Design of implementation-efficient channel estimation filter for wireless OFDM transmission

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Abstract — The detection performance of coherent OFDM receiver 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). Linear minimum mean-squared error (MMSE) filter is optimum as the CEF, but it may not practical due to 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 multi-path intensity profile and SNR of the channel. 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.

I. 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]. However, there exists fundamental challenge 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. 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]. The accuracy of channel estimation can be improved by properly post-processing the channel information extracted from the pilot symbol. 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 to 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 process. Furthermore, it may not be easy to optimize the filter coefficient in real-time 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 can analytically be 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-varying. 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 Section II. The proposed RSW CEF is analytically designed in Section III. We also consider the adaptation of the filter coefficient in real time. The analytic design of the proposed RSW CEF is verified by computer simulation in Section IV. Finally, Conclusions are summarized in Section V.

II. SYSTEM MODEL

Consider an OFDM transmission system whose baseband

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equivalent model is depicted in Fig. 1. In transmitter, N QAM symbols $\{X_k\}$ are converted into the time domain by the IFFT. A cyclic prefix (CP) is added to preserve the orthogonality between the sub-carriers and to eliminate the interference between the adjacent OFDM symbols. 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

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

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.,

$$E\{h_l h_m^*\}=A_l \delta(l-m) \quad (2)$$

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

$$Y_k = X_k H_k + Z_k \quad (3)$$

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

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

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

$$\mathbf{Y}=\mathbf{XH}+\mathbf{Z} \quad (5)$$

where $\mathbf{Y}=[Y_0 Y_1 \cdots Y_{N-1}]^T$, $\mathbf{X}=diag(X_0, X_1, \cdots, X_{N-1})$, $\mathbf{H}=[H_0 H_1 \cdots H_{N-1}]^T$ and $\mathbf{Z}=[Z_0 Z_1 \cdots Z_{N-1}]^T$. 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

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

III. DESIGN OF THE CHANNEL ESTIMATION FILTER

A. *Linear MMSE CEF*

The linear MMSE channel estimate is obtained by [5,9]

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

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

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

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

$$\hat{H}_{k,LS}=Y_k/X_k=H_k+Z_k',\ k=0,1,\cdots,N-1 \quad (9)$$

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.

B. *Real-valued symmetrically weighted (RSW) CEF*

Although the linear MMSE estimate provides the optimum performance, 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). Fig. 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

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

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

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

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.

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

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

where $\lambda_{l,m}=\cos(2\pi\tau_l m/N)-1$ and $\boldsymbol{\alpha}=[\alpha_1\alpha_1\cdots\alpha_M]^T$ represents the coefficient vector. The variance of the LS estimation error is

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

Then, the total estimation error can be represented as

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

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

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

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

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

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

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

and $\mathbf{c}$ is an $(M \times 1)$ column vector whose *i-th* element is $c_i = \sigma_Z^2$. It can be seen that the optimum coefficient is mainly determined by the channel multi-path intensity profile and SNR.

C. *Adaptive RSW CEF*

In real communication environment, the information on the channel characteristics may not be available since it is location-dependent and time-varying. As a practical solution to this problem, we consider the design of an adaptive RSW CEF that can adjust its coefficient according to the estimated channel condition. Since the root-mean-squared (RMS) delay spread τ_{rms} has a one-to-one relationship with the multi-path intensity profile, we use τ_{rms} as a measure of the channel condition.

Assuming an exponentially decaying multi-path intensity profile, the RMS delay spread τ_{rms} can be estimated from the channel gain difference between the adjacent sub-carriers. Let Λ_m be the normalized LS channel gain difference between the sub-carriers differed by m sub-carriers,

$$\Lambda_m = \frac{\sum_{k=m}^{N-1} |\hat{H}_{k,LS} - \hat{H}_{k-m,LS}|^2}{\sum_{k=m}^{N-1} |\hat{H}_{k,LS}|^2} \tag{18}$$

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

$$\overline{\Lambda}_m = \frac{E\left\{\sum_{k=m}^{K-1} |\hat{H}_{k,LS} - \hat{H}_{k-m,LS}|^2\right\}}{E\left\{\sum_{k=m}^{K-1} |\hat{H}_{k,LS}|^2\right\}} = \frac{\sum_{l=0}^{L-1} A_l |1 - e^{j2\pi m\tau_l/N}|^2 + 2\sigma_Z^2}{\sum_{l=0}^{L-1} A_l + \sigma_Z^2} \tag{19}$$

Note that $\overline{\Lambda}_m$ 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 $\overline{\Lambda}_m$ increases as τ_{rms} increases or the SNR decreases. Since τ_{rms} and the SNR affect $\overline{\Lambda}_m$, they can simultaneously be estimated using two values of Λ_m.

The RMS delay spread τ_{rms} can be estimated from $\Lambda_{m,d} = \Lambda_{m+d} - \Lambda_m$. The ensemble average of $\Lambda_{m,d}$ can be expressed as

$$\overline{\Lambda}_{m,d} = \frac{\sum_{l=0}^{L-1} A_l \{|1 - e^{j2\pi(m+d)\tau_l/N}|^2 - |1 - e^{j2\pi m\tau_l/N}|^2\}}{\sum_{l=0}^{L-1} A_l + \sigma_Z^2} \tag{20}$$

It can be seen that, unless the SNR is too low, $\overline{\Lambda}_{m,d}$ becomes quite insensitive to the SNR, making the estimation of τ_{rms} independent of the SNR. The estimation of τ_{rms} from $\Lambda_{m,d}$ can be implemented using a look-up table. Then, the normalized noise variance is estimated by

$$\hat{\sigma}_Z^2 = \frac{\Lambda_{m+d} - \chi_{m+d}(\hat{\tau}_{rms}) + \Lambda_m - \chi_m(\hat{\tau}_{rms})}{4} \tag{21}$$

where $\hat{\tau}_{rms}$ is the estimated RMS delay spread and $\chi_m(\cdot)$ is a function of τ_{rms} whose value is $\overline{\Lambda}_m$ when $\sigma_Z^2 = 0$. Using $\hat{\tau}_{rms}$ and $\hat{\sigma}_Z^2$, the coefficient of the RSW CEF can be adjusted adaptively using the pre-computed coefficient table.

D. *RSW CEF in the presence of timing offset*

The performance of the OFDM receiver is known to be quite insensitive to the timing offset [5, 14]. If the amount of the 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 in order to provide a margin for the variation of timing estimate [12]. However, the timing offset causes a phase rotation of the channel linearly proportional to the frequency index. This phase rotation may result in erroneous estimate of τ_{rms}, leading to inappropriate CEF selection. Therefore, the timing offset should be compensated before channel estimation.

Let $\Delta \cdot T_s$ be the timing offset introduced in the receiver. Then, the channel gain becomes

$$H_{k,\Delta} = \sum_{l=0}^{L-1} h_l e^{-j2\pi k(\tau_l+\Delta)/N} = H_k e^{-j2\pi k\Delta/N} \tag{22}$$

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

$$E\{H_{k,\Delta} H^*_{k-m,\Delta}\} = \sum_{l=0}^{L-1} A_l e^{-j2\pi m(\tau_l+\Delta)/N}. \tag{23}$$

It can easily be shown that the linear MMSE estimate with timing offset $\Delta \cdot T_s$ becomes

$$\begin{aligned} \hat{H}_{k,\Delta,LMMSE} &= \mathbf{p}_k^H (\mathbf{R}_{hh} + \sigma_Z^2 \mathbf{I})^{-1} \boldsymbol{\theta} \hat{\mathbf{H}}_{LS} \\ &= \mathbf{p}_k^H (\mathbf{R}_{hh} + \sigma_Z^2 \mathbf{I})^{-1} \hat{\mathbf{H}}_{LS,\Delta} \end{aligned} \tag{24}$$

where

$$\boldsymbol{\theta} = diag(e^{-j2\pi M\Delta/N}, e^{-j2\pi(M-1)\Delta/N}, \cdots, e^{j2\pi(M-1)\Delta/N}, e^{j2\pi M\Delta/N}) \tag{25}$$

and $\hat{\mathbf{H}}_{LS,\Delta} = \boldsymbol{\theta}\hat{\mathbf{H}}_{LS}$ is the LS channel estimate after timing offset compensation. It can be said that the coefficient of a linear MMSE CEF is unchanged as long as the phase rotation caused by the timing offset is compensated before filtering.

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

$$\begin{aligned} \hat{H}_{k,RSW} &= \left[\left(1 - \sum_{m=1}^{M} \alpha_m\right) H_k + \frac{1}{2}\sum_{m=1}^{M} \alpha_m (H_{k-m} e^{j2\pi m\Delta/N} + H_{k+m} e^{-j2\pi m\Delta/N})\right] e^{-j2\pi k\Delta/N} \\ &\quad + (1 - \sum_{m=1}^{M} \alpha_m) Z'_k + \frac{1}{2}\sum_{m=1}^{M} \alpha_m (Z'_{k-m} + Z'_{k+m}) \end{aligned} \tag{26}$$

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 filtering. Fig. 2 depicts an RSW CEF with timing offset compensation.

When the timing offset is not compensated, the phase rotation of the channel gain due to 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 propose to estimate the timing offset by finding the phase correction that minimizes the channel gain difference between the adjacent sub-carriers. For an assumed timing offset Δ', we can define the phase-corrected channel gain difference by

$$\Lambda_m^{\Delta'} = \frac{\sum_{k=m}^{K-1} |\hat{H}_{k,LS} - \hat{H}_{k-m,LS} e^{j2\pi m\Delta'/N}|^2}{\sum_{k=m}^{K-1} |\hat{H}_{k,LS}|^2} \tag{27}$$

Since the ensemble average of $\Lambda_m^{\Delta'}$

$$\overline{\Lambda_m^{\Delta'}} = \frac{\sum_{l=0}^{L-1} A_l \,|1 - e^{j2\pi m(\tau_l+\Delta-\Delta')/N}|^2 + 2\sigma_Z^2}{\sum_{l=0}^{L-1} A_l + \sigma_Z^2} \tag{28}$$

is minimum when $\Delta' = \Delta$, the timing offset can be estimated by finding $\hat{\Delta}$ that minimizes $\Lambda_m^{\Delta'}$, i.e.,

$$\hat{\Delta} = \arg\min_{\Delta'} \Lambda_m^{\Delta'}. \tag{29}$$

IV. PERFORMANEC EVALUATION

To verify the design and performance of the proposed RSW CEF, we consider the transmission of OFDM signal over a Rayleigh fading channel with an exponentially decaying multi-path intensity profile. The parameters of the OFDM signal are summarized in Table 1. We consider the use of a BPSK modulated pilot signal for channel estimation. 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 $10ns$ to $160ns$, corresponding to typical delay spread values in indoor wireless channel [1].

Fig. 3 depicts the MSE and coefficient of the optimum 3-tap RSW CEF (*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.

Fig. 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 40ns or 160ns. For performance comparison, we also consider the use of 7-tap linear MMSE CEF and adaptive CEF proposed by Onizawa [11]. Note that the performance with the use of ideal channel estimate and LS channel estimate is also depicted as the 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 2.3~2.6dB for $\tau_{rms} = 40ns$ and of 0.8~1.3dB for $\tau_{rms} = 160ns$ over the LS estimator at PER=0.01.

Fig. 5 compares the PER performance of the proposed adaptive RSW CEF with that of the optimum RSW CEF where $\Lambda_{m,d}$ with $m=1$ and $d=1$ 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 the performance improvement over the one in [11] increases as τ_{rms} increases.

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

Fig. 7 depicts the required SNR for the adaptive RSW CEF to achieve a PER of 0.01 in the presence of timing offset when $\tau_{rms} = 40ns$. 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. When the timing offset is corrected, the performance of the adaptive RSW CEF becomes robust to the timing offset.

V. CONCLUSIONS

In this paper, we have designed a frequency-domain CEF with real-valued and symmetrically weighted coefficient for wireless transmission of OFDM signal. The coefficient of the proposed RSW CEF is analytically determined to minimize the MSE of the channel estimate. The proposed CEF is analytically designed 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 the linear MMSE CEF, while requiring less computational complexity. We have also proposed an adaptive RSW CEF where coefficient is determined 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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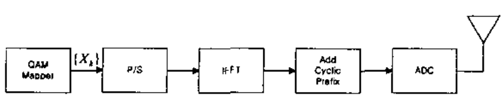
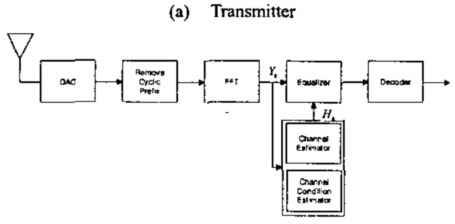
TABLE I. SIMULATION PARAMETERS

Parameters	Values
FFT duration	3.2 μs
CP duration	0.8 μs
FFT size	64
Sub-carrier spacing	312.5kHz
Sub-carrier modulation	64-QAM
FEC	Convolutional coding (R=1/2 or 2/3 K=7)
Data rate	24Mbps, 48Mbps
Packet length	500 Byte



(a) Transmitter



(b) Receiver

Fig. 1. OFDM transmission system with channel estimation

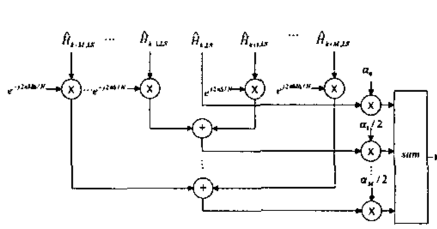
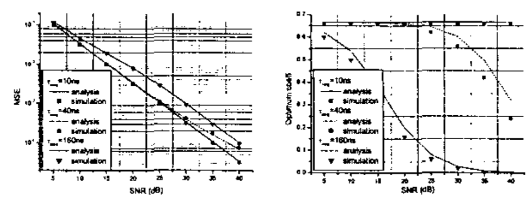
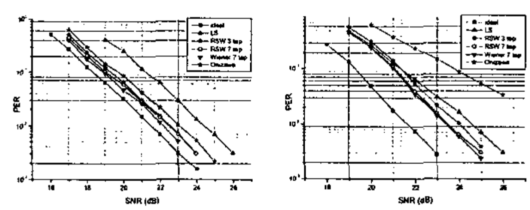


Fig. 2. Structure of the RSW CEF with timing offset correction



(a) MSE (b) Optimum tap coefficient

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



(a) $\tau_{rms} = 40ns$ (b) $\tau_{rms} = 160ns$

Fig. 4. PER performance in frequency selective fading channel

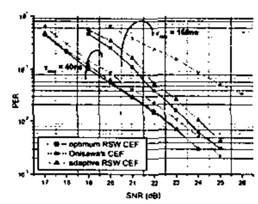


Fig. 5. PER performance of the adaptive RSW CEF

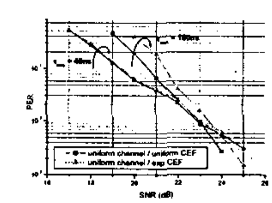


Fig. 6. Effect of the multi-path intensity profile model

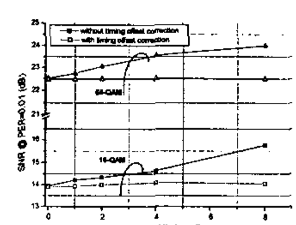


Fig. 7. Required SNR for PER of 0.01 in the presence of timing offset