Low-complexity Mean Delay Estimation for OFDM Systems

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Abstract—In this letter, we propose two simple OFDM mean delay estimation schemes, channel impulse response (CIR)-based and differential schemes, which are applicable for adaptive algorithms including improved channel estimation and fast handover. Instead of time-domain processing requiring additional inverse fast Fourier transform (IFFT), the mean delay can be obtained in the proposed CIR-based scheme using the estimated frequency-domain CIR. When cell-specific preamble sequences are employed, the mean delay can also be estimated without CIR estimation in the proposed differential scheme using cross-correlation of differentially correlated received signals, enabling further complexity reduction and performance improvement when the CIR estimation is not sufficiently accurate. The performance of the proposed schemes is verified by computer simulation.

I. INTRODUCTION

The use of adaptive algorithms in orthogonal frequency division multiplexing (OFDM) systems can enhance system capacity and utilize available resources in an efficient manner [1]. It requires a form of accurate parameter measurements. One of the key parameters in adaptation of OFDM systems is time dispersion or delay spread of the channel [2]. Once the maximum channel delay information is available, the spectral efficiency can be enhanced by adapting the length of the guard interval (GI) [3]. Adaptive filtering for channel impulse response estimation (CE) is another area where the time dispersion information of the channel is useful [4], [5], [6].

A mean delay is also one of the indicators manifesting the channel delay spread used for adaptive processing. For example, the performance of frequency-domain interpolation can be improved by shifting the mean delay before interpolation, followed by re-shifting after the interpolation [6]. For cellular systems, handover can be performed fast by prioritizing a base station (BS) having smaller mean delay and stronger received signal power for candidate cell selection [9].

The mean delay can be conventionally measured by taking the inverse fast Fourier transform (IFFT) of the frequency domain channel impulse response (CIR) which is calculated using received pilot subcarriers [4], [7]. Although a cyclic prefix (CP) can also be used for delay spread estimation as in [8], it cannot distinguish multiple cells and may suffer from degradation since only the CP portion is exploited for correlation instead of a full OFDM symbol.

In this letter, we propose two low-complexity mean delay estimation schemes. The proposed CIR-based scheme directly uses the estimated frequency-domain CIR without IFFT operation based on a new approximation on the mean delay expression. We also propose the differential scheme when cell specific pseudo-noise sequences are modulated on the preamble signal as in IEEE 802.16e [10]. Using the low cross-correlation property of the preamble sequences, the differential scheme can further reduce implementation complexity by skipping the CE procedure and moreover improve the performance when the estimated CIR is not quite accurate.

Following Introduction, the system model is described in Section II. Section III describes the conventional and proposed delay estimation schemes followed by performance and complexity evaluation in Section IV. Finally, conclusions are drawn in Section V.

II. SYSTEM MODEL

Consider an OFDM transmitter at the \( i \)-th BS (or cell) employing \( N \) subcarriers where a predetermined preamble sequence \( X_i[k], k=0, 1, 2, \ldots, N-1 \), is transmitted at the \( k \)-th subcarrier in the first OFDM symbol of each frame. The preamble symbol is converted into a time domain signal by the IFFT process followed by CP insertion. In order to make the preamble signal composed of \( q \) repeated signals in time for easier synchronization [10], the cell-specific preamble sequence \( X_i[k] \) is given by

\[
X_i[k] = \begin{cases} 
\pm 1, & \text{mod}(k, q) = p_{s,i} \\
0, & \text{otherwise}
\end{cases}
\]  

where \( \text{mod}() \) is a modular function and \( p_{s,i} \) is a segment index of the \( i \)-th BS, \( p_{s,i} \in \{0, 1, \ldots, q-1\} \). Here, \( q \) is equal to the subcarrier interval of the non-zero preamble subcarrier (e.g., \( q = 3 \) for 802.16e [10]). The transmit signals from multiple BS’s are synchronized and each BS has its own distinct preamble sequence and segment index, i.e., BS index \( i \) identifies the preamble sequence and the segment index [10].

Modeling a continuous-time channel with a sample-spaced taped delay line considering sampling after lowpass filtering at the receiver [12], [13], we can represent the received time-domain signal in a discrete form as

\[
y[n] = \sum_{i=0}^{L_i-1} \left( g_i \sum_{m=0}^{L_i-1} h_i[n, m] x_i[n - m] \right) + z[n]
\]  

where \( x_i[n], g_i \) and \( L_i \) are the time-domain transmit preamble signal at time \( n \), quasi-static large-scale gain due to path
loss and scattering, and the number of multipaths of the \( i \)-th channel link between the \( i \)-th BS and the mobile station (MS), respectively. \( f \) is the total number of adjacent BS’s (or their preamble sequences), where \( i = 0 \) denotes an index of the serving BS. \( z[n] \) is the background noise plus interference at time \( n \), which can be approximated as a zero mean additive white Gaussian noise (AWGN) with variance \( \sigma^2 \). \( h_i[n, m] \) is the CIR of the \( m \)-th sample delay at time \( n \) which is normalized to \( E \left\{ \sum_{m=0}^{L_i-1} |h_i[n, m]|^2 \right\} = 1 \). Omitting the sample index \( n \) for simple description with the assumption of negligible change of the CIR during an OFDM symbol, the mean delay of the \( i \)-th channel link can be given by

\[
\tau_i = \frac{\sum_{m=0}^{N-1} m |h_i[m]|^2}{\sum_{m=0}^{N-1} |h_i[m]|^2} \quad \text{and} \quad \tau_i = \frac{\sum_{m=0}^{L_i-1} m |h_i[m]|^2}{\sum_{m=0}^{L_i-1} |h_i[m]|^2} \tag{3}
\]

where the maximum channel delay \( L_i \) is assumed to be equal to or smaller than the length of GI, \( N_g \), i.e., \( L_i \leq N_g \) [4], [14].

Removing the CP followed by the FFT process at the MS receiver and omitting the OFDM symbol index for brevity, the received preamble symbol of the \( k \)-th subcarrier of the preamble symbol can be represented by

\[
Y[k] = \sum_{i=0}^{l-1} g_i H_i[k] X_i[k] + Z[k] \tag{4}
\]

where \( H_i[k] \) is the frequency-domain CIR at the \( k \)-th subcarrier of the \( i \)-th channel link and \( Z[k] \) is the background noise plus interference term.

III. CONVENTIONAL AND PROPOSED MEAN DELAY ESTIMATION SCHEMES

A. Conventional CIR-based scheme

The mean delay estimate for the \( j \)-th BS is conventionally obtained by substituting the time-domain CIR estimate \( \hat{h}_j[m] \) into (3). Here, \( \hat{h}_j[m] \) is calculated by the IFFT of the estimated frequency-domain CIR \( H_j[k] \), which is usually obtained by filtering initial channel estimates as

\[
\hat{H}_j[k] = \sum_{m \in P_j} a_{k,m} \tilde{H}_j[m] \tag{5}
\]

where \( a_{k,m} \) is the filter coefficient for the \( k \)-th subcarrier output from the \( m \)-th subcarrier input. The initial channel estimate \( \tilde{H}_j[m] \) is equal to \( Y[m] / X_j[m] \), \( m \in P_j \) where \( P_j \) is a set of the subcarrier indices used for preamble sequence \( j \), i.e., \( P_j \in \{ lq + p_{x,j} | l = 0, 1, \ldots, N/q - 1 \} \) [4], [5].

Here, \( p_{x,j} \) is a segment index for the \( j \)-th BS and \( (N/q) \) is assumed to be an integer. Various linear filters can be used for (5), where we consider linear minimum mean square error (LMMSE) and simple local moving average (MA) estimation [1]. The LMMSE filter, whose tap size is equal to the length of the preamble sequence (=\( N/q \)), is used as a reference for the optimum performance while less practical due to substantial complexity burden. In addition to a large number of calculations with filtering, the LMMSE CE also requires information on the signal to noise power ratio (SNR) and the channel correlation in the frequency domain which is equivalent to an average power-delay profile (PDP) in the time domain. Note that this average PDP is not easy to obtain since it needs to know the mean power of all the paths. On the other hand, the MA filter having equal filter coefficients is widely used in practice [6] since it provides relatively good performance, is easily implementable with some addition operations and only a single multiplication for normalization, and requires no extra information such as the SNR or the frequency-domain channel correlation.

As an example of applications, the mean delay can be used for improved CE through shifting the time-domain CIR response by the mean delay [6]. It can be realized by rotating phase of \( \hat{H}_j[m] \) by \( \exp(j2\pi m \alpha \tau_j / N) \) before frequency-domain filtering (5) and by de-rotating the phase after the filtering.

B. Proposed CIR-based scheme

On the contrary to the above conventional method, the first scheme that we propose can calculate the mean delay from the \( j \)-th BS without the IFFT operation of the frequency-domain CIR using a newly developed approximation as follows.

The phase of frequency-domain CIR correlation can be written by

\[
\begin{align*}
\frac{N}{2\pi q} \zeta &= \left( \sum_{k \in P_j} H_j[k] H_j^*[k + q] \right) \\
\frac{N}{2\pi q} \zeta &= \left( \sum_{m=0}^{N-1} \sum_{n=0}^{N-1} h_j^*[m] h_j[n] e^{j2\pi mq/N} \left( \sum_{k \in P_j} e^{j2\pi mkq/N} \right) \right) \\
\frac{N}{2\pi q} \zeta &= \left( \sum_{m=0}^{N-1} \sum_{n=0}^{N-1} h_j^*[m] h_j[n] e^{j2\pi mq/N} \frac{N}{q} \delta(m - n) \right) \\
\frac{N}{2\pi q} \zeta &= \left( \sum_{m=0}^{L_j-1} |h_j[m]|^2 e^{j2\pi mq/N} \right) \tag{6}
\end{align*}
\]

where \( \tilde{P}_j \) is a set of \( P_j \) excluding the highest subcarrier index and \( \zeta X \) is the phase of complex variable \( X \). Using the following second order Taylor series approximation around the mean delay

\[
e^{j2\pi m \alpha \tau_j} \approx e^{j\alpha \tau_j} + j\alpha e^{j\alpha \tau_j} (m - \tau_j) - \frac{\alpha^2 e^{j\alpha \tau_j}}{2} (m - \tau_j)^2, \quad \text{for } |m - \tau_j| \leq 1 \tag{7}
\]

where \( \alpha = 2\pi q / N \), (6) can be represented as (8) at the top of the next page, where the equality from the second row to the third holds based on (3).

The approximation (8) becomes more accurate for smaller \( |2\pi(m - \tau_j)q/N| \). Even if this constraint is not satisfied for some path delays of \( m \) far from the mean delay, the approximation is usually accurate as justified in the following.

Note that the mean delay is the center of mass of the PDP, implying that the phase difference between \( m \)-th delay and the mean delay \( \tau_j \) is reduced as the \( m \)-th delay approaches
the mean delay. When there is a single path in the PDP, the mean delay is equal to the path delay, resulting no error in the approximation. When there are multiple paths in the PDP which normally happens, the paths around the mean delay contribute little to the approximation error due to small $|2\pi(m - \tau_j)q/N|$ and only the paths far from the mean delay mainly increase the error. If these paths have small channel power, the effect of the approximation (7) on the total error in (8) is reduced due to low $|h_j[m]|^2$ of the paths in spite of large $|2\pi(m - \tau_j)q/N|$. As a result, the approximation error is maximized when there are multiple large-powered paths scattered across the PDP. In order to evaluate the approximation error in this worst case, we consider a Rayleigh fading channel having uniform PDP $(E\{|h_j[m]|^2\} = 1/L_j = 1/N_g$, for all $m$) where all the paths in the PDP are likely to have large power. The standard deviation (STD) of the approximation error (8) of the mean delay is plotted in terms of maximum channel delay $L_j$ in Fig. 1 (a) where the channel is generated $10^4$ times. It can be seen that the error radically increases after a threshold of $L_j$ for a given $q$ and that the threshold decreases for larger $q$ since $(m - \tau_j)q$ is a metric determining the error. Since the OFDM system experiences severe performance degradation when the maximum delay is larger than the CP duration due to intercarrier and intersymbol interference, the approximation is acceptable if the error is small up to a given CP ratio. It can be seen from Fig. 1 (a) that the approximation is satisfactory in practice. For example, the STD of the error is less than 0.2 sample for the IEEE 802.16e WiMAX system where $q=3$ and the mandatory CP ratio is 1/8 [10], [11]. The STD is also plotted in Fig. 1 (b) when the PDP has exponential power distribution with a path exponent of -0.25dB. Even for this small exponent (i.e., power decreases slowly as the path delay increases), it can be seen that the approximation error is significantly smaller than that of the uniform PDP, implying that the approximation is quite accurate in practice. Therefore, unlike the conventional scheme requiring the IDFT and calculation of (3), the mean delay can be directly estimated using the estimated frequency-domain CIR based on the approximation (8) by

$$\hat{\tau}_j \approx \frac{N}{2\pi q} \zeta \left( \sum_{k \in P_j} \tilde{H}_j[k] \tilde{H}_j^*[k + q] \right). \quad (9)$$

Note that the mean delays from BS’s other than the $j$-th BS can be similarly estimated based on (3) or (9) in the conventional and proposed CIR-based scheme, respectively, through demodulation of the corresponding preamble sequences, i.e., $\tilde{H}_j[m] = Y[k]/X_i[m], m \in P_i$ for $i \neq j$.

C. Proposed differential scheme

Note that the performances of the conventional and proposed CIR-based schemes are affected by the accuracy of CE, thus experiencing large errors when CE degrades, e.g., in deep fading or severe frequency selective fading, or under large inter-cell interference. Moreover, the receiver has to estimate CIR for the whole frequency band even if it is unnecessary for coherent detection. Thus, the CIR-based mean delay estimation has a risk of performance degradation and requires extra complexity, especially for cellular systems where CIR may need to be tracked for all the candidate cells. This drawback can be mitigated by exploiting the low cross-correlation property of the differential transmit preamble sequence $D_j[k] = X_j[k]X_j^*[k + q]$ as follows.

The differentially correlated signal of the received preamble subcarrier can be computed by

$$M[k] = Y[k]Y^*[k + q]$$

$$\quad = \sum_{i=0}^{l-1} g_i^2 H_i[k]H_i^*[k + q]D_i[k] + \tilde{Z}[k]$$

$$\quad + \sum_{i=0}^{l-1} \sum_{p=0, p \neq i}^{l-1} g_i g_p H_i[k]H_p^*[k + q]X_i[k]X_p^*[k + q] \quad (10)$$

where $\tilde{Z}[k]$ is the noise and interference term given by

$$\tilde{Z}[k] = Z[k] \sum_{p=0}^{l-1} g_p H_p^*[k + q]X_p^*[k + q]$$

$$\quad + Z^*[k + q] \sum_{i=0}^{l-1} g_i H_i[k]X_i[k] + Z[k]Z^*[k + q] \quad (11)$$

Then, the cross-correlation of $M[k]$ and $D_j[k]$ can be repre-
In addition, the last term can be neglected since the noise correlated term in the second line of (12) becomes negligible.

\[
\sum_{k \in P_j} M[k] D^*_j[k] \approx \sum_{k \in P_j} \left( g_j^2 H_j[k] H^*_j[k + q] D_j[k] D^*_j[k] \right) + \sum_{i=0, i \neq j}^{I-1} \left( \sum_{k \in P_j} \left( g_i^2 H_i[k] H^*_i[k + q] D_i[k] D^*_j[k] \right) \right). \quad (13)
\]

Note that the maximum cross-correlation of \( D_j[k] \) is relatively low due to the pseudo noise characteristics of \( X_j[k] \). For example, it has the maximum value for 1024 FFT [10] as

\[
\max_{j \neq i} \left( \frac{\sum_{k \in P_j} D_i[k] D^*_j[k]}{\sum_{k \in P_j} |D_j[k]|^2} \right) \approx 0.1731 \quad (14)
\]

where the correlation output is usually much less than the maximum value. Thus, using \( |D_j[k]| = 1 \), (13) can be represented by

\[
\sum_{k \in P_j} M[k] D^*_j[k] \approx \sum_{k \in P_j} g_j^2 H_j[k] H^*_j[k + q] \quad (15)
\]

when the effect of the received signal from the desired cell is dominant over those of other cells. Note that the approximation (15) is quite accurate for single-cell systems (\( I = 1 \)). However, it may have a noticeable error in cellular systems when inter-cell interference power is relatively strong. It occurs when either the signal from the desired \( j \)-th BS is in deep fading (i.e., low \( |H_j[k]| \)) or other cells severely interfere with high power (i.e., large \( g_i \), \( i \neq j \)). Assuming that the effect of the inter-cell interference is negligible until the next subsection, the mean delay can be estimated by

\[
\hat{\tau}_j = \frac{N}{2\pi q} \left( \sum_{k \in P_j} M[k] D^*_j[k] \right). \quad (16)
\]

D. Mitigation of inter-cell interference effect in cellular systems

Although the proposed differential scheme can provide good performance in single-cell systems, the mean delay in (16) may become too erroneous when there is significant inter-cell interference. The same problem occurs for the CIR-based estimation schemes (3) and (9) due to inaccurate CE. In order to mitigate this performance degradation, the mean delay can be calculated only when the output is reliable. As a simple metric indicating desired-cell signal power, \( \hat{\gamma}_j \) is considered as

\[
\hat{\gamma}_j = \left| \sum_{k \in P_j} M[k] D^*_j[k] \right|. \quad (17)
\]

Note that (17) shares the same operation as (16) except magnitude calculation. When the link from the \( j \)-th BS is not reliable due to strong inter-cell interference, \( \hat{\gamma}_j \) is likely to be much smaller than other \( \hat{\gamma}_i \)'s, \( i \neq j \). Thus, we calculate the
mean delay of the $j$-th BS only when the following condition holds

$$\hat{\gamma}_j > \gamma_{\text{max}} - \eta$$  \hspace{2cm} (18)

where $\eta$ is a given threshold and $\gamma_{\text{max}}$ is the maximum $\hat{\gamma}_i$, i.e., $\arg\max \hat{\gamma}_i$, $i = 0, 1, \ldots, I-1$. Finally, the delays of the desired BS’s can be obtained by averaging the outputs of (3), (9) or (16) within a given time window.

IV. EVALUATION OF THE COMPLEXITY AND PERFORMANCE

A. Complexity comparison

In this subsection, implementation complexity needed at the preamble symbol is compared between the conventional and proposed schemes. As seen in Table 1, the CIR-based schemes need CE operation for the whole band whose complexity varies according to a deployed CE scheme. The LMMSE CE involves $N^2/q$ multiplication for filtering in addition to calculation of the SNR, frequency-domain channel correlation, and matrix inversion whenever its filter coefficients are updated. However, the MA CE only requires $N/q$ multiplications with less complicated additions. The conventional CIR-based scheme additionally requires $N \log N$ multiplications for IFFT, and $3N_q$ multiplications and one division for (3), whereas the proposed CIR-based scheme only requires $N/q$ multiplications and one phase calculation for (9). The differential scheme can further reduce the computational burden since it can avoid CE operation and just requires $N/q$ multiplications and one phase calculation for (16) where the calculation of $D_j[k]$ can be available by sign change of binary sequences $X_j[k]$. As an example, the total number of multiplications is also shown in Table 1 for the IEEE 802.16e WiMAX system [10], [11] where $N=1024$, $q=3$, and $N_q/N$ (CP ratio)=1/8. It can be seen that the CE (especially for LMMSE-CE) and IFFT require lots of computation and thus the proposed schemes can reduce the complexity burden remarkably.

B. Performance comparison

The performance of the proposed schemes is evaluated using computer simulation under the IEEE 802.16e WiMAX system [10], [11]. The system parameters are summarized in Table 2. Each BS has its own preamble sequence and segment index $p_s, p_s \in \{0, 1, 2\}$ [10]. In order to evaluate the effect of multicell interference, we assume that there are at most two BS’s (named BS0 and BS1) having the same $p_s$ since nearby cells usually avoid intercell preamble interference by choosing different subcarrier sets, i.e., assigning different $p_s$. Two values of signal to interference power ratio (SIR), which is the ratio of received mean signal power of BS0 to BS1, are considered: SIR=30dB is close to single BS environments while SIR=0dB is taken for considering large multicell interference.

For single-cell environments (i.e., SIR=30dB), the normalized mean square errors (MSE’s) of the mean delay estimates of the conventional CIR-based scheme (3), the proposed CIR-based scheme (9) and the proposed differential scheme (16) are plotted in Fig. 2 without threshold comparison of (18), i.e., $\eta = \infty$. For conventional and proposed CIR-based schemes, both the LMMSE-CE and simple 3-tap MA CE, $a_k,m=1/3$, $m \in \{k - q, k, k + q\}$ are simulated for CE. Assuming that the average estimation error being equal to or less than one sample (MSE$\leq 1$) is satisfactory for proper operations of CE improvement or handoff [14], all the schemes provide good performance. Note that the performance of the conventional and proposed CIR-based schemes is close for both the LMMSE and MA-CE except for the LMMSE-CE at high SNR, validating the approximation of (8) at the MSE of interest. Although the LMMSE-CE CIR-based schemes provide the best performance, the differential scheme has better performance than MA-CE based schemes except for low SNR (around 0dB) where noise enhancement occurs in the differential scheme due to differential operations. This implies that CE accuracy is critical to mean delay estimation. That is, unless the estimated CIR is sufficiently accurate with the aid of complicated CE methods such as LMMSE CE, the CIR-based scheme may experience performance degradation in mean delay estimation, being inferior to the differential scheme in respect to the performance as well as the complexity. In other words, it implies that the differential scheme, which solely estimates the mean delay itself, has advantages over the CIR-based schemes whose performance of mean delay estimation relies on the CE performance. Note that although the performance under the ITU Pedestrian-B channel [13] is only plotted, similar tendency can be observed even for longer delay spread channels including the uniform PDP channel $(E|\{h_j[m]\}|^2 = 1/L_j = 1/N_g)$, which validates the applicability of the proposed schemes.

For multi-cell environments, all the conventional and proposed schemes experience degradation as seen in Fig. 3 (a) due to intercell interference. This can be significantly mitigated with threshold comparison (18) as seen in Fig. 3 (b), where $\eta$ is set to 2dB. Although the curves for other $\eta$’s are not included due to limited space, all the schemes provide
TABLE I
 COMPLEXITY IN RESPECT TO THE NUMBER OF COMPLEX MULTIPLICATIONS

<table>
<thead>
<tr>
<th>Scheme</th>
<th>CE</th>
<th>IFFT</th>
<th>Others</th>
<th>Sample values for WiMAX</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>LMMSE</td>
<td>MA</td>
<td>N log N, 3 N_p (+1 division)</td>
<td>360,149, 10,965</td>
</tr>
<tr>
<td>Conventional CIR-based</td>
<td>N^2/q</td>
<td>N/q</td>
<td>0</td>
<td>349,867, 682</td>
</tr>
<tr>
<td>Proposed CIR-based</td>
<td></td>
<td></td>
<td>0</td>
<td>341, 341</td>
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<tr>
<td>Proposed differential (No CE)</td>
<td>0</td>
<td>0</td>
<td>N/q (+1 phase calculation)</td>
<td></td>
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</table>

TABLE II SYSTEM PARAMETERS

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bandwidth</td>
<td>10 MHz</td>
</tr>
<tr>
<td>FFT Size</td>
<td>1024 (86 guard subcarr. on each side)</td>
</tr>
<tr>
<td>Guard interval ratio</td>
<td>1/8</td>
</tr>
<tr>
<td>Number of BS’s (I)</td>
<td>2</td>
</tr>
<tr>
<td>having the same segment index</td>
<td></td>
</tr>
<tr>
<td>Channel</td>
<td>Pedestrian-B</td>
</tr>
<tr>
<td>Cell configuration</td>
<td>Cell radius: 1km, path loss exp.: 3</td>
</tr>
<tr>
<td>SIR (dB) = (g_0/g_1)^2</td>
<td>30, 0</td>
</tr>
<tr>
<td>Number of preamble sequences</td>
<td>38 (=114/3) per cell</td>
</tr>
<tr>
<td>Preamble power boosting</td>
<td>8 (9.03dB)</td>
</tr>
</tbody>
</table>

similar performance for 1.5dB ≤ η ≤ 4dB. Similar to the single-cell case, the proposed CIR-based scheme experiences slight performance degradation over the conventional CIR-based scheme. Again, the LMMSE-CE CIR-based schemes provide the best performance followed by the differential scheme and then by the MA-CE CIR-based schemes. Note that the performance superiority of the LMMSE-CE may not be allowed in practice considering the complexity burden as described in the previous subsection. Thus, the proposed CIR-based and differential schemes can be good candidates in practice considering the complexity burden and estimation performance.

V. CONCLUSION

In this letter, we have proposed two low-complexity mean delay estimation schemes for OFDM systems. The proposed CIR-based scheme has an advantage of skipping the IFFT operation over the conventional scheme with slight performance degradation. The proposed differential scheme can further reduce the complexity without IFFT and CE operation, and can also provide better performance when the CIR estimation is not sufficiently accurate in the conventional scheme. The proposed schemes can be applied to multi-cell as well as single-cell systems, thus enabling the use at various applications including fast handoff and improved channel estimation.

REFERENCES

Fig. 3. Performance for two-cell environments (SIR=0dB)

(a) Without threshold setting

(b) With threshold setting of $\eta = 2\text{dB}$