A space-frequency coding scheme providing high level of diversity and spectrum efficiency for non-coherent frequency-selective MIMO-OFDM systems

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Abstract—This paper proposes a space-frequency (SF) coding scheme for non-coherent (NC) Multiple Input Multiple Output (MIMO)-Orthogonal Frequency Division Multiplexing (OFDM) fading links, where neither the transmitter nor the receiver knows the channel. Our strategy consists in applying a convolutional encoder and an interleaver before applying a SF encoder. This SF encoder distributes the encoded and interleaved bits over the different OFDM symbols. In order to reduce the decoding complexity, each OFDM symbol is divided into several groups. Within each group, the obtained bits are broken into two substreams. The first substream is used to construct a systematic NC codeword which leads to a simple decoding rule over the multipath channel. The second substream is mapped to a Grassmannian NC codeword obtained via an exponential map and that was proposed for NC space-time (ST) coding based system and has the advantage of using all the degrees of freedom of this system. We show through asymptotic pairwise error probability (PEP) analysis and simulation results that our encoding strategy can provide full diversity gain and achieves better performance in terms of spectrum efficiency and bit error rate than both the systematic NC-SF coding and our proposed SF coding without convolutional encoding and interleaving.

Index Terms—OFDM, MIMO, Space-Frequency, NC, Grassmannian, PEP analysis.

I. INTRODUCTION

The combination of OFDM and MIMO increases the spectrum efficiency and combats frequency-selective fading in wireless channels. The MIMO-OFDM transmission over frequency-selective channels can provide the spatial and the frequency diversity. In order to achieve these two forms of diversity, an SF code design is required. Most of the previous works introduced the SF code in coherent MIMO-OFDM systems, i.e assume a perfect knowledge of the channel state information (CSI) at the receiver [1]. In [1], some full-diversity SF codes are obtained from ST codes for arbitrary power delay profile (PDP) by a simple mapping. However, it is sometimes impossible to estimate the channel coefficients and especially in the frequency-selective case due to the presence of multipath fading. To combat this drawback, NC communication, where neither the transmitter nor the receiver has CSI, was considered [2]. In [2], Borgmann et al address code design for NC frequency-selective MIMO-OFDM fading channels. They propose design criteria for SF unitary codebook and show that unlike in the coherent case, NC-ST codes designed to achieve full spatial diversity in the frequency-flat fading case can fail completely to exploit spatial/frequency diversity over frequency-selective channels. They propose a simple unitary systematic constructions of NC-SF code which can achieve the full spatial/frequency diversity over unknown frequency-selective MIMO-OFDM channels. However, the decoding complexity of this code increases exponentially with the number of subcarriers. In order to reduce this complexity, the OFDM symbol is divided into several groups where each group has a small number of subcarriers in [3] and [4].

In this paper, we propose a unitary SF code design suited for NC MIMO-OFDM links. The proposed SF code is based on the combination of the unitary systematic NC-SF code proposed in [2], which achieves a full spatial/frequency diversity for frequency-selective channels, and unitary Grassmann NC-ST codes proposed in [5] for frequency-flat fading channels. The ST codes introduced in [5] are obtained via an exponential map from coherent codes. In order to reduce the decoding complexity, we divide the OFDM symbol into several groups. In our study, the information bits are convolutional encoded and interleaved before being mapped onto the transmit antennas, the OFDM tones and the OFDM symbols. The convolutional coding is used together with interleaving to separate adjacent coded bits in the frequency domain as much as possible longer than the coherence bandwidth of the channel in order to get the maximum space/frequency diversity. We show through asymptotic PEP analysis and simulation results that our encoding strategy can provide full diversity gain and achieves better performance in terms of spectrum efficiency and bit error rate than both the systematic NC-SF coding proposed in [2] and our composed coding without convolutional encoding and interleaving.

The rest of this paper is organized as follows. In Section II, we introduce the system model, describe our NC-SF MIMO-OFDM scheme, present our composed code and detail each component. In section III, we derive the asymptotic PEP for our system and deduce the diversity gain that can be achieved. In section IV, we provide some simulation results. Finally, section V contains our conclusions.

II. SYSTEM MODEL

A. Frequency selective MIMO channel Model

In this paper, we consider a MIMO-OFDM system with $M_t$ transmit antennas, $M_r$ receive antennas and $N$ subcarriers. Suppose that the Rayleigh fading channels between each pair of transmit and receive antennas have $L$ independent delay
paths. The frequency response of the length-$L$ MIMO channel from the $i$-th transmit antenna and the $j$-th receive antenna for the $n$-th subcarrier is defined as

$$[H_n]_{i,j} = \sum_{l=0}^{L-1} h_{i,j}(l) e^{-j \frac{2\pi}{L} ln},$$

where $h_{i,j}(l)$ is the path gain coefficient of the $l$-th path between the transmit antenna $i$ and the receive antenna $j$. We note that the elements of $h_{i,j}(l)$, $l = 0, ..., L - 1$ are i.i.d. circularly symmetric zero-mean complex Gaussian with variance $\sigma^2_l$. Note that $\sigma^2_l$ is normalized such that $\sum_l \sigma^2_l = 1$.

**B. NSC-MIMO-OFDM system model**

During transmission, the information bits are encoded by a convolutional code of rate $R_c$. The output bits of the encoder are interleaved by a bit interleaver $\pi$. The interleaved bit stream is divided into $b$ bit long segments and each segment is mapped onto a $N_g \times M_t$ SF codeword, with $N_g = N/N_g$ and $N_g$ is the number of groups. The OFDM modulator applies a $N$-point inverse fast Fourier transform (IFFT) and add a cyclic prefix (CP) to the parallel-to-serial-converted OFDM symbol. In this paper, we consider that the channel order is lower than or equal to the length of the CP, $L \leq L_{CP}$. The receiver discards the cyclic prefix, and then applies an $N$-point FFT to each of the $M_r$ received signals. The $1 \times M_r$ received signal vector on the $n$-th subcarrier is given by

$$y_n = \sqrt{\rho} x_n H_n + w_n,$$

where the factor $\sqrt{\rho}$ scales the signal energy to ensure that the average signal-to-noise ratio (SNR) at each receive antenna is $\rho$ and $x_n \in 1 \times M_t$ is the transmitted vector on the $n$-th subcarrier. We note that $w_n$ is an $1 \times M_r$ additive complex Gaussian noise vector with zero mean and unit variance satisfying $E[|w_n|^2] = 1$. By dividing the $N$ subcarriers into $N_g$ groups with each group containing $N_b$ subcarriers, the received signal corresponding to the $k$-th group can be written as

$$Y^{(k)} = \sqrt{\rho} \sum_{l=0}^{L-1} X^{(k)}_{l} \tilde{H}^l + W^{(k)} = \sqrt{\rho} F_{X}^{(k)} \tilde{H} + W^{(k)}.$$  

In order to not destroy the unitarity constraint, we consider the unitary blockwise scheme proposed in [3] by adding the blocks MPFI (maximum possible frequency domain interleaving) at the transmitter and MPFD (maximum possible frequency domain deinterleaving) at the receiver. So, the $N_b \times N_b$ diagonal matrix $D_k$ is given by

$$D_k = e^{-j2\pi k/N} \text{diag} \{ e^{-j2\pi n/N_b}, n=0,1,...,N_b-1 \}$$

and $\tilde{H} = (h_{i,j}(l)), l = 0, ..., L - 1$ is the $M_t \times M_r$ channel impulse response taps matrix defined as $H = (h_{i,j}(l)), l = 1, ..., M_t$ and $j = 1, ..., M_r$. Here $W^{(k)}$ is the $N_b \times M_r$ additive complex Gaussian noise matrix. We define the $N_b \times L M_t$ SF pseudo-codeword as $X^{(k)} = [X^{(k)}, X^{(k)} D X^{(k)} , ..., D^{L-1} X^{(k)}]$ and the $L M_t \times M_r$ stacked channel impulse response taps matrix as $\tilde{H} = [H_0, H_1, ..., H_{L-1}]$.

**C. Proposed composed codes**

Within each group, the obtained interleaved bits are broken into two substreams. The first substream is used to construct the systematic NC codeword. The second substream is mapped to a NC codeword obtained via an exponential map and will be called EP NC codeword. So, the transmitted codeword over the $k$-th group is given by the following expression

$$X^{(k)} = X^{(k)}_{sys} X^{(k)}_G,$$

with $X^{(k)}_{sys}$ is a $N_b \times M_t$ matrix and $N_b$ is the number of subcarriers for each group. Thereafter, we detail the description of each component codeword.

1) The component systematic NC codeword: In [2], the authors proposed a unitary SF code which achieves the full space-frequency diversity $L M_t M_r$ over frequency-selective MIMO-OFDM channel. Their construction is inspired by the systematic design of unitary ST codes proposed in [6] over flat fading-channel. In [2], coding is performed among all the subcarriers but in our work coding is considered for each group. Our codeword constructed, as in [2], for each group is called systematic NC codeword and given by

$$X^{(k)}_{sys} = \Phi^{(k)} \{ f_{p_1}, f_{p_2}, ..., f_{p_T} \},$$

where $X^{(k)}_{sys}$ is a $N_b T \times T$ matrix, $T$ is used in this paper to set the dimension of the two codewords but isn’t considered as the time period and $0 \leq m \leq L_{sys} - 1$, with $L_{sys}$ is the cardinality of systematic NC codebook. The vectors $f_{p_t}, t = 1, ..., T$, are the $p_t$-th columns of the $N_b \times N_b$ fast fourier transform (FFT) matrix, chosen to have $F_{X^{(k)}}$, a unitary pseudo-codeword. $\Phi$ is a $N_b \times N_b$ matrix defined as

$$\Phi = \text{diag} \{ e^{-j2\pi u_n/L_{sys}} \},$$

with $0 \leq u_n \leq L_{sys} - 1$ have to be chosen to optimize the chordal distance between the subspaces spanned by the columns of the two different codewords.

In order to achieve the full diversity $N_b$ should be chosen as $N_b \geq 2 TL$ [2].

2) Component EP NC codeword: The $T \times M_r$ matrix $X^{(k)}_G$ will be taken as a codeword from the Grassmann codebook constructed as follows [5]

$$X^{(k)}_G = \left[ \exp \left( \begin{array}{c} 0 \\ \alpha V^{(k)}_G \\ -\alpha V^{(k)\dagger}_G \\ 0 \end{array} \right) \right] I_{T,M_r},$$

where $V^{(k)}_G$ is a $M_t \times (T - M_t)$ matrix obtained from $M_t (T - M_t)$ information QAM symbols [5]. The common positive scalar $\alpha$, which is called homothetic factor, is chosen to satisfy the invertibility of the exponential map and to optimize the performance of the code, at the same time [5].

**D. Simplified decoding**

The received symbol corresponding to the $k$-th transmitted group can be rewritten as follows

$$Y^{(k)} = \sqrt{\rho} F_{X^{(k)}_{sys}} G_{X^{(k)}} \tilde{H} + W^{(k)}.$$
We define the $N_b \times LT$ SF pseudo-codeword as $F_{X_{sys}}^{(k)} = [X_{sys}^{(k)}, D_X X_{sys}^{(k)}, \ldots, D_X^{L-1} X_{sys}^{(k)}]$ and the $LT \times LT$ matrix $G_{X_G} = I_L \otimes X_{sys}^{(k)}$. So, the $LT \times M_r$ matrix $H_{G}^{(k)} = G_{X_G} H$ have the same probability distribution as $H$ and it can be considered as a pseudo-channel. In this case, at the receiver, the decoding of the combined codeword can be divided into two sequentially processes described as follows:

1) Decoding the systematic NC codeword: For the $k^{th}$ group, the decision on the systematic NC codeword is made according to the Generalized Likelihood Ratio Test (GLRT), which doesn’t require the knowledge of channel statistics since it is defined as

$$\hat{X}_{sys}^{(k)} = \arg \max_{X_{sys} \in C_1} p(Y^{(k)} | X_{sys}^{(k)}, H_G^{(k)}).$$

In our case, that is i.i.d. fading and unitary codebook, the GLRT is equivalent to the ML criterion, and it takes the form

$$\hat{X}_{sys}^{(k)} = \arg \max_{X_{sys} \in C_1} \text{Trace} \left\{ Y^{(k)} F_{X_{sys}}^{(k)} F_{X_{sys}}^{(k)\dagger} Y^{(k)} \right\}.$$  (11)

Let $Y^{(k)} = \text{vec}(Y^{(k)})$, where the vector operator $\text{vec}(A)$ stacks all columns of the matrix $A$ on top of each other, from left to right. The GLRT receiver in (11) can be written as [7]

$$X_{sys}^{(k)} = \arg \min_{X_{sys} \in C_1} \| Y^{(k)} - F_X^{(k)} h_G^{(k)} \|_F^2,$$  (12)

where $\| \cdot \|_F$ is the Frobenius norm. $F_X^{(k)} = \sqrt{\rho}(I_{M_r} \otimes F_X^{(k)})$ is the Kronecker tensor product and $h_G^{(k)} = (F_X^{(k)} F_X^{(k)\dagger})^{-1} F_X^{(k)\dagger} Y^{(k)}$.

2) Decoding the EP NC codeword: Since we obtain $X_{sys}^{(k)}$, we can decode the EP codeword. We note that the received signal $\tilde{Y}$ corresponding to the transmitted EP NC codeword is given by

$$\tilde{Y}^{(k)} = F_{X_{sys}}^{(k)} Y^{(k)}/\sqrt{\rho} = G_{X_G} H + W^{(k)}.$$  (13)

We note that $F_{X_{sys}}^{(k)}$ is unitary. So, $\tilde{W}^{(k)} = F_{X_{sys}}^{(k)} W^{(k)}/\sqrt{\rho}$ has the same distribution as $W^{(k)}$. At the receiver the decision of the NC Grassmann codeword is made according to the GLRT

$$X_G^{(k)} = \arg \max_{X_G \in C_G} \text{Trace} \left( Y^{(k)} F_{X_G}^{(k)} G_{X_G}^{(k)} F_{X_G}^{(k)\dagger} Y^{(k)} \right).$$  (14)

As we have seen, the decoding of the composed codeword requires two sequentially GLRT decoders. For the systematic SF code with rate $R_1$ and size $N_b \times T$, the GLRT decoder (implemented with exhaustive search) should search among $2^{R_1 N_b}$ possible codewords. The complexity of this decoder increases exponentially with $R_1$ and $N_b$. So, it is clear that by dividing the OFDM symbol into several groups, the decoding complexity of the first component is significantly reduced since the search is made on $2^{R_1 N_b}$ codewords instead of $2^{R_1 N_b}$ ones. Nevertheless, systematic SF codewords can be interpreted as subspaces in the Grassmannian manifold and the distance between these subspaces is improved by increasing the number $N_b$ of subcarriers [4]. We will present in section IV some simulation results related to the complexity-performance tradeoff in the choice of number $N_b$. For the EP NC codeword with rate $R_2$ and size $T \times M_r$, the GLRT decoder with exhaustive search should search among $2^{R_2 T}$ possible codewords. We note that to reduce the complexity of an exhaustive search, simplified GLRT receivers proposed in [5] and [8] can be implemented to decode the systematic and the EP NC codewords.

E. Decision bit metric

The bit interleaver can be modeled as $\pi : k' \rightarrow (k, i')$, where $k'$ denotes the original ordering of the coded bits $c_{k'}$, $k$ denotes the ordering of the group and $i'$ indicates the position of $c_{k'}$ in the $b$ bits segment corresponding to the group. Since a GLRT receiver is considered to decode the systematic NC codeword, the bit metric for bit $c_{k'}$ at the $i'^{th}$ location of the $k^{th}$ group can be given by

$$\lambda_i^{(k)} (c_{k'}) = \log \sum_{X_{sys} \in C_1} p(Y^{(k)} | X_{sys})$$

$$= \log \sum_{X_{sys} \in C_1} \exp \left\{ -\|y^{(k)} - F_X^{(k)} h_G^{(k)}\|_F^2 \right\}$$

$$= \min_{X_{sys} \in C_1} \|y^{(k)} - F_X^{(k)} h_G^{(k)}\|_F^2,$$  (15)

where $Y = E [w^{(k)} w^{(k)\dagger}] = I_{N_b M_r}$ with $w^{(k)} = \text{vec}(W^{(k)})$ and $c_{k'}$ is the subset of all signals whose $\ell^{th}$ $(X) = c_{k'}$, with $\ell^{th}$ $(X)$ is the $i'^{th}$ bit of the segment corresponding to codeword $X$.

Then, the metrics associated to the interleaved bits are used by the Viterbi decoder to decode the information bits by finding the shortest path in the trellis according to

$$\hat{c} = \arg \min_{c \in C} \left( \sum_{k'} \lambda_i^{(k)} (c_{k'}) \right).$$  (16)

III. ASYMPTOTIC PEP ANALYSIS

In this section, we use the same notations as [9] when the PEP is given for bit interleaved coded modulation (BICM) OFDM system using the orthogonal ST block codes (STBC) by transmitting each codeword over one subcarrier. In our work, we give the PEP of our SF coding scheme preceded by the convolutional encoder and bit interleaver. Assume that binary codeword $\xi$ is transmitted and $\hat{\xi}$ is detected in MIMO-OFDM system over NC frequency-selective channel with uniform PDP $\sigma^2 = 1/L, l = 0, \ldots, L - 1$. Then, the PEP can be written as

$$P(\xi \rightarrow \hat{\xi}) = P \left( \sum_{k'} \lambda_i^{(k')} (c_{k'}) \geq \sum_{k'} \lambda_i^{(k')} (\hat{c}_{k'}) \right)$$

$$= P \left( \sum_{X_{sys} \in C_1} \min_{X_{sys} \in C_1} \|y^{(k)} - F_X^{(k)} h_G^{(k)}\|_F^2 \geq \sum_{X_{sys} \in C_1} \min_{X_{sys} \in C_1} \|y^{(k)} - F_X^{(k)} h_G^{(k)}\|_F^2 \right)$$

$$= \left( \sum_{X_{sys} \in C_1} \min_{X_{sys} \in C_1} \|y^{(k)} - F_X^{(k)} h_G^{(k)}\|_F^2 \right).$$  (17)
Let $dfree$ be the minimum Hamming distance of the convolutional code. We assume that the distance between the incorrect path associated to $\xi$ and the correct one associated to $\xi'$ is $dfree$. The sets $X_{k'}$ and $X_{k''}$ are equal to one other for all $k'$ except for $dfree$ distinct values of $k'$. Then, only $dfree$ terms are different in the inequality (16).

By denoting $\tilde{F}(k) = \arg\min_{X(k')} \| y_1(k') + \tilde{F}(k') \|_F^2$ and $F(k) = \arg\min_{X(k')} \| y(k) - F(k) h_G \|_F^2$, the PEP can be written as

$$P(\xi \rightarrow \xi') = P \left( \sum_{k', dfree} \left\| y(k) - F(k') h_G \right\|_F^2 \geq \sum_{k, dfree} \left\| y(k) - \tilde{F}(k') h_G \right\|_F^2 \right).$$

By using the two equalities $h_G(k) = (F(k) F(k'))^{-1} F(k) F(k') y(k)$ and $y(k) = F(k) h_G + w(k)$, with $h_G(k) = \text{vec}(h_G(k))$ we can write the PEP as follows

$$P(\xi \rightarrow \xi') = P \left( \sum_{k', dfree} \left( w(k) (\Pi_\xi(k) - \Pi_\xi(k')) w(k) \right) \geq \sum_{k, dfree} \chi_k^2 \right),$$

where $\Pi(\cdot)$ stands for the real part of the complex number, $\chi_k^2 = \text{vec}(\Pi_\xi(k) F(k) F(k') h_G)$, with $\Pi_\xi(k) = \Pi_{N, M} - F(k) F(k') - F(k') F(k) - \chi_k^2$. For a fair comparison between our proposed scheme and the systematic NC-SF coding, the cardinality of the codebook for each rank is full, the maximal diversity of our scheme can achieve $dfree L M_r M_r$. We notice that $\rho_m$ are the singular values of the matrix $\bar{F}(\bar{k}) \bar{F}(\bar{k}) / \rho$. We can easily demonstrate that the full-rank of each matrix $\bar{F}(\bar{k}) \bar{F}(\bar{k}) / \rho$ can reach $r_{free} = LM_r M_r$. So, if all ranks are full, the maximal diversity of our scheme can achieve $dfree L M_r M_r$.

IV. SIMULATION RESULTS

We simulate the proposed SF coding for a MIMO-OFDM system with $M_t = 2$ transmit antennas, $M_r = 1$ receive antennas, $T = 4$ and $N = 128$ subcarriers. For the component EP codeword, we consider this construction of the matrix $V$

$$V = \left( \begin{array}{c} \phi_r(s_1 + r s_2) \\ \phi_r(s_3 + r s_4) \\ \phi_r(s_1 + r s_2) \end{array} \right),$$

where $r = (1 + \sqrt{5})/2$, $r = (1 - \sqrt{5})/2$, $\phi_r = 1 + i(1 - r)$, $\phi_r = 1 + i(1 - r)$ and symbols $s_i$ are drawn from a $4 - QAM$ constellation. We prove via numerical simulations that the optimal homothetic factor is $\alpha = 0.39$. In this case, the transmitted bits over SF composed code are broken into two substreams of length 8 bits. Figure 1 shows the performance comparison results of our proposed composed NC-SF code with the systematic NC-SF code with and without convolutional encoder and bit interleaver for MIMO-OFDM systems. For the convolutional encoder, we use the encoder [2, 3] with rate $R_c = 1/2$, 4 states and $dfree = 3$. In this paper, we are interested to unitary SF codes. For this reason, we don’t make comparison with a EP NC-SF coding based on applying the EP NC-ST code with replacing the time with frequency since it leads to a non unitary $F(k)$ matrix. For a fair comparison between our proposed scheme and the systematic NC-SF coding, the cardinality of the codebook for the systematic NC-SF code is chosen $K = 2^{16}$. From Figure 1, we can conclude that the performance achieved with our proposed composed SF code is better than the performance obtained with the systematic NC-SF code. Moreover, Figure 1 shows that our proposed SF code with convolutional encoder and bit interleaver get a significant diversity gain as predicted. Figure 2 shows the BER performance of our composed code for two numbers of subcarriers per group, $N_0 = 16$ and $N_0 = 32$. For the case of $N_0 = 32$, the number of bits transmitted by the codeword is divided into two substreams of length 16 bits. For this, the symbols $s_i$ transmitted over component EP NC codeword, are drawn from a $16 - QAM$ constellation and the cardinality of the codebook for the systematic NC-SF code is chosen $K = 2^{16}$. The two schemes without and with convolutional encoder and bit interleaver are also considered in Figure 2. We observe a performance enhancement when the number $N_0$ of subcarriers per group increases, for both cases with and without convolutional encoder and interleaver. For a further characterization of the performance of our proposed scheme, we compare in Figure 3 the BER of our scheme with convolutional encoder and bit interleaver for different convolutional encoders ($[1, 2]$ with $dfree = 2$, [2, 3] with $dfree = 3$ and [5, 7] with $dfree = 5$) and for high frequency-selective channel with two and three paths and considering
Fig. 1. Comparison performances of systematic NC-SF coding with our proposed NC-SF coding without convolutional encoding and interleaving and of our NC-SF code with and without convolutional encoding and interleaving. We take $M_t = 2$, $M_r = 1$, $N_b = 16$ and $N_g = 8$.

Fig. 2. Effect of the length group on the performance of our coding without and with convolutional encoder and bit interleaver. We take $M_t = 2$, $M_r = 1$, $N = 128$ and $L = 2$.

uniform and exponential PDP. We show that the diversity order becomes larger by increasing the distance $d_{\text{free}}$ of the convolutional encoder and when the selectivity of the channel becomes more important. For example, the BER reached in the case of exponential PDP when $L = 3$ and $d_{\text{free}} = 5$ is about $10^{-8}$ at $SNR = 7dB$. In this paper, we don’t consider the case of very selective channels ($L > 3$) since the constraint $N_b \geq 2TL$ must be satisfied and the search of the coefficients $u_n$ becomes more difficult.

V. CONCLUSION

In this paper, we have proposed a NC-SF code for MIMO-OFDM systems. This code is constructed as a combination of a NC systematic codeword and a Grassmannian codeword generated by the exponential map. We consider that our SF coding is preceded by a convolutional encoder and a bit interleaver. At the receiver, on each group, we decoded the component systematic NC codeword and EP NC codeword subsequently with the GLRT detection. The expression of the asymptotic PEP was derived for this scheme and we find the maximal diversity order that can be achieved. We showed that under some conditions the new NC-SF code with convolutional encoder and bit interleaver can get the maximum available diversity gains in frequency-selective fading channels. We demonstrated that the diversity order increases with the distance $d_{\text{free}}$ of convolutional encoder. Compared with the systematic NC-SF code, the proposed code can achieve larger spectrum efficiency with better performance in terms of bit error rate.

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