Turbo Coded OFDM for Reducing PAPR and Error Rates

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***Abstract*— A selective-mapping (SLM) scheme which does not require the transmission of side information and can reduce the peak to average power ratio (PAPR) in turbo coded orthogonal frequency-division multiplexing (OFDM) systems is proposed. The candidates of the proposed SLM are respectively generated by a turbo encoder using various interleavers. The waiver of side information can avoid the degradation of error rate performance which results from the incorrect recovery of side information at receiver in the conventional SLM OFDM system.**

***Index Terms*— Orthogonal frequency-division multiplexing (OFDM), peak-to-average power ratio (PAPR), selective-mapping (SLM), turbo code.**

I. Introduction

ORTHOGONAL frequency-division multiplexing (OFDM) is a popular modulation choice for digital transmissions. The occasional occurrence of high peak-to-average power ratio (PAPR) is a well-known disadvantage of the OFDM systems. By now, many techniques have been proposed for relieving the PAPR problem in the OFDM, which can be roughly divided into two classes, the distortion-based techniques and the redundancy-based techniques.

The distortion-based techniques reduce the PAPR of the OFDM symbol with the price of adding distortion to the signal points in the subcarriers. Direct clipping [1] simply suppresses the time-domain OFDM signals of which the signal powers exceed a certain threshold. The penalty is the significant increase of out-of-band energy. Peak windowing [2] or filtering after direct clipping [3] can be used to reduce the out-of-band energy. After the filtering operation, the peak of the time-domain signal may regrow. Hence, recursive clipping and filtering (RCF) [4] can be used to suppress both the out-of-band energy and the PAPR. RCF can be modified by restricting the region of distortion [5] to obtain improved error performance. On the other hand, estimation of the clipping noise at the receiver [6] can be used to improve the error performance of direct clipping or RCF.

The redundancy-based technique includes coding, selective-mapping (SLM), partial transmit sequences, tone reservation and tone injection [7]–[13], etc. For the redundancy-based technique, the undesired effects occurring to the distortion-based techniques can be alleviated while the penalty is the reduced transmission rate or increased average power due to the introduction of redundancy.

The basic idea of SLM technique is to generate several OFDM symbols as candidates and then select the one with the lowest PAPR for actual transmission. Conventionally, the transmission of side information is needed so that the receiver can use the side information to tell which candidate is selected in the transmission. In [14] and [15], the side information for a channel coded SLM appears explicitly in the data sequence to be encoded so that the side information is protected by the same channel code. The advantage of such an arrangement is that no additional protection is needed for side information and the rate loss due to the side information is small. However, once the side information is incorrectly decoded, the number of error bits in the erroneously decoded codeword can be great. In [16], an SLM technique (for either coded or uncoded cases) which does not need the transmission of side information was proposed, where the discrimination of the desired candidate against the undesired candidates is obtained by specially arranging the constellations for the subcarriers of each candidate so that the modulated signal points for the subcarriers of each pair of candidates are widely different.

In this letter, we propose a side-information free SLM scheme to reduce the PAPR of turbo coded OFDM, for which the turbo encoder uses distinct interleavers to generate distinct candidates. The receiver of the proposed scheme uses the maximum a posterior probability (MAP) decoder for the turbo code to calculate the reliability of each candidate. Although side information is not available, the reliability of the decoded results will be high and the receiver can recover the correct codeword in case that the candidate chosen by the receiver is correct. If the candidate is not the one chosen by the transmitter, the reliability of the decoded results will be very low and the receiver needs to try another candidate.

In Section II of this letter, we will provide some basics about the SLM technique. In Section III, a side-information free SLM scheme for turbo coded OFDM is proposed. In section IV, simulation results and performance evaluation of the proposed scheme are given. To reduce the decoding complexity, a variation of the proposed SLM is given in Section V. Conclusions are given in Section VI.

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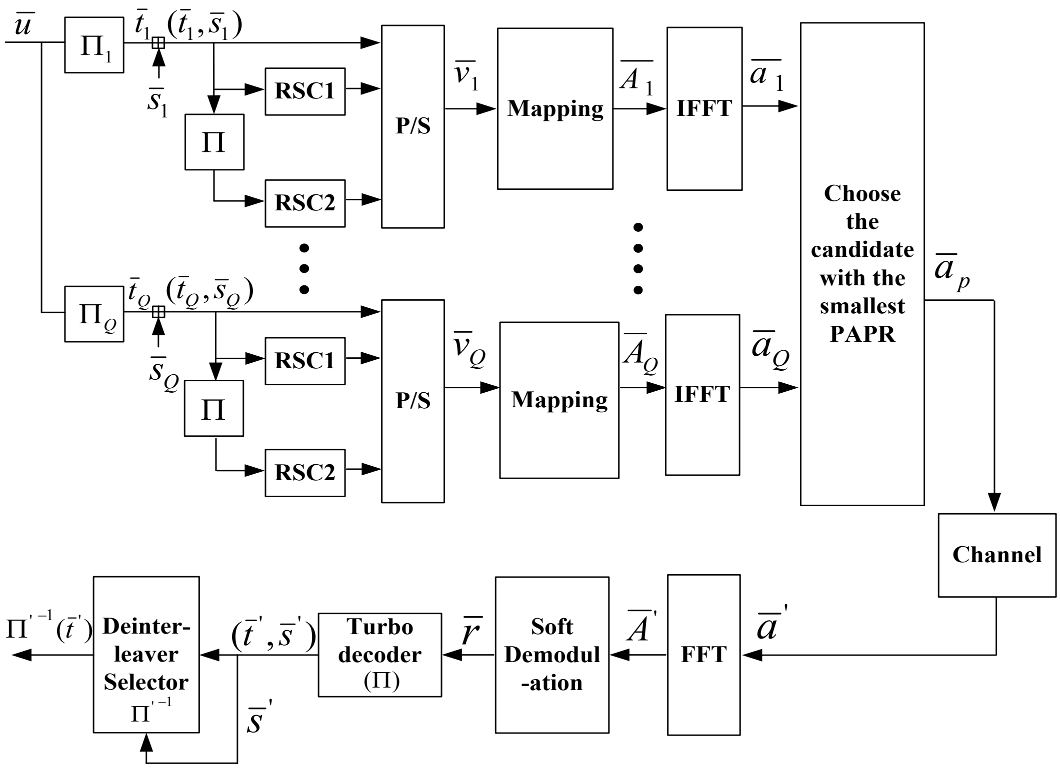

Fig. 1. The transmitter and receiver of Scheme I.

II. Selective-Mapping for PAPR Reduction

Consider an OFDM system with N subcarriers, for which the time-domain OFDM symbol $a(t)$ is represented by

$$a(t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} A_k e^{jk2\pi\Delta f t}, \quad 0 \leq t < T \qquad (1)$$

where T is the symbol interval, $\Delta f = 1/T$ is the frequency spacing between adjacent subcarriers and A_k is the complex value carried by the kth subcarrier of the OFDM symbol. The peak power of $a(t)$ may be significantly greater than the average power of the transmitted OFDM symbols. The effect is measured by the peak-to-average power ratio (PAPR). The definition of PAPR is

$$PAPR(a(t)) = \frac{\max_{0 \leq t \leq T} \mid a(t) \mid^2}{\frac{1}{T} E\{\int_o^T \mid a(t) \mid^2 dt\}}, \qquad (2)$$

where the expectation $E\{\cdot\}$ is taken over all the possible transmitted OFDM symbols $a(t)$. The complementary cumulative distribution function (CCDF) for PAPR is the probability of OFDM symbols with PAPR exceeding some threshold λ. We can approximate $a(t)$, $0 \leq t < T$, by N points, i.e, $\bar{a} = (a_0, a_1, \cdots, a_{N-1})$. In case that $N > 100$ and $\lambda \gg 1/2$ [17], we can well approximate the CCDF for PAPR associated to $\bar{a}$ by

$$P(PAPR > \lambda) \approx 1 - exp[-\sqrt{\frac{\pi}{3}} N \sqrt{\lambda} e^{-\lambda}]. \qquad (3)$$

The SLM [9] technique for reducing PAPR is operated as follows. Each message to be transmitted is assigned with Q possible OFDM symbols for transmission, where each OFDM symbol is called a candidate. The transmitter selects the candidate with the smallest PAPR for transmission. Under the assumption that the Q candidates generated from the transmitter are statistically independent OFDM symbols, the CCDF of PAPR is reduced to

$$P(PAPR > \lambda) \approx (1 - exp[-\sqrt{\frac{\pi}{3}} N \sqrt{\lambda} e^{-\lambda}])^Q. \qquad (4)$$

In order to recover the transmitted message, the receiver requires the knowledge about which candidate is selected at the transmitter. We may embed the $\lceil log_2 Q \rceil$-bit side information in the transmitted symbol so that the receiver can recover the side information and hence the associated candidate. Consider a SLM turbo coded OFDM scheme shown in Fig. 1, denoted scheme I, which is similar to the scheme in [14]. Scheme I will be used for comparison with the scheme proposed in Section III. For the transmitter of scheme I, the message sequence $\bar{u}$ is processed by Q distinct interleavers. The q-th interleaver, Π_q, $q = 1, 2, \cdots, Q$ converts $\bar{u}$ into a sequence $\bar{t}_q = \Pi_q(\bar{u})$. The sequence $(\bar{t}_q, \bar{s}_q)$, that is obtained by padding the $\lceil log_2 Q \rceil$-bit side information $\bar{s}_q$ to $\bar{t}_q$, is encoded into a binary turbo codeword $\bar{v}_q = (v_{q,0}, \cdots, v_{q,mN-1})$, which is mapped into $\bar{A}_q = (A_{q,0}, \cdots, A_{q,N-1})$, where for $k = 0, 1, \ldots, N-1$, $A_{q,k}$ is the constellation point mapped from a binary m-tuple $(v_{q,km}, \cdots, v_{q,km+m-1})$. Each $\bar{A}_q$ is then converted to $\bar{a}_q$ through the IFFT operation. Finally, the sequence $\bar{a}_p$ with the lowest PAPR among the Q output sequences, i.e., $\bar{a}_1$, $\bar{a}_2$, $\cdots$, $\bar{a}_Q$, is selected for transmission, where $\bar{a}_p$ = IFFT($\bar{A}_p$).

At the receiving side, the signal vector $\bar{a}'$ is converted into $\bar{A}' = (A'_0, A'_1, \ldots, A'_{N-1})$ through the FFT operation, where for $k = 0, \cdots, N-1$, $A'_k = h_k A_{p,k} + \eta_k$, h_k is the channel gain, and η_k is the complex noise with Gaussian distribution of zero mean and variance σ_n^2 for either the real part or the

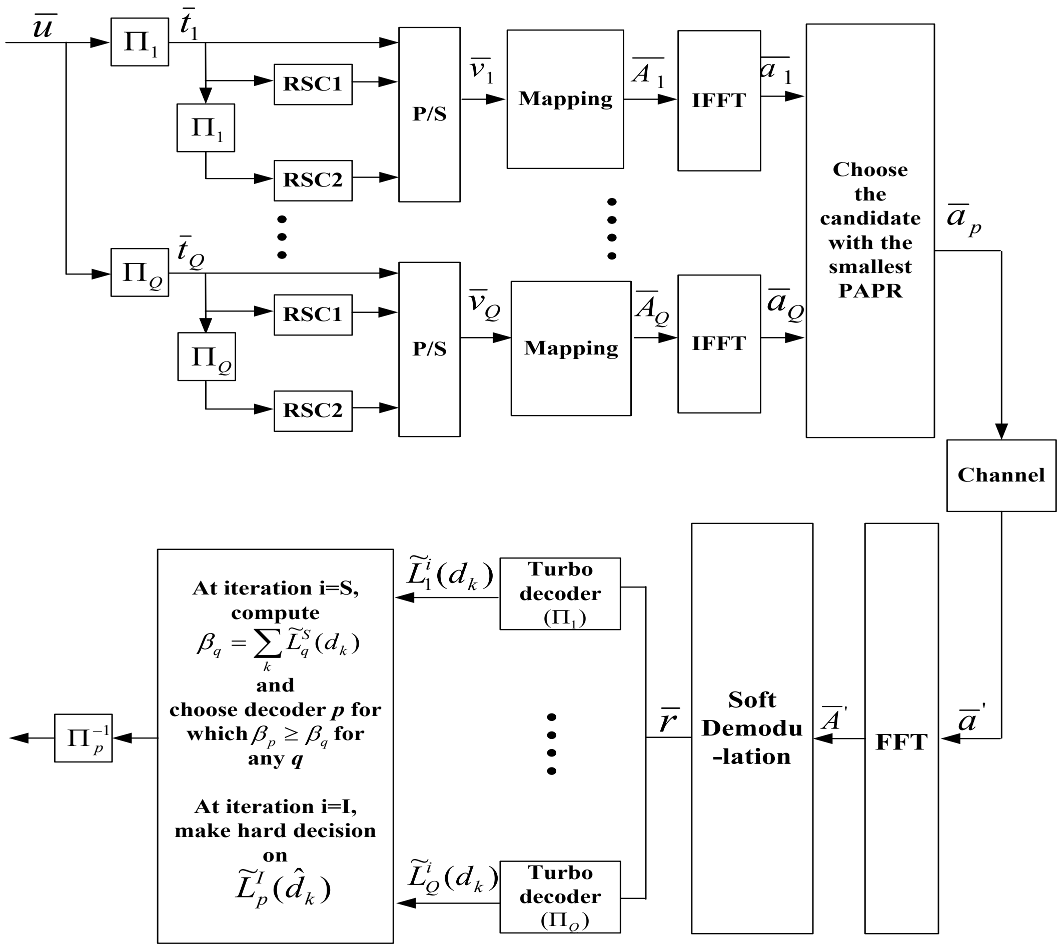

Fig. 2. The transmitter and receiver of Scheme II

imaginary part. We assume $h_k = 1$ for the additive white Gaussian noise (AWGN) channel. We also assume that h_k is a normalized Rayleigh random variable and h_i is independent of h_j for $i \neq j$ for the memoryless Rayleigh fading channel [18]. For the soft demodulation [18], the LLR for a code bit v_j, $j \in \{km, \cdots, km+m-1\}$ is

$$\wedge(v_j) = K_c \log \frac{prob(v_j = 1|A'_k)}{prob(v_j = 0|A'_k)}, \tag{5}$$

where K_c is a constant. We then have the vector of soft demodulation given by

$$\bar{r} = (r_0, r_1, \cdots, r_{mN-1}) = (\wedge(v_0), \wedge(v_1), \cdots, \wedge(v_{mN-1})) \tag{6}$$

which is fed to the binary turbo decoder to recover ($\bar{t}'$,$\bar{s}'$). From $\bar{s}'$, we find the associated interleaver Π'. Finally, we have $\bar{u}' = \Pi'^{-1}(\bar{t}')$ as the decoded message.

III. Selective-Mapping Turbo Coded OFDM Schemes without Side Information

In scheme I, if the side information is decoded incorrectly, the whole message block will be in serious error. Hence, the incorrectly decoded side information will significantly increase the bit error rate. In this section, we propose a SLM turbo coded OFDM scheme, denoted scheme II, that can reduce PAPR without the need of side information.

The transmitter of scheme II is shown in Fig. 2. For $q = 1, 2, \cdots, Q$, the $K - bit$ message sequence $\bar{u}$ is processed by the interleaver Π_q to obtain the output $\bar{t}_q$, which is then encoded by the q-th turbo encoder to yield a turbo coded sequence $\bar{v}_q$, that is mapped to $\bar{A}_q$ and is then converted into $\bar{a}_q$ through the IFFT operation. Note that the for the q-th turbo encoder, the input to the first component convolutional code RSC1 is $\bar{t}_q = \Pi_q(\bar{u})$, while the input to the second component convolutional code RSC2 is $\Pi_q(\bar{t}_q) = \Pi_q(\Pi_q(\bar{u}))$. Then, $\bar{a}_p$ is selected for transmission if $\bar{a}_p$ has the lowest PAPR among all the $\bar{a}_q, q \in \{1, 2, \cdots, Q\}$.

The structure of the receiver is shown in Fig. 2, where the input $\bar{a}'$ is converted into $\bar{A}'$ and the the vector of soft demodulation, $\bar{r}$, is obtained from (6). Then, $\bar{r}$ is sent to the q-th binary turbo decoder for $q = 1, 2, \cdots, Q$. At the i-th iteration, $1 \leq i \leq S \leq I$, the q-th decoder, $1 \leq q \leq Q$, computes the reliability

$$\tilde{L}_q^i(d_j) = \log[\frac{prob(d_j = 1 \mid \bar{r}, \tilde{L}_q^{i-1}(d_l), l \neq j)}{prob(d_j = 0 \mid \bar{r}, \tilde{L}_q^{i-1}(d_l), l \neq j)}] \tag{7}$$

for the j-th message bit d_j, where S is the number of iterations for the receiver to determine which candidate is the most possible one and I is the total number of iterations of complete turbo decoding. At the S-th iteration of decoding, $\beta_q = \sum_j \tilde{L}_q^S(d_j)$ is computed for each q.

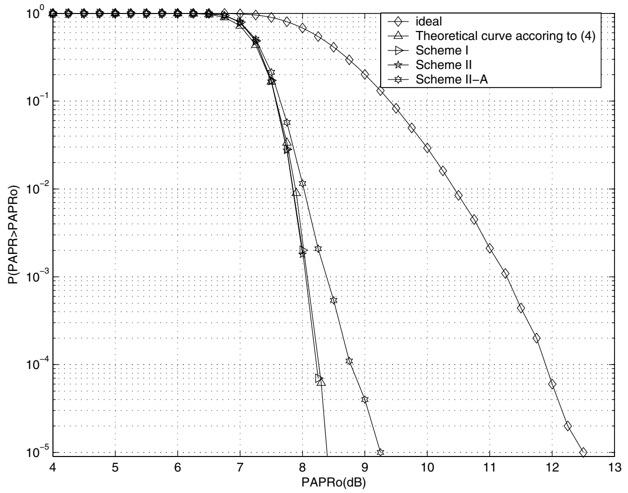

Fig. 3. CCDF of rate 1/2 16QAM Turbo-coded 256-tone OFDM systems, Q=16.

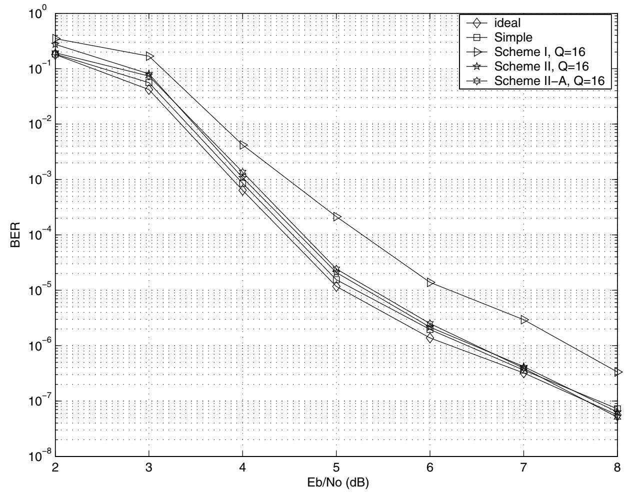

Fig. 4. BER for rate 1/2 16QAM Turbo-coded 256-tone OFDM systems over AWGN channel under IBO of 6dB.

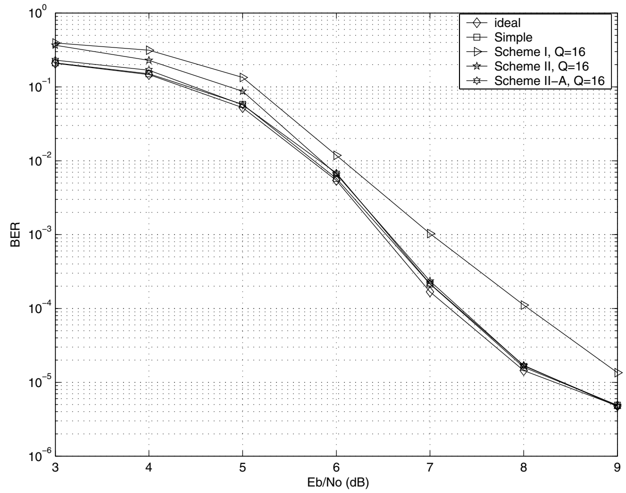

Fig. 5. BER for rate 1/2 16QAM Turbo-coded 256-tone OFDM systems over memoryless Rayleigh fading channel under IBO of 6dB.

The p-th decoder is selected if β_p is not less than any β_q for any $q \neq p$. Only the p-th decoder needs to continue its decoding until the I-th iteration. The hard decision on $\Pi_p^{-1}[\tilde{L}_p^I(d_0), \cdots, \tilde{L}_p^I(d_{K-1})]$ is the decoded output. For scheme II, we employ the redundancy in the turbo codeword for both error protection and discriminating the right candidate from others. Hence, attaching additional side information to the transmitted turbo codeword is not needed.

IV. Simulation Results and Performance Analysis

We consider the 256-subcarrier turbo coded OFDM using 16QAM modulation with Gray mapping. Perfect synchronization is assumed. We use a rate 1/2 binary turbo code of length 1024. The turbo encoder is formed by two 4-state recursive systematic convolutional (RSC) codes. Each rate 2/3 RSC code is obtained from regularly puncturing the rate 1/2 RSC code using generators $(1, \frac{1+X^2}{1+X+X^2})$. In the simulation, we use only one pre-iteration, i.e, $S = 1$ and the number of complete decoding iterations is $I = 7$. The number of candidates is $Q = 16$. For the transmitter, we model the power amplifier as a soft limiter [13] with input power backoff (IBO) given by 6 dB. The oversampling factor is set to 4. Simulation results are obtained for schemes I and II, the ideal turbo coded OFDM (i.e., no PAPR reduction is performed and ideal power amplifier is used) and the simple turbo-coded OFDM (i.e., Q=1, no PAPR reduction is performed and a power amplifier with the given IBO is used).

A. PAPR Distribution

We see from Fig. 3 that the curves of CCDF for schemes I and II are very close to the curve obtained by (4).

B. Complexity

In the transmitter of each of schemes I and II, Q candidates need to be generated. The generation of each candidate requires an IFFT operation. Hence, a total of Q IFFT operations are needed. For the receiver, the price to pay for eliminating the need of side information for scheme II is the increased decoding complexity as compared to scheme I. For scheme II, Q turbo decoders corresponding to Q candidates are needed, while scheme I needs only one decoder. Note that S iterations are needed for all the Q turbo decoders, while only one turbo decoder needs to complete I iterations. Hence, the complexity of the receiver in scheme II is about $[(Q-1)S+I]/I$ times of scheme I. Since we set $S = 1, I = 7$, and $Q = 16$, the decoding complexity of scheme II is about 3.14 times of that of scheme I.

C. Error Performance

The error performances of turbo coded OFDM systems over the AWGN channel and the memoryless Rayleigh fading channel [18] are given by the bit-error-rate (BER) curves in Fig. 4 and Fig. 5 respectively, where the IBO of power amplifier is 6 dB. We see that scheme II has error performance better than scheme I. The inferior error performance of scheme I is due to the serious error burst which results from the incorrectly decoded side information. Note that the decoding results of scheme II are close to those of the ideal turbo coded

OFDM. In the following, we will give a heuristic explanation about the error performance for scheme II.

Let C_q denote the turbo code associated to the q-th turbo encoder, for $1 \leq q \leq Q$. The weight spectrum of C_q can be represented by $\{N_q(w) : 0 \leq w \leq mN\}$, where $N_q(w)$ is the number of weight-w codewords in C_q. Suppose that $\bar{u}$ is transmitted through an interleaver Π_i. The associated turbo codeword is $\bar{v}_i(\bar{u})$ = $[\Pi_i(\bar{u}), P_a(\Pi_i(\bar{u}), P_b(\Pi_i(\Pi_i(\bar{u})))]$, where $P_a(\Pi_i(\bar{u}))$ is the parity sequence obtained from RSC 1, $P_b(\Pi_i(\Pi_i(\bar{u})))$ is the parity sequence obtained from RSC 2. We now check the Hamming distance, $D(\bar{v}_i(\bar{u}), \bar{v}_j(\bar{u}'))$, between $\bar{v}_i(\bar{u})$ and $\bar{v}_j(\bar{u}')$ = $[\Pi_j(\bar{u}'), P_a(\Pi_j(\bar{u}')), P_b(\Pi_j(\Pi_j(\bar{u}')))]$, where $\bar{u} \neq \bar{u}'$ and $i \neq j$. For simplicity, we may replace $\Pi_i(\bar{u}_i)$ by $\bar{u}_i$ and replace $\Pi_j(\bar{u}_j)$ by $\bar{u}_j$ since we consider arbitrary message sequences. Hence, we consider $\bar{v}_i(\bar{u}) = [\bar{u}, P_a(\bar{u}), P_b(\Pi_i(\bar{u}))]$ and $\bar{v}_j(\bar{u}') = [\bar{u}', P_a(\bar{u}'), P_b(\Pi_j(\bar{u}'))]$. We have

$$\begin{aligned} D(\bar{v}_i(\bar{u}), \bar{v}_j(\bar{u}')) &= w(\bar{u} \oplus \bar{u}') + w(P_a(\bar{u} \oplus \bar{u}')) \\ &+ w(P_b(\Pi_i(\bar{u}) \oplus \Pi_j(\bar{u}'))) \end{aligned} \quad (8)$$

where $w(\bar{x})$ is the Hamming weight of $\bar{x}$. We are interested in the distance spectrum between $\bar{v}_i(\bar{u})$ and C_j, represented by $\{N_{i,j,\bar{u}}(D) : 0 \leq D \leq mN\}$, where $N_{i,j,\bar{u}}(D)$ is the number of codewords in C_j at a distance D from $\bar{v}_i(\bar{u})$. Consider the following cases for evaluating $D(\bar{v}_i(\bar{u}), \bar{v}_j(\bar{u}'))$.

Case 1: Suppose that $\bar{u} = \bar{0}$, where $\bar{0}$ is the all zero vector. We have $D(\bar{v}_i(\bar{u}), \bar{v}_j(\bar{u}')) = w(\bar{u}') + w(P_a(\bar{u}')) + w(P_b(\Pi_j(\bar{u}')))$. We see that $N_{i,j,\bar{u}}(D) = N_j(w)$ for $0 \leq D = w \leq mN$. Assume the weight spectrum of C_i is similar to that of C_j for $1 \leq i, j \leq Q$.

Then, for scheme II, the error rate for $\bar{u} = \bar{0}$ will be the sum of probabilities of decoding $\bar{A}'$ into nonzero codewords in C_1, C_2, $\cdots, C_Q$ respectively, which will be increased as Q increases. Through simulation, we find that the error rate for $\bar{u} = \bar{0}$ is indeed roughly proportional to Q.

Case 2 : Suppose that the weight of $\bar{u}$ is around $\frac{K}{2}$. In (8), $w(\bar{u} \oplus \bar{u}') + w(P_a(\bar{u} \oplus \bar{u}'))$ is the weight associated to RSC1. The numbers of codewords in RSC 1 with weight w is denoted $N_{RSC1}(w)$. The weight spectrum of RSC 1 starting from a given state and a given time t, denoted, $N_{RSC1,t}(w)$, can be obtained from [19], where it is found that its free distance is 3 and $N_{RSC1,t}(w)$ are 1, 4, 14, 40, 116, 339, 991, 2897, 8468, 24752 for w = 3, 4, 5, 6, 7, 8, 9, 10, 11, 12 respectively. Since for RSC 1 of our turbo code, we restrict t to be in $\{1, 2, \cdots, K = 512\}$, then we have $N_{RSC1}(w) \leq K \times N_{RSC1,t}(w)$.

Assume Π_i and Π_j to be independent and uniform interleavers [20]. Then, $\Pi_i(\bar{u}) \oplus \Pi_j(\bar{u}')$ can be modelled as a fair binary independently and identically distributed (i.i.d.) random sequence. For a large K, it is also extremely likely that $w(\Pi_i(\bar{u}) \oplus \Pi_j(\bar{u}'))$ is close to $\frac{K}{2}$. With this assumption, the associated state sequence will be random and the output parity bit of the rate 2/3 RSC encoder for each state will be either 0 or 1 with equal probability, i.e., 0.5. Hence, the weight of $P_b(\Pi_i(\bar{u}) \oplus \Pi_j(\bar{u}'))$, i.e., $w(P_b(\Pi_i(\bar{u}) \oplus \Pi_j(\bar{u}')))$, occurs with probability

$$\begin{aligned} P(w(P_b(\Pi_i(\bar{u}) \oplus \Pi_j(\bar{u}'))) = x) &= C_x^{\frac{K}{2}} (\frac{1}{2})^x (\frac{1}{2})^{\frac{K}{2}-x} \\ &= C_x^{\frac{K}{2}} (\frac{1}{2})^{\frac{K}{2}} \end{aligned} \quad (9)$$

It can be shown that $C_x^{\frac{K}{2}} \leq 2^{(K/2)H(\frac{x}{K/2})}$, where $H(p) = -plog_2 p - (1-p) log_2(1-p)$. For K = 512, the probability of RSC 2 with small $w(P_b(\Pi_i(\bar{u}) \oplus \Pi_j(\bar{u}')))$ is extremely small. Let $x = 32$ in (9). We have $P(w(P_b(\Pi_i(\bar{u}) \oplus \Pi_j(\bar{u}'))) = 32)$ = $C_{32}^{256} (\frac{1}{2})^{256} \leq 2^{256H(32/256)} 2^{-256} \approx 2^{-116.85} \approx 6.7 \times 10^{-36}$. The weight spectrum $\{N_{i,j,\bar{u}}(D) : 0 \leq D \leq mN\}$ can be obtained by

$$N_{i,j,\bar{u}}(D) = \sum_w N_{RSC1}(w) \times C_{D-w}^{\frac{K}{2}} (\frac{1}{2})^{\frac{K}{2}} \quad (10)$$

For $D = 35$, we have

$$\begin{aligned} N_{i,j,\bar{u}}(D) &< 1 \times 512 \times 2^{256H(32/256)} 2^{-256} \\ &+ 4 \times 512 \times 2^{256H(31/256)} 2^{-256} \\ &+ \cdots + 24752 \times 512 \times 2^{256H(23/256)} 2^{-256} \\ &+ \cdots + N_{RSC1,t}(35) \times 512 \times 2^{-256}. \end{aligned} \quad (11)$$

From [19], we only have $N_{RSC1,t}(w)$ for $3 \leq w \leq 12$. Note that $N_{RSC1,t}(w)/N_{RSC1,t}(w-1) \approx 2.923$ for $w = 8, 9, 10, 11, 12$. Hence, we make a reasonable conjecture that $N_{RSC1,t}(w)/N_{RSC1,t}(w-1) \leq 4$. With this conjecture, we can check that the largest term on the right-hand side of (11) is the first term. Hence, we have $N_{i,j,\bar{u}}(35) \leq 32 \times 512 \times 2^{-116}$ $< 2^{-102}$. We now see that the spectrum $\{N_{i,j,\bar{u}}(D)\}$ is extremely thin for D up to 35. If maximum likelihood decoding and interleavers which are uniform and independent are used, the probability of erroneously decoding an error-corrupted turbo codeword associated to a candidate into a turbo codeword associated to another candidate is extremely small for high signal-to-noise-ratio (SNR) conditions. Since we use the suboptimal iterative decoding and the interleavers may not be perfectly uniform and independent, this assertion may not necessarily be true. However, as indicated in Fig. 6, the simulation for scheme II is consistent with the assertion, where the error rates for $Q = 1$ and $Q = 16$ are close if E_b/N_0 is greater than 4 dB; and the error rates for $Q = 1, 16, 64, 128$, and $Q = 256$ are close if E_b/N_0 is greater than 5 dB.

If we consider the message sequence $\bar{u}$ to be a fair binary random sequence, the typical sequences which have weight close to $K/2$ will occur with probability close to 1, while the nontypical sequences $\bar{u}$ will appear with extremely small probability. In practical applications, zero message sequence or some message sequences with weight quite different from $\frac{K}{2}$ may frequently occur. To avoid such cases, we can add a fixed random sequence, such as the maximum length sequence to the message sequences in the transmitter.

V. A Variation of The Proposed Scheme

We can modify scheme II by deleting the interleaver at the input of each turbo encoder. That means we use $\bar{u}$ at the input of the q-th turbo encoder instead of $\Pi_q(\bar{u})$ = t_q. We denote this modified scheme as scheme II-A. At the receiver side of scheme II-A, for the first iteration, the decoding of RSC1 for all the Q turbo decoders are the same, while the decoding of RSC2 for all the Q turbo decoders are distinct. Hence, for S = 1, Q = 16 and I = 7, the decoding complexity of scheme II-A

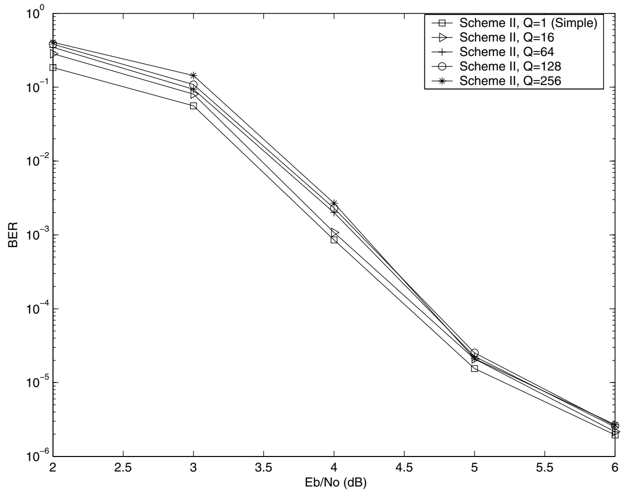

Fig. 6. BER for rate 1/2 16QAM Turbo-coded 256-tone OFDM systems (Scheme II) over AWGN channel under IBO of 6dB for various candidate numbers.

is only $[1+Q+2(I-1)]/2I$ = 2.07 times of that of scheme I. As indicated in Fig. 3, Fig. 4 and Fig. 5, the capability of PAPR reduction for scheme II-A is slightly inferior to that of scheme II, while the error rate is similar.

VI. Conclusions

We propose a side-information free SLM type turbo coded OFDM scheme for PAPR reduction. The reason that side information can be waived is due to the powerful discriminating capability of turbo decoding against the incorrect interleaver. Compared to the turbo coded OFDM using side information, the side-information free scheme has similar PAPR reduction capability and better error rates, while requires higher decoding complexity for the receiver. Like the side-information free SLM in [16], the proposed side-information free SLM needs Q decoders at the receiving side, where Q is the number of candidates. For the receiver, if channel coding is used, the ratio of the complexity of side-information free SLM to the complexity of SLM using side information is about Q to 1. However, the nature of iterative decoding for the turbo decoder allows us to reduce the decoding complexity. We show an example of the proposed scheme with $Q=16$ for which the ratio can be reduced to 3.14 to 1 without performance loss. We also show a variation of the proposed SLM for which the ratio is reduced to only about two to 1.

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