Frequency-Averaged MMSE Channel Estimator for Multicarrier Transmission Schemes

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***Abstract*— A new frequency-averaged MMSE estimator is proposed for multicarrier transmission schemes. The proposed estimator employs a new criterion of minimizing the mean square errors (MSE) over subcarriers, which provides an accurate channel estimation using a few pilot OFDM symbols without a priori channel knowledge. The proposed estimator has low complexity, since it uses high correlated adjacent subcarriers and neglects low correlated distant subcarriers. In the estimation process, intersymbol and intersubcarrier interference due to excess delay paths beyond guard interval is considered to achieve good performance. The simulation results showed that the frequency-averaged MMSE estimator can effectively reduce estimation error and keeps better performance than the conventional robust estimator in any channel conditions.**

I. Introduction

Multicarrier transmission schemes have been intensively investigated for high data rate wireless communications. In particular, pilot-assisted transmission schemes with coherent detection are attractive for fourth generation mobile communications and wireless local area networks (W-LANs), where pilot signals are inserted in all subcarriers of an orthogonal frequency division multiplexing (OFDM) symbol. Generally, signals are propagated through multipath delay channels and the channel frequency response at the receiver is different for each subcarrier. An efficient receiver requires accurate channel estimation for each subcarrier and, practically, it is essential to estimate the channel frequency response using a small number of pilot symbols within a limited complexity under any channel conditions.

So far, several channel estimation schemes have been proposed for multicarrier transmission schemes[1]–[10]. In [1][2], the minimum mean squared error (MMSE) estimator has been demonstrated by exploiting the correlation of the frequency response at the different frequencies. Although the optimal MMSE estimator has high efficiency, it requires large complexity and a priori knowledge of multipath channel statistics. One solution for these problems has been presented by using fixed weights[2][3], where an estimator is designed for a fixed multipath delay channel. However, the fixed-weight approaches increase the channel estimation errors, especially when the actual delay spread is larger than the designed one. In such a situation, a practical channel estimator applicable under any channel conditions is required to improve the receiver performance.

In this paper, we propose a new channel estimation method, which provides an accurate estimation using a few pilot OFDM symbols without a priori knowledge of multipath channel statistics. The proposed estimator determines the combining weights for correlated subcarriers, based on a new criterion of minimizing the mean square errors (MSE) over subcarriers, in place of the conventional time-averaged MMSE criterion. Since the proposed estimator has averaging process over subcarriers in an OFDM symbol, it requires only a few pilot OFDM symbols for weight convergence and does not require a priori channel knowledge. Moreover, the proposed estimator has low complexity, using only high correlated adjacent subcarriers and neglecting low correlated distant subcarriers. Since all signal processing in the proposed estimator is performed in the frequency domain, intersymbol and intersubcarrier interference from excess delay paths is properly considered in the estimation process.

This paper is organized as follows. In section **II**, we propose a new frequency-averaged MMSE estimator for multicarrier transmission. In sections **III** and **IV**, simulation results are shown for a W-LAN system [11] and for a multicarrier code division multiple access (MC-CDMA) [12][13][14] system, respectively.

II. Frequency-averaged MMSE estimator

In this section, we propose a new frequency-averaged MMSE estimator for multicarrier transmission.

A. Channel Estimation Procedure

Similarly to the traditional approaches [1]–[7], our channel estimation consists of two estimation steps: preliminary and secondary estimations. In the preliminary estimation, channel frequency responses are temporally estimated individually for different subcarriers, correlating the received signal by time pilot signals. The secondary estimation improves the accuracy by exploiting the correlation of the channel frequency responses at different subcarriers.

B. Preliminary Estimation

Consider an OFDM system with N subcarriers for transmission. The n-th transmission subcarrier in the q-th OFDM symbol is modulated by a signal $x_n(q)(E[|x_n(q)|^2] = 1)$ and the OFDM symbol is transmitted over a multipath fading channel after guard interval insertion. At the receiver, the received signal is fed to fast Fourier transform (FFT) after guard interval removal. Then, an output of FFT for the n-th subcarrier can be expressed as

$$y_n(q) \quad = \quad h_n x_n(q) + z_n(q),$$

where h_n is the channel frequency response for the n-th subcarrier and $z_n(q)$ is the interference-plus-noise with variance P_z ($E[|z_n(q)|^2] = P_z$). The frequency response h_n includes not only the effect of multipath delay channels but also the effect of transmitter and receiver filters [7]. The interference component included in $z_n(q)$ usually occurs due to intersymbol interference, intersubcarrier interference, and interference from other terminals.

In the preliminary estimation, the channel frequency response h_n is estimated by correlating the received signal by time pilot signals as

$$\tilde{h}_n \quad = \quad \frac{1}{q_0}\sum_{q=1}^{q_0} y_n(q)x_n(q)^* = h_n + z_n^{'}$$
$$z_n^{'} \quad = \quad \frac{1}{q_0}\sum_{q=1}^{q_0} z_n(q)x_n(q)^*$$

where q_0 is the number of pilot OFDM symbols and $z_n(q)^{'}$ is the output interference-plus-noise of preliminary estimator.

In some cases as in IEEE802.11a[11], there is a subcarrier with no data transmission among data transmitted subcarriers. In this case, a channel parameter for the subcarrier with no data transmission is estimated using parameters $\tilde{h}_{n-1}$ and $\tilde{h}_{n+1}$ for adjacent data transmitted subcarriers as

$$\tilde{h}_n \quad = \quad \frac{\tilde{h}_{n-1} + \tilde{h}_{n+1}}{2}. \qquad (1)$$

It should be noted that the preliminary estimation is applied for data transmitted subcarriers and a subcarrier with no data transmission which lies between data transmitted subcarriers. Hence, in exact sense, N represents the number of subcarriers for the preliminary estimation.

C. Secondary Estimation

Secondary estimation exploits the correlation of the channel frequency response at different subcarriers. The channel frequency response for the n-th subcarrier is estimated by linearly combining a block of the $(2n_0 + 1)$ preliminary estimation values as

$$\hat{h}_n \quad = \quad \begin{cases} \boldsymbol{w}^{\dagger}_{n-n_0-1}\tilde{\boldsymbol{h}}_{n_0+1} & n = 1, ..., n_0 \\ \boldsymbol{w}^{\dagger}_0\tilde{\boldsymbol{h}}_n & n = 1 + n_0, ..., N - n_0 \\ \boldsymbol{w}^{\dagger}_{n+n_0-N}\tilde{\boldsymbol{h}}_{N-n_0} & n = N - n_0 + 1, ..., N \end{cases} \qquad (2)$$

where $\tilde{\boldsymbol{h}}_n = [\tilde{h}_{n-n_0}, ..., \tilde{h}_{n+n_0}]^T$ is the preliminary channel estimation vector, $\boldsymbol{w}_k = [w_{k,-n_0}, ..., w_{k,n_0}]^T$ ($k = -n_0, ..., n_0$) is the weight vector, which is different according to k. In (2), a channel frequency response for the n-th subcarrier with $n \leq n_0$ or $n \geq N - n_0 + 1$ is estimated by the preliminary estimation values for the first or the last $(2n_0 + 1)$ subcarriers, respectively. The channel frequency response for the n-th subcarrier with $1 + n_0 \leq n \leq N - n_0$ is estimated by the preliminary estimation values for the adjacent n_0 subcarriers on both sides.

The secondary estimation has error $e_n = \hat{h}_n - h_n$ and it is required to determine a weight to minimize errors. Therefore, we consider a weight $\boldsymbol{w}_k$ to meet the demand of the minimum mean squared error(MMSE) over subcarriers, which can be put mathematically as

$$E_n[|e_n|^2] \equiv \frac{1}{N - 2n_0}\sum_{n=n_0+1}^{N-n_0} |\boldsymbol{w}_k^{\dagger}\tilde{\boldsymbol{h}}_n - h_{n+k}|^2 \rightarrow min \qquad (3)$$

where $E_n[\cdot]$ denotes the average over subcarriers. The MMSE demand (3) is individually adopted for different k.The proposed estimator minimizes the squared error $|e_n|^2$ averaged over subcarriers. It should be noted that the MMSE estimator described in the previous papers [1]–[8] minimizes the squared error averaged over long time duration. Therefore, the definition of mean squared error in our paper is different from that in the previous papers.

The MMSE weight $\boldsymbol{w}_k$ to meet (3) is given by

$$\boldsymbol{w}_k \quad = \quad \boldsymbol{\Phi}^{-1}\boldsymbol{v}_k \qquad (4)$$
$$\boldsymbol{\Phi} \quad = \quad \frac{1}{N - 2n_0}\sum_{n=n_0+1}^{N-n_0} \tilde{\boldsymbol{h}}_n\tilde{\boldsymbol{h}}_n^{\dagger} \qquad (5)$$
$$\boldsymbol{v}_k \quad = \quad \frac{1}{N - 2n_0}\sum_{n=n_0+1}^{N-n_0} \tilde{\boldsymbol{h}}_n h^*_{n+k}. \qquad (6)$$

Since, in practice, it is impossible to use a real channel parameter h_{n+k} in (6), we use an approximation as

$$[\boldsymbol{v}_{-n_0}, ..., \boldsymbol{v}_{n_0}] = \frac{1}{N - 2n_0}\sum_{n=n_0+1}^{N-n_0} \tilde{\boldsymbol{h}}_n(\tilde{\boldsymbol{h}}_n - \boldsymbol{z}_n^{'})^{\dagger} \qquad (7)$$
$$= \boldsymbol{\Phi} - \frac{1}{N - 2n_0}\sum_{n=n_0+1}^{N-n_0} (\boldsymbol{h}_n\boldsymbol{z}_n{'}^{\dagger} + \boldsymbol{z}_n{'}\boldsymbol{z}_n{'}^{\dagger}) \qquad (8)$$
$$\simeq \boldsymbol{\Phi} - (\tilde{P}_z/q_0)\boldsymbol{I} \qquad (9)$$

where $\boldsymbol{h}_n = [h_{n-n_0}, ..., h_{n+n_0}]^T$, $\boldsymbol{z}_n^{'} = [z_{n-n_0}^{'}, ..., z_{n+n_0}^{'}]^T$, and $\tilde{P}_z$ is the estimated value of the interference-plus-noise power P_z. In approximating (8) into (9), we consider the property of $E[\boldsymbol{h}_n\boldsymbol{z}_n^{'\dagger}] = \boldsymbol{0}$ and $E[\boldsymbol{z}_n^{'}\boldsymbol{z}_n^{'\dagger}] = (P_z/q_0)\boldsymbol{I}$ (appendix I). The estimated interference-plus-noise power $\tilde{P}_z$ can be usually obtained by

$$\tilde{P}_z = \frac{q_0}{q_0 - 1}\left(\frac{1}{q_0 N}\sum_{q=1}^{q_0}\sum_{n=1}^{N} |y_n(q)|^2 - \frac{1}{N}\sum_{n=1}^{N} |\tilde{h}_n|^2\right). \qquad (10)$$

From (4) and (9), the MMSE weight $\boldsymbol{w}_k$ is given by

$$[\boldsymbol{w}_{-n_0}, ..., \boldsymbol{w}_{n_0}] \quad \simeq \quad \boldsymbol{I} - (\tilde{P}_z/q_0)\boldsymbol{\Phi}^{-1}. \qquad (11)$$

The secondary channel estimation $\hat{h}_n$ can be obtained by substituting $\boldsymbol{w}_k$ into (2).

D. Frequency Averaged and Time Averaged MMSE Estimators

To get insight into the behavior of the frequency-averaged MMSE estimator, we discuss theoretical difference between frequency-averaged and time-averaged channel estimators.

Eqn.(11) has substantially the similar expression with (3) in [2], except that the correlation matrix $\mathbf{\Phi}$ has a frequency-average operation in place of a time-average operation. To understand the effect of different average operations, let us consider a simple example of a multipath delay channel, where the maximum excess delay is shorter than the guard interval duration. Then, the channel impulse response $a(t)$ is characterized by

$$a(t) \;=\; \sum_{l=1}^{GI} a_l \delta(t-(l-1)T_s) \tag{12}$$

where a_l is the amplitude of the l-th path, GI is the number of samples for the guard interval, and T_s is the sample duration.

Under this multipath delay channel condition, the channel frequency response h_n for the n-th subcarrier is given by

$$h_n \;=\; \sum_{l=1}^{GI} a_l e^{-j(2\pi/N)(l-1)n}$$

Assuming a large number of subcarriers N and using $E_n\left[e^{-j(2\pi/N)(l_1-l_2)n}\right]=1$ ($l_1=l_2$) and 0($l_1 \neq l_2$), we have the frequency-averaged correlation as

$$E_n[h_n h_{n+k}^*] \;=\; \sum_{l=1}^{GI} |a_l|^2 e^{j(2\pi/N)(l-1)k} \tag{13}$$

From (13), it is found that $E_n[h_{n_1} h_{n_2}^*]$ is determined by instantaneous power $|a_l|^2$.

Next, define $E_t[\cdot]$ as average over very long time duration, which is equivalent to ensemble average $E[\cdot]$. Then, using the property of $E_t[a_{l_1} a_{l_2}^*]=0$ ($l_1 \neq l_2$), the time-averaged correlation can be expressed as

$$E_t[h_n h_{n+k}^*] \;=\; \sum_{l=1}^{GI} E_t[|a_l|^2] e^{j(2\pi/L)(l-1)k} \tag{14}$$

From (14), $E_t[h_n h_{n+k}^*]$ is determined by time-averaged power $E_t[|a_l|^2]$.

Consequently, the frequency-averaged channel correlation $E_n[h_{n_1} h_{n_2}^*]$ is determined by instantaneous power $|a_l|^2$, which may vary at different time instants. Meanwhile, time-averaged channel correlation $E_t[h_n h_{n+k}^*]$ is determined by average power $E[|a_l|^2]$.

E. Advantage of Proposed Method

The proposed frequency-averaged MMSE estimator can obtain an accurate correlation matrix $\mathbf{\Phi}$, averaging a large number of subcarriers in an OFDM pilot symbols. Accordingly, the proposed estimator can estimate the channel frequency response instantaneously, using a few OFDM pilot symbols in a frame. In the estimation process, the proposed estimator does not require a priori knowledge for multipath channel statistics, although the conventional estimators require it. All signal processing is performed in frequency domain and the effect of interference due to the excess delay paths beyond the guard interval is taken into account. Therefore, the proposed estimator is applicable for a burst packet frame under any channel conditions.

The proposed estimator exploits high channel correlation with the adjacent n_0 subcarriers on both sides and neglects low channel correlation with other distant subcarriers. Therefore, an accurate channel estimation can be achieved in a limited complexity based on a low rank matrix computation. Also, the number of multiplications to compute a correlation matrix can be reduced according to a method shown in appendix II. Therefore, the complexity for the proposed estimator is considered as reasonable to implement.

TABLE I

SIMULATION PARAMETERS FOR IEEE802.11A

Number of Subcarriers	52 (#7-32, #34-59)
Number of pilot OFDM symbols	2
Modulation(Pilot/Data)	QPSK/QPSK
FFT/Guard Interval length	64/16 samples
Number of combining taps	$3,5,7$ ($n_0=1,2,3$)
Channel	L-path Rayleigh channel
RMS delay spread	$\tau_d=0.025T_d, 0.1T_d, 0.2T_d$

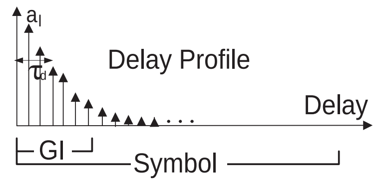


Fig. 1. Multipath delay profile.

III. PERFORMANCE EVALUATION FOR IEEE802.11A

In this section, we demonstrate the performance of the frequency-averaged MMSE estimator for an OFDM system.

A. System Parameters

The OFDM system parameters used for simulations are summarized in Table I. The subcarrier arrangement in this paper is similar to that in the high-rate wireless LAN: IEEE802.11a [11]. One OFDM symbol is composed of 80 samples, where the guard interval length is $GI=16$ samples and the effective symbol length is 64 samples. The OFDM symbol is generated with the 64-point Inverse FFT (IFFT), where only 52 subcarriers convey QPSK modulated data and pilot signals and the remaining 12 subcarriers have zero energy as guard subcarriers.

Fig. 1 shows multipath delay channel profile. We assume a quasi static L-path Rayleigh channel model, where the average power of a delay path decays exponentially according to the excess delay. In case of $L > GI$, there are some delay paths with excess delay beyond the guard interval duration. The root-mean-square (RMS) delay spread of the channel is denoted as

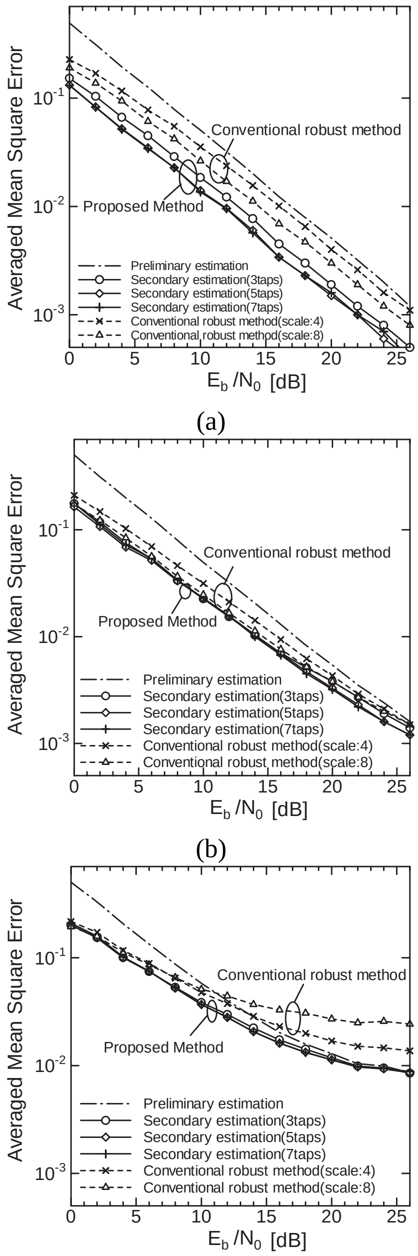


Fig. 2. Average MSE versus E_b/N_0 under $L = N$ (a) $\tau_d = 0.025T_d$ (b) $\tau_d = 0.1T_d$ (c) $\tau_d = 0.2T_d$.

τ_d. The multipath channel is independently generated in every frame. We discuss the performance of channel estimations under various values of RMS delay spread τ_d to effective OFDM symbol duation T_d.

At the receiver, the channel estimation is performed using 2 pilot OFDM symbols in a preamble of a frame. Perfect synchronization is assumed in order to evaluate the channel estimation alone. The mean square error (MSE) of the channel

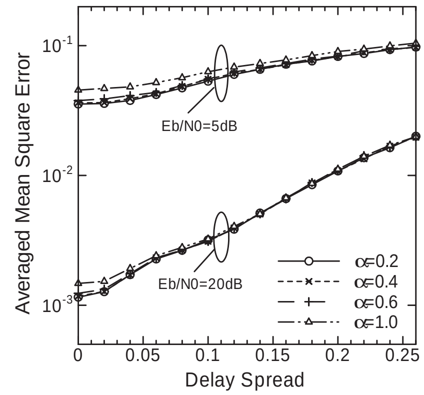


Fig. 3. Average MSE versus delay spread for frequency-averaged MMSE channel estimator $\tau_d(L = N)$.

estimation for a simulation trial is given by

$$MSE = \frac{1}{N}\sum_{n=1}^{N}|\hat{h}_n - h_n|^2. \quad (15)$$

The MSE is averaged over simulation trials. The simulation trials are performed 1000 times for different channel data.

For comparison purpose, we evaluate the conventional robust estimator (appendix III) designed for the uniform channel power delay profile within guard interval.

B. Mean Squared Error

Fig. 2 shows MSE performance under $L = N$ and $\tau_d = 0.025T_d,\ 0.1T_d,\ 0.2T_d$. It is seen that the proposed estimator can significantly reduce MSE and improve the channel estimation accuracy under a small delay spread of $\tau_d = 0.025T_d$. The conventional estimator has worse performance, due to mismatch of the estimator to the channel statistics. The performance difference between the proposed estimator and the conventional estimator decreases once τ_d increases up to $0.1T_d$. However, when the delay spread further increases up to $\tau_d = 0.2T_d$, the conventional method has worse performance than the preliminary estimation, due to unexpected interference by excess delay paths beyond guard interval. In this case, more combining taps degrades the performance of the conventional estimator due to incorrect combining weight. The MSE performance in the proposed estimator also deteriorates as the delay spread increases. However, the proposed estimation always keeps better performance than the preliminary estimation even at a large delay spread of $\tau_d = 0.2T_d$. This is because the proposed estimator performs all signal processing in frequency domain and the effect of intersymbol and intersubcarrier interference are taken into account.

C. Effects of Time Averaging on Proposed Estimator

As described before, the proposed estimator's MMSE weights are basically determined by an frequency averaging operation over subcarriers in a packet frame. Additionally, it may be possible to use time averaging operation for several blocks of pilot OFDM symbols inserted in different packet frames, in case of continuous frame transmission. Then, the correlation matrix $\mathbf{\Phi}_k$ for the k-th frame can be computed recursively as

$$\mathbf{\Phi}_k = \alpha\mathbf{\Phi} + (1-\alpha)\mathbf{\Phi}_{k-1} \qquad (16)$$

where α is the forgetting factor and $\mathbf{\Phi}$ is the pure frequency-averaged correlation matrix in the k-th frame.

In practice, multipath delay channels at different frames have a time correlation according to the Doppler frequency. Here, the multipath channels with a certain time correlation is considered as an intermediate condition of perfect correlation and no correlation. The performance under a perfect time correlation will be almost the same with a formerly described frequency-averaged MMSE estimator, because the MMSE weights in both two estimators are determined by an instantaneous delay profile. Therefore, the performance evaluation under an extreme case of no time correlation will help us to expect the performance under any time correlations. According to the above considerations, in the simulations, we evaluate the proposed estimator with time averaging operations under multipath delay channels with no time correlation between frames.

Fig. 3 shows MSE performance of the proposed estimator with time averaging operations versus delay spread τ_d under various α. In this figure, the forgetting factor α has very little effect on the MSE performance. From the results, we can see that the performance of the proposed estimator is insensitive to the difference of instantaneous delay profiles, if the delay spread is the same. Rather, the performance is more sensitive to mismatch of the delay spread.

From the simulation results, the proposed estimator can additionally use the time averaging without any performance deterioration. In other sense, a single correlation matrix can be used under various instantaneous multipath channels within the same delay spread. This characteristic permits the proposed estimator to have longer update duration for correlation matrix computation and to have lower computational complexity.

IV. Performance Evaluation for MC-CDMA

In this section, we evaluate the performance of the proposed estimator for an MC-CDMA system [12][13][14], which has larger number of subcarriers than IEEE802.11a.

A. System Parameters

System model used for simulations is almost the same as in the previous section, except some parameters. The system parameters to evaluate an MC-CDMA systems are summarized in Table II [13][14]. In an MC-CDMA scenario, the channel estimation is performed using 4 pilot OFDM symbols in a preamble of a frame. A pilot OFDM symbol is made by 1024 size FFT and 768 subcarriers are used for data and pilot transmission. The remaining 128 subcarriers on both edges have no energy as guard subcarriers.

TABLE II

Simulation parameters for MC-CDMA system

Number of sub-carriers	$768(\#129-896)$
Number of pilot OFDM symbols	4
Modulation(Pilot/Data)	QPSK/QPSK
FFT/Guard Interval length	1024/256 samples
Size of MMSE combiner	$3, 7, 11$ ($n_0 = 1, 3, 5$)
Channel	L-path Rayleigh channel
Delay spread	$\tau_d = 0.025T_d, 0.2T_d$

B. Mean Squared Error

Fig. 4 shows MSE performance under $L = N$ and $\tau_d = 0.025T_d,\ 0.2T_d$. Similarly to the case of IEEE802.11a, the proposed estimator always achieves good performance. It is seen that, in case of 7 combining taps($n_0 = 3$), the proposed estimator has about one-tenth lower MSE than the preliminary estimator only.

Since an MC-CDMA system has highly correlated subcarriers, a larger number of combining taps can be effectively adopted and the correlation matrix $\mathbf{\Phi}$ can be estimated accurately using a larger number of subcarriers. Therefore, in an MC-CDMA system, the proposed channel estimator achieves better accuracy compared to the conventional robust estimator than in a IEEE802.11a system.

C. Weight convergence

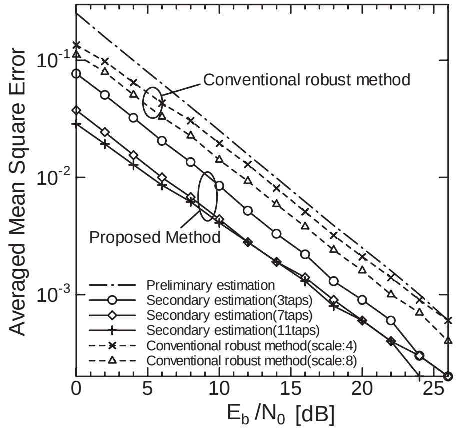
In the previous simulations, we have computed the correlation matrix by averaging over all available subcarriers. However, since an averaging operation for a lot of subcarriers gives rise to increase in computational burden, it is important to determine a reasonable number of averaged subcarriers. Therefore, we evaluate the weight convergence of MMSE weights considering a various number of averaged subcarriers for the correlation matrix.

Fig. 5 shows MSE performance under a various number of averaged subcarriers for the correlation matrix, under $\tau_d = 0.025T_d$ and $\alpha = 1$. From this figure, MSE is mostly converged by averaging 100 subcarriers under $E_b/N_0 = 0, 13, 26$dB. Therefore, averaging 100 subcarriers will be a suitable choice to meet both the complexity and MSE performance. We also evaluated the performance under different delay spread $\tau_d = 0.1T_d, 0.2T_d$, which gave us the same trend as in $\tau_d = 0.025T_d$.

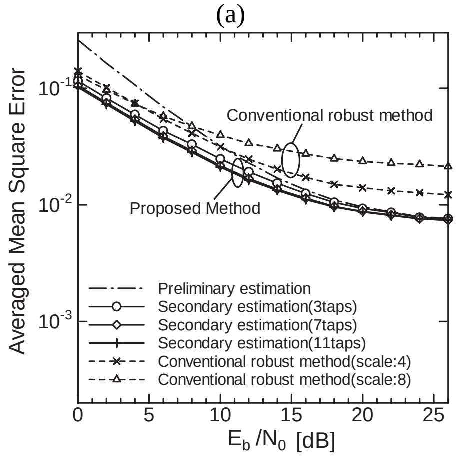
Fig. 6 shows MSE performance under a various number of averaged subcarriers for the correlation matrix, under $\tau_d = 0.025T_d$ and $\alpha = 0.4$. Using time averaging, we have better MSE performance even in a small number of averaged subcarriers. Thus, the combination with time averaging is effective in case of continuous frame transmission. In practice, the use of time-averaging will depend on whether the packet is burst or not.

V. Conclusion

In this paper, we have proposed frequency-averaged MMSE channel estimator, which improves channel estimation perfor-



(a)



(b)

Fig. 4. Average MSE versus E_b/N_0 under $L = N$ (a) $\tau_d = 0.025T_d$ (b) $\tau_d = 0.2T_d$.

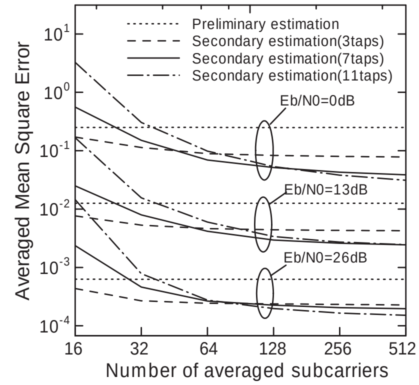


Fig. 5. Convergence of frequency-averaged MMSE weight ($\alpha = 1.0, \tau_d = 0.025T_d$).

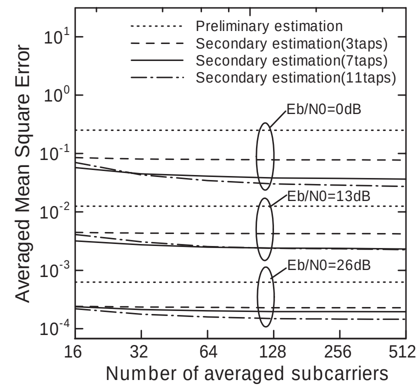


Fig. 6. Convergence of frequency-averaged MMSE weight ($\alpha = 0.4, \tau_d = 0.025T_d$).

mance without a priori knowledge of multipath delay channels. The proposed estimator can determine a combining weight for correlated subcarriers using a few pilot symbols, which is applicable to a burst packet. In addition, since the proposed estimator considers the effect of intersymbol interference and intersubcarrier interference in the frequency domain, the estimator always keeps good performance under any channel conditions.

The simulation results showed that the frequency-averaged MMSE estimator can effectively reduce estimation error and keeps better performance than the conventional robust estimator in any channel conditions. From practical point of view, the proposed estimator can achieve low complexity, choosing a suitable number of averaged subcarriers. Therefore, the proposed estimator is considered as a practical method for multicarrier channel estimator.

APPENDIX I
CORRELATION MATRIX OF $\boldsymbol{z}_n^{'}$

Let us consider the correlation between $z_n^{'}$ and $z_{n+\Delta n}^{'}$. Considering that $x_n(q)$ and $x_{n+\Delta n}(q)$ are statistically independent under $\Delta n \neq 0$, the correlation can be represented by

$$E[z_n^{'} z_{n+\Delta n}^{'*}] = \frac{1}{q_0^2}\sum_{q=1}^{q_0} E[z_n(q) z_{n+\Delta n}(q)^*] E[x_n(q)^* x_{n+\Delta n}(q)] = \begin{cases} P_z/q_0 & \Delta n = 0 \\ 0 & \text{otherwise} \end{cases} \quad (17)$$

In (17), $z_n^{'}$ and $z_{n+\Delta n}^{'}$ ($\Delta n \neq 0$) have no correlation, even if $z_n(q)$ and $z_{n+\Delta n}(q)$ have nonzero correlation. Therefore, we have $E_n[\boldsymbol{z}_n^{'} \boldsymbol{z}_n^{'\dagger}] = (P_z/q_0)\boldsymbol{I}$

Next, we consider estimation of interference-plus-noise power. From the relationship of

$$\begin{aligned} E[|y_n(q)|^2] &= E[|h_n x_n(q) + z_n(q)|^2] = |h_n|^2 + P_z \\ E[|\tilde{h}_n|^2] &= E[|h_n + z_n(q)^{'}|^2] = |h_n|^2 + P_z/q_0, \end{aligned}$$

we have

$$P_z = \frac{E[|y_n(q)|^2] - E[|\tilde{h}_n|^2]}{1 - 1/q_0}. \quad (18)$$

Considering the property of (18), we estimated the interference-plus-noise power under a finite number of averaged subcarriers as (10).

Appendix II
Multiplications to compute a correlation matrix

We consider a method to reduce multiplications to compute a correlation matrix $\mathbf{\Phi}$. From (5), (r_1, r_2) element of $\mathbf{\Phi}$ is given by

$$[\mathbf{\Phi}]_{r_1,r_2} = \frac{1}{N-2n_0}\sum_{n=n_0+1}^{N-n_0} \tilde{h}_{n-n_0-1+r_1}\tilde{h}^*_{n-n_0-1+r_2} \quad (19)$$

According to (19), $[\mathbf{\Phi}]_{r_1,1}$ $(r_1 = 1, ..., 2n_0+1)$ is initially computed. Then, complexity can be reduced by calculating other elements successively using the follwing equations:

$$[\mathbf{\Phi}]_{r_1+1,r_2+1} = [\mathbf{\Phi}]_{r_1,r_2} + \frac{\tilde{h}_{N-2n_0+r_1}\tilde{h}^*_{N-2n_0+r_2} - \tilde{h}_{r_1}\tilde{h}^*_{r_2}}{N-2n_0}$$
$$[\mathbf{\Phi}]_{r_1,r_2} = [\mathbf{\Phi}]^*_{r_2,r_1} \quad (20)$$

Based on this method, the total number of multiplications T for the correlation matrix is reduced to

$$T = N(2n_0+1). \quad (21)$$

Appendix III
Conventional robust estimator

A channel estimator robust to channel statistics can be composed by assuming uniform power-delay profile within guard interval [1]–[3]. To avoid a large dimension of matrix computation, total N subcarriers are divided into K equally sized blocks. The conventional robust estimator estimates the channel frequency response for a block of subcarriers as

$$\hat{\boldsymbol{h}}_n = \boldsymbol{R}_{hh}(\boldsymbol{R}_{hh} + (P_z/q_0)\boldsymbol{I})^{-1}\tilde{\boldsymbol{h}}_n$$

where the correlation matrix $\boldsymbol{R}_{hh} = E[\boldsymbol{h}\boldsymbol{h}^\dagger]$ is computed by uniform power-delay profile within guard interval.

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