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PAPER

Design of Implementation-Efficient Channel Estimation Filters for Wireless Transmission of OFDM Signal

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SUMMARY The detection performance of coherent OFDM receivers significantly depends on the accuracy of channel estimation. The accuracy of channel estimation can be improved by properly post-processing the channel estimate using a so-called channel estimation filter (CEF). Minimum mean-squared error (MMSE) filter is known optimum as the CEF, but it may not be practical due to its implementation complexity. We consider the use of a reduced-complexity CEF whose tap coefficient is real-valued and symmetrically weighted (RSW). The optimum RSW CEF is analytically designed using the SNR and multi-path intensity profile of the channel. For further improvement, we also propose a method to adapt the coefficient of the RSW CEF to the channel condition. Numerical results show that the proposed RSW CEF can provide channel estimation performance comparable to that of linear MMSE filter, while significantly reducing the computational complexity. In addition, the proposed RSW CEF can provide performance robust to unknown timing offset with a fractional dB loss compared to the optimum one.

key words: *OFDM, coherent detection, channel estimation, adaptive CEF, timing offset*

1. Introduction

As the demand for data communications rapidly increases with explosive growth of the Internet access and interactive multimedia services, broadband wireless access techniques are emerging as an attractive and economical alternative to wireline access technologies [1], [2]. The use of multi-level QAM signals has been considered to maximize the spectral efficiency, providing a data rate of tens of Mbps. However, there exist fundamental challenges in high data rate transmission over a radio propagation channel. The receiver should mitigate the inter-symbol interference (ISI) due to multi-path propagation.

Orthogonal frequency division multiplexing (OFDM) is a promising modulation technique that can mitigate the ISI caused by multi-path propagation [3], [4]. The detrimental effect of ISI can be avoided by inserting a cyclic prefix (CP) whose span is longer than that of the channel impulse response (CIR). The OFDM signal can be detected differentially or coherently at the receiver. Coherent detection is generally preferred in modern OFDM transmission systems due to its better detection performance [5] and it can easily be combined with the use of advanced multi-antenna schemes such as the space-time coding (STC) or multi-input multi-output (MIMO) system [6], [7]. However, the performance of coherent detection significantly depends upon the accuracy of channel information.

There have been a number of research activities on channel estimation method, including the scattered pilot-aided, pilot symbol-aided and data-aided method [1], [5], [8]. In continuous signal transmission mode such as the European terrestrial digital video broadcasting (DVB-T) system, the pilot sub-carriers are regularly scattered over the time and frequency grid [5]. The channel gain of other sub-carriers can be obtained by interpolating the channel information obtained at the pilot location. In the packet transmission mode such as the wireless local area network (WLAN) system, a dedicated pilot symbol is transmitted at the head of data burst to simultaneously extract the channel information of all sub-carriers [1]. In the data-aided channel estimation method, the channel information is extracted by removing the data information from the received signal [8]. It is usually employed as a complement to the pilot-aided channel estimation method, when the initial channel estimate needs to be fine-tuned or when slow channel variation is tracked. In this paper, we consider a pilot symbol-aided channel estimation method in packet mode transmission of OFDM signal.

The accuracy of channel estimation can be improved by properly post-processing the channel information extracted from the pilot symbol. The simplest method is to transmit multiple pilot symbols and to average the channel information extracted from each pilot symbol. However this method degrades the transmission efficiency, particularly when the packet length is short. An alternative method is to average the channel information of the adjacent sub-carriers in the frequency-domain. Hoeher proposed a frequency domain linear optimum filter in the minimum mean squared-error (MMSE) sense [9]. Beek proposed a channel estimation scheme that first transforms the frequency domain estimate into a time-domain one and then scales it with a weight associated with the signal-to-noise power ratio (SNR) [10]. However, these schemes may require large computational complexity for filtering and additional fast Fourier transform (FFT) and inverse FFT processes. Furthermore, it may not be easy to optimize the filter coefficient in real-time

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as the number of filter taps increases.

To reduce the implementation complexity, we consider the use of a CEF whose tap coefficient is real-valued and symmetrically weighted (RSW). A special form of the RSW CEF was considered in [11], where the filter coefficient is determined in a form of 2^{-k} to simplify the multiplication process. The CEF is selected among a set of pre-designed filters based on the measured channel characteristics. However, this CEF was designed in a heuristic manner, leaving a large room for further improvement.

We analytically design the RSW CEF so as to minimize the mean-squared-error (MSE) of the channel estimate by deriving the MSE as a function of the CEF coefficient. Assuming that the scattering characteristics of the channel are uncorrelated wide-sense stationary, the optimum RSW CEF is analytically designed in terms of the multi-path intensity profile and SNR of the channel.

The information on the channel characteristics may not be available in real environment since it is location-dependent and time-variant. When a CEF with fixed coefficient is employed, it is usually designed under the worst channel condition, making it robust to a wide range of channel condition [12], [13]. However, such a channel estimator may not fully exploit the correlation between the adjacent sub-carriers. We design an adaptive RSW CEF whose coefficient is adjusted in real time by exploiting the channel gain difference between the adjacent sub-carriers.

Following Introduction, we model the wireless OFDM transmission system in Sect. 2. The proposed RSW CEF is analytically designed in Sect. 3. We also consider the adaptation of the filter coefficient according to the channel condition. The analytic design of the proposed RSW CEF is verified by computer simulation in Sect. 4. Finally, conclusions are summarized in Sect. 5.

2. System Model

Consider an OFDM transmission system whose baseband equivalent model is depicted in Fig. 1. In the transmitter, N QAM symbols $\{X_k\}$ are converted into the time domain by the IFFT. A cyclic prefix (CP) is added to protect the OFDM signal from the ISI due to multi-path propagation. Preamble symbols are included at the head of each packet for the purpose of synchronization and channel estimation.

We assume a wireless channel consisting of L multi-paths

$$h(\tau) = \sum_{l=0}^{L-1} h_l \delta(\tau - \tau_l T_s) \tag{1}$$

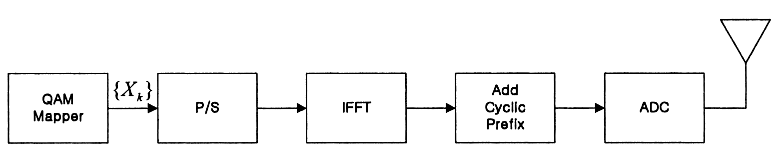
where h_l is the gain of the l-th path which can be modeled as a complex Gaussian random variable, T_s is the sampling interval, $\tau_l T_s$is the delay of the l-th path and $\delta(\cdot)$ is the Kronecker delta function. It is also assumed that the channel characteristics are not changed during the transmission of a single packet but can be changed randomly between the packets. The channel impulse responses with different delay are assumed to be mutually independent, i.e.,

$$E\{h_l h_{l'}^*\} = A_l \delta(l - l') \tag{2}$$

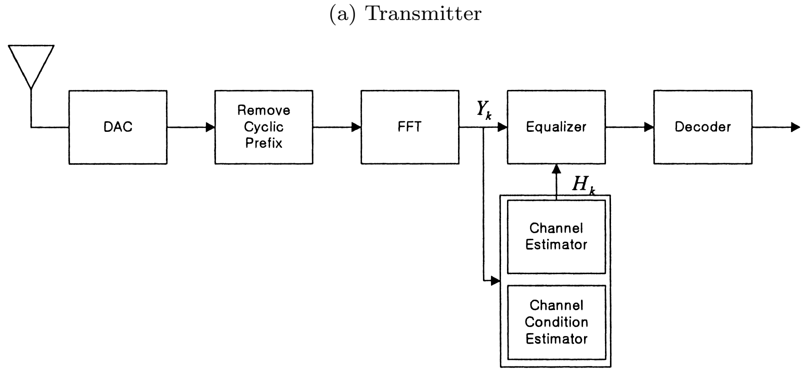
where $E\{X\}$ denotes the expectation of X. The multi-path intensity profile of the channel is defined by $\{A_l, \tau_l\}$.

The CP is removed before the FFT process in the receiver. Assuming ideal synchronization at the receiver, the signal at the k-th sub-carrier after the FFT can be represented by

$$Y_k = X_k + Z_k \tag{3}$$



(a) Transmitter



(b) Receiver

Fig. 1 OFDM transmission system with channel estimation.

where H_k is the channel gain of the k-th sub-carrier

$$H_k = \sum_{l=0}^{L-1} h_l e^{-j2\pi k\tau_l/N} \tag{4}$$

and Z_k is zero mean additive white Gaussian noise (AWGN) with variance σ_Z^2. In vector notation, it can be re-written as

$$\mathbf{Y} = \mathbf{XH} + \mathbf{Z} \tag{5}$$

where $\mathbf{Y} = [Y_0, \cdots, Y_{N-1}]^T$, $\mathbf{X} = diag(X_0, \cdots, X_{N-1})$ and $\mathbf{H} = [H_0, \cdots, H_{N-1}]^T$. Here, $diag(\cdot)$ denotes the diagonal matrix and the superscript T denotes the transpose of a vector. By using (2), it can be shown that the channel correlation function between the sub-carriers is represented as

$$E\{H_k H_{k-m}^*\} = \sum_{l=0}^{L-1} A_l e^{-j2\pi m\tau_l/N}. \tag{6}$$

3. Design of the Channel Estimation Filter

3.1 Linear MMSE CEF

The linear MMSE channel estimate is obtained by [5], [10]

$$\hat{H}_{k,LMMSE} = \mathbf{p}_k^H (\mathbf{R}_{hh} + \sigma_Z^2 \mathbf{I})^{-1} \hat{\mathbf{H}}_{LS} \tag{7}$$

where $\mathbf{R}_{hh} = E\{\mathbf{H}\mathbf{H}^H\}$, $\mathbf{p}_k = E\{H_k^* \mathbf{H}\}$ and $\hat{\mathbf{H}}_{LS} = [\hat{H}_{0,LS}, \cdots, \hat{H}_{N-1,LS}]^T$ is the least-square (LS) estimate of the channel. Here, the superscript H denotes the Hermitian transpose. The corresponding MSE is given by

$$\sigma_{LMMSE}^2 = E\{|H_k|^2\} - \mathbf{p}_k^H (\mathbf{R}_{hh} + \sigma_Z^2 \mathbf{I})^{-1} \mathbf{p}_k. \tag{8}$$

The LS estimate of the channel gain can be obtained by [10]

$$\hat{H}_{k,LS} = Y_k / X_k = H_k + Z_k' \tag{9}$$

where $Z_k' = Z_k / X_k$ denotes the LS estimation error. Assuming that $|X_k|^2 = 1$ for all k, Z_k' is also zero mean AWGN with variance σ_Z^2.

3.2 Real-Valued Symmetrically Weighted (RSW) CEF

Although linear MMSE filter can provide the best estimation performance among linear CEFs, it requires a large number of computations such as inversion of a complex-valued matrix. As an alternative CEF, we consider a reduced-complexity CEF whose tap coefficient is real-valued and symmetrically weighted (RSW). Figure 2 depicts the structure of the proposed RSW CEF. Without timing offset correction, the channel gain of the k-th sub-carrier can be estimated by

$$\begin{aligned}\hat{H}_{k,RSW} &= \left(1 - \sum_{m=1}^{M} \alpha_m\right) \hat{H}_{k,LS} \\ &\quad + \frac{1}{2} \sum_{m=1}^{M} \alpha_m (\hat{H}_{k-m,LS} + \hat{H}_{k+m,LS})\end{aligned} \tag{10}$$

where $\alpha_m, m = 1, 2, \cdots, M$, is real-valued filter coefficient and $(2M+1)$ is the filter tap size. Using (4) and (9), (10) can be rewritten as

$$\begin{aligned}\hat{H}_{k,RSW} &= H_k + \sum_{m=1}^{M} \alpha_m \sum_{l=0}^{L-1} h_l \lambda_{l,m} e^{-j2\pi k\tau_l/N} \\ &\quad + \left(1 - \sum_{m=1}^{M} \alpha_m\right) Z_k' \\ &\quad + \frac{1}{2} \sum_{m=1}^{M} \alpha_m (Z_{k-m}' + Z_{k+m}')\end{aligned} \tag{11}$$

where the first term is the desired channel gain, the second term is the interference from the adjacent sub-carriers and the third and fourth terms are the averaged LS estimation error. Here, $\lambda_{l,m} = \cos(2\pi\tau_l m/N) - 1$.

Assuming the channel condition (2), it can easily be shown that the variance of the interference is

$$\sigma_i^2(\underline{\alpha}) = \sum_{l=0}^{L-1} A_l \left[\sum_{m=1}^{M} \alpha_m \lambda_{l,m}\right]^2 \tag{12}$$

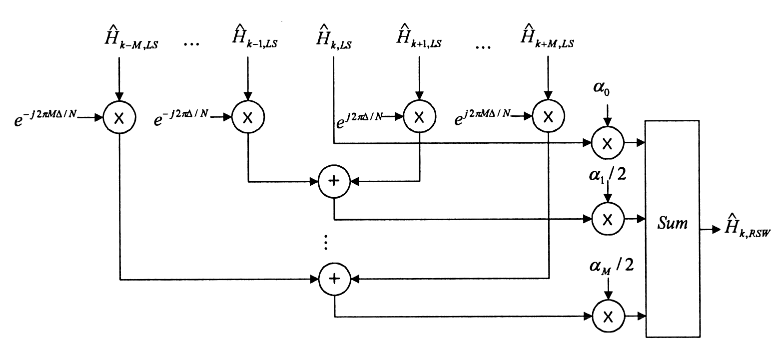


Fig. 2 Structure of the RSW CEF with timing offset correction.

where $\underline{\alpha} = [\alpha_1, \alpha_2, \cdots, \alpha_M]^T$ denotes the coefficient vector. The variance of the LS estimation error is

$$\sigma_e^2(\underline{\alpha}) = \left[\left(1 - \sum_{m=1}^{M} \alpha_m \right)^2 + \frac{1}{2} \sum_{m=1}^{M} \alpha_m^2 \right] \sigma_Z^2. \quad (13)$$

Then, the total estimation error is given by

$$\sigma_t^2(\underline{\alpha}) = \sigma_i^2(\underline{\alpha}) + \sigma_e^2(\underline{\alpha}) \quad (14)$$

The optimum RSW CEF can be designed by determining the coefficient $\underline{\alpha}$ so as to minimize $\sigma_t^2(\underline{\alpha})$. Since $\sigma_t^2(\underline{\alpha})$ is a quadratic function of $\underline{\alpha}$, the optimum coefficient can uniquely be determined by solving

$$\begin{aligned} \left. \frac{\partial \sigma_t^2(\underline{\alpha})}{\partial \alpha_p} \right|_{\underline{\alpha}=\hat{\underline{\alpha}}} &= \sum_{m=1}^{M} \hat{\alpha}_m \sum_{l=0}^{L-1} 2A_l \lambda_{l,m} \lambda_{l,p} \\ &+ \left[-2 \left(1 - \sum_{m=1}^{M} \hat{\alpha}_m \right) + \hat{\alpha_p} \right] \sigma_Z^2 = 0 \end{aligned} \quad (15)$$

for $p = 1, 2, \cdots, M$. It can be shown that the optimum RSW CEF coefficient is given by

$$\hat{\underline{\alpha}}_{opt} = \mathbf{\Phi^{-1} c} \quad (16)$$

where $\mathbf{\Phi}$ is an $(M \times M)$ matrix whose (i, j)-th element is

$$\phi_{i,j} = \begin{cases} \sigma_Z^2 + \sum_{l=0}^{L-1} A_l \lambda_{l,i} \lambda_{l,j}, & i \neq j \\ 1.5\sigma_Z^2 + \sum_{l=0}^{L-1} A_l \lambda_{l,i}^2, & i = j \end{cases} \quad (17)$$

and $\mathbf{c}$ is an $(M \times 1)$ column vector whose i-th element is $c_i^2 = \sigma_Z^2$.

It can be seen that the optimum coefficient is mainly determined by the channel multi-path intensity profile $\{A_l, \tau_l\}$ and the noise variance σ_Z^2. Note that a $(2M+1)$-th order RSW CEF can be designed by solving an $(M \times M)$ real-valued matrix equation, while the equivalent linear MMSE filter is designed by solving a $(2M \times 2M)$ complex-valued one.

3.3 Adaptive RSW CEF

In real communication environment, the information on the channel characteristics may not be available since it is location-dependent and time-variant. As a practical solution to this problem, we design the RSW CEF whose coefficient is adjusted according to the estimated channel condition. The channel characteristics (e.g., the channel multi-path intensity profile and SNR) can be estimated from the CIR. However, it may require an additional correlation block in the time-domain to estimate the CIR. Moreover, the calculation of the filter coefficient by (16) requires matrix inversion. Since the root-mean-squared (RMS) delay spread τ_{rms} has close relationship with the multi-path intensity profile, we use τ_{rms} as a measure of the channel condition.

In an exponentially decaying channel whose multipath intensity profile is given by

$$A(\tau) = \frac{1}{\tau_{rms}} e^{-\tau/\tau_{rms}}, \quad (18)$$

the RMS delay spread τ_{rms} can be estimated from the channel gain difference between the adjacent subcarriers. Let Λ_m be the normalized LS channel gain difference between the sub-carriers differed by m subcarriers,

$$\Lambda_m = \frac{\sum_{k=m}^{N-1} \left| \hat{H}_{k,LS} - \hat{H}_{k-m,LS} \right|^2}{\sum_{k=m}^{N-1} \left| \hat{H}_{k,LS} \right|^2}. \quad (19)$$

Since the LS estimation error at each sub-carrier is uncorrelated, the ensemble average of Λ_m can be written as

$$\begin{aligned} \bar{\Lambda}_m &\approx \frac{E\left\{ \sum_{k=m}^{N-1} \left| \hat{H}_{k,LS} - \hat{H}_{k-m,LS} \right|^2 \right\}}{E\left\{ \sum_{k=m}^{N-1} \left| \hat{H}_{k,LS} \right|^2 \right\}} \\ &= \frac{\sum_{l=0}^{L-1} A_l \left| 1 - e^{j2\pi m \tau_l / N} \right|^2 + 2\sigma_Z^2}{\sum_{l=0}^{L-1} A_l + \sigma_Z^2} \end{aligned} \quad (20)$$

Here, the expectation of the rational is approximated by the ratio of the expectation of the numerator and denominator. Note that $\bar{\Lambda}_m$ can be represented as a function of the channel multi-path intensity profile $\{A_l, \tau_l\}$ and σ_Z^2, or equivalently τ_{rms} and σ_Z^2. Numerical results indicate that $\bar{\Lambda}_m$ increases as τ_{rms} increases or the SNR decreases. Since τ_{rms} and SNR affect $\bar{\Lambda}_m$, they can simultaneously be estimated using two values of Λ_m.

Note that τ_{rms} and σ_Z^2 can be estimated separately from the property of the difference between Λ_{m+d} and Λ_m. Let us define $\Lambda_{m,d} = \Lambda_{m+d} - \Lambda_m$. Then, the ensemble average of $\Lambda_{m,d}$

$$\bar{\Lambda}_{m,d} \approx \frac{\sum_{l=0}^{L-1} A_l \left\{ \left| 1 - e^{j2\pi(m+d)\tau_l/N} \right|^2 - \left| 1 - e^{j2\pi m \tau_l / N} \right|^2 \right\}}{\sum_{l=0}^{L-1} A_l + \sigma_Z^2} \quad (21)$$

Since $\bar{\Lambda}_{m,d}$ is almost independent of the SNR unless the SNR is too low, τ_{rms} can be estimated from $\Lambda_{m,d}$. Then, the normalized noise variance is estimated by

$$\hat{\sigma}_Z^2 = \frac{\Lambda_{m+d} - \chi_{m+d}(\hat{\tau}_{rms}) + \Lambda_m - \chi_m(\hat{\tau}_{rms})}{4} \quad (22)$$

where $\hat{\tau}_{rms}$ is the estimated RMS delay spread and $\chi_m(\cdot)$ is a function of τ_{rms}, equal to $\bar{\Lambda}_m$ when $\sigma_Z^2 = 0$. Based on $\hat{\tau}_{rms}$ and $\hat{\sigma}_Z^2$, the coefficient of the RSW CEF can be adjusted adaptively.

3.4 RSW CEF in the Presence of Timing Offset

The performance of the OFDM receiver is known quite insensitive to the timing offset [5], [14]. If the amount of timing offset is less than the span of the CP uncorrupted by ISI, it does not cause any performance degradation. In practice, conventional timing synchronization methods add a small amount of delay to the estimated timing epoch to provide a margin for variation of the timing estimate [14]. However, the timing offset causes a phase rotation of the channel, which is linearly proportional to the frequency index. This phase rotation may result in erroneous estimate of τ_{rms}, leading to improper CEF selection. Therefore, the timing offset should be compensated before the channel estimation.

Let ΔT_s be the timing offset introduced in the receiver. Then, the channel gain becomes

$$H_{k,\Delta} = \sum_{l=0}^{L-1} h_l e^{-j2\pi k(\tau_l+\Delta)}/N = H_k e^{-j2\pi k\Delta/N} \tag{23}$$

and the channel correlation between the sub-carriers differed by m becomes

$$E\left\{H_{k,\Delta}H^*_{k-m,\Delta}\right\} = \sum_{l=0}^{L-1} A_l e^{-j2\pi m(\tau_l+\Delta)/N}. \tag{24}$$

It can easily be shown that the linear MMSE estimate with timing offset ΔT_s becomes

$$\begin{aligned}\hat{H}_{k,\Delta,LMMSE} &= (\mathbf{R}_{hh} + \sigma_Z^2\mathbf{I})^{-1}\mathbf{q}\hat{\mathbf{H}}_{LS} \\ &= (\mathbf{R}_{hh} + \sigma_Z^2\mathbf{I})^{-1}\hat{\mathbf{H}}_{LS,\Delta}\end{aligned} \tag{25}$$

where

$$\begin{aligned}\mathbf{q} = diag\Big(&e^{-j2\pi M\Delta/N}, e^{-j2\pi(M-1)\Delta/N}, \cdots, \\ &e^{j2\pi(M-1)\Delta/N}, e^{j2\pi M\Delta/N}\Big)\end{aligned} \tag{26}$$

and $\hat{\mathbf{H}}_{LS,\Delta}\left(=\mathbf{q}\hat{\mathbf{H}}_{LS}\right)$ is the LS channel estimate after the timing offset compensation. It can be seen that the coefficient of a linear MMSE CEF is unchanged as long as the phase rotation due to the timing offset is compensated before the filtering.

The output of the RSW CEF in the presence of timing offset can be represented as

$$\begin{aligned}\hat{H}_{k,RSW} = &\Bigg[\left(1-\sum_{m=1}^{M}\alpha_m\right)H_k \\ &+\frac{1}{2}\sum_{m=1}^{M}\alpha_m\left(H_{k-m}e^{j2\pi m\Delta/N} + H_{k+m}e^{-j2\pi m\Delta/N}\right)\Bigg] \\ &\cdot e^{-j2\pi k\Delta/N} + \left(1-\sum_{m=1}^{M}\alpha_m\right)Z'_k \\ &+\frac{1}{2}\sum_{m=1}^{M}(Z'_{k-m} + Z'_{k+m})\end{aligned} \tag{27}$$

where $e^{\pm j2\pi m\Delta/N}$ in the second summation represents the phase rotation due to timing offset. It can be inferred that, similar to the linear MMSE CEF, the coefficient of the optimum RSW CEF is preserved if these phase rotations are compensated before the filtering. Figure 2 depicts the RSW CEF with timing offset compensation.

When the timing offset is not compensated, the phase rotation of the channel gain due to the timing offset may not be distinguished from the channel variation due to large delay spread. As a result, even though the actual delay spread is small, the RSW CEF coefficient can be determined for a channel with a large delay spread.

We can estimate the timing offset by finding the phase correction that minimizes the channel gain difference between the adjacent sub-carriers. For a presumed timing offset Δ', we define the phase-corrected channel gain difference by

$$\Lambda_m^{\Delta'} \approx \frac{\sum_{k=m}^{N-1}\left|\hat{H}_{k,LS} - \hat{H}_{k-m,LS}e^{j2\pi m\Delta'/N}\right|^2}{\sum_{k=m}^{N-1}\left|\hat{H}_{k,LS}\right|^2} \tag{28}$$

The ensemble average of $\bar{\Lambda}_m^{\Delta'}$ can be approximated as

$$\begin{aligned}\bar{\Lambda}_m^{\Delta'} &= \frac{E\left\{\sum_{k=m}^{N-1}\left|\hat{H}_{k,LS} - \hat{H}_{k-m,LS}e^{j2\pi m\Delta'/N}\right|^2\right\}}{E\left\{\sum_{k=m}^{N-1}\left|\hat{H}_{k,LS}\right|^2\right\}} \\ &= \frac{\sum_{l=0}^{L-1}A_l\left|1-e^{j2\pi m(\tau_l+\Delta-\Delta')}\right|^2 + 2\sigma_Z^2}{\sum_{l=0}^{L-1}A_l + \sigma_Z^2}\end{aligned} \tag{29}$$

Comparing (29) to (20), it can be seen that the phase correction has the effect of shifting the delay of the l-th path from $\tau_l + \Delta$ to $\tau_l + \Delta - \Delta'$. Since all A_ls are real and positive, and $\left|1-e^{j2\pi m(\tau_l+\Delta-\Delta')/N}\right|^2$ has its minimum at $\Delta' = \Delta$, $\bar{\Lambda}_m^{\Delta'}$ has its minimum when $\Delta' = \Delta$. Therefore, the timing offset can be estimated by finding that minimizes $\Lambda_m^{\Delta'}$, i.e.,

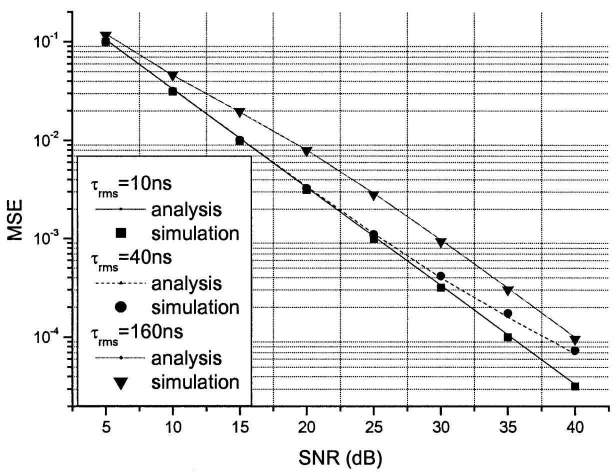
$$\hat{\Delta} = \arg\min_{\Delta'}\Lambda_m^{\Delta'} \tag{30}$$

4. Performance Evaluation

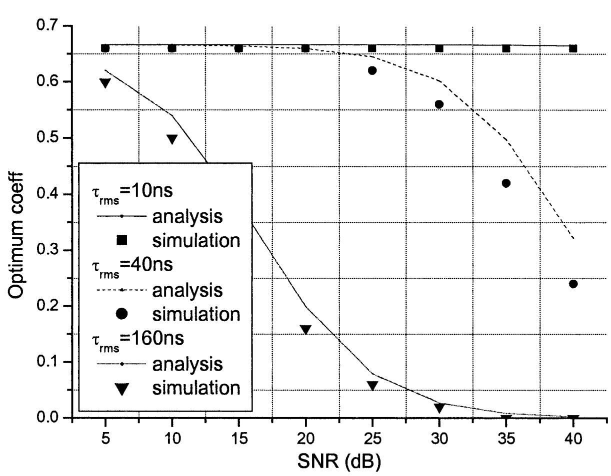
To verify the performance of the proposed RSW CEF, we consider the transmission of OFDM signals over a Rayleigh fading channel with an exponentially decaying multi-path intensity profile. The parameters of

Table 1 Simulation parameters.

Parameters	values
FFT duration	$3.2\,\mu$s
CP duration	$0.8\,\mu$s
FFT size	64
Sub-carrier spacing	312.5 kHz
Sub-carrier modulation	16-QAM, 64-QAM
FEC	Convolutional code ($R = 1/2$, $K = 7$)
Data rate	24 Mbps, 48 Mbps
Packet length	500 Byte
Pilot	Single dedicated pilot symbol



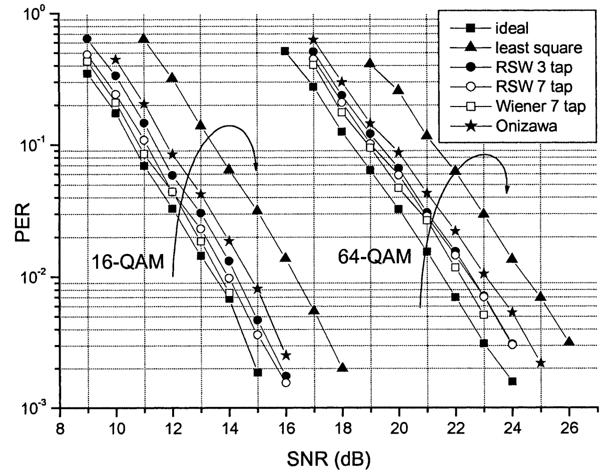
(a) MSE



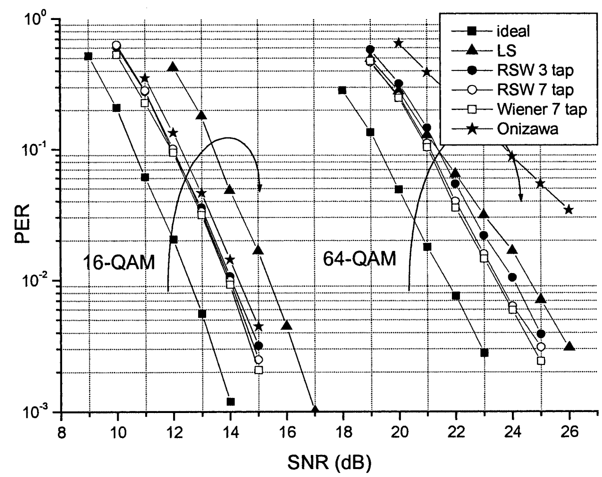
(b) Optimum tap coefficient

Fig. 3 Performance of the optimum 3-tap RSW CEF.

the OFDM signal are summarized in Table 1. We use one dedicated pilot symbol located at the head of each packet. The channel information extracted from the pilot symbol is used for demodulation of the signal. It is assumed that the channel gain of the corner sub-carriers is perfectly estimated for the purpose of performance comparison. Since the optimum RSW CEF is mainly determined by the multi-path intensity profile of the channel, we evaluate the receiver performance for a range of τ_{rms} from 10 ns to 160 ns corresponding to typical delay spread values in indoor wireless channel



(a) $\tau_{rms} = 40\,$ns



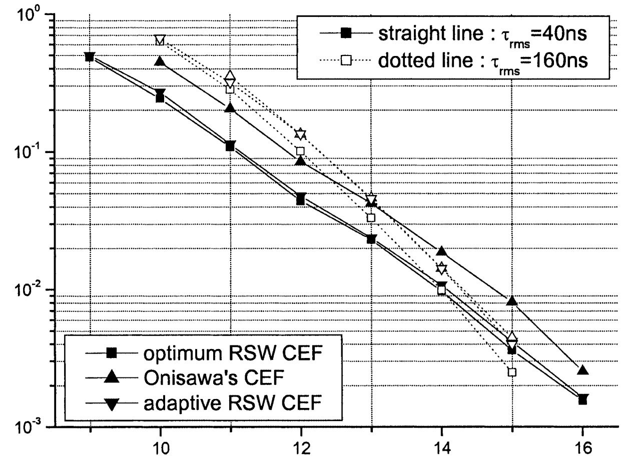
(b) $\tau_{rms} = 160\,$ns

Fig. 4 PER performance in frequency selective fading channel.

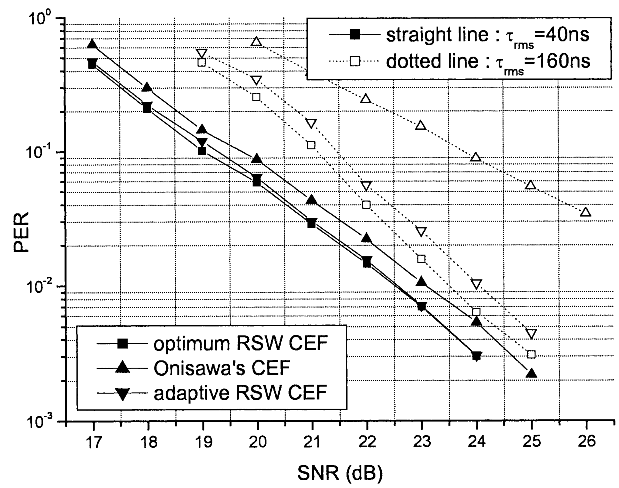
[1].

Figure 3 depicts the MSE and coefficient of the optimum 3-tap RSW CEF (i.e., $M = 1$). It can be seen that the analytic results agree quite well with the simulation results. When τ_{rms} is small, the optimum tap coefficient is around 0.67 irrespective of the SNR, approximately equal to the average of three sub-carriers with equal weight. When τ_{rms} is large, the magnitude of the optimum tap coefficient decreases as the SNR increases since the interference from the adjacent sub-carriers becomes significant compared to the noise term. Numerical results show that, unless the delay spread is too small, it suffices to use a CEF with a small number of taps. As an example, when $\tau_{rms} = 80\,$ns, it suffices to use a 7-tap CEF at low SNR and 3-tap CEF at high SNR to achieve near optimum performance.

Figure 4 depicts the receiver performance with the use of the proposed RSW CEF in terms of the packet error rate (PER) when the channel has a RMS delay spread of 40 ns or 160 ns. For performance comparison, we also consider the use of 7-tap linear MMSE CEF and



(a) 16-QAM



(b) 64-QAM

Fig. 5 PER performance of the adaptive RSW CEF.

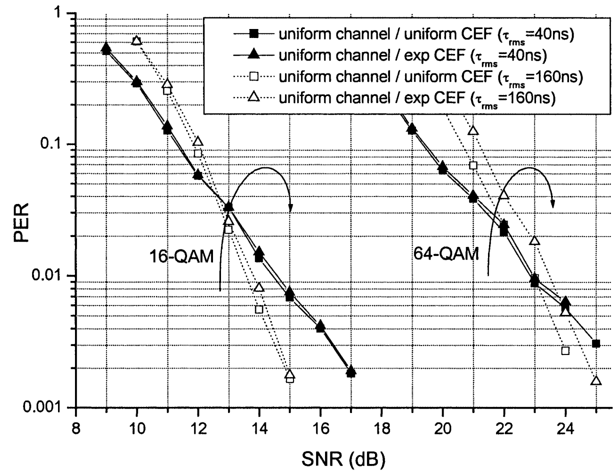


Fig. 6 PER performance of the adaptive RSW CEF with mismatch in multi-path intensity profile model.

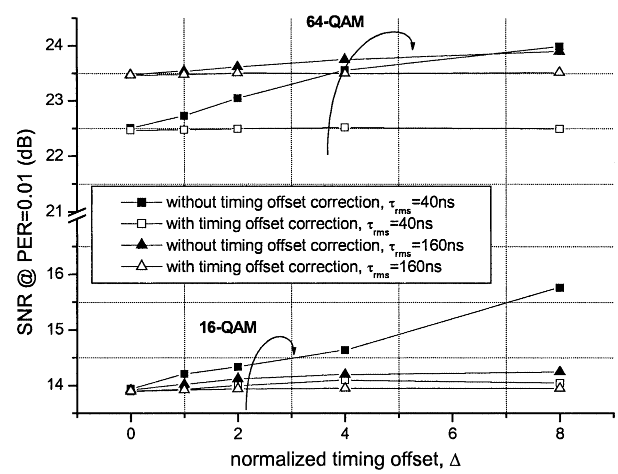


Fig. 7 Required SNR for PER of 0.01 in the presence of timing offset.

adaptive CEF proposed by Onizawa [11]. The tap size of the Onizawa's filer is adaptively selected according to the rule in [11]. Note that the performance with the use of ideal and LS channel estimate is also depicted as an upper and lower bound of the receiver performance. It can be seen that the use of the proposed RSW CEF results in performance degradation of a fractional dB compared to the use of linear MMSE filter and that it provides an SNR gain of 2.3–2.6 dB for $\tau_{rms} = 40$ ns and of 0.8–1.3 dB for $\tau_{rms} = 160$ ns over the LS estimator at PER=0.01.

Figure 5 compares the PER performance of the proposed adaptive RSW CEF with that of the optimum RSW CEF where $\Lambda_{1,1}$ is used for estimation of τ_{rms}. It can be observed the proposed scheme provides performance comparable to the optimum RSW CEF unless τ_{rms} is too large and that it outperforms Onizawa's filter particularly when the SNR is high and τ_{rms} is large.

When there is a mismatch in the multi-path intensity profile model, there can be performance degradation. Figure 6 depicts the PER performance when an adaptive RSW CEF designed for the channel with an exponentially decaying multi-path intensity profile is applied to a channel with a uniform multi-path intensity profile. It can be seen that the mismatch in the multi-path intensity profile model results in a loss of up to 0.4 dB depending on the magnitude of τ_{rms} and SNR. Since most of channels have an exponentially decaying channel profile, it can be acceptable in practice to use an RSW CEF designed for an exponentially decaying channel model.

Figure 7 depicts the required SNR for the adaptive RSW CEF to achieve a PER of 0.01 in the presence of timing offset, which is normalized by the sampling interval. When the timing offset is not compensated, the PER performance degrades as the timing offset increases since the CEF cannot be chosen properly due to the phase rotation. With timing offset correction, the performance of the adaptive RSW CEF becomes quite stable.

5. Conclusions

In this paper, we have designed a frequency-domain CEF with real-valued and symmetrically weighted coefficient for wireless transmission of OFDM signals. The coefficient of the proposed RSW CEF is analytically designed to minimize the MSE of the channel estimate by considering the multi-path intensity profile and SNR of the channel. Simulation results show that the optimum RSW CEF provides PER performance comparable to that of linear MMSE CEF, while requiring less computational complexity. We have also proposed an adaptive RSW CEF whose coefficient is adjusted based on the magnitude of the channel gain difference between the adjacent sub-carriers. By employing a simple timing offset estimation and correction method, the proposed RSW CEF can provide stable performance in the presence of timing offset.

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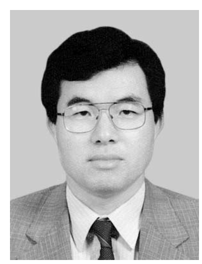

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