Channel Estimation Scheme for WLAN Systems with Backward Compatibility

Jee-Hoon Kim, Heejung Yu, and Sok-Kyu Lee

IEEE 802.11n standards introduced a mixed-mode format frame structure to achieve higher throughput with multiple antennas while providing backward compatibility with legacy systems. Although multi-input multi-output channel estimation was possible only with high-throughput long training fields (HT-LTFs), the proposed scheme utilizes a legacy LTF as well as HT-LTFs in a decision feedback manner to improve the accuracy of the estimates. It was verified through theoretical analysis and simulations that the proposed scheme effectively enhances the mean square error performance.

Keywords: WLAN, channel estimation, LTF, decision feedback, backward compatibility.

I. Introduction

High-throughput wireless local area network (HT-WLAN) systems that follow IEEE 802.11n standards [1] provide backward compatibility with legacy orthogonal frequency division multiplexing (OFDM) systems such as IEEE 802.11a/g. This compatibility of IEEE 802.11n with higher throughput enables a successful market expansion; however, legacy parts at the beginning of an HT-WLAN mixed-mode format (HT-MF) frame, such as legacy training and signaling fields (SIGs) for protection from legacy devices, seem only to burden HT devices.

In the WLAN preamble, the long training field (LTF) is used for fine carrier frequency offset synchronization, fine time synchronization, and channel estimation. The HT-MF adopts a legacy LTF (L-LTF) to comply with the legacy structure, which is not orthogonally designed between transmit streams. On the contrary, high-throughput LTFs (HT-LTFs) in HT-MF hold orthogonality for multi-input multi-output (MIMO) channel estimation although the total power of training fields for estimating each channel is generally reduced. Consequently, the channel estimation performance is decreased compared with legacy systems. In this letter, we propose utilizing the L-LTF with decision feedback [2] for an improvement of channel estimation accuracy in the HT-MF.

II. Proposed Channel Estimation

In the HT-MF transmission, cyclic shifts are considered for multiple transmit chains to prevent unintentional beamforming. In other words, different phase rotations for a certain subcarrier are adopted between transmit antennas to avoid undesirable beamforming. The values for the cyclic shifts that are applied in legacy and HT portions, which include short training fields (STFs), LTFs, and SIGs, are different since large delay spreads degrade a legacy receiver. For example, when the number of space-time streams, N_{STS}, is 2, the cyclic shift values for the legacy fields and HT fields are –200 ns and –400 ns, respectively, which are shown in Fig. 1. More specific information about HT-WLAN can be found in [3]. It is noted that the cyclic shift for L-LTF is different from those of HT-LTFs, and L-LTF is used for single-stream channel estimation to equalize the SIGs (L-SIG, HT-SIG1, HT-SIG2) modulated with the legacy format. On the other hand, HT-LTFs with orthogonal property are inserted for MIMO channel estimation to detect multi-stream data. This design makes the L-LTF hard to utilize for channel estimation in the HT-WLAN. Therefore, such approaches as in [4], [5] are not effective.

For a simple signal representation, let $e^{j\theta n}$ be the phase

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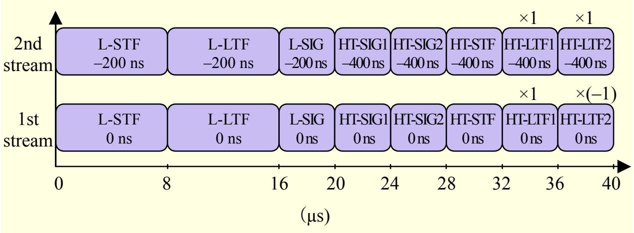

Fig. 1. Structure of HT-MF preamble when N_{STS} equals 2.

rotation of the n-th subcarrier in the L-LTF, and the phase rotation of the n-th subcarrier in the HT-LTFs is then given by $e^{j2\theta n}$ when N_{STS} equals 2. Based on the preamble structure for HT-MF with 2 streams, which is shown in Fig. 1, the frequency domain representations of the received L-LTF and HT-LTFs at the i-th receive antenna are as follows:

$$Y_0(n) = H_{i1}(n)P_{\text{L}}(n) + H_{i2}e^{j\theta n}P_{\text{L}}(n) + N_0(n), \qquad n \in S_{\text{L}}, \quad (1)$$

$$Y_1(n) = H_{i1}(n)P_{\text{L}}(n) + H_{i2}e^{j\theta n}P_{\text{L}}(n) + N_1(n), \qquad n \in S_{\text{L}}, \quad (2)$$

$$Y_2(n) = H_{i1}(n)P_{\text{HT}}(n) + H_{i2}e^{j2\theta n}P_{\text{HT}}(n) + N_2(n), \quad n \in S_{\text{HT}}, \quad (3)$$

$$Y_3(n) = -H_{i1}(n)P_{\text{HT}}(n) + H_{i2}e^{j2\theta n}P_{\text{HT}}(n) + N_3(n), n \in S_{\text{HT}}, \quad (4)$$

where $H_{ij}(n)$ denotes a frequency response of a channel at the n-th subcarrier from the j-th transmit antenna to i-th receive antenna, $P_{\text{L}}(n)$ and $P_{\text{HT}}(n)$ are n-th subcarrier symbols of the L-LTF and HT-LTFs, respectively, and $N_m(n)$ for $m \in \{0, 1, 2, 3\}$ represents complex Gaussian random noise with mean zero and variance $2\sigma^2$. Additionally, S_{L} and S_{HT} are sets of subcarrier indices with non-zero values in the frequency domains, L-LTF and HT-LTF, respectively. It is noted that two received symbols, $Y_0(n)$ and $Y_1(n)$, are considered for L-LTF since L-LTF is double the symbol length of the HT-LTF. On the other hand, there are two transmission times for the HT-LTFs (HT-LTF1 and HT-LTF2). Consequently, $Y_2(n)$ and $Y_3(n)$ are presented as received signals for them. Due to the orthogonal composition of the HT-LTFs in $Y_2(n)$ and $Y_3(n)$, the MIMO channel can be estimated as follows:

$$\hat{H}_{i1}(n) = \frac{Y_2(n) - Y_3(n)}{2P_{\text{HT}}(n)}, \quad n \in S_{\text{HT}}, \quad (5)$$

$$\hat{H}_{i2}(n) = \frac{Y_2(n) + Y_3(n)}{2P_{\text{HT}}(n)e^{j2\theta n}}, \quad n \in S_{\text{HT}}. \quad (6)$$

This is an ordinary method to estimate MIMO channels. On the contrary, the proposed scheme for $n \in S_{\text{L}} \cup S_{\text{HT}}$ additionally utilizes L-LTF in $Y_0(n)$ and $Y_1(n)$, where the general procedure of the proposed scheme is as follows. When $N_{\text{STS}}=1$, $H_{i1}(n)$ is estimated through the weighted sum of $Z(n)=(Y_0(n)+Y_1(n))/2$ and $Y_2(n)$, where the terms related to the second stream do not exist in the equations. If N_{STS} is larger than 1, the procedure is divided into two phases. In the first phase, $H_{ij}(n)$ for $j \in J = \{1, \ldots, N_{\text{STS}}\}$ is estimated using $Y_{j+1}(n)$, which is the same as an ordinary method. In the second phase, the estimated values of $H_{ik}(n)$, where $k \in J \backslash \{j\}$, are fed back to $Z(n)$ to acquire new estimates of $H_{ij}(n)$ and are updated through the weighted sum of the initial estimates in the first phase and their own values. It was noted that the combining weights are in inverse proportion to the standard deviation of noise of each estimate for optimal performance. The second phase may be repeated for maximum performance if any non-updated values of $H_{ik}(n)$, where $k \in J \backslash \{j\}$, are used to estimate $H_{ij}(n)$. On the other hand, for subcarrier indices where the L-LTF is not defined, that is, $n \in S_{\text{HT}} \backslash S_{\text{L}}$, the ordinary method should be used. The procedure of the proposed scheme when N_{STS} equals 2 is illustrated in Fig. 2.

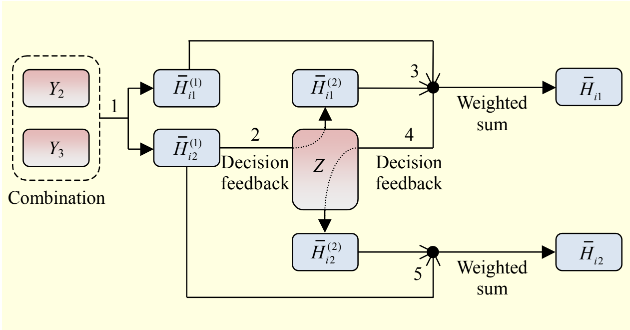

Fig. 2. Procedure of proposed scheme when N_{STS} is 2.

III. Performance Evaluation and Discussion

In this section, the proposed channel estimation scheme is confirmed when N_{STS} is 2 in terms of the mean square error (MSE), which is the most common measure used to evaluate the quality of the estimator. To simplify the notation, we omitted the subcarrier index because the analysis does not depend on the subcarrier. Additionally, we assume the HT-WLAN system is operating in the 20 MHz bandwidth. Since cyclic shifts do not affect the noise variance in MSE analysis, the averaged L-LTF, that is, Z and the first estimates of H_{ij}, that is, (5) and (6), can be rewritten as

$$Z' = \sqrt{\frac{56}{52}}(H_{i1} + H_{i2}) + \frac{N_0 + N_1}{2}, \quad (7)$$

$$\bar{H}_{i1}^{(1)} = H_{i1} + \frac{1}{\sqrt{2}}N_{2,3}, \quad (8)$$

$$\bar{H}_{i2}^{(1)} = H_{i2} + \frac{1}{\sqrt{2}}N'_{2,3}, \quad (9)$$

where $N_{2,3} = 1/\sqrt{2}(N_2 - N_3)$ and $N'_{2,3} = 1/\sqrt{2}(N_2 + N_3)$, and their variances are still $2\sigma^2$. Since the transmit power of each effective subcarrier for the L-LTF is larger than that of HT-LTF as a result of power normalization in transmission, a power scaling factor of $\sqrt{56/52}$ is included. From (7)-(9), the second estimate of H_{i1} is given by

$$\bar{H}_{i1}^{(2)} = \sqrt{\frac{52}{56}}Z' - \bar{H}_{i2}^{(1)} = H_{i1} + \frac{\sqrt{13}(N_0+N_1) - \sqrt{28}N'_{2,3}}{\sqrt{56}}. \quad (10)$$

It was noted that MSE equals the inverse of the signal-to-noise ratio (SNR), that is, SNR^{-1} for a Gaussian distribution. Since the noise components in (10) are independent of each other, the MSE of $\bar{H}_{i1}^{(2)}$ is calculated as

$$\mathrm{MSE}(\bar{H}_{i1}^{(2)}) = 2\sigma^2\left(\frac{2\left(\sqrt{13}\right)^2 + \left(\sqrt{28}\right)^2}{\left(\sqrt{56}\right)^2}\right) = \frac{27}{14}\sigma^2. \quad (11)$$

This means the combining weights for $\bar{H}_{i1}^{(1)}$ and $\bar{H}_{i1}^{(2)}$ are determined as $\sqrt{27/14}$ and 1, respectively. In the next step, $\bar{H}_{i1}$ is obtained as follows:

$$\begin{aligned}\bar{H}_{i1} &= \left(\sqrt{\tfrac{27}{14}}\bar{H}_{i1}^{(1)} + \bar{H}_{i1}^{(2)}\right)\Big/\left(1+\sqrt{\tfrac{27}{14}}\right)\\ &= H_{i1} + \frac{\sqrt{13}(N_0+N_1) + \sqrt{54}N_{2,3} - \sqrt{28}N'_{2,3}}{\sqrt{56}+\sqrt{108}}. \quad (12)\end{aligned}$$

Similarly,

$$\begin{aligned}\mathrm{MSE}(\bar{H}_{i1}) &= 2\sigma^2\left(\frac{2\left(\sqrt{13}\right)^2 + \left(\sqrt{54}\right)^2 + \left(\sqrt{28}\right)^2}{\left(\sqrt{56}+\sqrt{108}\right)^2}\right)\\ &\approx 0.676\sigma^2. \quad (13)\end{aligned}$$

For a simple representation, we round off the fractions to four decimal places. On the other hand, $\bar{H}_{i2}^{(2)}$ is estimated as follows:

$$\begin{aligned}\bar{H}_{i2}^{(2)} &= \sqrt{\frac{52}{56}}Z' - \bar{H}_{i1}\\ &= H_{i2} + \left(\sqrt{\tfrac{13}{56}\cdot\tfrac{27}{14}}(N_0+N_1) - \sqrt{\tfrac{27}{28}}N_{2,3} + \sqrt{\tfrac{1}{2}}N'_{2,3}\right)\Big/\left(1+\sqrt{\tfrac{27}{14}}\right)\\ &\approx H_{i2} + 0.2801(N_0+N_1) - 0.4111N_{2,3} + 0.296N'_{2,3}. \quad (14)\end{aligned}$$

The combining weights for $\bar{H}_{i2}^{(1)}$ and $\bar{H}_{i2}^{(2)}$ are $\sqrt{0.8271}$ and 1, respectively, since

$$\begin{aligned}\mathrm{MSE}(\bar{H}_{i2}^{(2)}) &= 2\sigma^2\left(2(0.2801)^2 + (0.4111)^2 + (0.296)^2\right)\\ &\approx 0.8271\sigma^2. \quad (15)\end{aligned}$$

Consequently, $\bar{H}_{i2}$ is then

$$\begin{aligned}\bar{H}_{i2} &= \left(\sqrt{0.8271}\bar{H}_{i2}^{(1)} + \bar{H}_{i2}^{(2)}\right)\Big/\left(1+\sqrt{0.8271}\right)\\ &\approx H_{i2} + 0.1467(N_0+N_1) - 0.2153N_{2,3} + 0.4918N'_{2,3}. \quad (16)\end{aligned}$$

Finally, the MSE of $\bar{H}_{i2}$ is given by

$$\begin{aligned}\mathrm{MSE}(\bar{H}_{i2}) &= 2\sigma^2\left(2(0.1467)^2 + (0.2153)^2 + (0.4918)^2\right)\\ &\approx 0.6625\sigma^2. \quad (17)\end{aligned}$$

Table 1. Comparison of analyzed average MSEs for 20 MHz bandwidth usage.

N_{STS}	Ordinary legacy	Ordinary HT-MF	Proposed HT-MF ($n \in S_L$)	Proposed HT-MF ($n \in S_L \cup S_{HT}$)
1	$0.4643\sigma^2$	σ^2	$0.317\sigma^2$	$0.3658\sigma^2$
2	–	σ^2	$0.676\sigma^2$ for H_{i1} $0.6625\sigma^2$ for H_{i2}	$0.6991\sigma^2$ for H_{i1} $0.6866\sigma^2$ for H_{i2}

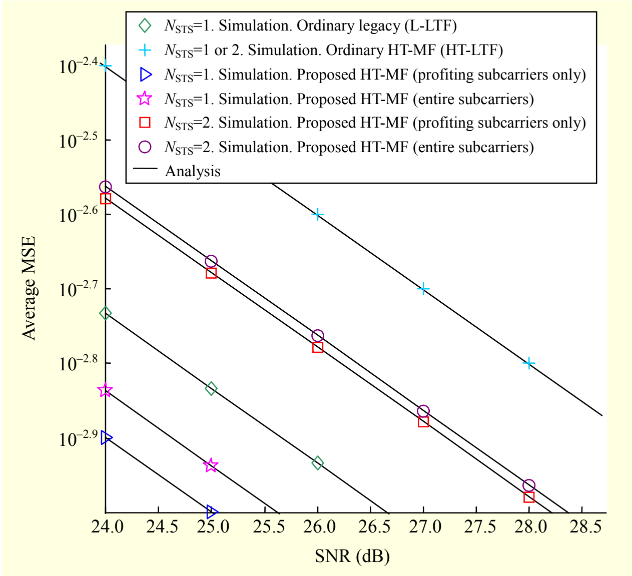

Fig. 3. Average MSE performance versus SNR comparison.

As mentioned in the previous section, if this process is repeated, the MSEs of two estimates will be decreased and equal, but the gain will be too small relative to the effort. Some meaningful MSEs are analyzed and given in Table 1.

Figure 3 shows the results of the MSE analysis and simulations of six cases with IEEE 802.11n operating in the 20 MHz bandwidth.[1)] It is remarked that the results of the MSE simulation coincided with those of the MSE analysis for all considered cases. In Fig. 3, the first and second cases are the results of the ordinary method for the legacy and HT-MF preambles, respectively. The others are for the proposed scheme according to N_{STS} and by considering the subcarriers. It was noted that the estimate of H_2 with one decision feedback is considered when N_{STS} equals 2 for the proposed scheme. The legacy uses a double-length LTF compared with the HT-MF, and the power is divided into 52 subcarriers instead of 56. This is why there is about a 3.4 dB performance gap in SNR between them. On the other hand, when only profiting subcarriers, that is, $n \in S_L$, are considered, the gain of the

1) Since the MSE gaps are always the same regardless of SNR for all cases, an arbitrary point is magnified to display the differences clearly.

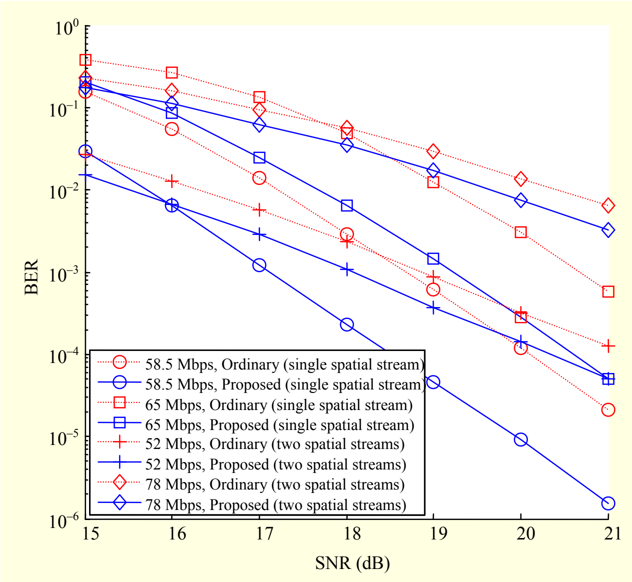

Fig. 4. BER performance versus SNR comparison.

proposed scheme is about 5 dB for one stream and about 1.8 dB for two streams. In addition, when all subcarriers, that is, $n \in S_L \cup S_{HT}$, are regarded, the gain is reduced to about 4.4 dB for one stream and about 1.65 dB for two streams. The advantage obtained by the proposed scheme is diminished as N_{STS} increases. This can be explained by the fact that in the proposed scheme, imperfectly estimated channels are fed back to Z for new estimates, which results in noise. This generated noise is then enlarged as the number of individual channels increases.

Simulation results of the bit error rate (BER) performance are presented in Fig. 4. We considered several cases for the HT-WLANs with single and two spatial streams operating in the 20 MHz bandwidth in IEEE 802.11n channel model B [3]. It is confirmed that the proposed scheme generally outperforms a non-L-LTF-aided method about 1.7 dB in SNR in the case of HT-WLANs utilizing single spatial stream and 1 dB in SNR in the other case.

Theoretically, increased signal processing for the proposed scheme could prevent a station from instantly detecting a signal. However, this potential problem can be addressed with the immediate feedback of estimates: If signal detection is required before finishing the channel estimation process, the last estimate for the moment is utilized, and the updated estimate is used as soon as it is available. Even though the first few data blocks do not profit from the proposed scheme, the overall performance is effectively enhanced.

IV. Conclusion

In this letter, we proposed an efficient channel estimation scheme for an HT-MF, where an L-LTF is employed for backward compatibility. A weak point of the proposed scheme is that a few subcarriers do not benefit from the use of L-LTF. This can be overcome through an interleaver and appropriate weights according to the variance of noise for the soft demapper.

On the other hand, it was noted that the proposed scheme can also be applied to the other WLAN standards having a similar structure, such as IEEE 802.11TGac and IEEE 802.11TGaf. Moreover, the performance of MIMO detection [6] and beamforming schemes [7], where channel estimation accuracy is important, can be improved through the proposed scheme.

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