Shaped offset 8PSK and coded shaped offset 8PSK with improved spectrum efficiency for satellite communication

Yue Liu¹ | Ning Cao¹ | Minghe Mao¹ | Gang Li²

¹ School of Computer and Information, Hohai University, Nanjing, China
² Information Center, Ministry of Water Resources of the People’s Republic of China, Beijing, China

Abstract
The development of satellite communication (SATCOM) puts forward demanding requirements for the modulation methods’ spectral efficiency (SE) which is mainly subject to the pulse shaping and the utilisation of signal space. To date, there is rare consideration on the modification of high order phase shift keying (HOPSK) to gain higher SE in this literature. In this work, shaped offset 8PSK (SO8PSK) is developed which splits 8PSK bit transmission into three offset paths and then loads the shaping pulse onto them. The obtained modulation with continuous constrained phase and constant envelope is shown to increase SE compared to 8PSK by about 43% and is applicable for the nonlinear and band-limited SATCOM environment. To address the HOPSK inherent complexity, a simplified receiver is provided to reduce about two-third matched filters. Additionally, the iterative encoding of SO8PSK is presented and it is used as the inner code of serial concatenated coding (SCC). This modification method can be seen as a starting point to the better use of HOPSK for SATCOM.

1 INTRODUCTION
Satellite communication (SATCOM) is one of the key components of space segments. It can provide the most comprehensive wireless coverage, especially for low-density population areas, which can act as a complement of terrestrial cellular networks. Hence, the development of SATCOM should keep step with the quick development of wireless terrestrial communications to satisfy the requirement of 5G wireless communications [1–3].

With the growth in date rate and increasing demand for quality of service (QoS) in 5G systems, the issue of enhancing spectral efficiency (SE) is important in SATCOM. [4] In order to realise this goal, communication modulation, the key part of the physical layer, should be equipped with higher SE. [5, 6] Besides, SATCOM channel has two properties. First, to minimise the interference to adjacent channels, the signal needed to be bandwidth limited. Second, to maximise the use of the scarce power, the satellite high power amplifier (HPA) needs to be operated close to the saturation. [7, 8] These requirements motivate researchers to design various competitive modulation schemes.

At present, there are two main common schemes. On one hand, more spectrally compact shaping pulse functions such as rectangular, raised cosine and so on, are widely used in low order modulation. For example, minimum shift keying (MSK), a form of quadrature phase shift keying (QPSK) with half-sinusoidal pulse shaping, is treated as potential future global navigation satellite systems (GNSS) signal modulation schemes in [9]. Gaussian filtered minimum shift keying (GMSK) with a filter having Gaussian impulse response prior to frequency modulation of the carrier is provided as the signal option for SATCOM in [10]. Shaped offset quadrature phase shift keying (SOQPSK), a hybrid of OQPSK and MSK, has also attracted wide interest in this literature. Burst-mode synchronisation of SOQPSK for the next-generation aeronautical telemetry system is discussed in [11, 12] also uses SOQPSK to solve the two-antenna problem of aeronautical telemetry. Field programmable gate array (FPGA) implementation of SOQPSK for deep-space, telemetry and unmanned aerial vehicle (UAV) links is presented in [13]. Nevertheless, the information contained in the low order modulations is too limited to satisfy the 5G high rate transmission.
On the other hand, the SE can be improved using multi-level encoding to increase the data rate. For one thing, the transmission throughput can be reasonably increased by making full use of space geometry, that is, signal space dimensions. Quadrature-carrier amplitude PSK (APSK) is proposed in [8] using two orthogonal carriers to convey ordinary APSK to double the transmission rate for deep space SATCOM. 4 dimensional 8PSK trellis coded modulation (4D-8PSK-TCM) has been used for Earth Observation (EO) satellite mission. [14] Although the increment of dimensions always improves SE considerably, employing APSK or PSK may generate undulating band-limited channel of SATCOM. [15, 16]

For another, the most well-known and common used approaches to achieve higher SE in communication literature is to increase the order of modulation scheme. A wide range of modulations, including QPSK, 8PSK, 16APSK, 32APSK, 64APSK, 128APSK are defined in the frame of DVB-S2 and DVB-S2x. [17, 6] considers transmitting higher-order modulation signals in a satellite-based Internet of Things (IoT) to improve the SE. The process of FPGA on 8PSK for SATCOM is given in [18]. The performance characteristics of the inter SATCOM link with 8PSK and 16 quadrature amplitude modulation (QAM) along with Saleh, Ghorbani and Rapp amplifier models are described in [19]. Bandwidth-efficient multiple carriers with multiple phase shift keying (MPSK) modulation are illustrated in [20]. However, abrupt phase changes in these modulations result in the unwanted spectral side lobes. The part of high side lobes will be filtered off by band-limiting filters. This severely distorts the original signal and then the constant envelope will not be maintained. When the signal with non-constant envelope enters the nonlinear HPA, the filtered components are almost recovered again. These regenerated side lobes will be filtered off by the filter after HPA, which leads to loss of signal energy and then degrades the system performance.

Therefore, the modulations with both higher SE and robustness to the band-limited and nonlinear environment of SATCOM are needed to be developed. In this paper, we propose shaped offset 8PSK (SO8PSK) with the hope of sacrificing less spectrum resources and making 8PSK, a representation of HOPSK, be more competitive modulation in the SATCOM. First of all, to make it applicable to the nonlinear and band-limited channel, we constrain the phase shift via dividing 8PSK bit transmission into three parallel, offsetting the paths by 1 bit and pulse shaping them. The obtained SO8PSK has continuous constrained phase and constant envelope. Both of them are popular characteristics in the nonlinear and band-limited channel. Next, to deal with the complexity of HOPSK, we provide the simplified receiver based on waveform similarities to reduce about two-third needed matched filters at the cost of about 0.3 dB bit error rate (BER). Furthermore, we present the iterative encoding of SO8PSK and use it to be the inner code of serial concatenated coding (SCC). This modification method can be seen as a starting point to the better use of HOPSK for SATCOM.

In summary, our main contributions are:

- For the first time, the concept of SO8PSK is proposed and the modification on the HOPSK for SATCOM is conducted. By means of bits offset and pulse shaping, SO8PSK not only exhibits about 43% SE increment than 8PSK but also is more suitable for nonlinear and band-limited SATCOM relying on its constant envelope and continuous constrained phase.
- Given the common high complexity in HOPSK, the simplified receiver reducing the required number of matched filters from 32 to 10 based on signal similarities is proposed.
- The iterative encoding of SO8PSK is also presented and used to be the inner code of serial concatenated coding (SCC). In a word, these findings provide the encouragement for the more feasible implementation of HOPSK for SATCOM.

The rest of this paper is organised as follows. The signal model of SO8PSK and the eight-state bit-interval trellis diagram are given in Section 2. In Section 3, the time invariant symbol interval trellis representation of SO8PSK is described and the corresponding transmitter is proposed. Section 4 expresses SO8PSK as trellis coded modulation. The optimal and simplified receivers are given in Section 5. In Section 6, power spectral density (PSD) and BER performance of SO8PSK are simulated and compared to those of 8PSK. Coded SO8PSK with iterative demodulation is illustrated in Section 7. Finally, we conclude the full paper in Section 8.

### 2 PROPOSAL OF SO8PSK

In this section, we provide the signal model of SO8PSK and its bit-interval trellis diagram. According to the Gray code mapping of 8PSK, the π phase shift may happen if two bits in one state transition both change. Specifically, the current phase \( \pi \) (110) changes \( \pi \) from the forward phase state 0 (000), which will bring about the bandwidth expansion. To constrain the phase shift in each time interval, we divide time interval into three kinds \( n_1, n_2, \) and \( n_3 \), where \( n_1 = 3k + 1 \), \( n_2 = 3k + 2 \), \( n_3 = 3k + 3 \), \( k \in \mathbb{N} \). The successive input bits will be alternately assigned to these three intervals. To reduce the phase change, we stagger the three kinds of intervals by one bit duration \( T_b \), which is shown in Figure 1. Then, we can constrain the phase change within...
3π/4 through controlling two bits unchanged in one state transition. To further band limit, SO8PSK introduces rectangular pulse shaping into this offset 8PSK.

2.1 Signal model

We represent SO8PSK in the form of a continuous phase modulation (CPM) signal, 

\[ s(t) = \sqrt{\frac{3E_b}{T_b}} \cos(2\pi f_c t + \phi(s, ff) + \phi_c), \]  

(1)

where \( E_b \) is the energy of a bit; \( T_b \) is the duration of a bit; \( f_c \) is the carrier frequency; \( \phi \) is an arbitrary phase constant. \( \phi(s, \alpha) \) is the phase-modulation process, 

\[ \phi(t, \alpha) = 2\pi B \sum_{i=1}^{9} \alpha_i g(t - iT_b), \]  

(2)

where for SO8PSK, the modulation index \( b = 1/4 \), the phase pulse \( g(t) \) varies linearly with time over each bit interval. Frequency pulse \( g(t) = d(q(t))/dt \) is a rectangular frequency pulse in each bit interval, that is, \( g(t) = 1/2T_b \). The symbol alphabet \( \{ff\} \in \{0, \pm 1, \pm 3\} \) is related to the true input binary data sequence \( \{a_i\} \) by 

\[ \alpha_i = \begin{cases} 
-1 \cdot (a_{i-3} + a_{i-2} - 1) |a_i - a_{i-3}| \cdot (a_{i-1} + 1), & i \in n_1 \\
-1 \cdot (a_{i-3} + a_{i-1}) |a_i - a_{i-3}| \cdot (-2a_{i-2} + 3), & i \in n_2 \\
-1 \cdot (a_{i-3} + a_{i-2} + a_{i-1}) |a_i - a_{i-3}|, & i \in n_3 
\end{cases} \]  

(3)

Detailed explanation on how (3) can be obtained is described in the next subsection.

2.2 Bit-interval trellis diagram of SO8PSK

We now describe an eight-state bit-interval trellis diagram of SO8PSK based on the aforementioned three path transmission, which can be used to obtain (3) as well as to provide a means for demodulation of SO8PSK using viterbi algorithm (VA).

The three bits of a trellis state correspond to the current phase, which is presented by the current \( n_1 n_2 n_3 \) bits. Specifically, assuming a conventional Gray code mapping, the phase states \( \pi/8, 3\pi/8, 5\pi/8, 7\pi/8, 9\pi/8, 11\pi/8, 13\pi/8, 15\pi/8 \), are assigned the bit mappings (in the form of “\( n_1 n_2 n_3 \)”): 000, 001, 011, 010, 110, 111, 101, 100, respectively. The trellis is divided into three kinds of intervals, that is, \( n_1, n_2, n_3 \). The successive input bits \( a_i \) will be alternately assigned to these three kinds of intervals. Note that due to path offset, at most one bit (\( n_1, n_2 \) or \( n_3 \)) of the phase state can change during each state transition, that is, the phase change is constrained to be \( \pm \pi/4 \) rad, \( \pm 3\pi/4 \) rad or 0 in each interval. For example, the phase state 000 (\( \pi/8 \)) can only undergo a transition to phase states 100 (15\( \pi/8 \)), 010 (7\( \pi/8 \)), 001 (3\( \pi/8 \)) or remain in phase state 000 (\( \pi/8 \)) depending on whether the incoming bit is \( n_1, n_2 \) or \( n_3 \) bit, and whether its value is 1 or 0. Figure 2 is the eight-state bit-interval trellis diagram illustrating the transitions from state to state in accordance with the above. The leftmost three bits denote the initial phase state. Above the branches of the trellis are labelled with the value \( a_i \). The input bits \( a_i \) emerging from each trellis state are “00” and “11”, respectively. \( \alpha_i \) is the value that corresponds to the phase-state transition. Specifically, assuming \( n_2 = 5 \), the current phase state is 001 (3\( \pi/8 \)), that is, \( a_{n_2-1} = 0; a_{n_2-3} = 0; a_{n_2-2} = 1 \), the input bit \( a_{n_2} \) is 1. The phase state converts to 011 (5\( \pi/8 \)). From (2), in each interval, the phase changes \( \pi/\alpha_i \). Thus, in this condition, \( \alpha_i = 1 \). We summarise all the situations in Figure 2 and obtain Equation (3).

The obtained SO8PSK with constant envelope and continuous constrained phase is assumed to be robust to the nonlinear and band-limited channel. We expect that these characteristics can make SO8PSK yield high SE.

3 TRANSMITTER DESIGN

In this section, to design a transmitter for SO8PSK, we firstly describe the time invariant symbol interval trellis representation of SO8PSK based upon Figure 2.

3.1 Time invariant symbol interval trellis representation of SO8PSK

Without loss of generality, we assume that the first bit in each symbol interval is always \( n_1 \) bit. Based on Figure 2, we provide the extended trellis coding in a symbol interval of SO8PSK in Figure 3. Each transition lasts three-bit interval and is relevant to a group of precoder output \{\( \alpha_i \), \( \alpha_{i+1} \), \( \alpha_{i+2} \)\}. For a fixed group \{\( \alpha_i \), \( \alpha_{i+1} \), \( \alpha_{i+2} \)\}, there exists a pair of waveforms transmitted in I and Q channels: \( s_I(t) = \cos(2\pi f_c t + \phi(s, \alpha_i, \alpha_{i+1}, \alpha_{i+2}) + \phi_c) \) and \( s_Q(t) = \sin(2\pi f_c t + \phi(s, \alpha_i, \alpha_{i+1}, \alpha_{i+2}) + \phi_c) \). Here, \( \phi_c \) is the initial phase of each transition, and for each transition interval 

\[ \phi(t, \alpha_i, \alpha_{i+1}, \alpha_{i+2}) = \begin{cases} 
\frac{\pi \alpha_i}{4T_b}, & 0 \leq t \leq T_b \\
\frac{\pi \alpha_i}{4} + \frac{\pi \alpha_{i+1}}{4T_b} (t - T_b), & T_b \leq t \leq 2T_b \\
\frac{\pi \alpha_i}{4} + \frac{\pi \alpha_{i+1}}{4} + \frac{\pi \alpha_{i+2}}{4T_b} (t - 2T_b), & 2T_b \leq t \leq 3T_b.
\end{cases} \]  

(4)

Given 8 possible initial phase states \( \phi \), and 64 possible combinations of output pair \{\( \alpha_i \), \( \alpha_{i+1} \), \( \alpha_{i+2} \)\}, it is easy to conclude that there are 32 kinds of signal waveforms denoted as \( s_i(t) \), \( i = 0, 1, ..., 31 \) in channel I and there are 32 kinds of signal waveforms denoted as \( s_j(t) \), \( j = 0, 1, ..., 31 \) in channel Q. The 32 signal waveforms of I channel have the relationship \( s_{i+4}(t) = -s_i(t), i = 0, 1, ..., 15 \). The waveforms of the Q channel is
the sin function of the corresponding phase. The waveform expression is easy to be obtained through substituting \( (\alpha_i, \alpha_{i+1}, \alpha_{i+2}) \) and \( \phi_i \) of each branch (Figure 3) into (4) and (1). Considering the space constraints, we no longer list the detailed signals’ representations.

### 3.2 Transmitter design for SO8PSK

Based on Figure 3 and the labelling of I, Q waveforms, we can represent the indexes of the transmitted signal \( s_I(t) \) and \( s_Q(t) \) in each symbol interval using the corresponding three \( \alpha \) values and the initial phase. Specifically, in the symbol interval \( iT_b \leq t \leq (i + 3)T_b \) \( (i \in n_1) \), we have \( s_I(t) = s_{I_p}(t) \) and \( s_Q(t) = s'_{Q_p}(t) \). Noted here, if \( \alpha = 3 \) or \( -3 \), it will be converted to 1 and \( -1 \) in the computation, respectively. The index \( n \) and \( m \) are represented as,

\[
n = k_{I_1} * 2^4 + k_{I_2} * 2^3 + |\alpha_i| * 2^2 + |\alpha_{i+1}| * 2^1 + |\alpha_{i+2}| * 2^0,
\]

where

\[
k_{I_1,k_{I_2}} = \begin{cases} 
00 & \phi_i = \pm \pi / 8 \\
01 & \phi_i = \pm 3 \pi / 8 \\
11 & \phi_i = \pm 5 \pi / 8 \\
10 & \phi_i = \pm 7 \pi / 8.
\end{cases}
\]

And,

\[
m = k_{Q_1} * 2^4 + k_{Q_2} * 2^3 + |\alpha_i| * 2^2 + |\alpha_{i+1}| * 2^1 + |\alpha_{i+2}| * 2^0,
\]

where

\[
k_{Q_1,k_{Q_2}} = \begin{cases} 
00 & \phi_i = \pi / 8 \text{ or } 7 \pi / 8 \\
01 & \phi_i = 3 \pi / 8 \text{ or } 5 \pi / 8 \\
11 & \phi_i = -3 \pi / 8 \text{ or } -5 \pi / 8 \\
10 & \phi_i = -7 \pi / 8 \text{ or } -\pi / 8.
\end{cases}
\]
Therefore, the SO8PSK transmitter can be implemented with the precoder satisfying (3), a signal mapper for choosing I and Q signals based on the above computation, and a quadrature modulator. The block diagram of SO8PSK equivalent transmitter is shown in the Figure 4.

4 | INTERPRETATION OF SO8PSK AS TRELLIS CODED MODULATION

To conveniently propose simplified receiver and coded SO8PSK, in this section, we will show that the indexes of SO8PSK transmitted signal $s_I(t)$ and $s_Q(t)$ can be directly represented by the original binary inputs $\{0, 1\}$. Specifically, in each symbol interval $n$, we denote the $n_1, n_2, n_3$ input binary data as $D_{1,n}, D_{2,n}, D_{3,n}$. Note that the phase state in the $(n-1)$th symbol interval is simply “$D_{1,n-1}D_{2,n-1}D_{3,n-1}$” and it will become “$D_{1,n}D_{2,n}D_{3,n}$” in the $n$th symbol interval with the input data $D_{1,n}D_{2,n}D_{3,n}$. We replace $k_{I1}, k_{I2}, k_{Q1}$ and $k_{Q2}$ by $D_{1,n-1}, D_{2,n-1}, D_{3,n-1}$ in (6) and (8),

$$
\begin{align*}
        k_{I1} &= D_{2,n-1}, \\
        k_{I2} &= D_{3,n-1}, \\
        k_{Q1} &= D_{1,n-1}, \\
        k_{Q2} &= D_{3,n-1}. \\
\end{align*}
$$

The value of $\alpha$ reflects the change between the binary inputs of the adjacent same kind of intervals. Therefore, it is straightforward to express the indexes in terms of the input data. Ultimately, the indexes of output waveform pair $\{s_i(t), s_i'(t)\}$ in...
the Figure 3 can be represented by the binary inputs,
\[ i = I_4 \ast 2^4 + I_3 \ast 2^3 + I_2 \ast 2^2 + I_1 \ast 2^1 + I_0 \ast 2^0, \]
\[ j = Q_4 \ast 2^4 + Q_3 \ast 2^3 + Q_2 \ast 2^2 + Q_1 \ast 2^1 + Q_0 \ast 2^0, \quad (10) \]

where \( I_4 = D_{2,n-1}; \quad I_3 = D_{3,n-1}; \quad I_2 = D_{1,a} \oplus D_{1,a-1} = Q_2; \]
\( I_1 = D_{2,a} \oplus D_{2,a-1} = Q_1; \quad I_0 = D_{3,a} \oplus D_{3,a-1} = Q_0; \)
\( Q_4 = D_{1,a-1}; \quad Q_3 = D_{3,a-1}. \)

The implementation of SO8PSK based upon the above index mappings is illustrated in Figure 5. We can observe that SO8PSK can be decomposed into a eight-state trellis encoder and a memoryless signal mapper. This inherent eight-state trellis encoder has three binary inputs \( D_{1,a}, D_{2,a}, D_{3,a} \) and two waveform outputs \( s_j(t) \) and \( s_j'(t) \), where the trellis state is defined by the previous 3-bit \( D_{1,a-1}, D_{2,a-1}, D_{3,a-1} \). The trellis of this eight-state encoder is exactly the one shown in Figure 3. The decomposition of SO8PSK into trellis encoder and memoryless mapper allows SO8PSK to be used in coded systems with iterative decoding, which is shown in Section 7.

5 | RECEIVER DESIGN FOR SO8PSK

In this section, to decrease the complexity of HOPSK, a simplified receiver based on signal similarities is proposed. We assume that \( r(t) \) is the received signal, branch metrics for VA is \( (Z_j + Z_Q) \) and \( \pi = 2\pi f_c \).

5.1 | Optimal receiver

The optimal receiver adopting VA for maximum-likelihood sequence is shown in Figure 6. It contains 32 matched filters (I and Q need 16 matched filters, respectively) and a 8-state trellis demodulator. The demanding requirement for matched filters leads to the high complexity. For practical use, simplified receiver with less necessary filters is required to be considered.

5.2 | Simplified receiver

To reduce the complexity of optimal receiver and not seriously sacrifice the power efficiency, we propose a simplified receiver to group sets of waveforms based on their similarities. 32 waveforms in I channel are divided into four groups to demodulate \( a_i \) in \( n_1 \) interval and 32 waveforms in Q channel are divided into 16 groups to demodulate \( a_i \) in \( n_2 \) interval. For \( s_j(t) \), the first group consists of \( s_1(t), s_4(t), s_5(t), s_8(t), s_9(t), s_{12}(t), s_{13}(t) \) and the second group consists of \( s_2(t), s_3(t), s_6(t), s_7(t), s_{10}(t), s_{11}(t), s_{14}(t), s_{15}(t) \) while for \( s_Q(t) \), the \( j \)th group consists of \( s_{2j}(t), s_{2j+1}(t) \) when \( j = 0, 2, 4, 6 \) and \( s_{2j-1}(t), s_{2j-2}(t) \) when \( j = 1, 3, 5, 7 \). We can use average signal to replace \( s_j(t) \). For I channel, we define \( g_j(t) \) to denote the average value of \( s_j(t) \).

\[
q_0(t) = \frac{1}{8} [s_0(t) + s_1(t) + s_4(t) + s_5(t) + s_8(t) + s_9(t) + s_{12}(t) + s_{13}(t)],
\]
\[
q_1(t) = \frac{1}{8} [s_2(t) + s_3(t) + s_5(t) + s_6(t) + s_{10}(t) + s_{11}(t) + s_{14}(t) + s_{15}(t)],
\]
\[
q_2(t) = \frac{1}{8} [s_{16}(t) + s_{17}(t) + s_{20}(t) + s_{21}(t) + s_{24}(t) + s_{25}(t) + s_{28}(t) + s_{29}(t)],
\]
\[
q_3(t) = \frac{1}{8} [s_{18}(t) + s_{19}(t) + s_{22}(t) + s_{23}(t) + s_{26}(t) + s_{27}(t) + s_{30}(t) + s_{31}(t)].
\]
Considering the signal with opposite signs,

\[ q_2(t) = -q_0(t), \]
\[ q_3(t) = -q_1(t). \]  

We also use average signal to replace \( s_q(t) \) and define \( q'_i(t) \) as the average signal of \( s_q(t) \).

\[ q'_i(t) = \begin{cases} \frac{1}{2} (s'_{2i}(t) + s'_{2i+2}(t)), & i = 0, 2, 4, 6 \\ \frac{1}{2} (s'_{2i-1}(t) + s'_{2i+1}(t)), & i = 1, 3, 5, 7. \end{cases} \]  

Similarly, considering the signal with opposite signs,

\[ q'_{i+8}(t) = -q'_i(t), \]

where \( i = 0, 1, ..., 7 \). Since we have replaced waveforms by their average signals, we can drop \( I_0, I_2, I_3 \) and \( Q_1 \). This is because what distinguishes the waveforms in each group for \( q_i(t) \) is \( I_0, I_2, I_3 \) and for \( q'_i(t) \) is \( Q_1 \). If no distinction needs to be made in each group, we can drop \( I_0, I_2, I_3 \) and \( Q_1 \). Then, from Figure 5, the I, Q output waveform indexes has no connections any more. In the receiver, we can decouple the grid structure of the modulator into independent two-state grid for I channel and four-state grid for Q channel. Then, we can demodulate information of I and Q channel independently. The structure of this simplified receiver is shown in Figure 7.

In this simplified receiver, I and Q make decisions separately using energy-biased VAs. Noted here that each symbol energy of \( q_i(t) \) as well as \( q'_i(t) \) is different. Thus, we need to place energy bias on the output signal from matched filter. \( \overline{E}_i \) and \( \overline{E}'_i \) represent each symbol energy of \( q_i(t) \) and \( q'_i(t) \), respectively.

Compared with 32 correlations of optimal receiver, the simplified receiver only needs 10 correlations and greatly reduces complexity.

6 PERFORMANCE COMPARISON AND SIMULATION RESULTS

In this section, we provide the MATLAB numerical simulation results comparison between SO8PSK and 8PSK. We assume that the channel is a typical SATCOM channel composed of a band-limiter filter and an HPA close to saturation. The type of selected HPA is travel wave transfer amplifier (TWTA). Parameters of this HPA are: \( \alpha_a = 2.1587; \beta_a = 1.1517; \alpha_\phi = 4.0033; \beta_\phi = 9.1040 \). The considered channel model is AWGN channel. Without loss of generality, the bit rate is set as 100 bits/s. Each interval has eight sampling points. The product of bandwidth \( B \) and bit duration \( T_b \) in band-limited filter is \( BT_b = 2/3 \). The reported results are obtained by using Monte Carlo simulations averaged over \( 10^4 \) trials.

6.1 Signal waveform and PSD comparison

Figure 8 shows the signal waveforms of SO8PSK and 8PSK and Figure 9 shows the PSD performance of these two modulations. HPA works in the nonlinear region when input back off (IBO) < 7 dB. We consider IBO = 0 dB.

On one hand, continuous phase in SO8PSK enables its original PSD to have compact main lobe as well as lower and
smoothing side lobes. Small out-of-band energy will be filtered off by the band-limiting filter. Hence, the input signal to the HPA has nearly constant envelope. Minimal nonlinear AM/AM and AM/PM conversion effects of HPA will be produced. The final signal distortion is small and there is slight spectrum regeneration.

On the other hand, the abrupt phase changes (two examples are circled in Figure 8(b)) inherent in 8PSK result in the high side lobes in its original PSD. Large part of side lobes will be filtered off by the band-limiting filter. This creates signal envelope fluctuation at the input to HPA. Then, the AM/AM and AM/PM conversion effects of the HPA can generate the significant nonlinear distortion and cause the signal distortion. The side lobes in the HPA output PSD will regrow largely due to the amplitude and phase nonlinearities of the amplifier. This regrowth of side lobes could cause significant interference to the adjacent channel.

Here, SE is expressed as reciprocal of the bandwidth of each signal containing 99% signal power normalised to the data rate $B_{99\%}T_b$. We estimate $B_{99\%}T_b$ for two modulations via MATLAB’s built-in function obw. The original SE value is $B_{99\%}T_b \approx 1.30$ Hz s/bit for SO8PSK and $B_{99\%}T_b \approx 2.29$Hz s/bit for 8PSK. It is concluded that SO8PSK offers an increase of SE over 8PSK by approximately $(2.29 - 1.30) \times 100 \% = 43\%$. Besides, the SE after HPA is $B_{99\%}T_b \approx 1.30$ Hz s/bit for SO8PSK and $B_{99\%}T_b \approx 2.29$Hz s/bit for 8PSK. The higher SE and and less spectral regeneration enable SO8PSK to be compatible to the SATCOM.

6.2 BER performance

Firstly, we give the minimum Euclidean distance of SO8PSK based on symbol-by-symbol trellis representation (Figure 3). Assuming the initial phase is 0(000) and the transmitted sequence is all zeros, we can observe that there exits another path that starts from (000) and ends at (000), while is different from the all-zero path. The minimum Euclidean distance is computed as,

$$D_{\text{min}}^2 = \int_0^{T_s} |s_4(t) - s_0(t)|^2 + |s'_4(t) - s'_0(t)|^2 dt$$

$$+ |s_4(t) - s_0(t)|^2 + |s'_2(t) - s'_0(t)|^2 dt$$

$$= 0.5234 T_s.$$  

(15)

The average signal (I + Q) energy per symbol $E_{av}$ is obtained from

$$E_{av} = 3E_b = \frac{1}{32} \sum_{m=0}^{31} \int_0^{3T_b} |s_i(t)|^2 + |s'_i(t)|^2 dt = T_s.$$  

Therefore, the normalised minimum squared Euclidean distance $d_{\text{min}}^2$ is

$$d_{\text{min}}^2 = \frac{D_{\text{min}}^2}{E_{av}} = 0.5234.$$  

(16)
Compared with 8PSK with minimum Euclidean distance $= 0.5857$, SO8PSK suffers about 0.48 dB loss in BER performance.

Figure 10 shows the comparison of BERs between SO8PSK and 8PSK in AWGN channel. BER performance of SO8PSK is relatively poor with performance loss of around 0.5 dB than 8PSK at BER of $10^{-5}$. Nonetheless, it is worth noting that SO8PSK improves considerably in spectrum efficiency than 8PSK at the expense of minor BER performance differences. Figure 10 also compares the theoretical results, results obtained from the optimal receiver and the simplified receiver for SO8PSK. The performance of simplified receiver is slightly inferior than that of optimal receiver with about 0.3 dB at BER of $10^{-5}$ due to the simplified receiver is mismatched. Nevertheless, we believe that it is worthwhile to sacrifice about 0.3 dB for a reduction in required matched filters amount from 32 to 10.

Figure 11 shows the BER performance comparison between two modulations in the typical satellite channel. We consider IBO = 0 dB and 0.3 dB. SO8PSK outperforms 8PSK about 2 dB at BER of $10^{-4}$. The reason is that the filtered 8PSK signal input to HPA has more fluctuated envelope. The AM/AM and AM/PM conversion effects of HPA can generate more nonlinear distortion. In contrast, the filtered SO8PSK has almost constant envelope and then suffers from slight nonlinear distortion. Thus, SO8PSK is more robust to the nonlinear channel and can be seen a good candidate for SATCOM.

Total degradation (TD) is a performance measure in the nonlinear environment. TD indicates how, as compared to the AWGN channel, the BER performance degrades in the presence of a nonlinear HPA as a function of its operating
SO8PSK enables operating closer to the saturation point than 8PSK. The optimum IBO for 8PSK is 2 dB leading to TD of 1.25 dB. SO8PSK outperforms significantly than 8PSK. The reason is that the 8PSK signal input to HPA has more fluctuated envelope. This non-linear distortion can cause a significant degradation in the BER performance. In contrast, SO8PSK is more robust to the nonlinear channel and can be seen a good candidate for SATCOM. The performance of simplified receiver is slightly inferior than that of optimal receiver.

7 Coded SO8PSK with iterative decoding

To improve the BER performance of SO8PSK and benefited from its inherent trellis code (Figure 5) which can be viewed as the inner code of concatenated system, we apply SO8PSK to the SCC system with iterative demodulation. As was true for [22] and [23], I, Q inner encoder of the transmitter in Figure 5 should be replaced by their recursive equivalents to achieve coding gains in concatenated system. To achieve recursive encoder of SO8PSK, the original binary bit need to be differentially encoded and then converted to transmitted symbols by (3). Then, the recursive encoding of SO8PSK is derived,

\[ u_i = a_i \oplus u_{i-3}, \quad a_i \in \{0, 1\} \]

\[ \alpha_j = \begin{cases} 
(1 + 2^{-1})[u_i - u_{i-3}] & , \quad i \in \mathbb{R}_1 \\
(1 + 2^{-1})[u_i - u_{i-3}] \cdot (2u_{i-2} + 1) & , \quad i \in \mathbb{R}_2 \\
(1 - 2^{-1})[u_i - u_{i-3}] & , \quad i \in \mathbb{R}_3.
\end{cases} \]

The corresponding \( a_i/\alpha_i \) from recursive representation is labelled below each branches in Figure 2. We can see that the final transmitted symbol \( \alpha_i \) of each branch remains unchanged and some input binary bits \( a_i \) have changed. Figure 13 illustrates the implementation block diagram of the recursive SO8PSK.

We consider SCC system in Figure 14. The binary inputs are first encoded by the outer encoder, interleaved and then applied to I, Q channel under the recursive implementation in Figure 13. After AWGN or SATCOM channel, the received I, Q channel passes through matched filters of optimal or simplified receiver to generate correlation outputs. These outputs are used to be soft information \( \Lambda(\cdot;\cdot) \) for 8-state soft-input soft-output (SISO) iterative decoding as branch metrics.

Figure 15 shows the iterative demodulation results of the SCC SO8PSK system. In our simulation, the number of iteration is 4, the outer encoder is 1/3 rate convolutional code (cc) encoder with generator polynomial \((4,12,17)\) and the interleaver block size is chosen to be \( N = 2304 \) bits. Note that since \( N \) is small, for better performance, we have scaled the extrinsic information from the inner SISO and the outer SISO by a factor of 0.51. The iterative demodulation can reduce the \( E_{b}/N_0 \) required for SO8PSK about 5 dB at BER of \( 10^{-5} \) sacrificing two-third spectrum efficiency. The BER performance of coded SO8PSK is also simulated in the typical SATCOM channel. In the nonlinear channel, coded SO8PSK performs almost the same as it is in the AWGN channel. The reason is that the filtered SO8PSK has almost constant envelope, which suffers from slight nonlinear
distortion. This modification method can be seen as a starting point to the better use of HOPSK for SATCOM.

8 I CONCLUSION

We modify 8PSK to be more spectral efficiency and applicable for SATCOM by offsetting its bits transmission and controlling its phase shift. Compared with linear modulation, the advantages of SO8PSK are provided: (1) High spectral efficiency: Continuous phase of SO8PSK brings smoother and tighter spectrum characteristics, that is, narrower main lobe and lower side lobes. (2) Less interference to adjacent channels: Considering a multi-channel communication system, the lower side lobes of SO8PSK can make less interference to adjacent channels. (3) High power efficiency: The filtered SO8PSK waveform still has a nearly constant envelope. Therefore, it can be robust to the AM/AM and AM/PM effects of the HPA. HPA can work near the saturation point and has high power efficiency. (4) Less signal degradation: Filter after HPA just needs to filter off slight regenerated side lobes since SO8PSK’s insensitivity to the nonlinear effect. This can basically maintain the signal energy. Moreover, simplified receiver is given to lower its complexity and iterative encoding for SCC is also evaluated to improve BER. The results reveal that SO8PSK can be a wonderful candidate for SATCOM given its SE, BER and complexity. We believe that our findings may encourage the reform of HOPSK to for 5G SATCOM communication. Future work will entail refining our
work by modifying higher order modulations such as 16, 64, 256-ary modulations and so on. Besides, combining the proposed waveform and compressed sensing to reduce the system complexity will also be studied.

ACKNOWLEDGEMENTS
This work was supported by the National Nature Science Foundation of China under Grant No. 41830110 and the National Nature Science Foundation of China for Young Scholars under Grant No. 61701167 and the Fundamental Research Funds for the Central Universities, No. 2019B00814.

ORCID
Minghe Mao https://orcid.org/0000-0001-8424-566X

REFERENCES
1. Kuang, L., et al.: Radio resource management in future terrestrial-satellite communication networks. IEEE Wirel. Commun. 24, 81–87 (2017)
2. Rice, M., Gagakuma, E.: Approximate MLSE equalization of SOQPSKTG in aeronautical telemetry. IEEE Trans. Aerosp. Electron. Syst. 55, 769–784 (2019)
3. Zhang, Y., et al.: Optimal design of cascade LDPC-CPM system based on bionic swarm optimization algorithm. IEEE Trans. Broadcast. 64, 762–770 (2018)
4. Back, H., Lim, J.: Spectrum sharing for coexistence of fixed satellite services and frequency hopping tactical data link. IEEE J. Sel. Areas Commun. 34, 2642–2649 (2016)
5. Kim, H.: Coding and modulation techniques for high spectral efficiency transmission in 5G and satcom. In: 2015 23rd European Signal Processing Conference (EUSIPCO), pp. 2746–2750. IEEE, Piscataway (2015)
6. Hu, D., et al.: A novel forward-link multiplexed scheme in satellite-based internet of things. IEEE Internet Things J. 5, 1265–1274 (2018)
7. Morello, A., Reimers, U.: DVB-S2, the second generation standard for satellite broadcasting and unicoating. Int. J. Satell. Commun. Netw. 22, 249–268 (2004)
8. De Gaudenzi, R., et al.: Performance analysis of turbo-coded APSK modulations over nonlinear satellite channels. IEEE Trans. Wireless Commun. 5, 2396–2407 (2006)
9. Liu, X., et al.: Performance evaluation of MSK and OFDM modulations for future GNSS signals. Springer, New York (2014)
10. Downey, J., et al.: Bandwidth-efficient communication through 225 MHz Ka-band relay satellite channel. In: 34th AIAA International Communications Satellite Systems Conference, pp. 5737–5737. AIAA, Reston (2016)
11. Hosseini, E., Perrins, E.: Burst-mode synchronization for SOQPSK. IEEE Trans. Aerosp. Electron. Syst. 55, 2707–2718 (2019)
12. Rice, M., et al.: Space-time coding for aeronautical telemetry: Part ii—Decoder and system performance. IEEE Trans. Aerosp. Electron. Syst. 53, 1732–1754 (2017)
13. Rieth, D., et al.: FGPA implementation of shaped offset QPSK modulator. In: 2015 IEEE International Conference on Digital Signal Processing (DSP), pp. 790–793. IEEE, Piscataway (2015)
14. Addabbo, P., et al.: A review of spectrally efficient modulations for earth observation data downlink. In: 2014 IEEE Metrology for Aerospace (MetroAeroSpace), pp. 428–432. IEEE, Piscataway (2014)
15. Simon, M.K.: Bandwidth-Efficient Digital Modulation with Application to Deep Space Communications, vol. 2, John Wiley & Sons, New York (2005)
16. Proakis, J.G., Salehi, M.: Digital Communications, vol. 4. McGraw-hill, New York (2001)
17. ETSI, E.: Digital video broadcasting (DVB); second generation framing structure, channel coding and modulation systems for broadcasting, interactive services, news gathering and other broadband satellite applications. Technical report, ETSI, 2005
18. Sharma, S., et al.: FGPA implementation of m-PSK modulators for satellite communication. In: 2010 International Conference on Advances in Recent Technologies in Communication and Computing, pp. 136–139. IEEE, Piscataway (2010)
19. Ramarakula, M.: Performance analysis of various modulation schemes on inter satellite communication link. Int. J. Comput. Eng. Res. (IJCER) 8, 1–12 (2018)
20. Ghannam, H., Darwazah, I.: SEFDM over satellite systems with advanced interference cancellation. IET Commun. 12, 59–66 (2018)
21. Dalakas, V., et al.: BICMC and TD comparative performance study of 16-APSK signal variants for DVB-S2 systems. IEEE Commun. Lett. 19, 723–726 (2015)
22. Lifang, L., Simon, M.K.: Performance of coded OFDM and M-STD SOQPSK with iterative decoding. IEEE Trans. Commun. 52, 1890–1900 (2004)
23. Simon, M., Divsalar, D.: ICC 2001. IEEE International Conference on Communications. Conference Record (Cat. No.01CH37240), A reduced-complexity highly power/bandwidth-efficient coded Fisher-patented quadrature-phase-shift-keying system with iterative decoding, 10, 2963–2964 (2001)
24. Kim, D.W., et al.: A modified two-step sova-based turbo decoder with a fixed scaling factor. In: 2000 IEEE International Symposium on Circuits and Systems (ISCAS), vol. 4, pp. 37–40. IEEE, Piscataway (2000)