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Spectral Efficiency Maximization for Deliberate Clipping-Based Multicarrier Faster-Than-Nyquist Signaling

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ABSTRACT High peak-to-average power ratio (PAPR) is one of the common drawbacks of multicarrier signals at the transmitter. In this paper, we investigate the PAPR reduction of spectral efficient multicarrier faster-than-Nyquist (MFTN) signals, where the PAPR problem has still not been seriously considered yet in current literatures. Specifically, the low complexity deliberate clipping scheme is exploited in MFTN signaling system. In order to evaluate the practical system performance, we consider a memoryless high power amplifier PAPR-efficiency model, and the auxiliary channel model under low complexity symbol-by-symbol receiver is also derived. Moreover, the achievable spectral efficiency (ASE) which can be viewed as the low bound for any modulation and coding schemes is taken as a figure of merit throughout this paper. By jointly optimizing the time-frequency spacing and clipping ratio to maximization the ASE under the given shaping pulse and modulation format, we show that the ASE of clipping-based MFTN substantially outperforms conventional Nyquist signaling schemes.

INDEX TERMS Multicarrier transmission, faster-than-Nyquist, peak-to-average power ratio, clipping, spectral efficiency.

I. INTRODUCTION

In this bandwidth scarce world, the interest in spectral efficient transmission schemes is growing rapidly [1]–[4]. Instead of simply increasing the modulation level, an alternative scheme is to break the classical Nyquist transmission theory, which derives the faster-than-Nyquist (FTN) signaling technique [5], [6]. The multicarrier FTN (MFTN) is one of the members of this family [7]–[9]. The MFTN mainly relies on the time interval packing of adjacent symbols/pulses and/or frequency spacing packing between adjacent subcarriers for conventional Nyquist signaling systems. Although the intersymbol interference (ISI) and intercarrier interference (ICI) are introduced intentionally, it was shown that the spectral efficiency could be substantially improved than the ordinary modulation schemes at the same bit energy and error rate [8]. Due to this attractive advantage, MFTN is considered to be one of the candidate waveforms for 5G mobile communication systems, and it also attracts wide interests in other areas including satellite communication as well as optical communication systems [10]–[13].

Due to the ISI and ICI introduced by time-frequency packing in MFTN signaling system, current attentions about MFTN are mainly focused on the Euclidean distance (or Mazo limit) and practical signal detection schemes. In the pioneering works [7], the minimum Euclidean distance of MFTN signals under different time-frequency spacing was investigated, and it showed that the Mazo limit of MFTN is better than conventional single carrier FTN signaling system. In [14], the lattice staggering technique has been introduced into MFTN signaling system. By exploiting the optimal hexagonal lattice over ISI and ICI channels, the Mazo limit could be further improved than conventional rectangular MFTN signaling scheme. Moreover, the *maximum a posterior* (MAP) equalization combined with successive interference cancellation (SIC) has been presented in [8], and the ISI and ICI-free bit-error-rate (BER) performance could be approximately approached by the proposed scheme. Due to the complexity of MAP equalization is exponentially increased with the length of the channel memory as well as the modulation level, a low complexity

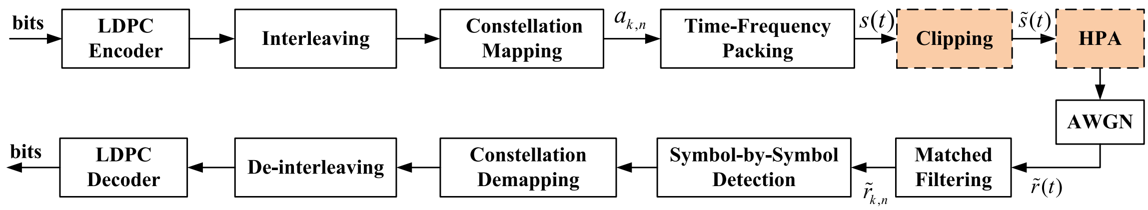

FIGURE 1. Block diagram of the clipping-based MFTN signaling system. In this figure, the brown blocks are where we mainly interested in this paper.

symbol-by-symbol receiver was proposed in [9], and it showed that the spectral efficiency of MFTN still substantially outperforms conventional Nyquist signaling systems. Due to this reason, the symbol-by-symbol receiver has been also widely employed in MFTN signaling system [1], [14].

On the other hand, it is known that the high peak-to-average power ratio (PAPR) is also one of the common drawbacks of multicarrier signals at the transmitter, and which results in the MFTN receiver's detection efficiency very sensitive to the nonlinear devices used in its signal processing loop, such as the digital-to-analog converter (DAC) and high power amplifier (HPA) [15]. Unfortunately, the PAPR problem of MFTN signals has still not been seriously considered yet in the literatures. In [16] and [17], the PAPR of single-carrier FTN has been investigated, and it was shown that the time packing will result in an increase of the envelope fluctuations. Furthermore, due to the time packing employed in MFTN signaling system, it was predicted in [8] that the MFTN signals will suffer from higher PAPR than conventional orthogonal frequency division multiplexing (OFDM) signals. However, no practical PAPR reduction scheme has been ever mentioned in it. Recently, a two-stage phase rotation factors optimized partial transmit sequence (PTS) scheme has been proposed for the PAPR reduction of MFTN signals in [18]. By making use of the discrete Fourier transform (DFT)-based implementation of the MFTN transmitter, it was shown that the PAPR reduction performance outperforms the direct application of conventional PTS scheme. However, the exhaustive search of the optimal phase rotation vector will result in extremely high computational complexity and the overhead of side information will also degrade the overall system throughput performance. In addition, although lots of PAPR reduction schemes have been proposed for conventional Nyquist multicarrier transmission systems, such as OFDM or filter-bank multicarrier (FBMC), they may difficult to be employed in MFTN signaling system due to its special signal structure [19]–[21] (i.e., time-frequency packing and pulse shaping).[1]

[1]We remark that the MFTN signaling is essentially a nonorthogonal transmission scheme, and the shaping pulse employed in MFTN signaling system results in the adjacent symbols overlapped with each other. Moreover, the frequency packing also degrades the orthogonality between adjacent subcarriers. Hence, the conventional fast Fourier transform (FFT) algorithm and the corresponding PAPR reduction schemes could not be directly employed in MFTN signaling system.

In this paper, we investigate the PAPR reduction of MFTN signals, and the contributions of this paper can be organized as follows.

- We consider the low complexity PAPR reduction for MFTN signals, and as far as we known, similar works are still not enough in the literatures.
- The deliberate clipping scheme is exploited in MFTN signaling system. Moreover, in order to evaluate the practical system performance, we consider a memoryless high power amplifier (HPA) PAPR-efficiency model, and derive the auxiliary channel model based on the low complexity symbol-by-symbol receiver.
- We jointly optimize the time-frequency spacing and clipping ratio to maximize the achievable spectral efficiency for MFTN signaling system.
- We demonstrate that the clipping-based MFTN outperforms conventional Nyquist signaling systems in terms of PAPR and spectral efficiency.

The reminder of this paper is organized as follows. In Section II, the system model of MFTN signaling as well as the basic principle of PAPR is introduced briefly. The proposed scheme is formulated in Section III, and following which, the numerical results are presented in Section IV. Finally, the conclusion is summarized in Section V.

II. SYSTEM MODEL AND PAPR

The general system model of MFTN signaling is shown in Fig. 1. Let us consider a baseband MFTN signaling system with K subcarriers, where the base pulse $g(t)$ is regularly shifted in the time and frequency domains. We will also assume a perfect synchronization among the data streams, and the LDPC encoded information bits are converted to the constellation mapping block after the random interleaving, the transmitted symbols $\{a_{l,k}\}$, where l is the time index and k is the subcarrier index, belonging to a given zeros-mean $\mathcal{M}$-th order complex constellation and are independent and identity distributed (i.i.d). The baseband transmitted signal in the downlink is thus given by

$$s(t) = \sqrt{E_s} \sum_{l} \sum_{k} a_{l,k} g(t - lT_{\Delta}) e^{j2\pi k F_{\Delta} t} \tag{1}$$

where E_s is the mean power per symbol, and $T_{\Delta} = \tau T$, $F_{\Delta} = \upsilon F$ are the packed time interval and frequency spacing. Moreover, T is the Nyquist time interval and F is the minimum orthogonal subcarrier spacing, respectively. For the root raised cosine (RRC) pulse with roll-off factor β, we have $F = (1+\beta)/T$. τ is the time interval packing factor and υ is the frequency spacing packing factor, respectively. For MFTN signaling system, we have $\tau \cdot \upsilon < 1$. Without loss of generality, we will also assume $T = 1$ in the following parts of this paper.

Under the assumption of transmission over an additive white Gaussian noise (AWGN) channel, the complex envelope of the continuous-time received signal is

$$r(t) = s(t) + w(t) \tag{2}$$

where $w(t)$ is the complex circularly symmetric zero-mean white Gaussian process with power spectral density (PSD) N_0.

The observed sample at the discrete time lT_Δ on the k-th subcarrier is

$$r_{l,k} = \int_{-\infty}^{\infty} r(t) g(t - lT_\Delta) e^{-j2\pi k F_\Delta t} dt \tag{3}$$

By joining (1) and (3), we eventually yield that

$$r_{l,k} = \sqrt{E_s} a_{l,k} + \sqrt{E_s} \sum_{(m',n') \neq (0,0)} \psi_{m',n'} a_{m',n'} + n_{l,k} \tag{4}$$

where we used $m' = l - m$, $n' = k - n$, and $\psi_{m',n'} = A_g(m'T_\Delta, n'F_\Delta)$ is the ISI and ICI introduced by time-frequency packing, $A_g(\tau, \upsilon)$ is the ambiguity function defined as

$$A_g(\tau, \upsilon) \triangleq \int_{-\infty}^{\infty} g(t) g(t - \tau) e^{-j2\pi \upsilon t} dt \tag{5}$$

and $n_{l,k} = \int_{-\infty}^{\infty} w(t) g(t - lT_\Delta) e^{-j2\pi k F_\Delta t} dt$.

For low complexity detection of the MFTN signals, the symbol-by-symbol receiver as given by [1], [9], and [10] will be considered in this paper, and in this regard, the interference introduced by time-frequency packing in (4) could be modeled as the zero-mean Gaussian process with PSD E_I and independent of the additive noise. We remark that this approximation is exploited only by the receiver, while in the actual channel the interference is clearly generated as in (4). Based on this observation, the signal model utilized by the receiver, i.e., the so-called *auxiliary channel* model, is therefore

$$r_{l,k} = \sqrt{E_s} a_{l,k} + \eta_{l,k} \tag{6}$$

where $\eta_{l,k}$ is the equivalent channel interference noise with variance $N_I = E_I + N_0$, and

$$E_I = E_s \sum_{(m',n') \neq (0,0)} \left| A_g(m'T_\Delta, n'F_\Delta) \right|^2 \tag{7}$$

where the independence of the transmitted symbols has been employed.

In general, the PAPR is a convenient metric to measure the sensitivity of a transmission scheme having a non-constant envelope. Hence, similar as [18], we will focus on the PAPR in per MFTN symbol period $T_\Delta = \tau T$, and the PAPR in the p-th symbol interval can be defined as

$$PAPR_p \triangleq \frac{\max\limits_{p\tau T \leq t < (p+1)\tau T} |s(t)|^2}{E[|s(t)|^2]} \tag{8}$$

where $E\{\cdot\}$ denotes the mathematic expectation.

In general, the PAPR performance can be evaluated by the complementary cumulative distribution function (CCDF) and which is defined as the probability that the PAPR exceeds a given level $PAPR_0 > 0$, i.e.,

$$CCDF = \Pr(PAPR > PAPR_0) \tag{9}$$

III. CLIPPING AND SPECTRAL EFFICIENCY OPTIMIZATION

Assume that $s[n] = s(nT_c)$, where T_c is the system sampling period, then, the discrete-time MFTN signal can be expressed as

$$s[n] = \sqrt{E_s} \sum_l \sum_k a_{l,k} g[n - l\tau N] e^{j2\pi k \upsilon n / K} \tag{10}$$

where N is the Nyquist discrete time symbol interval.

In order to reduce the PAPR of MFTN signals, the sample $s[n]$ will be clipped to a predefined threshold by using the soft limiter. The clipped signal $\tilde{s}[n]$ is thus given by

$$\tilde{s}[n] = \begin{cases} Ae^{j\phi[n]}, & if\ |s[n]| > A \\ s[n], & otherwise \end{cases} \tag{11}$$

where $A > 0$ represents the predefined threshold of the clipper, and $\phi[n]$ is the phase of input signal $s[n]$. The clipping ratio γ is defined as

$$\gamma = \frac{A}{\sqrt{P_{av}}} \tag{12}$$

where $P_{av} = E\{|s[n]|^2\}$ is the average power of the signal before clipping.

Assume that $s[n]$ is Gaussian distributed,[2] according to the Bussgang's theorem, we can write the output of the clipper as [24]

$$\tilde{s}[n] = \alpha s[n] + d[n] \tag{13}$$

where the distortion term $d[n]$ is uncorrelated with $s[n]$ and the attenuation factor α can be calculated for the clipper as

$$\alpha = \frac{E[s^*[n]\tilde{s}[n]]}{E[|s[n]|^2]} = 1 - e^{-\gamma} + \frac{\sqrt{\pi}\gamma}{2} erfc(\gamma) \tag{14}$$

where $\mathrm{erfc}(\gamma) = 2/\sqrt{\pi} \int_\gamma^\infty e^{-x^2} dx$ is the error function.

Hence, the variance of the clipped noise $d[n]$ is

$$\sigma_d^2 = P_{av}(1 - e^{-\gamma^2} - \alpha^2) \tag{15}$$

[2]We point out that according to the central limit theory, when the number of subcarriers K is larger enough, the distribution of $s(t)$ will approach (even not very accurate) the Gaussian distribution, and this approximation has been also widely used in general multicarrier transmission systems. For example, more details can be found in [22] and [23].

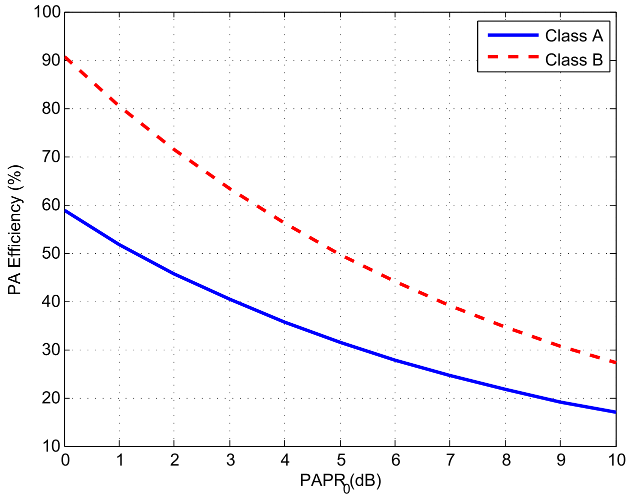

FIGURE 2. **PA efficiency with respect to the PAPR of input multicarrier signals.**

We point out that the deliberate clipping will result in the in-band distortion and out-of-band radiation, simultaneously. For the sake of simplicity, we will assume that the out-of-band radiation could be eliminated by the filter used at the transmitter front-end. Moreover, although the clipping operation will result in the signal distortion before passing through the HPA, it may also obtain much higher PA efficiency. Hence, in order to evaluate the practical system performance, we consider the theoretical efficiency upper limits for the classes A and B PA, and which are given by

$$\mu = G\exp(-\kappa \cdot PAPR_0) \tag{16}$$

where μ is the PA efficiency and depends on the PAPR $PAPR_0$ (in dB). For the clipped multicarrier system, we have $PAPR_0 = 10\log_{10}\gamma^2$. The values of G and κ can be found in [25]. As an example, Fig. 2 shows the PA efficiency curves with respect to the PAPR of input multicarrier signals. It is seen from this figure that for per 1dB PAPR reduction, the PA efficiency could be improved as much as 10%, and which further reveals the significance to reduce the PAPR of MFTN signals.

According to [25], the channel SNR depends on the PA efficiency as

$$SNR_c = \mu \cdot SNR_s = \frac{\mu \cdot \bar{P}_{av}}{N_0} \tag{17}$$

where $SNR_s = \bar{P}_{av}/N_0$ and $\bar{P}_{av} = E\{\tilde{s}[n]\tilde{s}^*[n]\}$ is the output average power of the clipper.

Substituting (16) into (17), the SNR_c can be rewritten as

$$SNR_c = \frac{G}{\gamma^{(20\kappa/\ln 10)}} \frac{\bar{P}_{av}}{N_0} \tag{18}$$

We assume that the clipped MFTN signals are all located in the linear region of HPA, i.e., no nonlinear distortion is introduced by HPA. Moreover, taking (13) into (2)-(4), and with small abuse of notations, we obtain the clipped observation sample

$$\tilde{r}_{l,k} = \alpha\sqrt{E_s}a_{l,k} + \alpha\sqrt{E_s}\sum_{(m',n')\neq(0,0)}\psi_{m',n'}a_{m',n'} + \tilde{d}_{l,k} + n_{l,k} \tag{19}$$

where $\tilde{d}_{l,k} = \int_{-\infty}^{\infty}\tilde{d}(t)g(t-lT_\Delta)e^{-j2\pi kF_\Delta t}dt$ is the clipping noise after matched-filtering at the receiver.

The auxiliary channel model (6) for clipped MFTN signaling system can be thus modified as

$$\tilde{r}_{l,k} = \alpha\sqrt{E_s}a_{l,k} + \tilde{\eta}_{l,k} \tag{20}$$

where $\tilde{\eta}_{l,k}$ is the equivalent channel noise with variance $\tilde{N}_I = \alpha^2 E_I + N_0 + \sigma_d^2$.

Then, we will turn to evaluate the system performance achievable by a symbol-by-symbol receiver designed for the auxiliary channel (20) when the actual channel is (19). Different from conventional system performance metrics such as PAPR or signal-to-distortion ratio (SDR) employed for Nyquist signaling systems [24], we pursue here a more general method for MFTN signaling system, i.e., the performance limit measured by achievable spectral efficiency (ASE),[3] and which is the amount of information transmitted per unit of time and per unit of bandwidth defined as

$$\eta \triangleq \frac{I(a_k; r_k)}{F_\Delta T_\Delta}\ bit/s/Hz \tag{21}$$

where a_k and r_k are the sequences of the transmitted and received samples for the k-th subcarrier. $I(a_k; r_k)$ is an achievable lower bound on the information rate measured in bit per channel use, and can be computed as

$$I(a_k; r_k) = \lim_{N_a\to\infty}\frac{1}{N_a}E\left\{\log\frac{p(r_k|a_k)}{\sum_{\tilde{a}}p(r_k|\tilde{a})}\right\} \tag{22}$$

where N_a is the length of the observation sequence. $p(r_k|a_k)$ is a Gaussian probability density function (pdf) of mean $\alpha\sqrt{E_s}a_k$ and variance $\tilde{N}_I$ in accordance with the auxiliary channel model (20). We point out that the time-frequency spacing T_Δ, F_Δ will have an impact on the ultimate PAPR distribution of MFTN signals (thus affecting the clipping noise) and also determines the amount of ISI and ICI introduced in MFTN signaling system (more details regarding T_Δ and F_Δ on the ASE can be found in [9] and [26]). On the other hand, the clipping ratio γ will also dominate the clipping noise as well as the PA efficiency (thus affecting the channel SNR). Hence, to determine the *ultimate* system performance, our target here is to maximize the ASE under the given shaping pulse and modulation format by jointly optimizing the time-frequency

[3]This can effectively avoid the trade-off between PAPR reduction and clipping distortion. Moreover, the ASE for a given transceiver scheme (modulation, shaping pulse and detector) shows the maximum spectral efficiency that can be achieved by any practical modulation and coding (MODCOD) scheme.

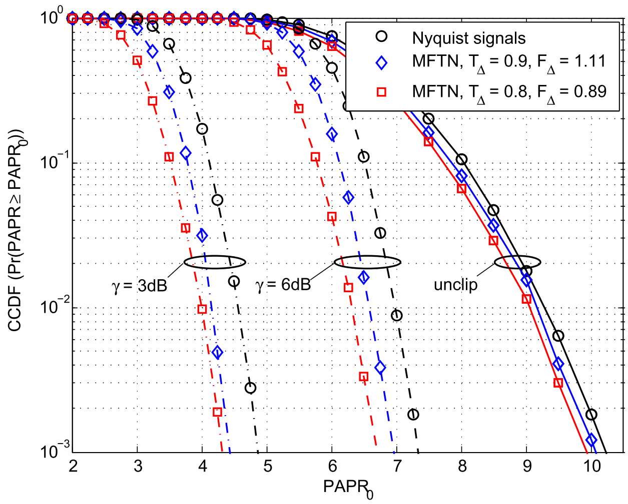

FIGURE 3. PAPR of MFTN and Nyquist signals under different time-frequency spacing and clipping ratios. In this figure, the modulation format is QPSK, and the shaping pulse is the RRC pulse with roll-off factor $\beta = 0.3$ and pulse length $L_g = 6T$.

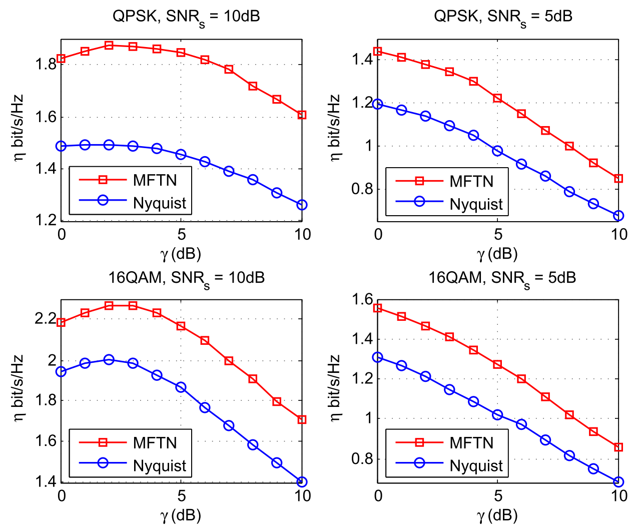

FIGURE 4. ASE as a function of clipping ratio γ for MFTN and Nyquist signals under different SNRs and clipping ratios.

spacing T_Δ, F_Δ and clipping ratio γ, which is given by

$$\begin{aligned} \eta_M &= \max \eta(T_\Delta, F_\Delta, \gamma) \\ &\quad \textit{subject to.}\ T_\Delta = \tau T, F_\Delta = \upsilon F, \\ &\quad\quad \tau \cdot \upsilon < 1, \\ &\quad\quad SNR_c = \frac{G}{\gamma^{(20\kappa/\ln 10)}} \frac{\bar{P}_{av}}{N_0} \end{aligned} \tag{23}$$

We point out that since the properties of the function η cannot be determined in closed form, the elegant theoretical derivation to solve (23) is an impractical tack. Hence, for the sake of simplicity, we will resort to the numerical analysis in next section.

IV. NUMERICAL RESULTS

In this section, we will focus on the numerical experiments to evaluate the system performance of clipping-based MFTN signaling system as shown in Fig. 1. In the simulations, the shaping pulse is fixed to the RRC pulse with roll-off factor $\beta = 0.3$ and pulse length $L_g = 6T$. Moreover, the number of subcarriers is $K = 60$, and the memoryless Class-B PA [25], i.e., $G = 90.7\%$ and $k = 0.1202$, is employed in the MFTN transmitter.

First, the PAPR distribution of MFTN and Nyquist signals under different time-frequency spacing T_Δ, F_Δ and clipping ratios γ is plotted in Fig. 3. In this figure, QPSK modulation is employed and the PAPR of MFTN signals under $T_\Delta = 0.9$, $F_\Delta = 1.11$ and $T_\Delta = 0.8, F_\Delta = 0.89$ as well as clipping ratios $\gamma = 3$dB and $\gamma = 6$dB has been considered. As a reference, the PAPR of original MFTN and Nyquist signals is also presented. It is seen from this figure that for the same modulation format, the MFTN could obtain lower PAPR than the Nyquist case. Moreover, we can also find that the PAPR could be efficiently reduced by the clipping scheme, and the time-frequency spacing as well as clipping ratio will jointly determine the ultimate PAPR reduction performance.

Then, we present the ASE as a function of the clipping ratio γ for MFTN and Nyquist signaling systems in Fig. 4, where the time-frequency spacing has been optimized for the unclipped case–the values of $T_\Delta = \tau T$ and $F_\Delta = \upsilon F$ are obtained by a coarse search followed by an interpolation of the obtained values (fine search) [9]. We remark that the spectral efficiency is also related to the SNR, even the dependence of the optimal spacing values on the SNR is usually limited. Hence, several SNRs, i.e., $SNR_s = 10dB, 5dB$, have been considered in the simulations. From this figure, we can see that for QPSK modulation and $SNR_s = 10$dB, $\gamma = 2$dB, the ASE of MFTN is about 0.35bit/s/Hz better than its counterpart, i.e., Nyquist signaling systems. While when $SNR_s = 5$dB, the optimal clipping ratio γ will be changed accordingly. Similar phenomenon can be also found for the case of 16QAM modulation. According to the simulation results presented in Fig. 4, we can conclude that in practical applications, the clipping ratio γ should be carefully chosen, and when the SNR is high enough, γ could be properly increased. However, when the SNR is small, γ should be reduced in order to obtain enough PA efficiency.

Based on Fig. 4, Fig. 5 shows the ASE as a function of SNR E_s/N_0 for MFTN and Nyquist signaling systems, where the clipping ratio γ have been further optimized (exhaustive search) for the selected cases. From this figure, we can find that for QPSK modulation, the spectral efficiency could be improved as much as 0.5bit/s/Hz than the Nyquist case. While for 16QAM modulation, there is still 0.2bit/s/Hz improvement can be achieved. The reason for this phenomenon is due to the high order modulations are more sensitive to the channel impairments and in this regard,

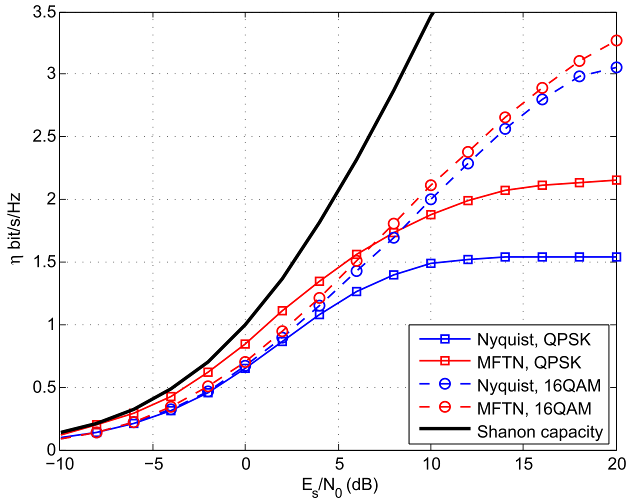

FIGURE 5. ASE of clipping-based MFTN and Nyqist signals. In this figure, the modulation formats are QPSK and 16QAM, and the time-frequency spacing as well as clipping ratio have been optimized for the selected cases.

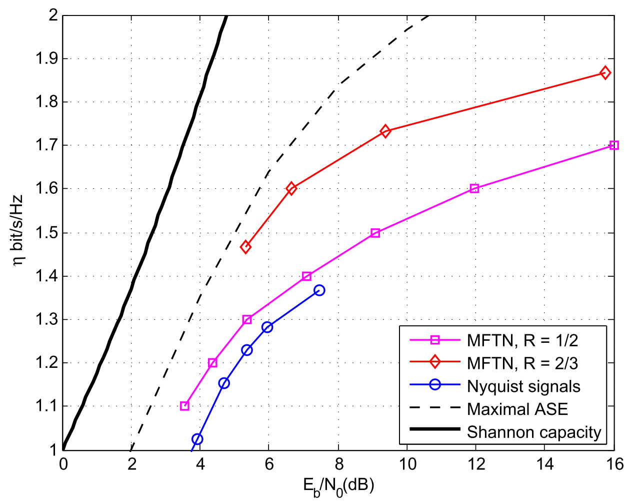

FIGURE 6. ASE versus required SNR for clipping-based MFTN and Nyquist signaling systems. In this figure, the BER is fixed to 10^{-5}, and the time-frequency spacing as well as clipping ratio have been optimized according to Fig. 4. Moreover, the modulation format is QPSK, and the LDPC codes from DVB-S2 with code length 64800 and 25 inner iterations have been employed.

the optimal time-frequency spacing will approach the Nyquist case.[4]

Finally, we consider the practical bit-error-rate (BER) performance of the aforementioned MFTN signaling system. In the simulations, the LDPC codes with rate $R = 1/2$ and $R = 2/3$ having code length 64800 bits of the DVB-S2 standard have been employed. Assuming a target BER of 10^{-5}, the required minimal E_b/N_0 for per given SE is shown in Fig. 6. As references, the Shannon capacity as well as the maximal ASE curves have been also presented. It can be observed that the information-theoretical results could be approached with only a slight loss of the SNR.[5]

V. CONCLUSION

In this paper, we investigated the PAPR reduction for the spectral efficient MFTN signals. For a low complexity system implementation, we exploited the deliberate clipping scheme, and the symbol-by-symbol receiver has been also considered. In order to evaluate the practical system performance, a PA PAPR-efficiency model was introduced and the achievable spectral efficiency has been taken as a figure of merit throughout this paper. We then proposed to jointly optimize the time-frequency spacing and clipping ratio to maximize the achievable spectral efficiency. Our numerical results validated that the MFTN signaling is not only a spectral efficient transmission scheme but also an energy efficient scheme than conventional Nyquist signaling systems.

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