Channel Estimation for Non-Line-of-Sight WiMax Communication System

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ABSTRACT - In this paper, we present an adaptive channel estimation scheme for IEEE 802.16-2004 Wireless Metropolitan Area Network (a.k.a. WiMAX) in the case that the channel impulse response is longer than the cyclic prefix (CP). The scheme is combined with time domain impulse response shortening technique and frequency domain channel equalization. Simulation results show that the shortening algorithm gives the shortening SNR (SSNR) of effective channel is at least 10 times better than the original channel impulse response. The performance of proposed method can satisfy the BER requirement of IEEE 802.16-2004 standard in the non-line-of-sight environment with the highest data rate (20MHz) transmission. We have also found that the better strategy to design a shortening filter is to over-constraint the desired length to much shorter than CP.

1. INTRODUCTION

WiMAX (Worldwide Interoperability for Microwave Access) is defined for broadband wireless access (BWA). It is increasingly gaining interests as an alternative “last-mile” technology to DSL lines and cable modems. This technology aims to provide broadband wireless access to residential and small business applications, as well as to enable Internet access in countries without any existing wired infrastructure in place. So it is also the best candidates for wireless links between Wireless Local Area Networks (WLAN) and the Internet. It specifies channel bandwidth from 1.75MHz up to 20MHz and cell distance up to 30 miles.

Wireless communication systems usually suffer from frequency selective and time variant channel. Orthogonal Frequency Division Multiplexing (OFDM) technique is widely adopted in those systems due to it’s robustness against multipath fading and simpler equalization scheme. In most of applications, for retaining the orthogonality of subcarriers and overcome intersymbol interference (ISI), a cyclic prefix (CP) is inserted instead of simply inserting guard interval. For CP length equal to L, the CP is the last L samples of original symbol. If the maximum delay of the multipath channel does not exceed the CP length, the OFDM system would be ISI free by removing the guarding interval. For WiMAX systems, its delay spread is typically over several micro-seconds which is easily longer than the guarding interval. Therefore, it is very challenging to maintain the system BER performance for non-line-of-sight (NLOS) channels at high data rate transmission with high bandwidth (e.g. 20MHz) since high sampling rate is needed and the CP length is often much shorter than the physical channel impulse response (CIR).

For the channel has longer delay spread than the CP length, we can apply a finite impulse response (FIR) shortening filter in the receiver. The purpose of this filter is to shorten the impulse response of the effective channel which is the convolution of the transmission filter, physical channel, receiver filters, and shortening filters. If the shortened response is less than CP length, the system would be ISI free.

This paper is organized as follows. In section 2 we will introduce the background of impulse response shortening and IEEE 802.16-2004 standard packet format. Then, a combination of LMS channel estimator and shortening impulse response filter (SIRF) architecture is presented in Section 3. Simulation results are next provided in Section 4. Finally, Section 5 gives the conclusion.

2. BACKGROUND

A. SUI Channels Models

Stanford University Interim (SUI) channel models provide the basis specifying channel for a given scenario. The channel parameters are related to terrain type, delay spread, and antenna directionality. It defines six channel models for different environment; each channel model has three taps with distinct K-factor and average power. The SUI channel models have the line-of-sight channels which are referred to SUI-1, SUI2, and SUI-6, and non-line-of-sight channels which include SUI-3, SUI-4, and SUI-5 [5]. Table 1 shows an example of time domain attribute of the SUI-4 channel, which is chosen to evaluate the proposed algorithm.

<table>
<thead>
<tr>
<th>Table 1: SUI-4 channel model</th>
</tr>
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<tbody>
<tr>
<td>SUI-4</td>
</tr>
<tr>
<td>-------</td>
</tr>
<tr>
<td>Delay</td>
</tr>
<tr>
<td>Power (00% K-factor)</td>
</tr>
<tr>
<td>Power (30% K-factor)</td>
</tr>
<tr>
<td>Normalization factor:</td>
</tr>
</tbody>
</table>

B. Channel Shortening

In OFDM system, when the maximum delay is longer
than the CP duration which is at most 1/4 symbol duration, the ISI effect is irreducible from the system and will cause the following blocks losing efficacy. If a shortened impulse response filter (SIRF) \( w(n) \) with length \( W \) is applied before removing CP block in the receiver, which is indicated in Fig. 1, the output of the SIRF can be expressed as

\[
y(n) = (h(n) \ast w(n)) \ast x(n) = h_{\text{eff}}(n) \ast x(n) \quad (1)
\]

where \( x(n) \) is the transmitted data, \( \ast \) denotes the convolution operator, and \( h_{\text{eff}}(n) \) is the effective channel impulse response after shortening.

On a point of view of adding a FIR to obtain the effective channel, it is generally not possible to shorten the impulse response perfectly. On the contrary, the effective channel is always longer than the physical channel due to convolution. A measurement using to check the shortening performance is called shortening SNR (SSNR), which is defined as the ratio of the energy lies within the desired length \( D \) to the energy lies out of \( D \). In OFDM application, the desired length is often set to be the length of CP\[3,4]\.

In our proposed scheme, we have observed that this is not the best case. Instead, it is better to have D shorter than the CP. We will discuss about this later.

\begin{figure}
\centering
\includegraphics[width=0.4\textwidth]{fig1}
\caption{Simplified WiMax transceiver diagram with SIRF}
\end{figure}

\section{C. Symbol description}

The OFDM symbol time domain structure and frequency domain description in WiMAX are illustrated in Fig. 2 and Fig. 3 respectively \[1\]. In Fig. 2, the useful symbol time is \( T_s \). A copy of the last \( T_s \) of the useful symbol period, termed CP, is used to combat multipath, while maintaining the orthogonality of each subcarrier. The CP length can be configured to 1/32, 1/16, 1/8, and 1/4 of the period \( T_s \). In Fig. 3, total of 256 subcarriers are spread to four parts. There are 192 subcarriers for data transmission, 8 subcarriers for pilot tone, 52 subcarriers for guard bands, and 1 subcarrier for DC in every OFDM symbol. The data modulation type can be BPSK, QPSK, 16-QAM, or 64-QAM. The constellations shall be normalized by multiplying the constellation point with the appropriate factor to achieve equal average power. The pilots are generated by Pseudo Random Binary Sequence (PRBS) and modulated by BPSK. These pilots are used for phase tracking and symbol re-timing estimation. The packet format is shown in Fig. 4. The first two consecutive OFDM symbols are short preamble and long preamble respectively. The short preamble uses only subcarriers the indices of which are a multiple of 4. As a result, the time domain waveform of the first symbol consists of four repetitions of 64-sample fragment, preceded by a CP. The long preamble utilizes only even subcarriers, resulting in time domain structure composed of two repetitions of a 128-sample fragment, preceded by a CP. The data symbols follow the long preamble. The preamble are used for symbol timing, synchronization, and channel estimation.

\begin{figure}
\centering
\includegraphics[width=0.4\textwidth]{fig2}
\caption{OFDM symbol time domain structure}
\end{figure}

\begin{figure}
\centering
\includegraphics[width=0.4\textwidth]{fig3}
\caption{OFDM frequency domain description}
\end{figure}

\begin{figure}
\centering
\includegraphics[width=0.4\textwidth]{fig4}
\caption{Packet format in time domain}
\end{figure}

\section{3. PROPOSED CHANNEL ESTIMATION METHOD}

The proposed channel estimation method consists two parts: 1) the time domain equalizer (TEQ) and 2) frequency domain equalizer (FEQ) as the block diagram is depicted in Fig. 5. The TEQ is designed to mitigate the ISI by shortening filter and then the FEQ follows to mitigate the multipath effect within the OFDM symbol.

In the following we describe the operation of the scheme.

\begin{figure}
\centering
\includegraphics[width=0.4\textwidth]{fig5}
\caption{The proposed channel estimation diagram}
\end{figure}

\section{A. Time Domain Equalizer}

The TEQ is composed of a LMS channel estimator, a shortening coefficient calculator, and two FIR filters as SIRF. When synchronization process is done, the time domain ISI-polluted preamble is sent into the LMS channel estimator. By choosing LMS step-size appropriately, we get the estimated CIR \( h'(n) \) and then see whether it is longer than the CP length or not. If so, the shortening calculator is enabled to calculate the tap-weight of the SIRF, otherwise the weights will be set to be \( [1, \ 0, \ 0 \ ...] \) as the trivial convolution operation. Finally, both the estimated CIR and time domain preamble are passed through the SIRF and FFT blocks parallel as the frequency domain channel response and
preamble respectively.

Let the physical channel impulse response \( h(n) \) has length \( M \) and the equalized channel impulse response \( h_{\text{eff}}(n) \) has length of \( M+W-1 \). We define \( h_{\text{wall}}(n) \) as the equalized impulse response in the window of desired length \( D \) and \( h_{\text{win}}(n) \) as the the equalized impulse response outside the window of desired length \( D \) as shown in Eq.(2).

\[
\begin{align*}
    h_{\text{wall}} &= [h_y(0), h_y(1), \ldots, h_y(D-1)]^T \\
    h_{\text{win}} &= [h_y(D), h_y(D+1), \ldots, h_y(M+W-1)]^T
\end{align*}
\]

Let \( H, H_{\text{wall}} \), and \( H_{\text{win}} \) be the convolutional matrices of the physical channel \( h(n) \), \( h_{\text{wall}}(n) \), and \( h_{\text{win}}(n) \) respectively. The energy outside and inside the window can be expressed as

\[
\begin{align*}
    h_{\text{wall}}^T H_{\text{wall}} &= w^T H_{\text{wall}}^T H_{\text{wall}} = w^T A w \\
    h_{\text{win}}^T h_{\text{win}} &= w^T H_{\text{win}}^T H_{\text{win}} = w^T B w
\end{align*}
\]

where \( A \) and \( B \) are symmetry and positive semidefinite matrices. Then the SSNR is defined as

\[
\text{SSNR} = \frac{w^T H_{\text{win}}^T H_{\text{win}} w}{w^T H_{\text{wall}}^T H_{\text{wall}} w} \tag{4}
\]

The optimum shortening algorithm tries to choose \( w \) to maximize the energy inside the DL (i.e. \( h_{\text{win}}^T h_{\text{win}} \)) with the constraint that the channel energy outside remains unity (i.e. \( h_{\text{wall}}^T h_{\text{wall}} = 1 \)). To reformulate the constraint, we define

\[
y = \sqrt{A}^T w
\]

such that

\[
y^T y = w^T \sqrt{A} \sqrt{A}^T w = w^T A w = 1 \tag{6}
\]

From the definition of \( y \), we have

\[
w = (\sqrt{A}^T)^{-1} y
\]

Therefore,

\[
h_{\text{win}}^T h_{\text{win}} = w^T B w
\]

\[
y^T = (\sqrt{A}^T)^{-1} B(\sqrt{A}^T)^{-1} y = y^T C y
\]

where \( C \) is defined as

\[
C = (\sqrt{A}^T)^{-1} B(\sqrt{A}^T)^{-1}
\]

The optimal shortening is thus expressed as finding \( y \) to maximize \( y^T C y \) while constraining \( y^T y = 1 \). The solution to this problem occurs when \( y = l_{\text{max}} \), where \( l_{\text{max}} \) is the unit-length eigenvector corresponding to the maximum eigenvalue \( \lambda_{\text{max}} \) of \( C \). The optimum solution of the SIRF coefficient is then

\[
w_{\text{opt}} = (\sqrt{A}^T)^{-1} l_{\text{max}} \tag{10}
\]

As a result, the corresponding SSNR is maximized.

To estimate the channel, the least square (LS) optimization is often used, but it is not suitable for non-stationary environments. For local-stationary environments, we need to estimate the autocorrelation in a short period when we use the LS optimization technique. The method of substitution is to iterate the filter weights to reach the optimal solution as well as to track the changes. Hence we have the least mean square (LMS) algorithm which is defined as in Eq. (11).

\[
f_{s+1} = f_s + \alpha e(n) r(n) \tag{11}
\]

\[
e(n) = x(n) - r(n) \tag{12}
\]

where \( f_s \) is the training filter coefficients for \( h'(n) \), \( \alpha \) is the step size, \( x(n) \) is the training data, \( r(n) \) is the received training data, and \( e(n) \) is defined as estimation error signal.

![Fig. 6 The block diagram of LMS channel estimator](image)

**B. Frequency Domain Equalizer (FEQ)**

The FEQ includes a simple one-tap equalizer and a LS estimator. The estimated channel response out of the LS estimator is then

\[
\hat{H}_k = P_k Z_k = \begin{bmatrix} Z_{k,1} & Z_{k,2} & \ldots & Z_{k,L} \end{bmatrix}
\]

where \( k \) is the index of OFDM symbol, \( l \) is the index of subcarrier, so that \( H_k \) is the whole channel response of the \( k \)-th symbol, \( Z_k \) is the received preamble, and \( P_k \) is the known preamble.

The equalizer eliminates the channel effect and LS estimator estimate the residual distortion on each subcarrier. For low SNR environment, the LS estimator provides a about 0.5dB BER gain. The final channel response used to compensate data is simply the multiplication of output of FFT and LS estimator.

**4. SIMULATION**

We have simulated the proposed scheme for WiMAX communication system specifications with channel bandwidth of 20MHz, 256 signal samples per OFDM symbol and CP length of 64 samples. It represents the highest speed with longest CP in WiMAX specs. This CP length is equivalent to 3.2\( \mu \)s. The SUI channel models are used as testing environment. Using SUI-4 channel model for simulation with DL equal to CP length, Fig. 7 shows that the SSNR improves step-wisely as the SIRF length
increases. The SUI-4 channel model has delay spread of 4μs which is longer than the CP length of the OFDM symbol.

When designing the SIRF filter, one thing we need to determine is the length of the filter. Intuitively, it is obvious that longer the SIRF length the better the system performance at the expense of implementation cost. The typical profiles of the effective CIR with SIRF filter length equal to 100, 150, 200, and 300 are shown in Fig. 8. It can be seen that the longer SIRF length, the more likely that signal energy concentrates inside of desired length window. However, we have observed that, given a fixed DL to the SIRF filter, the BER performance is not linear with the SIRF length. The BER literally goes down as the SIRF length increase, but also fluctuate locally. In Fig.8, the locally best performance is the second plot of the figure in which the SIRF length is 150, not the figure with SIRF length of 300.

We have observed that the SIRF length is not the only variable affecting the BER. We found from the shortening profile that the effective channel is shorten to the length of less of what we designated (i.e. DL = CP length). By choosing different DL other than CP, it actually could converge to better BER performance. As forcing DL =0, the SIRF would become a zero-forcing filter. Intuitively, we might think that the smaller the DL, the better the system BER performance. However, this is not true. Fig. 9 shows that the performance evaluation with different DL of 12, 22, 32, 42, and 80 while the target CP is still 64. The result shows that the case with D = 22 has the best BER performance instead of DL = 12. It indicates that the BER performance is not inversely proportional with DL. It is not the case that the shorter the DL, the better the system BER performance. There exists a saddle point of DL that gives the best result.
We have simulated the proposed scheme over various SUI channels as shown in Fig. 12. The most difficult channel condition is the SUI-4 model, which has very high BER if only FEQ is applied. This is because of the un-recoverable ISI in the channel. One can see that the performance of SUI-4 channel is improved effectively and satisfies the BER requirement listed in the standard.

5. CONCLUSION

In this paper, a channel estimation method for non-line-of-sight environment with high data rate transmission is presented. In many cases, it is usually to level down the transmission rate to combat the ominous environment[6]. The BER performance under the high rate transmission situation shows that the proposed method is capable and reliable to overcome the ISI problem without dropping the transmission rate. We have also found that the better strategy to design a shortening filter is to over-constraint the desired length to much shorter than CP.

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