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Adaptive Bit-Interleaved Coded OFDM With Reduced Feedback Information

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***Abstract*—If the channel is static and is perfectly known to both the transmitter and the receiver, the water-filling technique with adaptive modulation is known to be optimal (Gallager, 1968). However, for orthogonal frequency-division multiplexing (OFDM) systems, this requires intensive traffic overheads for reporting channel state information on all subcarriers to the transmitter. In this paper, we consider an adaptive modulation and coding scheme for bit-interleaved coded OFDM with reduced feedback information satisfying a specified quality of service level. We propose a rate adaptation scheme, which utilizes the estimated bit error rate for supportable transmission rates. In this scheme, a user equipment chooses a modulation and coding scheme (MCS) level, which can provide the maximum spectral efficiency based on one OFDM symbol rather than on all subchannels. Then the user needs to send back only the selected MCS level index. The proposed scheme does not require the water-filling procedure, and the amount of the feedback information reduces to a single integer value irrespective of the number of subcarriers. Simulation results show that the proposed scheme can significantly reduce the system complexity while minimizing the performance loss compared to the optimum water-filling scheme.**

***Index Terms*—Adaptive modulation and coding (AMC), bit-interleaved coded modulation (BICM), fading channels.**

I. Introduction

FOR HIGH-SPEED wireless packet communication systems, we need to combat the interference caused by frequency selective channels. By using orthogonal frequency-division multiplexing (OFDM), wideband transmission is possible over frequency selective fading channels without applying equalizers. In addition, bit-interleaved coded modulation (BICM) [2] offers good diversity gains with higher order modulation schemes using binary convolutional codes. Bit-interleaved coded OFDM (BIC-OFDM), which combines the OFDM system with the BICM, has been applied to a wide range of wireless standards such as the IEEE 802.11a wireless local area network (WLAN) [3].

In typical indoor wireless environments, the channel is assumed to be quasi-static. In such environments, adaptive modulation and coding (AMC) [4] is a powerful technique to enhance the link performance for packet transmission systems. In this paper, we consider the AMC scheme for downlink packet transmissions with a quality of service (QoS) constraint. The adaptive transmission with the QoS constraint requires accurate channel state information (CSI) estimated at the receiver and a reliable feedback path from the receiver to the transmitter. The basic idea behind the adaptive transmission is to adjust the transmission power level, channel coding rate, and/or constellation size to the current channel state. Thus, AMC can enhance the average throughput by transmitting at high data rates under favorable channel conditions and reducing the transmission rate as the channel state degrades. For single carrier systems, a variable-rate variable-power M-QAM modulation scheme with a bit error rate (BER) constraint was proposed in [4]. In this scheme, the signal-to-noise ratio (SNR) range is divided into several regions where different modulation levels are applied to each transmission.

In contrast, for OFDM modulation, the BER performance not only depends on the received SNR but also on the frequency selectivity of the channel [5]. For example, for the same received SNR, the BER decreases as a root mean square (RMS) delay spreads increase. Therefore, a usage of a fixed link table defined only by SNR ranges for supportable spectral efficiencies as in [4] does not work well for OFDM systems. In this case, the water-filling with adaptive modulations is known to be optimal [1].

When the coded modulation technique is considered for OFDM systems, an adaptive BIC-OFDM with full feedback information (FI) proposed in [6] yields the capacity-approaching throughput by employing bit loading and power adaptations based on the water-filling algorithm. We will refer to this method as the water-filling AMC schemes (WF-AMC). However, the WF-AMC requires intensive traffic overheads for the feedback channel to report information on all subcarriers. In practical wireless communication systems, the bandwidth allocated for the feedback channel is normally small to preserve the resource. Consequently, the whole FI amount for all subcarriers may be too big to be handled using the limited feedback bandwidth as the number of subcarriers grows.

In this paper, we propose an AMC scheme for the BIC-OFDM system with reduced FI. The proposed scheme significantly

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reduces the amount of feedback payloads by employing the rate adaptation based on an individual OFDM symbol, rather than on each subcarrier. In frequency selective channels, the received signal at each subcarrier normally exhibits different received SNRs due to the frequency selectivity of the channel. The BIC-OFDM mitigates the fluctuation by employing the AMC over the OFDM symbol. With the proposed scheme, a user equipment (UE) computes the maximum spectral efficiency satisfying the required BER constraint and reports the selected modulation and coding scheme (MCS) level index to the transmitter. To accomplish this, we will first propose a simple BER estimation scheme for the BIC-OFDM system, and then, we devise an MCS-level selection method, which does not require the water-filling process by utilizing the proposed BER estimation scheme for a given CSI. For the proposed AMC scheme, the error bounds are computed for candidate spectral efficiencies by averaging symbol error probabilities on each subcarrier. Thus, this results in much lower computational complexity than the water-filling procedure.

By adopting rate adaptation based on one OFDM symbol, the feedback overhead can be greatly reduced at the expense of a small performance loss. We will show in the simulation section that the proposed scheme significantly reduces the system complexity while minimizing the performance loss compared to the optimum WF-AMC scheme. Thus, the proposed scheme is beneficial for practical systems where the bandwidth for the feedback channel is limited.

The paper is organized as follows. In Section II, the system model for the adaptive BIC-OFDM is presented. The proposed AMC scheme with reduced FI is presented in Section III. Also, we derive a simple BER estimation method and apply it to the AMC scheme. Finally, the simulation results and a conclusion are presented in Sections IV and V, respectively.

II. System Model

We consider an OFDM system with N subcarriers, transmitting information sequences modulated by BICM [2]. The BICM achieves diversity gain in the frequency domain through channel coding in frequency-selective channels. Fig. 1 shows the system model of the adaptive BIC-OFDM system where thee error-free feedback channels are assumed for simplicity. The BIC-OFDM is constructed by concatenating a binary convolutional encoder with a memoryless mapper through a bit-level interleaver (Π). After the mapping, symbols in M-QAM are then serial-to-parallel converted and are modulated by the inverse fast Fourier transform (IFFT) block. As shown in Fig. 1, the transmission rates and power levels are controlled based on the FI obtained at the receiver.

Assuming that m bits are loaded at the nth subcarrier, the coded and bit interleaved bits $[d_n^1 d_n^2 \cdots d_n^m]$ are mapped into a symbol $s_n \in \mathcal{X}$ where $\mathcal{X}$ represents a constellation of size $M = 2^m$ with the minimum distance $d_{\min,m}$. For simplicity, one frame is assumed to consist of one OFDM symbol.

With the FFT operation at the demodulation, the received signal at the nth subcarrier at the lth time slot after removing

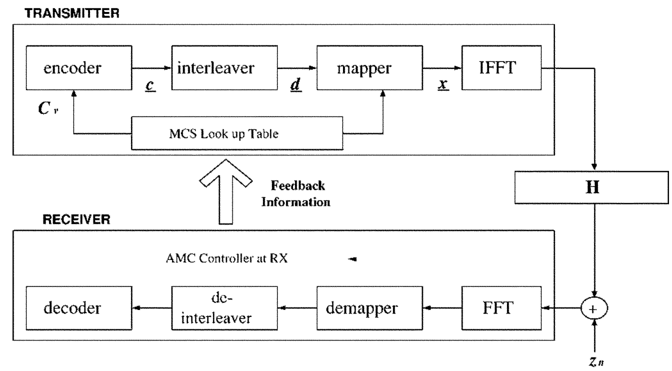

Fig. 1. System model for the adaptive BIC-OFDM.

the cyclic prefix (CP) is given by

$$y_{n,l} = H_{n,l} s_{n,l} + z_{n,l}$$

where $z_{n,l}$ denotes independent and identically distributed (i.i.d) complex additive Gaussian noise with variance σ^2 per complex dimension and $H_{n,l}$ represents the equivalent channel frequency response. Denoting the sampling period as T_s, the corresponding time-domain channel response can be modeled as a finite impulse response intersymbol-interference (FIR-ISI) channel

$$h(t,\tau) = \sum_{i=0}^{L-1} \bar{h}(i;t)\delta(\tau - iT_s)$$

where the channel coefficients $\bar{h}(i;t)$ are independent Gaussian variables with zero mean; $\delta(\cdot)$ represents the Dirac delta function; and L denotes the number of channel taps.

For the adaptive transmission, the CSI is assumed to remain unchanged during each packet transmission. Hence, we will omit the time indices l and t for simplicity. This gives rise to the channel frequency response at the nth subcarrier as

$$H_n = \sum_{i=0}^{L-1} \bar{h}(i) \exp\left(-j\frac{2\pi n i}{N}\right) \tag{1}$$

where $\bar{h}(i)$ represents the ith tap of the channel response. Throughout this paper, we assume that the CSI is perfectly known at the UE side.

III. Proposed Adaptive BIC-OFDM With Reduced Feedback Information

The WF-AMC scheme proposed in [6] requires full FI and the water-filling procedure to perform the rate adaptation. Thus, the overall FI amount and the computational complexity may be too high to be applied in practical wireless systems. In this section, we will propose a new AMC scheme for the BIC-OFDM system, which significantly reduces the required FI amount as well as the computational complexity.

One simple AMC scheme for reducing the FI amount is to employ a link table, which defines the SNR range assigned to each MCS level. For example, the UE reports the received SNR, and the BS selects the MCS level from the link table [3], [4]. However, this MCS-level selection technique does not work well in OFDM environments. For example, it is reported in [5] that the required SNR to achieve the BER of 10^{-4} varies by up to 5 dB as the RMS delay changes from 50 to 250 ns. This indicates that a different link table should be employed whenever the channel conditions such as the delay spread changes, and it is impractical to determine the link table in advance for all possible channels. Instead, in the proposed scheme, we utilize the estimated BER based on one OFDM symbol rather than each subcarrier. The UE computes the maximum achievable spectral efficiency satisfying the BER constraint, and then, sends back only the selected MCS-level index. The proposed estimation scheme makes use of the current CSI on each subcarrier in itself known at the UE. Thus, its performance depends mainly on the received SNR rather than on other parameters such as channel delay spreads or channel statistics.

In this section, we first propose a simple BER estimation method. Then, the rate adaptation algorithm based on one OFDM symbol will be proposed by utilizing the estimated BER.

A. Simple BER Estimation Scheme for BIC-OFDM

In the proposed scheme, it is important to estimate an instantaneous BER with the given CSI $\mathbf{H} = \{H_1, \ldots, H_N\}$ for variable constellation sizes and code rates. Our goal is to estimate the BER with reduced complexity for the AMC scheme.

Given an encoder ζ, $R_c(\zeta)$ represents the code rate, and $d_H(\zeta)$ stands for the minimum Hamming distance. Throughout this paper, we employ rate compatible punctured convolutional codes (RCPC) [7] where higher rate codes are punctured from a rate $1/2$ mother code with puncturing period p. Then, the instantaneous BER can be estimated as [7]

$$\hat{P}_b \simeq \frac{1}{p} \sum_{d=d_H(\zeta)}^{d_H(\zeta)+5} N(d) P(d, \mathbf{H}) \tag{2}$$

where $N(d)$ denotes the total input weight of error events at Hamming distance d of the code ζ, and $P(d, \mathbf{H})$ represents the average code word pairwise error probability (PEP) between the code words having Hamming distance d. $N(d)$ for various RCPC codes are tabulated in [8]. In computing the ealier estimated BER, we truncate the summation in (2) to $d_H(\zeta) + 5$ since the effect of the error events with Hamming distance d exponentially decays as the Hamming distance increases. The BER estimation in (2) involves high computational complexity associated with $P(d, \mathbf{H})$. Now, we focus on computing the average code word PEP $P(d, \mathbf{H})$ with lower complexity.

Suppose that the sequence of $\underline{c} = [c_1, \ldots, c_K]$ is sent, and the decoder chooses an erroneous sequence $\underline{\hat{c}} = [\hat{c}_1, \ldots, \hat{c}_K]$ where $\underline{c}$ and $\underline{\hat{c}}$ differ by d bits. Without loss of generality, we assume that $\underline{c}$ and $\underline{\hat{c}}$ differ in the first d consecutive bits. Let us define a sequence variable $\underline{S}$ as a Cartesian product of d such bit positions for $c_1, \ldots, c_d$ on one OFDM symbol after the bit-level interleaving $\underline{S} = (i_1, n_1) \times \cdots \times (i_d, n_d)$, where n_k and i_k denote the subcarrier index and the bit position of which c_k is mapped, respectively. We denote $\mathcal{X}_{\underline{c}}^{\underline{S}} = \mathcal{X}_{c_1}^{i_1} \times \cdots \times \mathcal{X}_{c_d}^{i_d}$ and $\mathcal{X}_{\underline{\hat{c}}}^{\underline{S}} = \mathcal{X}_{\hat{c}_1}^{i_1} \times \cdots \times \mathcal{X}_{\hat{c}_d}^{i_d}$ as the Cartesian product sequence of the signal subsets selected by the first d bits of $\underline{c}$ and $\underline{\hat{c}}$, respectively, where $\mathcal{X}_{c_k}^{i_k}$ stands for a subset of signal points whose i_kth bit is c_k. Denoting γ as the SNR of equivalent binary input BICM additive white Gaussian noise AWGN (BI-AWGN) channel [9], and $\bar{B}_M(\mathbf{H})$ as the Bhattacharyya factor for M-QAM signal which is defined as $E_k\left[P(c_k \to \hat{c}_k | \underline{S}, \mathbf{H})\right]$ [10], the standard union and Bhattacharyya bound of $P(d, \mathbf{H})$ can be written as [2], [10]

$$P(d, \mathbf{H}) \leq Q(\sqrt{2d\gamma}) \leq \left(\bar{B}_M(\mathbf{H})\right)^d \tag{3}$$

where $Q(x)$ is defined as $1/\sqrt{2\pi} \int_x^\infty e^{-z^2/2} dz$. For BI-AWGN channel, $\bar{B}_2(\mathbf{H})$ is equal to $e^{-\gamma}$ [10], and $\bar{B}_M(\mathbf{H})$ for $M > 2$ can also be approximated as $e^{-\gamma}$ where γ is given as a scaled version of the received SNR [9]. Thus, $P(d, \mathbf{H})$ in (3) can be rewritten as

$$P(d, \mathbf{H}) \leq Q\left(\sqrt{-2d \ln \bar{B}_M(\mathbf{H})}\right). \tag{4}$$

From the Chernoff bound of the Q-function, which is expressed as $Q(x) \leq e^{-x^2/2}$, it is obvious that $Q(\sqrt{-2d \ln \bar{B}_M(\mathbf{H})}) \leq \bar{B}_M(\mathbf{H})^d$. This implies that the approximation in (4) provides more tightly estimated BER than the standard Bhattacharyya bound using the instantaneous Bhattacharyya factor $\bar{B}_M(\mathbf{H})$.

For OFDM systems, the exact computation of $P(d, \mathbf{H})$ or $\bar{B}_M(\mathbf{H})$ in (3) is quite complex due to the frequency selectivity of the channel. In [2], an expurgated union bound has been proposed to compute $P(d, \mathbf{H})$ efficiently using the Gauss–Chebyshev quadrature method. This bound provides a tight BER bound by expurgating irrelevant sequences of erroneous bit sequences. It is to be noted that the expurgation is performed on error events, not on each symbol. Although the expurgated union bound approximates the BER accurately, channel coefficients are required to be uncorrelated in order for d coded bits errors to be considered as independent events. However, in practice, the channel coefficients of each subcarrier in the OFDM system are normally correlated in frequency domain. In addition, since the Gauss–Chebyshev quadrature method requires numerical integrations, the scheme in [2] can hardly be employed for practical AMC schemes due to its computational complexity. Instead, we propose a simple BER estimation scheme by employing an instantaneous Bhattacharyya union bound, which averages the transition probability of a coded bit. The proposed scheme does not depend on the channel statistics and the sequence of bit positions. Assuming a uniform interleaver [11], (3) can be simplified as

$$\begin{aligned} P(d, \mathbf{H}) &\simeq \prod_{k=1}^{d} E_{H_{n_k}} \\ &\quad \times \left[\frac{1}{m 2^{m-1}} \sum_{i=1}^{m} \sum_{x_{n_k} \in \mathcal{X}_0^i} \sum_{v_{n_k} \in \mathcal{X}_1^i} P(x_{n_k} \to v_{n_k} | H_{n_k}) \right] \\ &\simeq \left(E_k \left[\frac{1}{m 2^{m-1}} \sum_{i=1}^{m} \sum_{x_k \in \mathcal{X}_0^i} \sum_{v_k \in \mathcal{X}_1^i} P(x_k \to v_k | H_k) \right] \right)^d \\ &= (\bar{B}_M(\mathbf{H}))^d \end{aligned} \tag{5}$$

where the instantaneous Bhattacharyya factor $\bar{B}_M(\mathbf{H})$ can be computed as

$$\bar{B}_M(\mathbf{H}) = \frac{1}{N}\sum_{n=1}^{N}\frac{1}{m2^{m-1}}\sum_{i=1}^{m}\sum_{x_n\in\mathcal{X}_0^i}\sum_{v_n\in\mathcal{X}_1^i}P(x_n\to v_n|H_n)$$
$$\triangleq\frac{1}{N}\sum_{n=1}^{N}B_M(H_n). \tag{6}$$

Here, the symbol transition probability at the nth subcarrier $P(x_n \to v_n|H_n)$ can be written as

$$P(x_n\to v_n|H_n) = Q\left(\sqrt{\frac{|H_n|^2|x_n-v_n|^2}{4\sigma^2}}\right).$$

In contrast to (3), we have not employed the inequality in (5) since H_n are not uncorrelated, and the interleaver is not ideal in practical OFDM systems. It will be shown in the simulation section that actual BER results are usually poor than estimated BER values due to the correlations and the finite length of the interleaver if (4) is employed to estimate the BER in (2). However, the purpose of the BER estimation is to provide a decision metric for rate adaptations. Therefore, we will use inequalities in (3) to compensate the looseness of the standard Bhattacharyya union bound for OFDM systems.

The above computation can be further reduced by considering only the neighboring points. The most likely error events occur when the erroneous decision v_k is one of the neighboring points around x_k whose ith bit is different from x_k. When computing the symbol transition probability $P(x_n \to v_n|H_n)$, the contribution from the transition to points other than the neighboring ones becomes negligible especially at high SNR. For M-QAM constellations with Gray labeling, the number of such neighboring points is at most three. Fig. 2 shows an example of 16-QAM constellation for $i = 1, 3$. In Fig. 2(a), the shaded circles represent a set of points whose first bit position is 1, while the shaded circles in Fig. 2(b) indicate a set of points whose third bit position is 1. There are three types of transitions characterizing each error event. Point A1 has two neighboring points, which differ at the first bit position, whereas A2 has three such points. On the other hand, A3 does not have any neighboring points, which differ at the first bit position. We note that there are four such points for $i = 1$ and 2, while there exists no point that can be classified as A3 for $i = 3$ and 4. In contrast to the expurgated union bound [2], we also consider diagonal transitions such as the transition "0000" → "1000" in Fig. 2(a) to average erroneous bit transitions for obtaining the instantaneous Bhattacharyya factor since the PEP estimation in (3) comes from the Bhattacharyya union bound for computational simplicity.

Denote Q_1 and Q_2 as

$$Q_1 = Q\left(\sqrt{\frac{|H_n|^2 d_{\min,m}^2}{4\sigma^2}}\right)$$
$$Q_2 = Q\left(\sqrt{\frac{|H_n|^2(\sqrt{2}d_{\min,m})^2}{4\sigma^2}}\right). \tag{7}$$

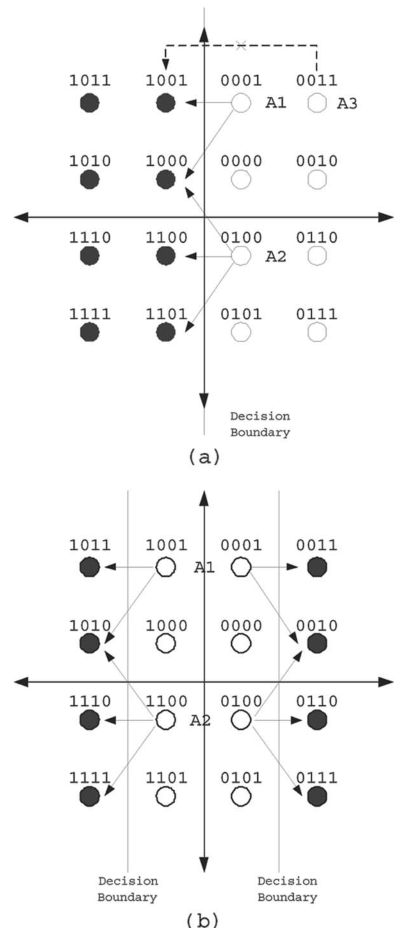

Fig. 2. Neighboring points of 16-QAM signal set with Gray mapping. (a) First bits. (b) Third bits.

The transition probability for A1 and A2 are $Q_1 + Q_2$ and $Q_1 + 2Q_2$, respectively. Considering only the points classified as A1 and A2, $B_M(H_n)$ in (6) for 16-QAM can be obtained as

$$B_{16}(H_n) = \frac{2}{4}\underbrace{\left(\frac{2}{8}(2Q_1+3Q_2)\right)}_{i=1,2} + \frac{2}{4}\underbrace{\left(\frac{4}{8}(2Q_1+3Q_2)\right)}_{i=3,4}$$
$$=\frac{3}{8}(2Q_1+3Q_2) \tag{8}$$

where $Q(x)$ can be approximated by [12]

$$Q(x)\approx\frac{1}{x(1-(1/\pi))+(1/\pi)\sqrt{x^2+2\pi}}\frac{1}{\sqrt{2\pi}}\exp\left(-\frac{x^2}{2}\right).$$

Similarly, $B_M(H_n)$ values for QPSK and 64-QAM can be computed as $Q_1 + Q_2$ and $(28Q_1 + 49Q_2)/48$, respectively. As $B_M(H_n)$ is readily computed as a function of H_n, the BER computation becomes quite simple. By applying $B_M(H_n)$ for each subcarrier to (6), we can compute the instantaneous Bhattacharyya factor $\bar{B}_M(\mathbf{H})$, which can be utilized for obtaining the PEP and the estimated BER in (4) and (2), respectively.

B. Proposed Block AMC Scheme for BIC-OFDM

In this section, we propose a rate adaptation scheme based on the OFDM symbol utilizing the estimated instantaneous BER described in Section III-A. This proposed scheme will be referred to as the block AMC (BL-AMC), since this is based on individual OFDM symbol rather than each subcarrier. The proposed BL-AMC significantly reduces the amount of FI while minimizing the performance loss compared to the WF-AMC. The objective of the adaptive transmission is to maintain the average BER below the BER constraint by varying the transmission rate according to the current CSI. Thus, the AMC provides high spectral efficiencies by transmitting at high speeds under favorable channel conditions and reducing the transmission rate as the channel degrades.

Each MCS level l $(l = 1, \ldots, l_{\max})$ consists of a convolutional encoder $\mathcal{C}_l \in \{\zeta_1, \ldots, \zeta_V\}$, and the M_l-QAM constellation where $\log_2 M_l = m(l) \in \{1, \ldots, m_{\max}\}$. The spectral efficiency supported by the MCS level l can be obtained as $R_T = R_c(\mathcal{C}_l) \log_2 M_l$. With the BER constraint P_e, the rate maximizing AMC for the proposed BL-AMC results in the following optimization problem:

$$\begin{aligned} &\underset{l}{\text{maximize}}\ R_T(l) = R_c(\mathcal{C}_l) \log_2 M_l \\ &\text{subject to } \hat{P}_b(l) \le P_e \end{aligned} \tag{9}$$

where $\hat{P}_b(l)$ denotes the estimated BER with respect to the MCS level l.

The proposed BL-AMC scheme maintains the constant average transmit energy over every subchannel while adapting the transmission rate to the channel conditions without degrading P_e at the receiver. Ignoring $N(d)$ in (2), we can obtain an acceptable BER estimation performance as long as $|H_n|^2 d_{\min}^2$ in (7) is kept above a certain level with small $d_{\min}$. This implies that higher modulation levels can be selected for the transmission under favorable channel conditions (i.e., high $|H_n|^2$). Using $d_{\min}^2 = 6E_s/(M-1)$ [13], Q_1 and Q_2 in (7) can be rewritten as

$$\begin{aligned} Q_1 &= Q\left(\sqrt{\frac{|H_n|^2 6E_s}{4\sigma^2(M-1)}}\right) \\ Q_2 &= Q\left(\sqrt{\frac{|H_n|^2 12E_s}{4\sigma^2(M-1)}}\right). \end{aligned} \tag{10}$$

Thus, we can estimate the BER as a function of SNR and M by utilizing (10) for obtaining $\bar{B}_M(\mathbf{H})$ in (6).

Now, by applying (4) to (2), the optimization problem in (9) can be rewritten as

$$\begin{aligned} &\underset{l}{\text{maximize}}\ \ R_T(l) = R_c(\mathcal{C}_l) \log_2 M_l \\ &\text{subject to } \frac{1}{p_l} \sum_{d=d_H(\mathcal{C}_l)}^{d_H(\mathcal{C}_l)+5} N_l(d) Q\left(\sqrt{-2d \ln \bar{B}_{M_l}(\mathbf{H})}\right) \le P_e \end{aligned} \tag{11}$$

where p_l represents the puncturing period for an RCPC code with a rate $R_c(\mathcal{C}_l)$, and $N_l(d)$ stands for the total input weight of error events at Hamming distance d of the code $\mathcal{C}_l$.

Then, UE reports the MCS-level index l, which maximizes the spectral efficiency in (11). A set of supportable spectral efficiencies is normally predetermined by the system specification.

For the proposed scheme, the amount of FI is independent of N and requires only $\lceil \log_2 l_{\max} \rceil$ bits, irrespective of N, which is the number of bits required to represent one of $l_{\max}$ MCS levels, where $\lceil a \rceil$ denotes the smallest integer no less than a. In contrast, for the WF-AMC case, the amount of FI increases as the number of subcarrier grows. Assuming that the water-filling optimization is carried out at the UE side to reduce the control channel overheads, the WF-AMC scheme requires $N(\lceil \log_2 m_{\max} \rceil + N_q) + \lceil \log_2 V \rceil$ bits in total for FI where N_q denotes the number of bits to represent the quantized power level.

Now, we consider the computational complexity of adaptation schemes. The WF-AMC scheme for the BIC-OFDM system needs to perform the Levin–Campello (LC) algorithm V times to select the $\mathcal{C}_v$ and m, which maximizes the throughput [6]. The LC algorithm performs the "efficientizing" and the "E-tightening" subroutines to obtain the optimal bit distribution for a given channel code [13]. Then, the total complexity for the WF-AMC becomes $O(V(2N + 2N_{EF} \cdot N + N_{ET} \cdot N))$, where $O(V(2N + 2N_{EF} \cdot N))$ and $O(V \cdot N_{ET} \cdot N)$ complexity account for the "efficientizing" subroutine and the "E-tightening" subroutine, respectively, where N_{ET} and N_{EF} depend on the number of iterative searches to obtain the solutions for the "efficientizing" and "E-tightening" subroutines, respectively.

In contrast, as shown in (6) and (11), the proposed BL-AMC scheme requires at most $O(l_{\max} N)$ operations; thus, the complexity reduction over the WF-AMC becomes $l_{\max}/(V(2 + 2N_{EF} + N_{ET}))$. Note that the terms N_{ET} and N_{EF} can grow up to N in the worst case, depending on the initial bit-loading and channel conditions [13] at each transmission. Therefore, the proposed scheme not only reduces the amount of FI, but also provides a substantially simpler computational algorithm. In addition, the proposed BL-AMC scheme does not require any memory space to store intermediate results as in the WF-AMC scheme [13].

IV. Simulation Results

In this section, we present the simulation results for the proposed BL-AMC scheme. For OFDM modulation, $N = 64$ subcarriers are used and the cyclic prefix length is set to 16 samples with the sampling rate $T_s = 50$ ns. The AMC table used for simulations is listed in Table I.

The channel is assumed to have exponentially decaying multipath Rayleigh fadings. To generate channel impulse responses with various root mean square (rms) delay spreads T_{rms}, $\bar{h}(i)$ in (1) is modeled as a complex Gaussian random variable with $\bar{h}(i) \sim \mathcal{CN}(0, \sigma_0 \exp(-iT_s/T_{\text{rms}}))$ where σ_0 is given as $\sigma_0 = 1 - \exp(T_s/T_{\text{rms}})$ [5]. For each convolutional code ζ_v, the information error weight $N_v(d)$ is tabulated in Table II [8].

TABLE I
AMC TABLE FOR SIMULATIONS

l	R_T(bps/Hz)	R_c	Modulation
1	1	1/2	QPSK
2	2	1/2	16-QAM
3	3	3/4	16-QAM
4	4	2/3	64-QAM
5	4.5	3/4	64-QAM
6	5	5/6	64-QAM

TABLE II
INFORMATION ERROR WEIGHT FOR RCPC CODES [8]

v	R_c	d_H	p	$N_v(d)$, $d = d_H, \cdots, d_H + 5$
1	1/2	10	3	108,0,633,0,4212,0
2	2/3	6	2	3,70,285,1276,6160,27128
3	3/4	5	3	42,201,1492,10469,62935,379546
4	5/6	4	5	92,528,8694,79453,791795,7369828

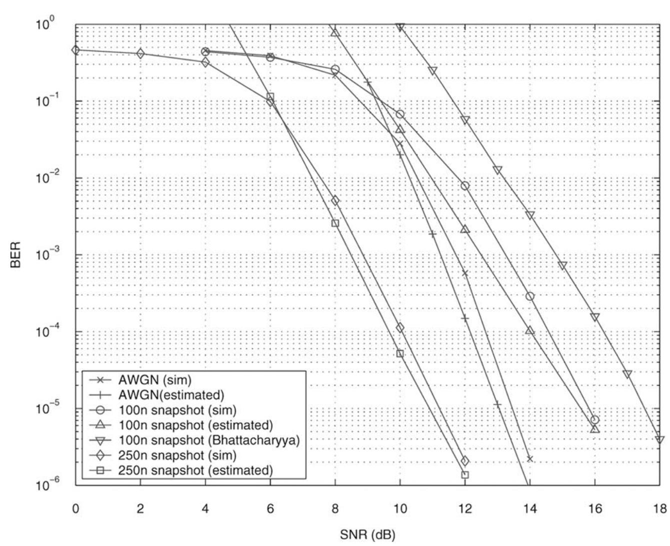

Fig. 3. Performance of the proposed BER estimation scheme.

Over 100,000 frame transmissions are simulated, and the target frame error rate (FER) is set to 1%; the corresponding BER constraint is around 10^{-4} as in [6].

In Fig. 3, we present the results of the BER estimation method proposed in Section III. In this figure, we plot the BER for 16-QAM obtained using (8) for $T_{\text{rms}} = 100$ and 250 ns. We generate snapshots at an arbitrary time for a given channel profile, and perform the Monte Carlo simulations to measure the instantaneous BER. For comparison, we also plot the average BER for an AWGN channel. As shown in the figure, the estimated BER is only 0.5 dB away from the actual simulated results at a BER

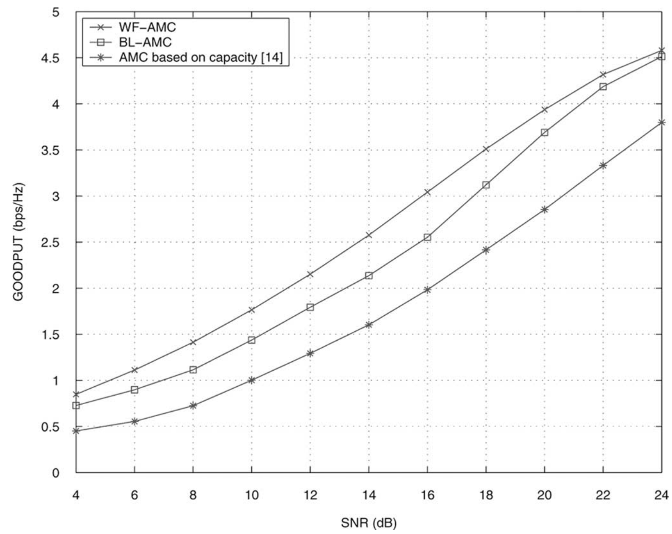

Fig. 4. Average goodput for the adaptive BIC-OFDM with 100 ns RMS delay spread.

of 10^{-4}. This plot shows that for the SNR range of interest, the proposed estimation scheme is accurate enough for the MCS-level decision process. The proposed scheme employs the PEP bound with the Gaussian approximation and excludes a large number of irrelevant terms by considering only the neighboring points. Hence, the proposed scheme exhibits a better bound with simpler computation compared to the standard Bhattacharyya union bound as shown in Fig. 3.

In evaluating the performance of the AMC schemes, we adopt the "goodput" [14] to measure the system throughput by counting the information bits in decoded frames with correct cyclic redundancy check (CRC) in the automatic repeat request (ARQ) mechanism. Fig. 4 presents the average throughput with the RMS delay spread of 100 ns. This plot shows that when comparing the proposed scheme with WF-AMC, the SNR gap is about 2 dB at the spectral efficiency of 2.5 b/s/Hz. It should, however, be emphasized that the total system complexity for the proposed scheme is much lower than the WF-AMC. Employing the AMC table in Table I, only 3 bits of FI are sufficient for the whole packet for the proposed BL-AMC scheme, while $64(3+6)+2 = 578$ bits should be reported for the WF-AMC scheme if $N_q = 6$ bits are assumed to quantize the loaded energy level for each subcarrier. Thus, the proposed scheme significantly reduces the FI payloads. For comparison, we also plot the throughput of the rate adaptation scheme based on the instantaneous capacity

$$R_T = \left[\frac{1}{N} \sum_{k=1}^{N} \log \left(1 + \frac{|H_k|^2 E_s}{\sigma^2 \Gamma} \right) \right]$$

proposed in [15]. This is one of the simplest AMC scheme but suffers a performance loss due to the suboptimum rate selection. Here, the channel gap Γ is set to 12 dB for 100 ns RMS delay spread to satisfy the 1% FER constraint.

In Fig. 5, we plot the FER of the proposed BL-AMC scheme. As shown in the figure, the proposed scheme satisfies the

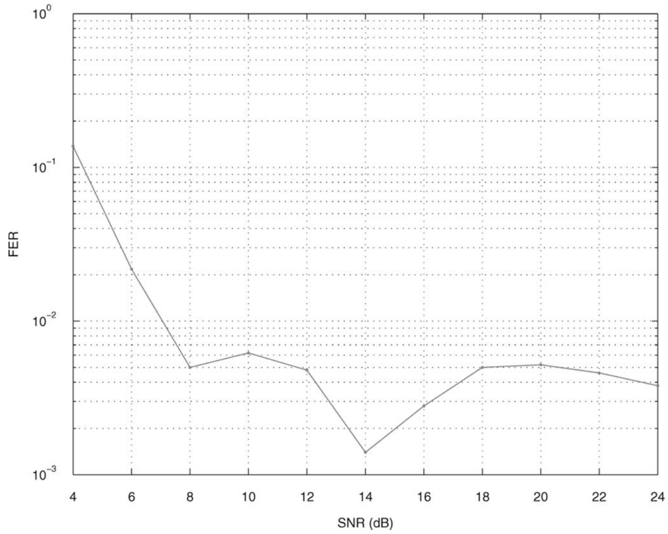

Fig. 5. FER performance of the proposed BL-AMC.

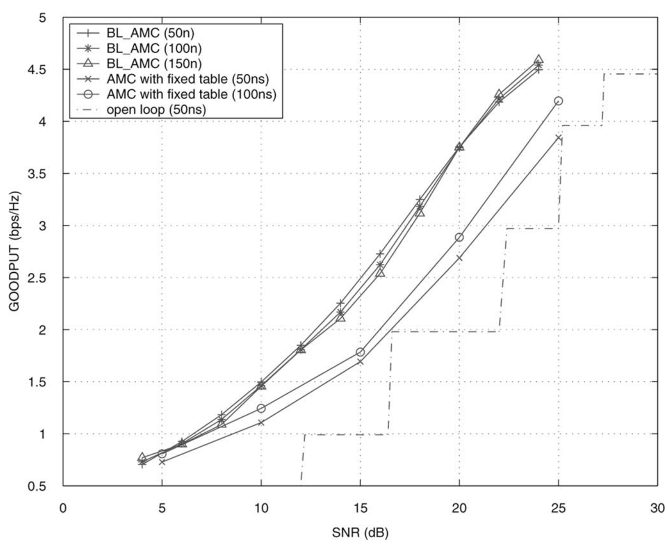

Fig. 6. Average goodput for the proposed BL-AMC with various RMS delay spreads.

constraint of $\mathrm{FER} = 1\%$ at SNRs above 8 dB. For SNRs below 8 dB, the FER constraint (11) cannot be satisfied since the minimum MCS level of 1 b/s/Hz exceeds the actual supportable rate for the SNR range.

As mentioned earlier, the performance of OFDM systems without rate adaptations depends not only on the received SNR but also the RMS delay spreads of the channel impulse response. As the delay spreads decrease, the performance gets degraded due to the reduced frequency diversity. This performance loss can be compensated by controlling the transmission rates. We present the goodput measurements of the rate adaptation schemes with various channel delay profiles in Fig. 6. In this simulation, we employ the link tables matched with the RMS delay spread at each case. Fig. 6 shows that the proposed AMC scheme is almost insensitive to different RMS delay spreads, and exhibits better performance than the conventional AMC schemes with fixed link tables. We also plot the throughput of the system without AMC satisfying the 1% FER constraint. This clearly illustrates that the performance gain of the AMC is significant over the non-AMC case.

V. Conclusion

In this paper, we propose a new AMC scheme for BIC-OFDM to reduce the overall system complexity by employing the rate adaptations based on one OFDM symbol. In this scheme, the required FI reduces to an MCS-level index number, whereas the optimal water-filling solution needs significantly higher feedback traffics to represent the bit loading information or channel coefficients for each subcarrier.

Simulation results show that the proposed method exhibits a significant throughput enhancement over the system without AMC. The performance of the proposed BL-AMC comes within 2 dB of the optimal WF-AMC with much reduced system complexity. We conclude that the proposed scheme is beneficial for systems where the bandwidth for the feedback channel is limited.

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