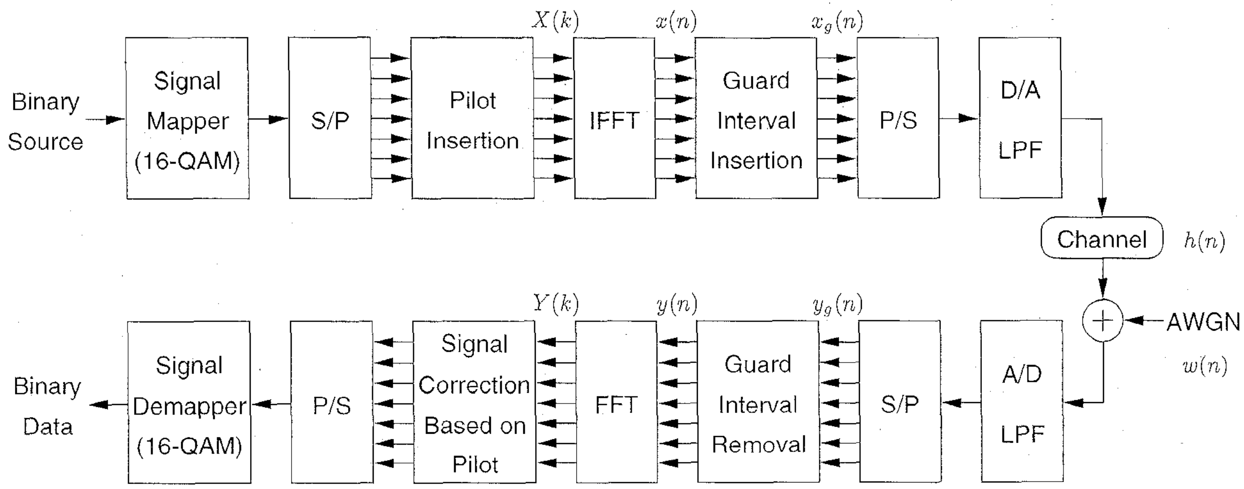


Fig. 1. Baseband model of a typical pilot-based OFDM system.

transformed and multiplexed into $\{x(n)\}$ as

$$x(n) = \text{IDFT}\{X(k)\} = \sum_{k=0}^{N-1} X(k) e^{j2\pi kn/N}, \quad n = 0, 1, \cdots, N-1, \tag{1}$$

where N is the number of subcarriers. The guard interval is inserted to prevent possible inter-symbol interference in OFDM systems, and the resultant samples $\{x_g(n)\}$ are

$$x_g(n) = \begin{cases} x(N+n), & n = -N_g, -N_g+1, \cdots, -1 \\ x(n), & n = 0, 1, \cdots, N-1 \end{cases} \tag{2}$$

where N_g is the number of samples in the guard interval. The transmitted signal is then sent to a frequency selective multi-path fading channel. The received signal can be represented by

$$y_g(n) = x_g(n) \otimes h(n) + w(n), \tag{3}$$

where $h(n)$ is the impulse response of channel and $w(n)$ is the additive white Gaussian noise. The channel impulse response $h(n)$ can be expressed as [4]

$$h(n) = \sum_{i=0}^{r-1} h_i e^{j2\pi f_{D_i} T \frac{n}{N}} \delta(\lambda - \tau_i), \quad 0 \leq n \leq N-1, \tag{4}$$

where r the total number of propagation paths, h_i the complex impulse response of the ith path, f_{D_i} the ith-path Doppler frequency shift which causes ICI of the received signals, λ the delay spread index, and τ_i the ith-path delay time normalized by sampling time.

After removing the guard interval from $y_g(n)$, the received samples $y(n)$ are sent to a DFT block to demultiplex the multi-carrier signals:

$$Y(k) = \text{DFT}\{y(n)\} = \frac{1}{N}\sum_{n=0}^{N-1} y(n) e^{-j2\pi kn/N}, \quad k = 0, 1, \cdots, N-1 \tag{5}$$

Suppose that the guard interval is longer than the length of channel impulse response, that is, there is no inter-symbol interference between OFDM symbols, the demultiplexed samples $Y(k)$ can be represented by [3]:

$$Y(k) = X(k)H(k) + I(k) + W(k), k = 0, 1, \cdots, N-1, \tag{6}$$

where

$$H(k) = \sum_{i=0}^{r-1} h_i e^{j\pi f_{D_i} T} \frac{\sin(\pi f_{D_i} T)}{\pi f_{D_i} T} e^{-j\frac{2\pi\tau_i}{N}k},$$

$$I(k) = \frac{1}{N}\sum_{i=0}^{r-1}\sum_{\substack{K=0 \\ K\neq k}}^{N-1} h_i X(K) \frac{1 - e^{j2\pi(f_{D_i}T - k + K)}}{1 - e^{j\frac{2\pi}{N}(f_{D_i}T - k + K)}} e^{-j\frac{2\pi\tau_i}{N}K},$$

and $W(k)$ is the Fourier transform of $w(n)$.

After that, the received pilot signals $\{Y_p(k)\}$ are extracted from $\{Y(k)\}$ and the channel transfer function $\{H(k)\}$ can be obtained from the information carried by $\{H_p(k)\}$. With the knowledge of the channel responses $\{H(k)\}$, the transmitted data samples $\{X(k)\}$ can be recovered by simply dividing the received signal by the channel response:

$$\hat{X}(k) = \frac{Y(k)}{\hat{H}(k)}, \quad k = 0, 1, \cdots, N-1, \tag{7}$$

where $\hat{H}(k)$ is an estimate of $H(k)$. After signal demapping, the source binary information data are re-constructed at the receiver output.

Based on the principle of OFDM transmission scheme, it is easy to assign the pilot both in time-domain and in frequency-domain. Several types of pilot arrangement have been studied [1]. Here we consider two major types of pilot arrangement as shown in Fig. 2. The first kind of pilot arrangement shown in Fig. 2(a) is denoted as block-type pilot arrangement. The pilot signal is assigned to a particular OFDM block, which is sent periodically in time-domain. This type of pilot arrangement is especially suitable for slow-fading radio channels. Because the training block contains all pilots, channel interpolation in frequency domain is not required. Therefore, this type of pilot arrangement is relatively insensitive to frequency selectivity. The estimation of channel response is usually obtained by LS or MMSE estimate of training pilots [5] [6].

The second kind of pilot arrangement shown in Fig. 2(b) is denoted as comb-type pilot arrangement. The pilot signals are uniformly distributed within each OFDM block. Assuming that the payloads of pilot signals of the two arrangements are the same, the comb-type pilot assignment has a higher re-transmission rate. Thus, the comb-type pilot arrangement system is provides better resistance to fast-fading channels. Since only some subcarriers contain the pilot signal, the channel response of non-pilot subcarriers will be estimated by interpolating neighboring pilot sub-channels [2] [3]. Thus, the comb-type pilot arrangement is sensitive to frequency selectivity when comparing to the block-type pilot arrangement system. That is, the pilot spacing $(\Delta f)_p$ must be much smaller than the coherence bandwidth of the channel $(\Delta f)_c$.

In this paper, we consider the radio channel which changes rapidly. In such environment, the channel transfer function changes significantly from one block to the next block. Thus, the channel estimation in the present block can not be used as the channel response of the next block. Therefore, the comb-type pilot subcarrier arrangement is adopted to estimate the channel transfer function in each block. As shown in Fig. 3, the pilot signals is first extracted from the received signal, and the channel transfer function is estimated from the received pilot signals and the known pilot signals. Then, the channel responses of subcarriers that carry data are interpolated by using the neighboring pilot channel responses. In the following sections, the pilot signal estimation and channel interpolation algorithms are discussed separately.

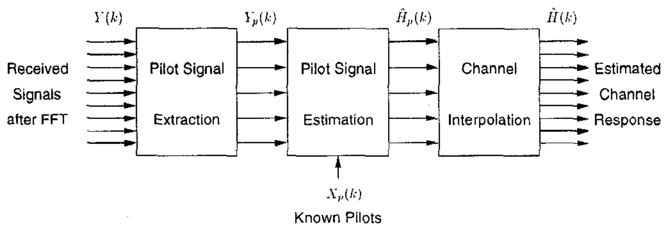


Fig. 3. Block diagram of channel estimation algorithm based on comb-type pilots.

III. Pilot Signal Estimation

For comb-type pilot subcarrier arrangement, the N_p pilot signals $X_p(m)$, $m = 0, 1, \cdots, N_p - 1$ are uniformly inserted into $X(k)$. That is, the total N subcarriers are divided into N_p groups, each with $L = N/N_p$ adjacent subcarriers. In each group, the first subcarrier is used to transmit pilot signal. The OFDM signal modulated on the kth subcarrier can be expressed as

$$\begin{aligned} X(k) &= X(mL+l) \\ &= \begin{cases} X_p(m), & l = 0, \\ \text{information data}, & l = 1, 2, \cdots, L-1. \end{cases} \end{aligned} \quad (8)$$

The pilot signals $\{X_p(m)\}$ can be either equal complex values c to reduce the computational complexity, or random generated data that can also be used for synchronization.

Let

$$\begin{aligned} \mathbf{H_p} &= [H_p(0)\ H_p(1)\ \cdots\ H_p(N_p-1)]^T \\ &= [H(0)\ H(L-1)\ \cdots\ H((N_p-1)\cdot L-1)]^T \end{aligned} \quad (9)$$

be the channel response of pilot subcarriers, and

$$\mathbf{Y_p} = [Y_p(0)\ Y_p(1)\ Y_p(N_p-1)]^T \quad (10)$$

be the vector of received pilot signals. The received pilot signal vector $\mathbf{Y_p}$ can be expressed as

$$\mathbf{Y_p} = \mathbf{X_p} \cdot \mathbf{H_p} + \mathbf{I_p} + \mathbf{W_p}, \quad (11)$$

where

$$\mathbf{X_p} = \begin{bmatrix} X_p(0) & & \mathbf{0} \\ & \ddots & \\ \mathbf{0} & & X_p(Np-1) \end{bmatrix},$$

$\mathbf{I_p}$ is the vector of ICI and $\mathbf{W_p}$ is the vector of Gaussian noise in pilot subcarriers.

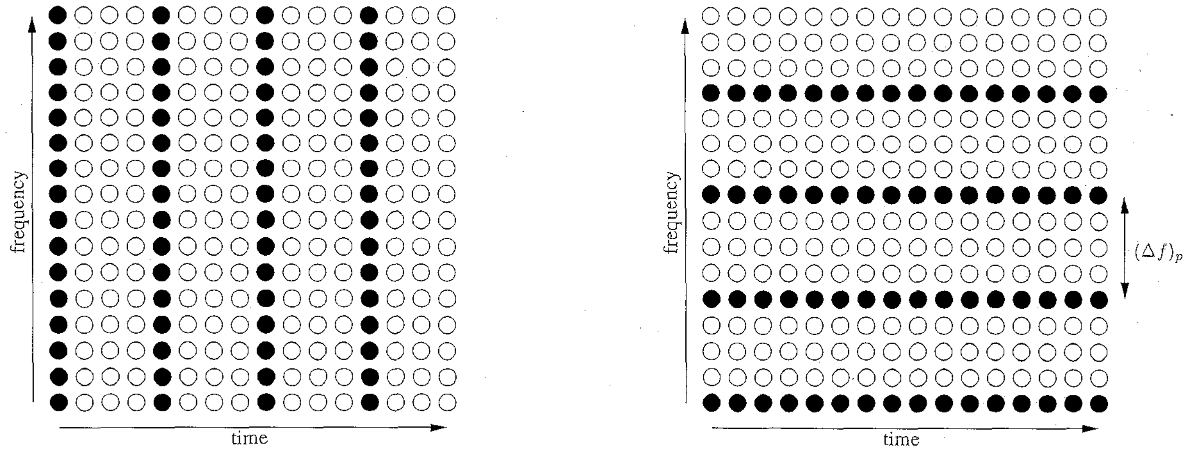
In conventional comb-type pilot based channel estimation methods, the estimate of pilot signals, based on least squares (LS) criterion, is given by:

$$\begin{aligned} \hat{\mathbf{H}}_{p,ls} &= [H_{p,ls}(0)\ H_{p,ls}(1)\ \cdots\ H_{p,ls}(N_p-1)]^T \\ &= \mathbf{X_p^{-1} Y_p} \\ &= \left[\frac{Y_p(0)}{X_p(0)}\ \frac{Y_p(1)}{X_p(1)}\ \cdots\ \frac{Y_p(N_p-1)}{X_p(N_p-1)}\right]^T. \end{aligned} \quad (12)$$

The LS estimate of $\mathbf{H_p}$ is susceptible to Gaussian noise and inter-carrier interference (ICI). Because the channel responses of data subcarriers are obtained by interpolating the channel responses of pilot subcarriers, the performance of OFDM system based on comb-type pilot arrangement is highly dependent on the rigorousness of estimate of pilot signals. Thus a estimate better than the LS estimate is required.

The minimum mean-square error (MMSE) estimate has been shown to be better than the LS estimate for channel estimation in OFDM systems based on block-type pilot arrangement [5]. Regarding the mean square error of the estimation shown in [5], the MMSE estimate has about 10–15 dB gain in SNR over the LS estimate for the same MSE values. The major drawback of the MMSE estimate is its high complexity, which grows exponentially with the observation samples. In [6], a low-rank approximation is applied to a linear minimum mean-squared error (LMMSE) estimator that uses the frequency correlation of the channel. In this paper, the same facilitating process is applied to estimate the comb-type pilot signals. The key idea to reduce the complexity is using the singular-value decomposition (SVD) to derive an optimal low-rank estimator, where performance is essentially preserved. The mathematical representation for MMSE estimator of pilot signals is as follows [6]:

$$\begin{aligned} \hat{\mathbf{H}}_{p,lmmse} &= \mathbf{R_{H_pH_{p,ls}}R^{-1}_{H_pH_{p,ls}}\hat{H}_{p,ls}} \quad (13) \\ &= \mathbf{R_{H_pH_p}}\left(\mathbf{R_{H_pH_p}} + \sigma_{\mathbf{n}}^2\left(\mathbf{X_pX_p^H}\right)^{-1}\right)^{-1}\mathbf{\hat{H}_{p,ls}}, \end{aligned}$$



(a) Block-type pilot sub-carrier arrangement

(b) Comb-type pilot sub-carrier arrangement

Fig. 2. Two different types of pilot sub-carrier arrangement.

where $\hat{\mathbf{H}}_{p,ls}$ is the least-squares estimate of $\mathbf{H_p}$ as shown in (12), σ_n^2 is the variance of $W(k)$, and the covariance matrices are defined by

$$\begin{aligned} \mathbf{R_{H_pH_p}} &= E\left\{\mathbf{H_p H_p^H}\right\}, \\ \mathbf{R_{H_pH_{p,ls}}} &= E\left\{\mathbf{H_p H_{p,ls}^H}\right\}, \\ \mathbf{R_{H_{p,ls}H_{p,ls}}} &= E\left\{\mathbf{H_{p,ls} H_{p,ls}^H}\right\}. \end{aligned}$$

Note that there is a matrix inverse involved in the MMSE estimator, which must be calculated every time. This problem can be solved by using static pilots such as $X_p(m) = c$ for $m = 0, 1, \cdots, N_p - 1$. A more generic solution is to average over the transmitted data, and a simplified linear MMSE estimator of pilot signals is obtained as [6]:

$$\hat{\mathbf{H}}_p = \mathbf{R_{H_pH_p}}\left(\mathbf{R_{H_pH_p}} + \frac{\beta}{\mathrm{SNR}}\mathbf{I}\right)^{-1}\hat{\mathbf{H}}_{\mathbf{p,ls}}, \qquad (14)$$

where $\mathrm{SNR} = E\left|X_p(k)\right|^2/\sigma_n^2$ is the average signal-to-noise ratio, and $\beta = E\left|X_p(k)\right|^2 E\left|1/X_p(k)\right|^2$ is a constant depending on the signal constellation. For 16-QAM transmission, $\beta = 17/9$. If the auto-correlation matrix $\mathbf{R_{H_pH_p}}$ and SNR are known in advance, $\mathbf{R_{H_pH_p}}\left(\mathbf{R_{H_pH_p}} + \frac{\beta}{\mathrm{SNR}}\mathbf{I}\right)^{-1}$ needs to be calculated only once. Although the simplified LMMSE estimator avoids the matrix inverse operation, the computational complexity is still very high. As shown in (14), the estimation requires N_p complex multiplications per pilot tone. To reduce the number of multiplication operations, a low-rank approximation using singular value decomposition is adopted [6]. The channel correlation matrix is first decomposed as

$$\mathbf{R_{H_pH_p}} = \mathbf{U\Lambda U^H} \qquad (15)$$

where $\mathbf{U}$ is a matrix with orthonormal columns $\mathbf{u_0}$, $\mathbf{u_1}$, ..., $\mathbf{u_{N_p-1}}$ and $\mathbf{\Lambda}$ is a diagonal matrix, containing the singular values $\lambda(0) \leq \lambda(1) \leq \ldots \leq \lambda(N_p - 1) \leq 0$ on its diagonal. The best rank-m approximation of the estimator in (14) then becomes

$$\hat{\mathbf{H}}_p = \mathbf{U}\begin{bmatrix} \mathbf{\Delta_m} & 0 \\ 0 & 0 \end{bmatrix}\mathbf{U^H}\hat{\mathbf{H}}_{\mathbf{p,ls}}, \qquad (16)$$

where $\mathbf{\Delta_m}$ is a diagonal matrix containing the values

$$\delta(k) = \frac{\lambda(k)}{\lambda(k) + \frac{\beta}{\mathrm{SNR}}}, \quad k = 0, 1, \ldots, m \qquad (17)$$

After some facilitations, the estimator in (16) thus requires $2mN_p$ multiplications and the total number of multiplications per pilot tone becomes $2m$. In general, the number of significant singular value m is much smaller than N_p, and the computational complexity is reduced considerably comparing to the full rank estimator (14).

The low-rank MMSE estimator is insensitive to parameter mismatch [6]. That is, it is possible to design a generic rank-m estimator for a wide range of channels.

IV. Channel Interpolation

After the estimation of the channel transfer functions of pilot tones, the channel responses of data tones can be interpolated according to adjacent pilot tones. The piecewise-linear interpolation method has been studied in [2], and is shown to be better than the piecewise-constant interpolation. In this paper, we consider a piecewise-linear and a piecewise second-order polynomial interpolation method which due to their inherent simplicity are easy to implement.

A. Linear interpolation method

In the linear interpolation algorithm, two successive pilot subcarriers are used to determine the channel response for data subcarriers that are located in between the pilots [2]. For data subcarrier k, $mL \leq k < (m+1)L$, the estimated channel response using linear interpolation method is given by

$$\begin{aligned} \hat{H}(k) &= \hat{H}(mL+l) = \left(1 - \frac{l}{L}\right)\hat{H}_p(m) + \frac{l}{L}\hat{H}_p(m+1) \\ &= \hat{H}_p(m) + \frac{l}{L}\left(\hat{H}_p(m+1) - \hat{H}_p(m)\right), 0 \leq l < L. \end{aligned} \qquad (18)$$

The linear channel interpolation can be implemented by using digital filtering such as the Farrow-structure [8]. Furthermore, by carefully inspecting (18), we find that if L is chosen as the power of 2, the multiplication operations involved in (18) can be replaced by shift operations, and therefore no multiplication operation is needed in the linear channel interpolation.

B. Second-order polynomial interpolation method

Theoretically, using higher-order polynomial interpolation will fit the channel response better than the linear interpolation. However, the computational complexity grows as the order is increased. Here we consider the second-order polynomial interpolation for its acceptable computational complexity. A piecewise second-order polynomial interpolation can be implemented as a linear time-invariant FIR filter [9]. The interpolator is given by

$$\begin{aligned} \hat{H}(k) &= \hat{H}(mL+l) \\ &= C_1\hat{H}_p(m-1) + C_0\hat{H}_p(m) + C_{-1}\hat{H}_p(m+1), \end{aligned} \tag{19}$$

where

$$\begin{cases} C_1 = \frac{\alpha(\alpha+1)}{2} \\ C_0 = -(\alpha-1)(\alpha+1) \\ C_{-1} = \frac{\alpha(\alpha-1)}{2} \end{cases}$$

and $\alpha = l/N$.

C. Phase compensated interpolation

For a realistic OFDM system, the frame position error will affect the channel estimation. Because of the low-pass filtering in the receiver, the sampled channel impulse is dispersed and introduces energy leaks, especially, when the path delay time is non-T-spaced [5]. Due to the multipath propagation and the energy leaks due to sampling, the guard interval will be affected by the preceding symbol. To avoid the inter-symbol interference, the frame synchronization must provide a portion $\{y(n)\}$ of the samples $\{y_g(n)\}$ of length N that is influenced by one transmitted symbol only. Let the correct position of the FFT window is $p = q(N_g + N)$ where q is a positive integer and p_ϵ the offset with respect to this position. In [10], a frame misalignment is identified as two situations as shown in Fig. 4.

Suppose that the start position of the FFT window is within region A, then no ISI occurs. The only effect suffered by the subchannel symbols is a change in phase that increases with the subcarrier index. If the start position is within regions B, the subcarrier symbols will suffer from ISI in addition to the phase rotation, and the orthogonality of the system is disturbed. In this paper, we assume that the coarse frame synchronization has pulled the frame start position p_ϵ within region A. Thus the received signal after FFT becomes [10]

$$\begin{aligned} Y(k) &= X(k)H'(k) + W'(k) \\ &= X(k)H(k)e^{-j2\pi(p_\epsilon - N_g)/N} \\ &\quad + W(k)e^{-j2\pi(p_\epsilon - N_g)/N}. \end{aligned} \tag{20}$$

From Fig. 4, we see that $p_\epsilon \leq N_g$ when p_ϵ is in region A. Therefore, a group delay exists at the OFDM receiver before demultiplexing. The group delay in time-domain introduces phase rotations after demultiplexing, which introduce model mismatch error for channel interpolation algorithms. As shown in Fig. 5, the linear interpolation algorithm introduces significant errors when the group delay is present. In addition, the model mismatch error grows as the group time delay increases. The model mismatch problem is more serious for the comb-type pilot system than the block-type pilot system, since the comb-type pilot system uses interpolation to determine the channel response of the data carriers, and the pilot carriers are separated more widely. In addition, the MMSE estimator discussed in Section III is sensitive to the timing error. The reason is that the MMSE estimator uses a pre-estimated channel statistics to estimate the pilot signals. If the frame position error is present, the MMSE estimator will fail due to mismatch of channel statistics.

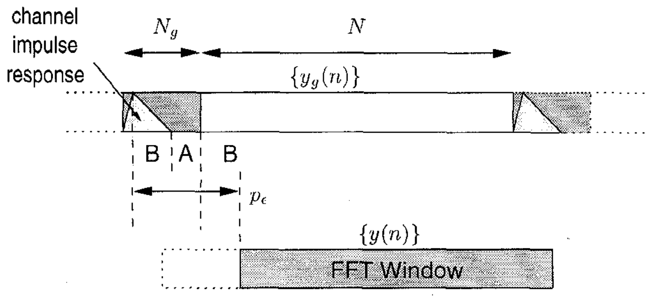


Fig. 4. Principle of frame synchronization [10].

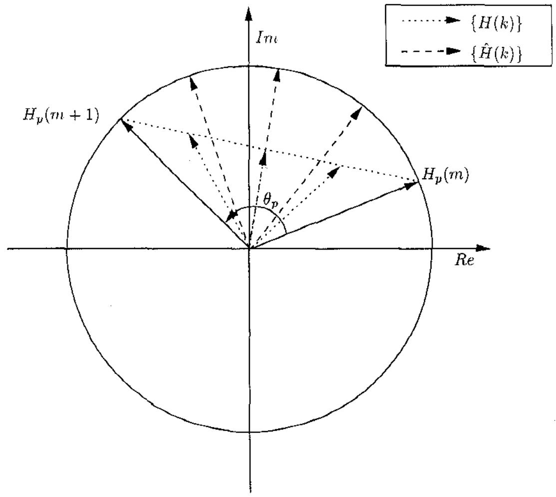


Fig. 5. The influence of frame position error to interpolations.

Here we propose a method to solve the model mismatch problem as shown in Fig. 6. The estimated channel transfer function of pilot signals when a group delay in time domain is present can be represented by

$$\hat{H}_p(m) = H_p(m) \cdot e^{-j\theta_p m} + E_{H_p}(m), \tag{21}$$

where θ_p is the change in phase caused by frame error and $E_{H_p}(m)$ is the estimation error. We can see that $\{\hat{H}_p(m)\}$ is a sequence that with a single frequency. Therefore, the estimators presented in [11] can be applied to estimate θ_p. The estimator in [11] that provides the best overall performance is

$$\hat{\theta}_p = \sum_{m=0}^{N_p-2} w_m \angle \hat{H}_p^*(m)\hat{H}_p(m+1) \tag{22}$$

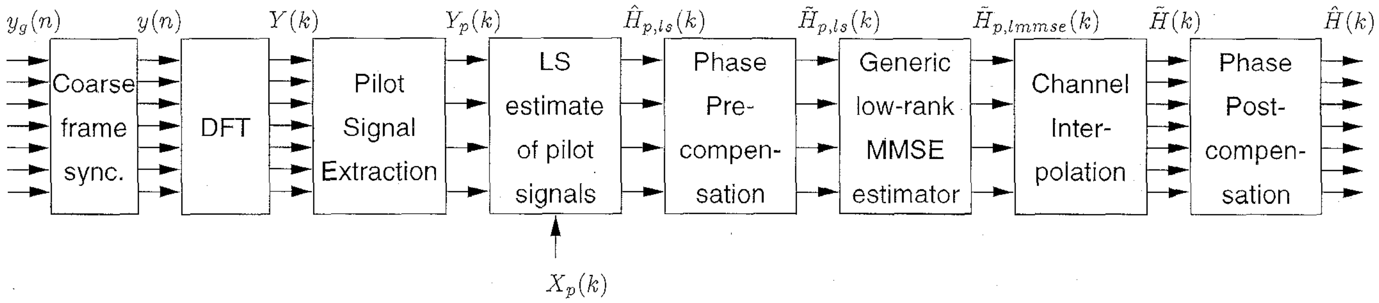


Fig. 6. Channel estimation with phase compensation.

where

$$w_m = \frac{\frac{3}{2}N_p}{N_p^2 - 1}\left\{1 - \left[\frac{m - \left(\frac{N_p}{2} - 1\right)}{\frac{N_p}{2}}\right]^2\right\}. \quad (23)$$

After $\hat{\theta}_p$ is obtained, the estimated channel transfer function of pilot subcarriers is compensated by it as:

$$\begin{aligned}\tilde{H}_p(m) &= \hat{H}_p(m) \cdot e^{j\hat{\theta}_p m} \\ &\approx H_p(m) + E'_{H_p}(m) \end{aligned} \quad (24)$$

where $E'_{H_p}(m) = E_{H_p}(m) \cdot e^{j\hat{\theta}_p m}$.

After the phase pre-compensation, the channel transfer functions of data carriers are interpolated by using linear or higher-order polynomial interpolations. Because the group delay is compensated by (24), the model mismatch error is reduced considerably. Then, after interpolation, the phase change is restored as:

$$\hat{H}(k) = \tilde{H}(k) \cdot e^{-jk\frac{\hat{\theta}_p}{L}}. \quad (25)$$

Fig. 7 shows the influence of frame position error on the channel estimation. The FFT size $N = 1024$, total number of pilot subcarriers N_p is 128, and E_b/N_0 equals 10 dB. The AWGN channel is adopted, with the delay time from 0 to 100 samples. From Fig. 7, we can see that interpolations without phase compensation will suffer from frame position error, even there is no ISI. The phase compensator is not affected by frame position error for $\theta_p < \pi$. When θ_p is larger than π (corresponding to the 64th sample shown in Fig. 7), the phase estimator suffers from phase ambiguity and therefore the compensator failed. However, since the phase change θ_p is more probable to be positive, the applicable range of the phase compensator can be enlarged by changing the range of $\hat{\theta}_p$ from $[-\pi, \pi)$ to $[-\pi + \phi, \pi + \phi)$ where $0 \leq \phi \leq \pi$.

From Fig. 7, we can see that the performances of linear and polynomial interpolations are not sensitive to frame position errors when the frame offset is small. This enlight us to use a simpler estimator of phase change instead of (22). We choose a simpler estimator from [11] and the phase estimator is

$$\hat{\theta}_p = \angle \frac{1}{N-1}\sum_{m=0}^{N_p-2} \hat{H}_p^*(m)\hat{H}_p(m+1). \quad (26)$$

According to our own simulations, the performances of channel estimator applying the phase compensators shown in (22) and (26) are almost the same.

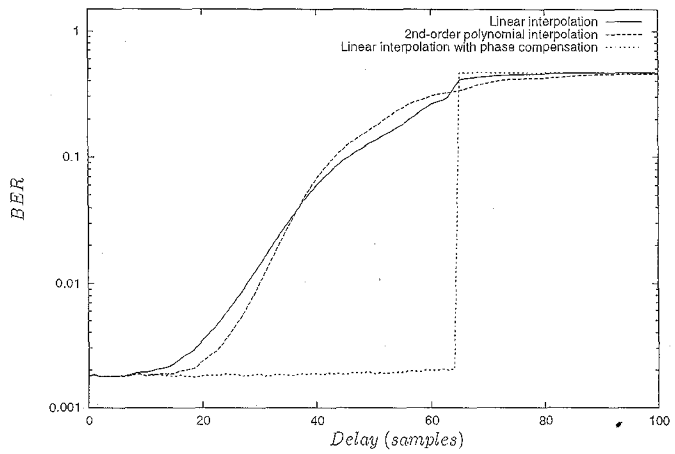


Fig. 7. The influence of group delay to the interpolation. E_b/N_0 = 10dB, N = 1024, N_p = 128.

V. Simulations

The proposed channel estimation method is evaluated under slow- and fast-fading radio channels. As in [3], the same simulation environments are used in this paper. An OFDM system with symbols modulated by 16-QAM is used. The carrier frequency is 1 GHz, and the bandwidth is 2 MHz. The total number of sub-channels is $N = 1024$, and the number of uniformly-distributed pilot carriers is $N_p = 128$. The channel models adopted are Rayleigh and Rician as recommended by GSM Recommendations 05.05 [12], with parameter shown in TABLE I. Here we assume that the guard intervals are longer than the maximum delay spread of the channel. To avoid ISI arising from energy leakages caused by sampling, a frame position offset of 30 samples is applied. The sampled channel impulse response in time-domain is shown in Fig. 8. We can see that there are energy leakages ahead the main path sample.

Several channel estimation methods, LS estimate of pilot with linear interpolation, MMSE estimate of pilot with linear interpolation, MMSE estimate of pilot with 2nd-order polynomial interpolation, MMSE estimate of pilot with phase compensated interpolations, and channel estimation using transform domain processing technique, are simulated and compared. The simulations are carried out for different noise and ICI levels, and the results are shown in Fig. 9–12. The horizontal variable is chosen as the signal energy per bit-to-noise power density ratio (E_b/N_0). The ICI levels are controlled by the vehicle speed.

	Rayleigh Channel		*Rician Channel*	
Path Number	Average Power (dB)	Delay (μs)	Average Power (dB)	Delay (μs)
1	-3.0	0.0	0.0	0.0
2	0.0	0.2	-4.0	0.1
3	-2.0	0.5	-8.0	0.2
4	-6.0	1.6	-12.0	0.3
5	-8.0	2.3	-16.8	0.4
6	-10.0	5.0	-20.0	0.5

TABLE I
PARAMETERS OF CHANNELS.

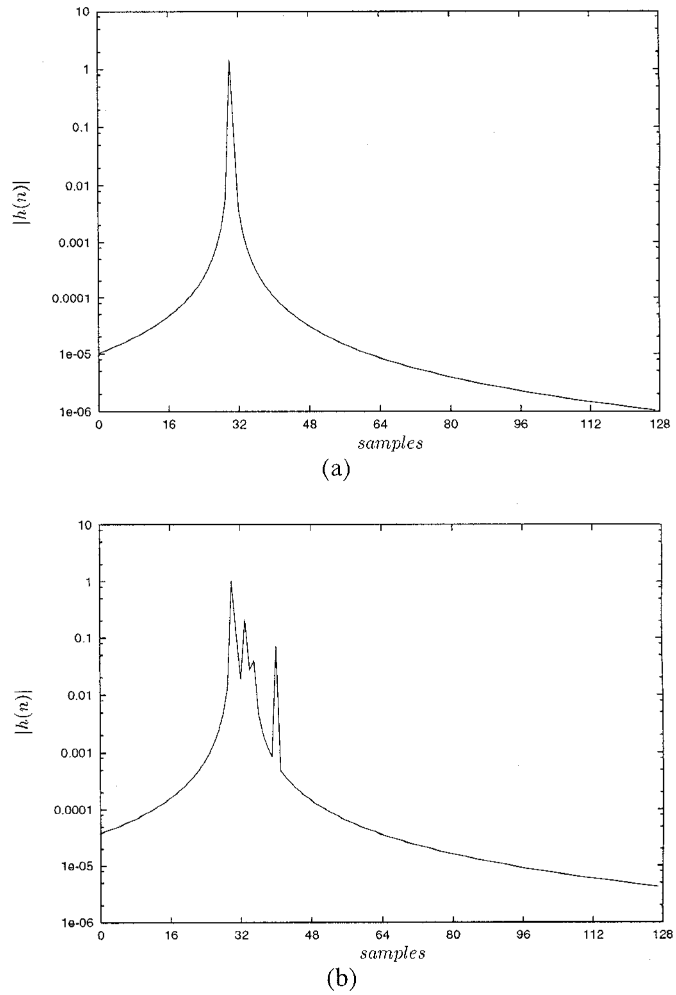


Fig. 8. Typical channel impulse of (a) Rician and (b) Rayleigh fading channels.

For low vehicle speed $V = 6$ km/hr, the corresponding maximum Doppler spread is 0.3 % of the subcarrier spacing. Thus, the ICI can be neglected. For high vehicle speed $V = 150$ km/hr, the corresponding maximum Doppler spread is then 7.5 % of the subcarrier spacing. Thus, the system performances are strongly affected by ICI. After decomposition of the autocorrelation matrices of Rayleigh and Rician channels, we find the sufficient rank of MMSE estimators are are 6 and 4, respectively. However, when we use the rank-6 estimator designed for Rayleigh fading channel as the pilot signal estimator of Rician fading channel, the performance degradation is neglectable. Thus, a generic low-rank estimator can be used for slow- and fast-fading Rayleigh and Rician fading channels.

From Figs. 9–12, we see that the performance is improved significantly when the low-rank MMSE estimator is applied. The phase compensated interpolation successfully solves the model mismatch problem, and gains a significant improvement in E_b/N_0 for the same BER values in all cases. We also notice that the higher-order polynomial interpolation only improves the performance little. It is suspectable to use higher-order polynomial interpolation instead of linear interpolation for channel with small delay spread. However, we expect that the improvement in performance will be more if the delay spread of channel is larger. The performance of Zhao and Huang's algorithm is limited by the low-pass filtering in transform domain. In higher SNR, the low-pass filtering operation discards some channel information and thus a BER floor is formed.

From the simulations shown above, we see for comb-type pilot assisted OFDM systems, the channel estimation based low-rank MMSE pilot estimator with phase compensated linear interpolation is suggested. The proposed method eliminate ICI and AWGN by using the low-rank MMSE pilot estimation, and the model mismatch error is reduced by using linear interpolation with phase compensation. It should be note that although the proposed algorithm provided better performance, the SNR and channel statistics have to be estimated in advance. Whereas the channel estimation based on transform domain processing needs not to know these informations in advance.

VI. COMPUTATIONAL COMPLEXITY

In this section, we compare the computational complexity of the proposed algorithm and the channel estimation algorithm presented in [3]. We evaluate the computational complexity roughly by calculate the number of complex multiplications used per tone to estimate the channel transfer function. The proposed method uses low-rank MMSE estimator, which requires $2mN_p$ complex multiplications where m is the rank of the estimator. The phase estimator requires N_p complex multiplications, and the phase compensator requires N complex multiplications. As shown in (18), the linear interpolation requires no multiplications. Thus, the proposed method totally requires $2mN_p + N_p + N$ complex multiplications, which is equivalent to $1 + \frac{(2m+1)N_p}{N}$ complex multiplications per tone. The channel estimator based on transform domain processing uses two FFTs to estimate the channel response, one is of size N_p and the other is of size N. By using the decimation-in-frequency radix-4 implementation of FFT, the total number complex multiplications required is roughly $N_p \log_4 N_p + N \log_4 N$. Note that about one or two multiplications can be saved for FFT because of some trivial multiplications. By using the parameters in Section V, the proposed channel estimator only requires about 2 complex multiplications per tone, while the channel estimator proposed by [3] requires at least 4–5 complex multiplications per tone by conservative estimate. Therefore, the proposed channel estimator is computational efficient and is attractive for

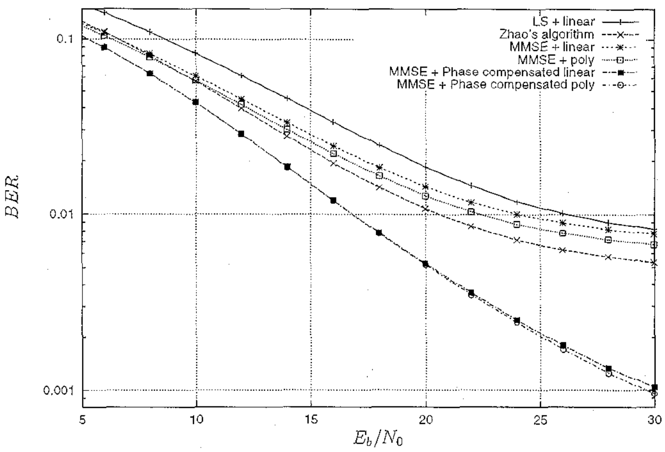


Fig. 9. Comparisons of BERs in slow Rayleigh fading channel (vehicle speed = 6 km/hr).

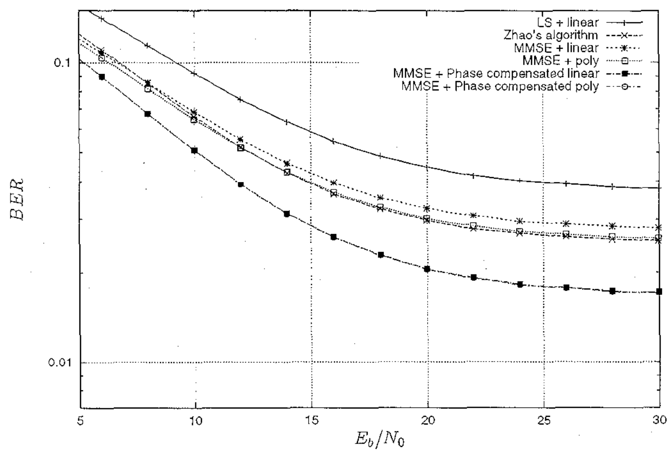


Fig. 10. Comparisons of BERs in fast Rayleigh fading channel (vehicle speed = 150 km/hr).

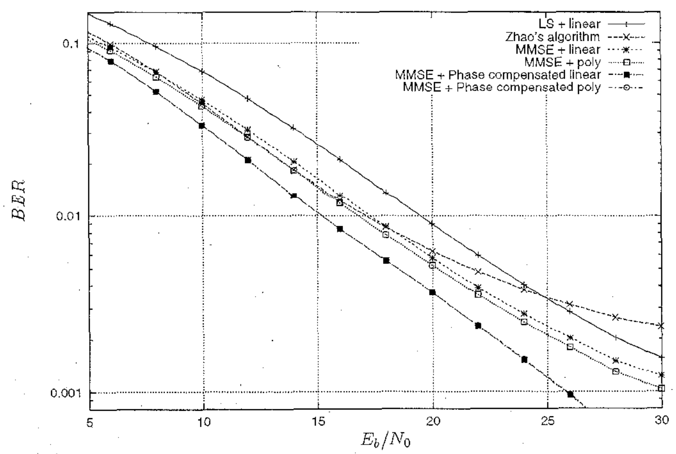


Fig. 11. Comparisons of BERs in slow Rician fading channel (vehicle speed = 6 km/hr).

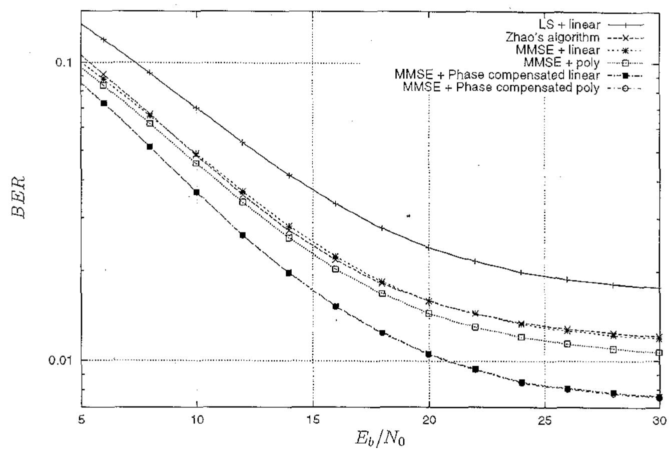


Fig. 12. Comparisons of BERs in fast Rician fading channel (vehicle speed = 150 km/hr).

OFDM systems in mobile communications.

VII. Conclusion

In this paper, a low-complexity channel estimator based on comb-type pilot arrangement is presented. The channel transfer function of pilot tones are estimated by using low-rank MMSE estimator, and the channel transfer function of data tones are interpolated by piecewise linear interpolation method with phase compensation. Simulations show that the performance of the proposed channel estimator outperforms the channel estimator based on transform domain processing presented by Zhao and Huang [3]. The computational complexity of the proposed channel estimator is also studied, and is lower than the previous channel estimator in [3]. In conclusion, the proposed channel estimator provides a practical channel estimation for OFDM systems based on comb-type pilot arrangement.

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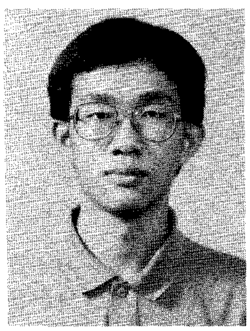
BIOGRAPHY



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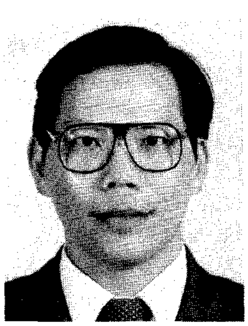

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