Generalized Space–Time–Frequency Index Modulation

Mohammad Ismat Kadir

***Abstract*— Generalized space–time–frequency index modulation (GSTFIM) is proposed for conveying extra information with the aid of the space, time, and frequency indices. GSTFIM subsumes the concept of generalized space–time shift keying (GSTSK), which flexibly activates Q' out of Q dispersion matrices, and the orthogonal frequency-division multiplexing (OFDM) with index modulation. The transceiver architecture of the proposed GSTFIM is outlined. GSTFIM is capable of providing higher transmission rate than the classic OFDM-based GSTSK. It also outperforms the classic OFDM-aided GSTSK in terms of the bit-error-probability.**

***Index Terms*— Orthogonal frequency division multiplexing (OFDM), index modulation (IM), generalized space-time shift keying (GSTSK).**

I. Introduction

INDEX modulation (IM) [1] is a radically new concept for the transmission of information. The idea of mapping information bits to the indices of different entities, namely space, time and/or frequency, was conceived in [2]–[4] and employed in different applications. The concept of spatial modulation (SM) [2] is to transmit classic constellation symbols using a single antenna element (AE) of a multiple-antenna transmitter, whereas space-shift keying (SSK) [5] utilizes only the index of the AE activated for the transmission of source information. The SM and the SSK may thus be regarded as the spatial IM schemes. Space-time shift keying (STSK) [4] employs the index of the dispersion matrix (DM) activated for conveying additional information and may be called the DM-IM or the space-time (ST) IM scheme. The employment of the DM index as the carrier of information has the additional advantage that STSK is capable of spreading the information bits to both the spatial- and time dimensions. A generalized SM (GSM) [6] and a generalized STSK (GSTSK) [7] scheme were also proposed for further increasing the achievable rate. Usually in an index modulated system, a single transmission entity, such as space or time or DM, is activated and the index of the activated entity facilitates the transmission of extra information. On the other hand, the term 'generalized' refers to a system where an arbitrary number of transmission entities may be activated and the spectral efficiency may be further increased. In GSM, any arbitrary number of transmit AEs are activated at a time [6]. Similarly, an arbitrary number of DMs are activated simultaneously in GSTSK [7]. Orthogonal frequency division multiplexing (OFDM) with the subcarrier-domain IM [3] has also emerged as a beneficial technique of information transmission. IM in different resource domains makes the transmit signal sparse, increases the achievable rate and improves the reliability. Motivated by the benefits of IM, generalized space- and frequency index modulation was proposed in [8] combining GSM and OFDM-IM. In spite of these developments on IM and on STSK-OFDM-IM [9], [10], generalized ST-IM [4], [11] amalgamated with the OFDM-IM of [3] still remains an unexplored open problem.

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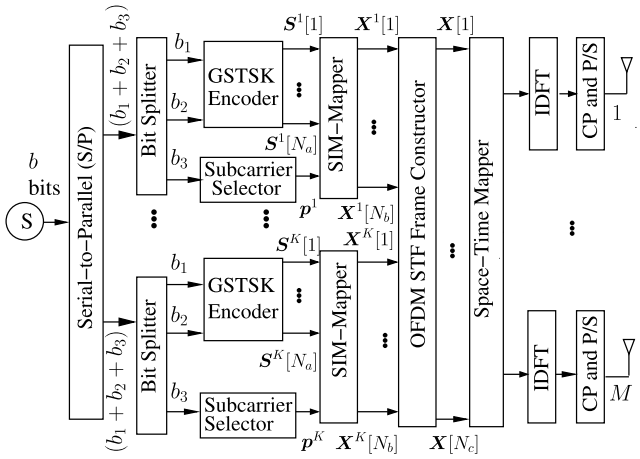


Fig. 1. Transmitter model of the proposed GSTFIM scheme. The b source bits are mapped to the AE-, time- and subcarrier-indices and the classic symbols to form STF frame. The frame symbols are transmitted by M transmit AEs.

Against this background, generalized space-time-frequency index modulation (GSTFIM) is conceived in this letter. The system activates Q' out of Q DMs according to the GSTSK philosophy and relies simultaneously on OFDM-IM for transmitting extra information with the aid of space, time as well as frequency indices. A single-stream maximum-likelihood (ML) detector and a soft-decision detector are proposed for the GSTFIM system. The achievable rate of the scheme and the complexity imposed by the detectors are quantified. Helpful design criteria for the GSTFIM are also provided.

The rest of the letter is organized as follows. In Section II, an overview of the proposed GSTFIM transceiver is provided. The achievable rate, complexity and design criteria of the scheme are highlighted in Section III. Section IV provides the numerical results. Finally, the letter is concluded in Section V.

II. System Overview

Fig. 1 portrays the schematic diagram of the transmitter of the proposed GSTFIM system, which employs M transmit- and N receive AEs for transmissions over T symbol durations with the aid of N_c parallel subcarriers.

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A. The Transmitter

In the proposed scheme, some of the source bits are first mapped to the GSTSK codewords [7], while the remaining bits are mapped to the index of the set of subcarriers activated according to the OFDM-IM philosophy of [3]. The GSTSK codewords are then mapped to the activated subcarriers by the 'SIM-Mapper' of Fig. 1. The 'OFDM STF Frame Constructor' of Fig. 1 accumulates the codewords of all the N_c subcarriers. After proper space-time mapping and appropriate OFDM processing, the codeword symbols are transmitted by the AEs.

To be specific, the b source bits of Fig. 1 are first partitioned into K blocks of $(b_1 + b_2 + b_3)$ bits each and the N_c subcarriers into K blocks of N_b subcarriers each. The b_1 and b_2 bits are mapped to N_a GSTSK codewords [7], while the b_3 bits are employed for activating N_a out of N_b subcarriers. The GSTSK codewords are generated by activating Q' out of Q DMs $\boldsymbol{A}_1^k, \boldsymbol{A}_2^k, \ldots, \boldsymbol{A}_Q^k$ in every parallel block k by the b_1 bits and by selecting Q' constellation symbols s^k with the aid of the b_2 bits [7]. More specifically, the GSTSK codewords generated by the 'GSTSK Encoder' of Fig. 1 are given by [7]: $\boldsymbol{S}^k = \sum_{q'=1}^{Q'} s_{q'}^k \boldsymbol{A}_{q'}^k \in \mathbb{C}^{M\times T}$. Hence, we have $b_1 = N_a \left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor$ and $b_2 = N_a Q' \log_2 \mathcal{L}$. On the other hand, the b_3 bits are mapped to the subcarrier activation vector $\boldsymbol{p}^k$ whose N_a elements are the indices of the subcarriers activated, so that $b_3 = \left\lfloor \log_2 \binom{N_b}{N_a} \right\rfloor$.

The DMs are $M \times T$ matrices, but they have only M' non-zero elements in each of the activated T' columns. Additionally, the DMs are appropriately optimized under the power contraint [7]: $\text{tr}(\boldsymbol{A}_q^H \boldsymbol{A}_q) = \frac{T}{Q'} \ \ \forall q$. Specifically, $\left\lfloor \log_2 \binom{M}{M'} \right\rfloor$ information bits are mapped to the indices of M' AEs activated, which may be termed as the space index modulation. Similarly, $\left\lfloor \log_2 \binom{T}{T'} \right\rfloor$ source bits are used for selecting the T' symbol durations, which may be referred to as the time index modulation. Furthermore, Q and Q' are so chosen that $\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor = Q' \left(\left\lfloor \log_2 \binom{M}{M'} \right\rfloor + \left\lfloor \log_2 \binom{T}{T'} \right\rfloor \right)$. The mapping of the index bits to the DM indices may thus be viewed as the space- as well as time-index modulation.

Example 1: Let us assume the DMs to be 2×2 matrices, while a single AE out of the two AEs and one out of the two symbol durations are activated. We have $M = 2$, $T = 2$, $M' = 1$, $T' = 1$. With $Q' = 1$ DM activated out of $Q = 4$, the mapping of the DM index bits to the AE- and time-index may be shown as follows:

$$00 \rightarrow \begin{matrix} \\ \text{AE ='0'} \end{matrix} \overset{\text{Time ='0'}}{\begin{bmatrix} a_{11} & 0 \\ 0 & 0 \end{bmatrix}}, \quad 01 \rightarrow \begin{matrix} \\ \text{AE ='0'} \end{matrix} \overset{\text{Time ='1'}}{\begin{bmatrix} 0 & a_{12} \\ 0 & 0 \end{bmatrix}},$$

$$10 \rightarrow \begin{matrix} \\ \text{AE ='1'} \end{matrix} \overset{\text{Time ='0'}}{\begin{bmatrix} 0 & 0 \\ a_{21} & 0 \end{bmatrix}}, \quad 11 \rightarrow \begin{matrix} \\ \text{AE ='1'} \end{matrix} \overset{\text{Time ='1'}}{\begin{bmatrix} 0 & 0 \\ 0 & a_{22} \end{bmatrix}}.$$

Following the bit mapping, the N_a codewords defined by $\boldsymbol{S}^k[1], \boldsymbol{S}^k[2], \ldots, \boldsymbol{S}^k[N_a]$ are mapped to the block matrix $\boldsymbol{X}^k\mathbf{:} = \left[\boldsymbol{X}^k[1], \boldsymbol{X}^k[2], \ldots \boldsymbol{X}^k[N_b]\right]^T$. The block matrices of all the parallel blocks generate the space-time-frequency (STF) frame represented by another block matix $\boldsymbol{X}\mathbf{:} = [\boldsymbol{X}[1], \boldsymbol{X}[2], \ldots, \boldsymbol{X}[N_c]]^T$. The STF frame $\boldsymbol{X}\mathbf{:}$ consists of N_c ST matrices, out of which $N_p = (K \times N_a)$ are the GSTSK codewords and the rest are the $M \times T$ zero matrices. As a result, the STF frame $\boldsymbol{X}\mathbf{:}$ is sparse in frequency domain, while each of the constituent matrices $\boldsymbol{S}_{q'}^k = s_{q'}^k \boldsymbol{A}_{q'}^k$ of the GSTSK codeword $\boldsymbol{S}^k$ is sparse in space and time domains. Thus the STF frame $\boldsymbol{X}\mathbf{:}$ holds index modulation in all the three domains and may be referred to as the GSTFIM codeword. The GSTFIM codeword is mapped to the M AEs over T symbol durations by the 'Space-Time Mapper' of Fig. 1. The frequency-domain (FD) symbols at each of the M AEs are transmitted after applying N_c-point inverse discrete Fourier transform (IDFT) and after adding cyclic prefix (CP).

B. The Receiver

Each link of the multiple-antenna aided system is assumed to be frequency-selective. After removing the CP, the N_c-point discrete Fourier transform (DFT) is first applied on the received signal. The resulting FD output corresponding to the subcarrier n_b $(n_a = 1, 2, \ldots, N_b)$ of the $k-$th parallel block may be expressed by the linearized model of [7] as:

$$\overline{\boldsymbol{Y}}^k[n_b] = \overline{\tilde{\boldsymbol{H}}}^k[n_b]\boldsymbol{\chi}\boldsymbol{\mathcal{K}}[n_b] + \overline{\boldsymbol{V}}^k[n_b], \tag{1}$$

where $\overline{\boldsymbol{Y}}^k[n_b] = vec(\boldsymbol{Y}^k[n_b])$, $\boldsymbol{\chi} = [vec(\boldsymbol{A}_1), \ldots, vec(\boldsymbol{A}_Q)]$, $\overline{\boldsymbol{V}}^k[n_b] = vec(\boldsymbol{V}^k[n_b])$ is the noise vector, $\overline{\tilde{\boldsymbol{H}}}^k[n_b] = \boldsymbol{I}_T \otimes \tilde{\boldsymbol{H}}^k[n_b]$ is the equivalent FD channel transfer matrix, $\boldsymbol{I}_T$ denotes the $T \times T$ identity matrix and $\otimes$ represents the Kronecker product. Still referring to (1), the equivalent transmit signal vector $\boldsymbol{\mathcal{K}}[n_b]$ is defined by [7]: $\boldsymbol{\mathcal{K}}[n_b] \triangleq [\kappa_1, \kappa_2, \ldots, \kappa_Q]^T$, where Q' of the Q elements $\kappa_1, \kappa_2, \ldots, \kappa_Q$ are the constellation symbols $s_{q'}^k$ and the rest are zeros. N_b subcarriers can be demapped to N_a activated subcarriers for any specific $\boldsymbol{p}^k$. The ML detector may thus be formulated as:

$$\hat{\boldsymbol{p}}^k, \hat{\boldsymbol{q}}^k, \hat{\boldsymbol{l}}^k = \underset{\boldsymbol{p}^k, \boldsymbol{q}^k, \boldsymbol{l}^k}{\arg\min} \sum_{n_a=1}^{N_a} \left\| \overline{\boldsymbol{Y}}^k[n_a] - \overline{\tilde{\boldsymbol{H}}}^k[n_a]\boldsymbol{\chi}\boldsymbol{\mathcal{K}}[n_a] \right\|^2 \tag{2}$$

where $\hat{\boldsymbol{p}}^k, \hat{\boldsymbol{q}}^k$ and $\hat{\boldsymbol{l}}^k$ denote the estimates of the subcarrier activation vector, of the DM indices and of the constellation symbols, respectively, correponding to the $k-$th block. The ML detector of (2) imposes a high decoding complexity especially for large values of N_a and $\binom{Q}{Q'}$. To avoid the computational burden, a logarithmic likelihood ratio (LLR)-based detector [3] is also proposed. We define the LLR $L(n_c)$ $(n_c = 1, 2, \ldots, N_c)$ of the n_c-th subcarrier as:

$$L(n_c) \triangleq \ln \frac{\sum_{\mu=1}^{\alpha} \Pr\left(\boldsymbol{\mathcal{K}}[n_c] = \boldsymbol{\mathcal{K}}_\mu \mid \overline{\boldsymbol{Y}}[n_c]\right)}{\Pr\left(\boldsymbol{\mathcal{K}}[n_c] = \mathbf{0} \mid \overline{\boldsymbol{Y}}[n_c]\right)}, \tag{3}$$

where $\boldsymbol{\mathcal{K}}_\mu$ belongs to the set of the $\alpha = 2^{\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor} \mathcal{L}^{Q'}$ values of the GSTFIM vector $\boldsymbol{\mathcal{K}}[n_c]$. Using the constraint $N_p = N_c/2$ for the LLR-based detector, the LLRs may be expressed as [3]:

$$L(n_c) = \frac{\left\|\overline{\boldsymbol{Y}}[n_c]\right\|^2}{2\sigma_n^2} + \text{jac}\left(\sum_{\mu=1}^{\alpha} e^{-\frac{\left\|\overline{\boldsymbol{Y}}[n_c] - \overline{\boldsymbol{H}}[n_c]\boldsymbol{\chi}\boldsymbol{\mathcal{K}}_\mu[n_c]\right\|^2}{2\sigma_n^2}} \right), \tag{4}$$

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where jac $(\bullet)$ and σ_n^2 denote the Jacobian logarithm of '$\bullet$' and the FD noise variance per dimension, respectively. Considering the subcarriers corresponding to the highest N_p LLR values to be activated, the source bits mapped to the DM indices and to the classic symbols may be estimated by the ML search:

$$\left(\hat{\boldsymbol{q}}, \hat{\boldsymbol{l}}\right) = \arg\min_{\boldsymbol{q},\boldsymbol{l}} \left\| \overline{\boldsymbol{Y}}[n_p] - \overline{\tilde{\boldsymbol{H}}}[n_p]\boldsymbol{\chi}\boldsymbol{\mathcal{K}}_{\mu}[n_p] \right\|^2. \quad (5)$$

Under specific system configurations, the proposed GSTFIM reduces to different IM schemes. Specifically, GSTFIM with $Q' = 1$ and $Q' = Q$ correspond to STSK-frequency index modulation (FIM) and linear dispersion code (LDC) [12]-FIM, respectively. STSK-FIM reduces to GSFIM and space-frequency IM (SFIM) under the conditions of $(T = 1)$ and $(T = 1,\ M' = 1)$, respectively. On the other hand, LDC-FIM reduces to spatial multiplexing-FIM (SMUXFIM) and time-frequency IM (TFIM) when $(T' = 1,\ M' = M)$ and $(T' > 1,\ M = 1)$, respectively.

III. GSTFIM: Performance Metrics

In this section, some insights on the achievable rate and the design criteria of the GSTFIM system and the complexity associated with the ML and LLR-based detectors are provided.

A. Achievable Rate of GSTFIM

The normalized throughput of GSTFIM is expressed as:

$$R_{\mathrm{G}} = \frac{KN_a}{N_cT}Q' \log_2 \mathcal{L} + \frac{KN_a}{N_cT}\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor + \frac{K}{N_cT}\left\lfloor \log_2 \binom{N_b}{N_a} \right\rfloor \quad \text{(bits/sec/Hz)} \quad (6)$$

The upper and the lower bounds of the normalized maximum achievable rate for a fixed value of Q' may be deduced as [8]:

$$R_{\mathrm{G}}^{\max} \leq \frac{K}{N_cT}\left[\log_2\left(\frac{e}{2\pi}\sqrt{\frac{N_b}{N_b - 1}} \right) + f_1\left(N_b, N_a^*\right) \right] + \frac{KN_a^*}{N_cT}\left[\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor + Q' \log_2 \mathcal{L} \right], \quad (7)$$

and

$$R_{\mathrm{G}}^{\max} \geq \frac{K}{N_cT}\left[\log_2\left(\frac{1}{e^2}\sqrt{\frac{2\pi}{N_b}} \right) + f_1\left(N_b, N_a^*\right) \right] + \frac{KN_a^*}{N_cT}\left[\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor + Q' \log_2 \mathcal{L} \right], \quad (8)$$

respectively, where $f_1\left(N_b, N_a^*\right) \triangleq N_b \log_2 N_b - N_a^* \log_2 N_a^* - \left(N_b - N_a^*\right) \log_2\left(N_b - N_a^*\right)$, and

$$N_a^* = \frac{N_b \cdot 2^{\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor} \mathcal{L}^{Q'}}{1 + 2^{\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor} \mathcal{L}^{Q'}}. \quad (9)$$

B. Design Criteria

1) Criterion 1: GSTFIM is capable of achieving rate higher than that achievable with OFDM-aided GSTSK, iff

$$N_b \geq 2^{\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor + 1} \mathcal{L}^{Q'}. \quad (10)$$

2) Criterion 2: GSTFIM can achieve rate greater than that achievable with LDC-FIM, iff $Q \geq 2\mathcal{L}$.

For $N_a = (N_b - 1)$, (6) reduces to

$$R_{\mathrm{G}} = \frac{KN_b}{N_cT}Q' \log_2 \mathcal{L} + \frac{KN_b}{N_cT}\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor + \frac{K}{N_cT}\left\lfloor \log_2 N_b \right\rfloor - \frac{K}{N_cT}Q' \log_2 \mathcal{L} - \frac{K}{N_cT}\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor. \quad (11)$$

When $N_b \geq 2^{\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor + 1} \mathcal{L}^{Q'}$, we have [8]:
$R_{\mathrm{G}} > \frac{1}{T}Q' \log_2 \mathcal{L} + \frac{1}{T}\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor$.

Conversely, $N_b < 2^{\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor + 1} \mathcal{L}^{Q'}$ can be expressed as

$$\log_2 N_b < Q' \log_2 \mathcal{L} + \left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor + 1. \quad (12)$$

Upon applying the relationship $\binom{N_b}{N_a} < \frac{N_b^{N_b - N_a}}{2^{N_b - N_a - 1}}$, we have [8]: $R_{\mathrm{G}} \leq \frac{1}{T}Q' \log_2 \mathcal{L} + \frac{1}{T}\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor$. Thus, Criterion 1 is proved. Criterion 2 can be proved in a similar manner.

C. Complexity of the Scheme

The complexity imposed by the ML detector of (2) evaluated in terms of real-valued multiplication operations (RMOs) required per bit of information transmitted may be given by

$$C_{\mathrm{ML}}^f = \frac{\left[4MNT^2Q + (4NTQ' + 2NT)\mathcal{L}^{Q'} \cdot 2^{\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor}\right] \times 2^{\left\lfloor \log_2 \binom{N_b}{N_a} \right\rfloor} N_a}{N_aQ' \log_2 \mathcal{L} + N_a\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor + \left\lfloor \log_2 \binom{N_b}{N_a} \right\rfloor}. \quad (13)$$

The complexity associated with the reduced-complexity soft-decision LLR detector of (5), taking the complexity of the LLR calculation in consideration, may be expressed as

$$C_{\mathrm{LLR}}^f = \frac{\left[4MNT^2Q + (4NTQ' + 2NT)\mathcal{L}^{Q'} \cdot 2^{\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor}\right] \times (N_c + 1) + N_c(2NT + Q\mathcal{L} + 1)}{KN_aQ' \log_2 \mathcal{L} + KN_a\left\lfloor \log_2 \binom{Q}{Q'} \right\rfloor + K\left\lfloor \log_2 \binom{N_b}{N_a} \right\rfloor}. \quad (14)$$

IV. Numerical Results

The bit-error-ratio (BER) performance, achievable rate and complexity of the GSTFIM scheme are numerically investigated and compared to those of the relevant benchmark schemes. Fig. 2 characterizes the BER performance of the ML and the soft-decision detector for the GSTFIM system. GSTSK $(2, 2, 2, 4)$ with QPSK modulation and $Q' = 2$ communicating over COST207-TU12 channel having a normalized Doppler frequency of $f_d = 0.01$ with the aid of $N_c = 64$ subcarriers is considered. OFDM-IM $(N_b = 8, N_a = 7, K = 8)$ for the ML and OFDM-IM $(N_b = 8, N_a = 4, K = 8)$ for the LLR detector are considered. Observe that both the detectors exhibit improved BER performance than the benchmark schemes, especially at high SNR. Since the GSTFIM codewords are sparse, advanced signal processing based detection method, such as compressed sensing [9], [13], may be employed for further improving the performance of the scheme.

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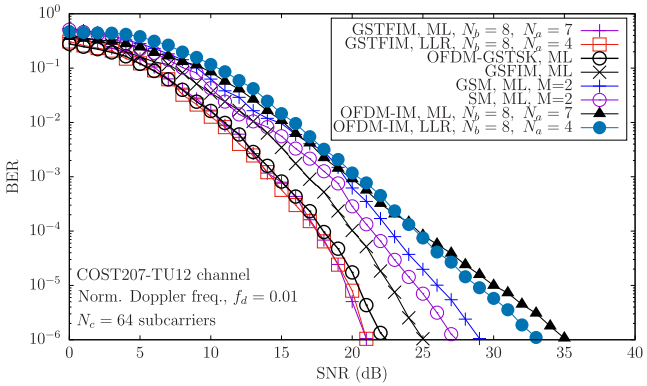


Fig. 2. BER performance of the GSTFIM with both the ML and the LLR detectors and of the classic OFDM-GSTSK with ML detector compared to the relevant benchmark schemes.

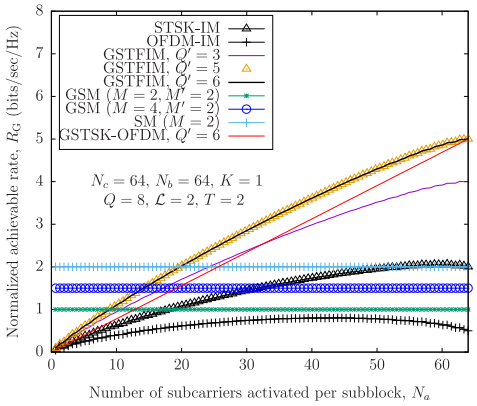


Fig. 3. Normalized achievable rate R_G as a function of N_a.

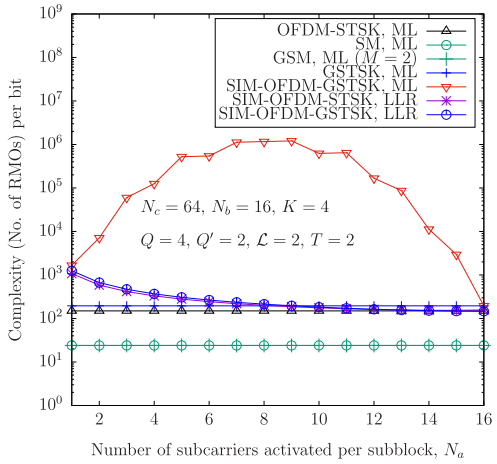


Fig. 4. Computational complexity of the relevant detectors.

The normalized achievable rate R_G of the GSTFIM scheme is characterized in Fig. 3 as a function of the number N_a of subcarriers activated for different numbers Q' of DMs activated. Note from Fig. 3 that the GSTFIM scheme is capable of achieving higher transmission rate than the relevant systems. Furthermore, the scheme achieves the maximum rate when less than the total number of subcarriers are activated.

To compare the computational complexity associated with the relevant detectors, Fig. 4 depicts the complexity as a function of the number of subcarriers activated N_a, while Q' is held constant at $Q' = 2$. Observe that the SM system imposes the lowest complexity. The ML detector of the classic-OFDM GSTSK imposes constant complexity with variations in N_a. The complexity associated with the ML detector is excessively high, when approximately 50% of the total subcarriers are activated. By contrast, the LLR-based detector imposes a substantially reduced complexity without any performance penalty. Thus the GSTFIM is capable of striking an appealing and flexible tradeoff between the rate achieved and the complexity imposed.

V. Conclusions

The concept of GSTFIM and its transceiver architecture have been proposed in this letter. The achievable rate of the proposed scheme has been analytically and numerically characterized. Important and helpful design criteria for the GSTFIM system have been provided. The complexity of the ML- as well as the soft-decision LLR-based detector has been quantified and numerically evaluated. The proposed system is capable of spreading the source information to space, time and frequency domains and of providing additional spectral efficiency with the aid of the multi-domain indices.

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