SCIENCE CHINA
Information Sciences

• LETTER •

August 2020, Vol. 63 180305:1–180305:3
https://doi.org/10.1007/s11432-020-2936-2

Special Focus on mmWave and Terahertz Wireless Communications for B5G/6G

Hybrid prefix OFDM with spatial modulation toward terahertz broadband transmission

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Received 1 January 2020/Revised 1 March 2020/Accepted 1 June 2020/Published online 7 July 2020

Citation Wang T B, Jing W P, Song W L. Hybrid prefix OFDM with spatial modulation toward terahertz broadband transmission. Sci China Inf Sci, 2020, 63(8): 180305, https://doi.org/10.1007/s11432-020-2936-2

Dear editor,

In the development of 6G, terahertz communications [1] have attracted much attention owing to their potential for providing wireless broadband connections with even higher frequency resources. Furthermore, new transmission technologies have been widely considered for supporting future 6G systems [2]. Specifically, the spatial modulation (SM) technique has been suggested in [3] for striking a balanced trade-off between spectral efficiency (SE) and transmission performance compared with traditional multiple-input multiple-output systems. SM can be also combined with orthogonal frequency division multiplexing (OFDM) for combating multi-path fading [4,5] while making flexible use of the frequency resources on the terahertz band in 6G wireless systems [6].

Conversely, in OFDM systems, a cyclic prefix (CP) is typically employed to counteract the multipath interferences. The CP length is usually equal to or longer than the maximal channel delay spread; accordingly, the SE is significantly reduced. Recently, to balance the SE and the system's performance, hybrid CP transmission was developed in [7], which aims at reducing the length of CP by 50%.

Against this background, hybrid prefix (HP) technology is employed in the SM-OFDM system to reduce the CP length, in the context of both low and high frequency transmissions.

The SM-HP-OFDM communications system considered is equipped with N_t transmitting antenna, N_r receiving antennae, N_u activated transmitting antennae, N subcarriers and M frames over a Rayleigh fading channel in each time slot. Thus, in the SM structure, there are $H = 2^{\lfloor \log_2(C_{N_t}^{N_u}) \rfloor}$ effective transmitting antenna combinations (TACs) employed to select the activated transmitting antenna, where $\lfloor \cdot \rfloor$ and C_a^b denote the floor function and the binomial coefficient, respectively. Thus $\log_2 H$ information bits are modulated by selecting the TAC index $\boldsymbol{I}_h$ $(h \in (1, H))$, while $N_u \log_2(Q)$ bits are modulated into N_u constellation symbols, where Q denotes the modulation order. Therefore, $b = N_u \log_2(Q) + \log_2 H = b_1 + b_2$ bits are assigned to modulate the SM symbols. According to the above SM structure, the SM symbol $\boldsymbol{t}_n^m \in \mathbb{C}^{N_t \times 1}$, which denotes the symbol on the n-th $(n \in \{1, 2, \ldots, N\})$ subcarrier and the m-th $(m \in \{1, 2, \ldots, M\})$ frame can be expressed as

$$\boldsymbol{t}_n^m = [\ldots, 0, s_{h_1}, 0, \ldots, s_{h_2}, \ldots, 0, s_{h_{N_u}}, 0, \ldots]^{\mathrm{T}} \in \mathbb{C}^{N_t \times 1}, \tag{1}$$

where $(h_1, \ldots, h_{N_u}) = \boldsymbol{I}_h$ and $s_{h_1}, s_{h_2}, s_{h_{N_u}}$ represent the activated transmitting antenna indices and the corresponding constellation symbols, respectively. For the sake of simplicity, the SM symbols $\boldsymbol{t}_n^m$ on the n-th subcarrier and the m-th frame are expressed as $\boldsymbol{t}_n^m = [t_{n,1}^m, t_{n,2}^m, \ldots, t_{n,N_t}^m]^{\mathrm{T}}$. Then the vector $\boldsymbol{X}_i^m \in \mathbb{C}^{N \times 1}$, which denotes the SM symbols on the i-th transmitting antenna and the m-th frame, is obtained by restoring the SM transmitting symbols $\boldsymbol{t}_n^m$ according to all the subcarriers

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as follows:

$$\boldsymbol{X}_i^m = \left[t_{1,i}^m,\ t_{2,i}^m,\ \ldots,\ t_{N,i}^m\right]^{\mathrm{T}} \in \mathbb{C}^{N\times 1}, \tag{2}$$

where $i \in \{1, 2, \ldots, N_t\}$.

Then the time-domain (TD) symbol $s_{n,i}^m$, which denotes the symbol on the i-th transmitting antenna, the n-th subcarrier and the m-th frame, is obtained by the inverse fast Fourier transform of $\boldsymbol{X}_i^m$. Thus, the TD symbols $\boldsymbol{s}_i^m$, which denote the symbols on the i-th transmitting antenna and the m-th frame, can be expressed as $\boldsymbol{s}_i^m = [s_{1,i}^m,\ s_{2,i}^m,\ \ldots,\ s_{N,i}^m]^{\mathrm{T}} \in \mathbb{C}^{N\times 1}$. Finally, the CP is added discontinuously to the TD symbols $\boldsymbol{s}_i^m$ for transmission as follows. The TD symbols $\boldsymbol{s}_i^m$ are divided into two parts according to the frame index m, in which even indices are regarded as $2m$ and odd ones are expressed as $2m-1$ (m is an integer), i.e., when the frame indices are even, the CP is added to the even frame TD symbols $\boldsymbol{s}_i^m$ according to the design of the HP-OFDM [7]. Otherwise, CP is saved at the odd frame.

After the TD symbols undergo the wireless channel, the symbols are obtained by the receiver. It is worth noting that with the aid of CP, the even frame OFDM symbols are still capable of achieving identical performance to the conventional OFDM. However, for the odd frame OFDM symbols, the absence of CP breaks the orthogonality of subcarriers and introduces the inter-symbol interference (ISI). Then the ISI can be constructed by the convolution of the even frame OFDM symbols and the channel state information

According to the mathematical model of HP-OFDM [7], the union-bound technique could be employed to analyse the bit-error rate (BER) performance in the context of general Rayleigh fading. The derivation is divided into two parts according to the frames of the SM-HP-OFDM systems in the following.

There are 2^b types of possible SM transmitting symbols $\boldsymbol{t}^{2m}$. If it is assumed that the true and estimated frequency domain transmitted signals are $\boldsymbol{X}_j^{2m}\left(j \in \{1, 2, \ldots, 2^b\}\right)$ and $\boldsymbol{X}_v^{2m}\left(v \in \{1, 2, \ldots, 2^b\}\right)$, respectively, the Euclidean distance metrics between $\boldsymbol{X}_j^{2m}$ and $\boldsymbol{X}_v^{2m}$ are defined as

$$E = ||\boldsymbol{H}^{2m}(\boldsymbol{X}_j^{2m} - \boldsymbol{X}_v^{2m})||_F^2. \tag{3}$$

A variable $\delta = \frac{E}{2\sigma^2}$ is defined for the sake of simplicity, where $H_{N_r,1}^{2m}$ denotes the channel impulse response on the $2m$-th frame. According to the theory of the SM system [8], the pairwise error probability (PEP) is expressed as

$$P_e\left(\boldsymbol{X}_j^{2m} \to \boldsymbol{X}_v^{2m}\right) = Q(\sqrt{\delta}), \tag{4}$$

where σ^2 denotes the variance of the AWGN $\boldsymbol{W}^{2m}$, and $Q(x)$ denotes the Q-function, as $Q(x) = (1/\sqrt{2\pi})\int_x^\infty \mathrm{e}^{-t^2/2}\mathrm{d}x$.

On the one hand, $\boldsymbol{H}^{2m}(\boldsymbol{X}_j^{2m} - \boldsymbol{X}_v^{2m})$ is a variable when $N_r = 1$ [8]. Therefore, the average PEP with $N_r = 1$ is given by

$$\bar{P}_e\left(\boldsymbol{X}_j^{2m} \to \boldsymbol{X}_v^{2m} | N_r = 1\right) = \frac{1}{2}\left(1 - \sqrt{\frac{\bar{\delta}}{1 + \bar{\delta}/2}}\right), \tag{5}$$

where $\bar{\delta}$ denotes the mean value of δ with $N_r = 1$.

On the other hand, $\boldsymbol{H}^{2m}(\boldsymbol{X}_j^{2m} - \boldsymbol{X}_v^{2m})$ is a vector when $N_r > 1$. The PEP with $N_r > 1$ can be expressed as

$$P_e\left(\boldsymbol{X}_j^{2m} \to \boldsymbol{X}_v^{2m} | N_r > 1\right) = Q\left(\sqrt{\sum_{n_r=1}^{N_r} \delta_{n_r}}\right). \tag{6}$$

In this situation, it is noted that the argument of the above Q-function can be represented as the summation of $2N_r$ squared Gaussian random variables according to [8], with zero mean and variance of 1. Thus the average PEP with $N_r > 1$ can be expressed as

$$\begin{aligned}&\bar{P}_e\left(\boldsymbol{X}_j^{2m} \to \boldsymbol{X}_v^{2m} | N_r > 1\right)\\ &= \left[\Delta(\bar{\delta})\right]^{N_r} \sum_{n_r=0}^{N_r-1} \begin{pmatrix} N_r - 1 + n_r \\ n_r \end{pmatrix} \left[1 - \Delta(\bar{\delta})\right]^{n_r},\end{aligned} \tag{7}$$

where $\Delta(\bar{\delta}) = \frac{1}{2}(1 - \sqrt{\frac{\bar{\delta}}{1+\bar{\delta}/2}})$ and $\bar{\delta}$ denotes the mean value of δ with $N_r = 1$.

Finally, from all the possible transmitting symbols, the average bit error probability (ABEP) for even frame symbols can be expressed as

$$\mathrm{ABEP}_{\mathrm{even}} = \sum_{\boldsymbol{X}_j^{2m}\in\Omega} \sum_{\boldsymbol{X}_v^{2m}\in\Omega} \frac{d_{j,v}\bar{P}_e\left(\boldsymbol{X}_j^{2m} \to \boldsymbol{X}_v^{2m}\right)}{b2^b}, \tag{8}$$

where $d_{j,v}$ is the number of bit errors corresponding to $\boldsymbol{X}_j^{2m}$ and $\boldsymbol{X}_v^{2m}$, Ω is the set of all the possible SM symbols. And $\bar{P}_e(\boldsymbol{X}_j^{2m} \to \boldsymbol{X}_v^{2m})$ is obtained by (5) and (7).

According to the HP-OFDM model [7], the interference Z_r^{2m-1} approximately follows a Gaussian distribution with a fixed mean and variance via both the central limit theorem and statistical characteristics of the channels, as follows:

$$E\left\{Z_r^{2m-1}\right\} = 0, \tag{9}$$

$$\begin{aligned}E\left\{\left|Z_r^{2m-1}\right|^2\right\} =& \frac{2N_u}{NN_t}\sum_{t=1}^{N_t}\sum_{n=0}^{L-2}\sum_{l=n+1}^{L-1} E\left\{\left|h_{r,t,l}^{2i-1}\right|^2\right\}\\ &+ \frac{L-1}{N}\sigma^2,\end{aligned} \tag{10}$$

where $h_{r,t,l}^{2m-1}$ represents the TD channel gain of path l from the rth transmitting antenna to the tth receiving antenna, and L denotes the number of distinguishable channel paths.

The interference and channel noise are uncorrelated; the average PEP with $N_r = 1$ of the true and estimated frequency domain transmitted signals $\boldsymbol{X}_j^{2m-1}$ and $\boldsymbol{X}_v^{2m-1}$ for the odd frame symbols are obtained using a similar process to that in (4)–(8).

Therefore, the formula of the overall BER is obtained as

$$\mathrm{ABEP} = \frac{1}{2}\left(\mathrm{ABEP}_{\mathrm{even}} + \mathrm{ABEP}_{\mathrm{odd}}\right). \tag{11}$$

According to [7], the interference caused by the symbols of the even frame $(2i-2)$ could be cancelled by reconstructing it. Thus the variance of Z_r^{2m-1} in (10) can be reduced as

$$E\left\{\left|Z_r^{2m-1}\right|^2\right\} = \frac{L-1}{N}\sigma^2, \tag{12}$$

i.e., in the above equation, only the interference from the even frame $(2i-2)$ exists. The variance of the interference is minimal because the length of the channel L is much shorter than the length of the carrier N. This means that the interference caused by the even frame $(2i-2)$ can be ignored; this has been confirmed via theoretical and simulated BER results of the recovered SM-HP-OFDM system.

Equipped with beamforming architectures, the above SM-HP-OFDM system is suitable for wireless communications in terahertz channels, where the length of the CP in the terahertz channel is much longer than that of the extended vehicular a model (EVA) channel. The parameters of the terahertz channel can be generated using Table 1 of [9]; therefore, the CP length at different bandwidths can be obtained. In the terahertz channel, the required length of the CP is at least 5 times compared with that in the EVA channel. As an SM-HP-OFDM system can save 50% of the CP length, it could be an efficient candidate for future terahertz band communications.

Moreover, Figure 1 presents the simulated BER results of the unrecovered and recovered SM-HP-OFDM systems with $N_u = 1$ utilizing a traditional ML detector. As expected, when the odd frame symbols are recovered by the decision feedback equalization of the even frame symbols [6], the simulated BER will also be effectively improved.

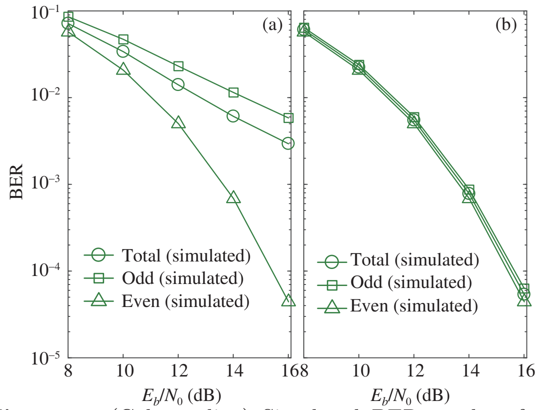

Figure 1 (Color online) Simulated BER results of the SM-HP-OFDM systems in QPSK and the terahertz channel, where antenna configuration is $N_t = 4$, $N_r = 2$ and $N_u = 1$. (a) Unrecovered, (b) recovered.

Therefore, a recovered SM-HP-OFDM system is generally suitable for supporting terahertz communications with improved SE.

Acknowledgements This work was supported by Fundamental Research Funds for the Central Universities (Grant No. 2572018BF16).

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