An Enhanced Link Adaptation for the MB-OFDM UWB System

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Abstract—In the paper, an improved link adaptation scheme is proposed for the WiMedia MB-OFDM UWB system, in which quality of service (QoS) support is provided. The proposed scheme consists of three functional blocks: link quality indicator (LQI) calculator, frame error rate (FER) estimator, and transmitter (TX) parameter selector. Instead of using the average receive SNR (ASNR) as LQI, a new LQI metric is defined based on a union bound analysis to capture the effects of both path loss and frequency selectivity of an instantaneous UWB channel. How to calculate LQI for each rate mode is investigated by analyzing the distribution of soft bit information. With the calculated LQI, the FER performance of each rate mode can be accurately estimated with a look-up table method, which is suitable for practical implementation. Using the estimated FERs, TX parameter selector can optimize TX rate mode to improve the communication throughput under QoS constraints. It is shown the proposed scheme can significantly improve the throughput while maintaining the required QoS compared to conventional ASNR based link adaptation schemes.

I. INTRODUCTION

Attracted by the huge potential market of ultra-wideband (UWB), leading companies including Intel, TI, and Philips, formed an industry consortium, the WiMedia Alliance, working on UWB standardization and successfully convinced ECMA International to accept the WiMedia multi-band orthogonal frequency division multiplexing (MB-OFDM) UWB PHY/MAC specification as an ISO-based international standard, namely ECMA-368 [1]. One promising application of the WiMedia UWB is video streaming, in which UWB is used to distribute real-time video streams in a home environment. Since video quality is sensitive to data transfer latency and frame error rate (FER), quality of service (QoS) support is critical. However, UWB link conditions can be significantly changed from time to time in such application, since users may move around in the vicinity of UWB devices.

In the literature, there are numerous link adaptation schemes available for OFDM based wireless systems. In [2], [3], the authors directly extended the average signal-to-noise ratio (ASNR) based scheme, which was original proposed for single carrier systems, to OFDM systems and showed considerable throughput improvement. However, as [6] pointed out, the ASNR based schemes may be not a good choice for OFDM systems since ASNR can only account for path loss and shadowing fading effects while the frequency selectivity of multi-path channels is not considered. It is well known that the channel frequency selectivity can significantly affect the FER performance. In [4], the authors introduced some extra parameters, such as sub-carriers’ SNR variance, to catch the frequency selectivity effect. However, a big (usually two-dimension) look-up table indexed by ASNR and those extra parameters is needed to estimate the FER performance. In [5], the authors proposed to calculate the FER estimation based on the instantaneous SNRs of all data sub-carriers directly and use the estimated FERs to do link adaptation. However, it needs to calculate a union bound of FER and is too complicated for practical implementation. All the above schemes are for general OFDM systems. There are very few link adaptation results specifically for the WiMedia UWB. [7] and [8] addressed this topic and got some interesting results. However, they are ASNR based schemes and more importantly, they were optimized only for throughput and no QoS support was considered. In this paper, we propose an improved link adaptation scheme for the WiMedia UWB that can achieve higher throughput than [7] and [8] while maintaining the required QoS.

The remainder of paper is organized as follows. In Section II, the system model is provided. In Section III, the proposed link adaptation scheme is discussed in detail. In Section V, some simulation results are presented to demonstrate the good performance of the proposed scheme. At last, some conclusion remarks are given in Section VI.

II. SYSTEM MODEL

A system model based on the WiMedia UWB PHY is illustrated in Fig. 1. The WiMedia PHY is a bit interleaved coded modulation (BICM) MB-OFDM system. The $3.1 \sim 10.6$ GHz UWB spectrum is divided into 14 frequency bands, each of which has a $528$ MHz bandwidth. These bands are grouped into several band groups and each of them has either 2 or 3 frequency bands. Data transmission is always conducted...
within a band group. A UWB device may transmit data always on one frequency band (referred to as fixed frequency interleaving, FFI) or on different bands of the same group at different OFDM symbol intervals (referred to as time-frequency interleaving, TFI). A time-frequency code (TFC) is used to determine the band hopping pattern.

The information bits are first scrambled and then protected by a punctured convolutional code. The encoded bits are partitioned into 6-OFDM-symbol blocks and then interleaved on a block base. The interleaved bits are mapped to symbols and transmitted on data sub-carriers. Each OFDM symbol has 128 sub-carriers and 100 of them are used to transmit data. At the receiver side, the received baseband signal first is transformed to frequency domain via FFT operations. Then the soft bit information, i.e. the log-likelihood ratio (LLR), is obtained. The function of TX parameter selector is to select the transmitter (transmitter) parameter selector, respectively.

### III. THE PROPOSED LINK ADAPTATION SCHEME

The proposed link adaptor consists of three functional blocks, which are LQI calculator, FER estimator, and TX parameter selector, respectively.

#### A. TX Parameter Selector

The function of TX parameter selector is to select the optimal TX rate mode based on the estimated FERs of all rate modes, where “optimal” is in the sense of maximizing the achieved throughput while maintaining the required QoS. In this work, a simplified throughput definition is used, which is given as below

\[ T_r = R \times (1 - FER_t) \]  (2)

The extension to the real MAC throughput, which takes all the PHY/MAC overhead into consideration, is straightforward and not discussed here.

#### B. FER estimator

FER estimator is responsible for accurately estimating the achieved FER for each rate mode. With the input LLR, \( \lambda_i \), the pairwise error event \( c \rightarrow \hat{c} \) in soft Viterbi decoding happens when

\[ \sum_{i \in I_{(c, \hat{c})}} (1 - 2c_i)\lambda_i < 0, \]  (3)

where \( I_{(c, \hat{c})} \) denotes the set of bit index \( i \) with \( c_i \neq \hat{c}_i \). \( (1 - 2c_i)\lambda_i \) is a random variable and its distribution depends on the used modulation/spreading scheme and the bit interleaver. If the distribution of \( (1 - 2c_i)\lambda_i \) is known, we can calculate the exact PEP. In the following subsection, it is shown that \( (1 - 2c_i)\lambda_i \) can be treated as a Gaussian distributed random

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**TABLE I**

<table>
<thead>
<tr>
<th>Data Rates (Mbps)</th>
<th>Modulation</th>
<th>Coding Rate</th>
<th>FDS</th>
<th>TDS</th>
</tr>
</thead>
<tbody>
<tr>
<td>53.3</td>
<td>QPSK</td>
<td>1/3</td>
<td>YES</td>
<td>YES</td>
</tr>
<tr>
<td>80</td>
<td>QPSK</td>
<td>1/2</td>
<td>YES</td>
<td>YES</td>
</tr>
<tr>
<td>106.7</td>
<td>QPSK</td>
<td>1/3</td>
<td>NO</td>
<td>YES</td>
</tr>
<tr>
<td>160</td>
<td>QPSK</td>
<td>1/2</td>
<td>NO</td>
<td>NO</td>
</tr>
<tr>
<td>200</td>
<td>QPSK</td>
<td>5/8</td>
<td>NO</td>
<td>YES</td>
</tr>
<tr>
<td>320</td>
<td>DCM</td>
<td>1/2</td>
<td>NO</td>
<td>NO</td>
</tr>
<tr>
<td>400</td>
<td>DCM</td>
<td>5/8</td>
<td>NO</td>
<td>NO</td>
</tr>
<tr>
<td>480</td>
<td>DCM</td>
<td>3/4</td>
<td>NO</td>
<td>NO</td>
</tr>
</tbody>
</table>

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**Fig. 1. System Model**
variable with mean of $M_i$ and variance of $2M_i$. The value of $M_i$ depends on where the $i$-th coded bit is transmitted, which can be described with $(n, k, g)$. $(n, k, g)$ denotes the $n$-th bit position of the binary label of the symbol (or symbol-pair when DCM is used) transmitted on the $k$-th sub-carrier channel of the $g$-th frequency band. The mapping between $i$ and $(n, k, g)$ is determined by the bit interleaving operation. In order to simplify the analysis, the bit interleaving operation is assumed to be perfect, which means bit $i$ can be mapped to any $(n, k, g)$ with equal probability and the mappings of different bits are independent. This assumption makes $M_i$ be an i.i.d. random variable with equal probability over all its possible values. Therefore,

$$P(e → \hat{c}) = E_{M_i}\{P(e → \hat{c}|M_i)\} = E_{M_i}\{Q\left(\frac{\sum_{i∈\{0,1\}} M_i}{2}\right)\} ≤ \frac{1}{2} E_{M_i}\{e^{-\frac{M_i}{2}}\}^d$$

(4)

where $(a)$ comes from $Q(x) ≤ \frac{1}{2} e^{-x^2}$ and the ideal bit interleaving assumption. $d$ denotes the hamming distance between $c$ and $\hat{c}$. Then, an upper bound of FER can be obtained based on the BICM union bound, as shown below

$$FER ≤ FER_u = \sum_{d=d_{free}}^{N_p} \frac{\alpha_d}{2} (E_{M_i}\{e^{-\frac{M_i}{2}}\})^d,$$

(5)

where $d_{free}$, $\alpha_d$, and $N_p$ denote the free distance of the used convolutional code, the number of codewords with weight of $d$, and the data frame length, respectively. The upper bound $FER_u$ can be used to estimate FER and only depends on $E_{M_i}\{e^{-\frac{M_i}{2}}\}$. We can generalize $E_{M_i}\{e^{-\frac{M_i}{2}}\}$ and define

$$L_M = -\beta\ln\{E_{M_i}\{e^{-\frac{M_i}{2}}\}\},$$

(6)

where $\beta$ is a design parameter that can be optimized for each rate mode to improve the precision of FER estimation. Instead of using (5) to estimate $FER$, we use the AWGN FER-versus-SNR performance as reference to estimate FER. Under AWGN channel, it can be proved that $M_i = \gamma t$, where $\gamma$ is the receive SNR and $t$ is a constant depending on the rate mode, and $L_M = t\gamma$. Therefore, we define the effective SNR as below and use it as LQI for link adaptation.

$$LQI = ESNR = \frac{L_M}{t} = \frac{-\beta\ln\{E_{M_i}\{e^{-\frac{M_i}{2}}\}\}}{t}.$$

(7)

In the proposed scheme, we only need to store $FER_{AWGN}(\gamma)$ curves as look-up tables and use $ESNR$ as index to get the estimated FER.

IV. LQI CALCULATOR

We divide all the rate modes into three groups to discuss the LQI calculation.

A. TDS, FDS, QPSK

This group includes 53.3Mbps and 80Mbps modes, in which TDS and FDS are used to exploit frequency diversity. 2 bits are mapped to a QPSK symbol $s_{k,l}, 0 ≤ k ≤ 49$, via gray mapping and then transmitted on the $k$-th and $(99 − k)$-th data sub-carriers of the $2t$-th and $(2t + 1)$-th OFDM symbols. $s_{k,l}$ is normalized to have unit symbol energy. Denote the used frequency hopping pattern as $\{g_0, g_1, g_2, g_3, g_4, g_5\}$, where $g_0, ..., g_5 ∈ \{1, 2, 3, \}$. $h_{k,g}$ denotes the channel parameter of the $k$-th data sub-carrier of band $g$. Let $f(l)$ denote the index of the band in which the $l$-th OFDM symbol is transmitted. Then after FFT, the received signal model can be described as below

$$y_{k,2l} = h_{k,f(2l)} s_{k,l} + w_{k,2l},$$

$$y_{g99−k,2l} = h_{g99−k,f(2l)} s_{g99−k,l} + w_{g99−k,2l},$$

(8)

where $w_{k,2l}$ denotes the additive white gaussian noise on the $k$-th data sub-carrier of the $l$-th OFDM symbol and they are i.i.d. and with zero mean and unit variance, $p_n$ is the $n$-th scramble symbol with value of 1 or −1. After the maximum ratio combining (MRC), we get

$$\bar{y}_{k,l} = \bar{h}_{k,l} s_{k,l} + \bar{w}_{k,l},$$

where $\bar{h}_{k,l}$ and $\bar{w}_{k,l}$ are an AWGN term with zero mean and unit variance. Since the QPSK is used, the LLR distribution of 2 bits are same and here we only consider the LLR of bit $c_l$. Based on (8), the LLR of $c_l$ can be calculated as $\lambda_l = \ln(P(\bar{y}_{k,l} | c_l = 0)) − \ln(P(\bar{y}_{k,l} | c_l = 1)) = −2\sqrt{2}\bar{h}_{k,l} Re(\bar{y}_{k,l})$ and $(1 − 2c_l)\lambda_l = 2\beta h_{k,l}^2 + w_{k,l}$ where

$$\bar{w}_{k,l} = −2\sqrt{2}(1 − 2c_l)h_{k,f(2l)} Re(\bar{w}_{k,l}).$$

This proves our previous claim that $(1 − 2c_l)\lambda_l$ is a Gaussian random variable with mean $\lambda_l = \beta h_{k,l}^2$ and variance of $2\lambda_l$. One can see $\lambda_l$ depends on data sub-carrier index and band indices $(f(2l), f(2l + 1))$. In order to calculate LQI, we need to calculate $M_i$ for all possible $k$ and $(f(2l), f(2l + 1))$. Under AWGN channel, $|h_{k,g}|^2$ is a constant and equal to the average SNR $\gamma$. Therefore, we have $M_i = \gamma t$, i.e. $t = 8$. Then, the LQI calculation (7) becomes

$$\text{LQI} = \frac{-\beta\ln\{E_{M_i}\{e^{-\frac{M_i}{2}}\}\}}{8}.$$ (9)

In non-hopping FFI modes, UWB transmission always stays on one band, i.e. $g_l = g$, and we only need to calculate 50 possible $M_i$ for $0 ≤ k ≤ 49$ with

$$M_k = 2(|h_{k,g}|^2 + |h_{g99−k,g}|^2 + |h_{k,g}|^2 + |h_{g99−k,g}|^2).$$

The LQI calculation (9) can be done as

$$\text{LQI} = \frac{-\beta\ln\{\frac{1}{M_i} \sum_{i=0}^{49} \{e^{-\frac{M_i}{2}}\}\}}{8} = \frac{\beta(-\ln(\frac{1}{M_i}) + \sum_{i=0}^{49} \frac{M_i}{8})}{8}.$$
where $\Theta$ denotes the log-sum operation defined as $a \Theta b = -\ln(e^{-a} + e^{-b})$. The log-sum operation can be simplified as below to ease its implementation:

$$a \Theta b \approx \min(a, b) - \max\left(\frac{5}{8} - \frac{|a - b|}{4}, 0\right).$$

While in band-hopping TFI modes, we need to calculate 300 possible $M_i$ for $0 \leq k \leq 49$ and $(f(2l), f(2l+1)) \in \{(g_0, g_1), (g_2, g_3), (g_4, g_5)\}$ with

$$M_k = 2(|h_{k,g_0}|^2 + |h_{99-k,g_1}|^2 + |h_{k,g_1}|^2 + |h_{99-k,g_2}|^2),$$
$$M_{(k+50)} = 2(|h_{k,g_2}|^2 + |h_{99-k,g_3}|^2 + |h_{k,g_3}|^2 + |h_{99-k,g_4}|^2),$$
$$M_{(k+100)} = 2(|h_{k,g_4}|^2 + |h_{99-k,g_5}|^2 + |h_{k,g_5}|^2 + |h_{99-k,g_6}|^2),$$

$$LQI = \frac{\beta(-\ln(\frac{1}{150}) + 149 M_i)}{8}.$$

**B. TDS, QPSK**

This group includes 106.7Mbps, 160Mbps and 200Mbps data rate modes, in which only TDS is used with QPSK modulation. With the maximum ratio combining (MRC), we get

$$\bar{y}_{k,l} = \hat{h}_{k,l} s_{k,l} + \tilde{w}_{k,l}$$

(10)

where $\hat{h}_{k,l} \triangleq \sqrt{|h_{k,f(l)}|^2 + |h_{99-k,f(l)+1}|^2}$ and $\tilde{w}_{k,l}$ is an AWGN term with zero mean and unit variance. Again, the LLR of 2 bits are same due to QPSK modulation. Following the same line as the previous section, we get

$$\lambda_{ci} = \ln(P(\bar{Y} = 0 | c_i = 0)) - \ln(P(\bar{Y} = 1 | c_i = 1))$$
$$= \ln(\sum_{S \in S_{ci}} e^{-\|\bar{Y} - HRS\|}) - \ln(\sum_{S \in S_{ci}} e^{-\|\bar{Y} - HRS\|^2})$$
$$\approx \ln(\max_{S \in S_{ci}} e^{-\|\bar{Y} - HRS\|^2}) - \ln(\max_{S \in S_{ci}} e^{-\|\bar{Y} - HRS\|^2}),$$

(15)

where $\bar{w}_{k,l}$ and $\tilde{w}_{k,l}$ are i.i.d. real gaussian noise terms with zero mean and unit variance, $H = \begin{bmatrix} |h_{k,f(l)}| & 0 \\ 0 & |h_{k+50,f(l)}| \end{bmatrix}$.

$S = \begin{bmatrix} Re\{s_{k,l}\} \\ Im\{s_{k,l}\} \end{bmatrix}$. From (15), one can see we need to demodulate $c_1$ (mapped to $Re\{s_{k,l}\}$) and $c_2$ (mapped to $Re\{s_{k+50,l}\}$) jointly. The LLR calculation of $c_i$, $i = 1, 2$, can be expressed as

(16)

### C. DCM

In high data rate modes, including 320Mbps, 400Mbps, and 480Mbps, DCM is adopted to jointly modulate two symbols, which are transmitted on the $k$-th data sub-carrier and $(k+50)$-th data sub-carrier of an OFDM symbol. Since DCM is a real linear unitary transform, we can separately demodulate I and Q components. The two bits corresponding to I have the same LLR distribution as the two bits corresponding to Q and here we only consider I components. After equalization, we get

$$\bar{Y} = \left[ \begin{array}{c} \bar{y}_{k,l} \\ \bar{y}_{k+50,l} \end{array} \right] = \left[ \begin{array}{c} \frac{Re\{s_{k,l}\}}{|h_{k,f(l)}|} \\ \frac{Re\{s_{k+50,l}\}}{|h_{k+50,f(l)}|} \end{array} \right]$$

(15)

where $\bar{w}_{k,l}$ and $\tilde{w}_{k+50,l}$ are i.i.d. real gaussian noise terms with zero mean and unit variance, $H = \begin{bmatrix} |h_{k,f(l)}| & 0 \\ 0 & |h_{k+50,f(l)}| \end{bmatrix}$.

$S = \begin{bmatrix} Re\{s_{k,l}\} \\ Im\{s_{k,l}\} \end{bmatrix}$. Let $c_1$ denote the bit corresponding to $Re\{s_{k,l}\}$ and $c_2$ denote the bit corresponding to $Re\{s_{k+50,l}\}$. From (15), one can see we need to demodulate $c_1$ (mapped to $Re\{s_{k,l}\}$) and $c_2$ (mapped to $Re\{s_{k+50,l}\}$) jointly. The LLR calculation of $c_i$, $i = 1, 2$, can be expressed as

$$\lambda_{ci} = \ln(P(\bar{Y} = 0 | c_i = 0)) - \ln(P(\bar{Y} = 1 | c_i = 1))$$
$$= \ln(\sum_{S \in S_{ci}} e^{-\|\bar{Y} - HRS\|}) - \ln(\sum_{S \in S_{ci}} e^{-\|\bar{Y} - HRS\|^2})$$
$$\approx \ln(\max_{S \in S_{ci}} e^{-\|\bar{Y} - HRS\|^2}) - \ln(\max_{S \in S_{ci}} e^{-\|\bar{Y} - HRS\|^2}),$$

where $S_{ci}$ denotes the subset of $S$ whose label has $c_i = b, b \in \{0, 1\}$. It is difficult to get the exact distribution of $\lambda_{ci}$. Instead, we study the conditional LLR distribution, $P(1 - 2c_i | \lambda_{ci} = S_{ci})$, which is the distribution under the condition $S_{ci}$ is transmitted.

As shown in [8], for non-hopping FFI modes, we need to calculate 200 possible $M_i$ for $0 \leq k \leq 49$ with $f(l) = g$. The LQI calculation (7) can be done as

$$LQI = \frac{\beta(-\ln(\frac{1}{200}) + 199 M_i)}{4}.$$
Further simplify the proposed scheme, we used the same $\beta$ for all rate modes in the same modulation/spreading group and the used $\beta$ values are given in Table II.

First we studied the accuracy of the effective SNR calculated by the proposed LQI calculator. For each channel realization, we take the the channel parameters corresponding to a specific FER as input to calculate the effective SNR and compare it to the actual SNR required to achieve the same FER under the AWGN channel. Here, we study several typical values of the target FER, including 0.01, 0.04, and 0.08. Due to the space limit, we only give the results of the 400Mbps mode with CM1 channels in Figure 2. Simulation results for all 8 rate modes are available online [9]. As one can see, the estimated effective SNR is in most cases within $\pm 0.3$dB of its actual value. For link adaptation, the achieved accuracy is reasonably good. More importantly, the performance is stable under different UWB channel sceneries (either line-of-sight or non-line-of-sight), which is important since usually no a prior knowledge of UWB channel statistic characteristics is available in practice.

![Graph 2](image1)

**Fig. 2. The effective SNR estimation performance: CM1, TFC=1, 400Mbps**

We then studied the achieved throughput under the QoS constraint as $FER_t = 0.08$. Fig. 3 compares the throughput performance achieved by the proposed scheme with the one in [8] (referred to as “Philips-MAC”), and the one in [7] (referred to as “Nokia”). The achieved throughput of the proposed scheme decreases and in some cases is worse than “Philips-MAC”. This is because it sacrifices the throughput to guarantee the required QoS. Fig. 4 shows the outage probability of each scheme, where “outage” means the achieved FER with the selected TX rate is larger than $FER_t$. The proposed scheme has zero outage probability when the average SNR is larger than 0dB, while for “Philips-MAC”, the outage probability can be as high as 90%, which is not acceptable for video streaming applications. “Nokia” has a good outage performance but with the worst throughput performance. More results are available online [9]. All the simulation results demonstrate the proposed scheme can significantly improve the achieved throughput while maintaining the required QoS under different channel sceneries.

![Graph 3](image2)

**Fig. 3. The achieved throughput under CM1/CM3, with $FER_t = 0.08$**

![Graph 4](image3)

**Fig. 4. The link outage probability under CM1/CM3, with $FER_t = 0.08$**

### VI. Conclusions

In this paper, we propose a link adaptation scheme for the WiMedia UWB. A “divide and conquer” methodology is used to simplify FER estimation, in which the distribution of soft bit information is studied to separate the effects of modulation/spreading and channel coding. A new LQI metric is defined and how to calculate LQI for each rate mode is given based on a Gaussian approximation of the LLR distribution. With the calculated LQI, the FER performance of each rate mode under an instantaneous channel can be accurately estimated and TX can adapt its rate accordingly. Compared to ASNR based schemes, the proposed scheme can significantly enhance the throughput performance while maintaining the required QoS.

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