MIMO/-SPACE TIME CODES FOR POLYNOMIAL PHASE MODULATION WIRELESS COMMUNICATIONS

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ABSTRACT

Multiple input multiple output (MIMO) space time codes are used to increase the diversity or the number of degrees of freedom for improving the capacity (data rate) and reliability (error rate) of wireless communication over fading channels. In this paper, we propose a new structure of space time code designed for non-stationary constant-amplitude modulation formats using Polynomial Phase Signals (PPS). In PPS modulation, the information is carried in the coefficients of polynomials that modulate the phase of a carrier. Here, Space Time Code (STC) structures differ from the conventional STC design as they are designed based on the coefficients of the phase polynomials instead of the Euclidean distance of the transmitted signals. These MIMO/STC/PPS based systems offers variable rates of transmission, better performance and easier implementation than conventional formats.

1. INTRODUCTION

Non-stationary constant-amplitude Polynomial Phase Signals (PPS) are commonly used in active sonar and radar systems. For example Pulse Linear FM (LFM) and Quadratic FM (QFM) are used by such systems. Many algorithms have been developed to estimate the parameters of the PPS, e.g., Polynomial Phase Transform (*PT*) [1], High-order Instantaneous Moments (HIM), Spatial HIM (SHIM), Wigner-Ville Distribution, Generalized Ambiguity Function (GAF), High-order Ambiguity (HAF) [2] etc, and are being used in many practical applications related to radar and sonar. These tools are used for demodulation of PPS in mobile/wireless communication systems [3]. This work improves on the performance of PPS modulation system by exploiting the available diversity resources introduced by the use of MIMO/STC. In this work we assume a slow and non-selective frequency fading channel and we will discuss how to use PPS modulation in MIMO communication systems.

the original works of Telatar [4] and G. Foschini [5, 6] have proved that MIMO (Multiple Input Multiple Output) communications systems are more spectrum-efficient than SISO (Single Input Single Output) systems. The MIMO system's capacity increases linearly with the number of transmit antennas M or receive antennas N. Space time coding has been introduced in [7] to take advantages of the spatial diversity provided by multiple antennas. Since then significant research have been done to find codes that maximize the coding and diversity gain in different channel conditions as frequency selective or non-selective, fast or slow fading. Examples of these are space time block codes (STBC), space time trellis code (STTC) or multilevel space time code, super orthogonal space time trellis code, D-BLAST, V-BLAST Differential unitary space-time modulation, Space Time Codes for OFDM modulation, etc, [5, 7-10].

A common characteristic among these codes is that all are designed based on the Euclidean distance of the modulation signals. In conventional modulation formats QAM and M-PSK, the information is contained in each modulated symbol chosen from a M-symbol constellation. At the receiver, the demodulator uses either matched-filters or correlators to output the decision variables to detectors. The decoding is based on computation of the Euclidean distance metrics for all possible codewords. This decoding principle makes us to design space time code directly on the modulation signals, and in this paper we refer to this type of code as traditional STC. However the design rules of traditional STC cannot be applied to the PPS modulation systems. The demodulation of PPS signals uses a different algorithm, which is based on the computation of coefficient distances. Our paper will discuss a new design for STC, which we call "*Module structure space time code*" in PPS modulation systems. The module structure STC makes adaptation of the STC easier and flexibly. Our paper is organized as follow. A brief description of PPS modulation is given in section 2. Section 3 presents a general structure of module MIMO/STC/PPS system. Section 4 considers the performance analysis of MIMO/STC/PPS systems using the Discrete Polynomial Phase Transform (*DPT*) algorithm in the demodulation process. Section 5 shows the performance results of STC/PPS. The last section is "Conclusion".

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2. POLYNOMIAL PHASE SIGNAL MODULATION

PPS modulation refers to a constant envelop modulation where the information bits are mapped to the coefficient of a polynomial that it is used to modulate the phase of a carrier. By varying the coefficient alphabet and polynomial degree, we can change the rate of transmission with minimum bandwidth expansions and/or additional power resources.

Let $\varphi(t)$ represents a polynomial phase of order L, with a coefficients set $\{a^{(L)}, \cdots, a^{(1)}, a^{(0)}\}$, where the $a^{(l)}$ take real values, for all l. The polynomial phase $\varphi(t)$ in one symbol interval can be written as:

$$\varphi(t) = a^{(L)}\left(\frac{t}{T}\right)^L + \cdots + a^{(1)}\left(\frac{t}{T}\right) + a^{(0)}, \quad 0 \le t \le T \tag{1}$$

where T is the duration of the signaling interval. Then the modulated signal $s(t)$ can be written as:

$$s(t) = \sum_n A\cos\big(\omega_c(t - nT) + \varphi_n(t - nT)\big) \tag{2}$$

$$\text{with} \quad \varphi_n(t) = \begin{cases} \displaystyle\sum_{l=0}^{L} a_n^{(l)}\left(\frac{t}{T}\right)^l, & 0 \le t \le T \\ 0 & otherwise \end{cases} \tag{3}$$

where ω_c is carrier frequency, A is the carrier amplitude. The $(L+1)$ coefficients $a^{(l)}$ in each symbol interval T contain the information. Each coefficient might have its own alphabet. The serial information bits are first converted to $(L+1)$ parallel paths, and then mapped to an appropriate coefficient value. The data rate of each path is proportional to the logarithm of A_l (alphabet size of the coefficient $a^{(l)}$).

At the receiver, the demodulation uses an iterative decoding algorithm called Polynomial Phase Transform (*PT*) that decodes the information by estimating the coefficients of the polynomial from the received signal [1]. This algorithm is flexible enough to handle a polynomial phase of any order and a coefficient alphabet within a given bound. This coefficient estimation algorithm can be implemented using a Discrete Polynomial Phase Transform (*DPT*) [11].

The basic principle of demodulation is to successively reduce the order of the polynomial phase signal to a first order polynomial (linear frequency modulation LFM [12]), then using the Fourier transform to retrieve the frequency component which is proportional to the value of the polynomial's highest coefficient.

As an example let consider a second order polynomial with coefficients $a^{(0)}, a^{(1)}, a^{(2)}$. The polynomial phase is:

$$\varphi(t) = a^{(2)}\left(\frac{t}{T}\right)^2 + a^{(1)}\left(\frac{t}{T}\right) + a^{(0)}$$

In the decoding process, we multiply the complex analytic form of the received signal by a synchronous carrier $\{e^{-j\omega_c t}\}$, which gives the signal $s(t) = Ae^{j\varphi(t)}$.

We next form the product of $s(t)$ and $s^*(t-\tau)$.

$$\begin{aligned} s_2(t) &= s(t)s^*(t-\tau) = A^2 e^{j[\varphi(t) - \varphi(t-\tau)]} \\ &= A^2 e^{j[2a^{(2)}\tau(t/T)^2 + a^{(1)}(\tau/T) - a^{(2)}(\tau/T)^2]} \end{aligned} \tag{4}$$

The Fourier transform of $s_2(t)$ will give a peak at the frequency value $\Omega_2 = 2a^{(2)}\tau/T^2$. The coefficient estimate $\hat{a}^{(2)}$ is then obtained as:

$$\hat{a}^{(2)} = \Omega_2 \frac{T^2}{2\tau} \tag{5}$$

We then eliminate $a^{(2)}$ to reduce the polynomial phase $\varphi(t)$ to a first order polynomial and continue applying the Fourier transform to retrieve the estimated $\hat{a}^{(1)}$.

The above procedure can be generalized to an arbitrary order polynomial phase signal.

Next we present some results of using the *DPT* algorithm in demodulation that are related to our analysis. The variance of the error of estimation of the highest order coefficient $a^{(L)}$ is given by [11]:

$$var(\hat{a}^{(L)}) \approx \Xi\{(a^{(L)} - \hat{a}^{(L)})^2\} \approx (k_L SNR)^{-1} \tag{6}$$

with SNR is Signal to Noise Ratio and the constant:

$$k_L{}^{-1} = \{(6L^{2L+1})\ \mathrm{div}\ (L!^2 K)\}\binom{2L-2}{L-1}$$

K: Number of samples taken in one symbol interval T.

div : Representing the division operation.

Then the Symbol Error Probability (SER) of $a^{(L)}$ is:

$$P_e(\hat{a}^{(L)}) = P(\hat{a}^{(L)} \ne a^{(L)}) = Q(\sqrt{k_L SNR}) \tag{7}$$

Assuming that coefficients of order higher than l have been decoded correctly

$$\boldsymbol{\Xi}^{\boldsymbol{l}} = \{\hat{a}^{(i)} = a^{(i)}\,;\ \forall\, i = (l+1) \div L\},$$

the approximation (6) can be applied for any coefficients $a^{(l)}$ as follows

$$\begin{aligned} var\{\hat{a}^{(l)} | \boldsymbol{\Xi}^{\boldsymbol{l}}\} &\approx \{(6l^{2l+1})\ \mathrm{div}\ (l!^2 K)\}\binom{2l-2}{l-1} SNR^{-1} \\ &= (k_l SNR)^{-1} \quad , \quad l = 0 \div L-1 \end{aligned} \tag{8}$$

with

$$k_l{}^{-1} = \{(6l^{2l+1})\ \mathrm{div}\ (l!^2 K)\}\binom{2l-2}{l-1}$$

In each stage of successive decoding of the polynomial coefficients, we use the same number of K samples and obtain the Symbol Error Rate of $a^{(l)}$ similar to (7), given the condition $\boldsymbol{\Xi}^{\boldsymbol{l}}$:

$$P_e(\hat{a}^{(l)} | \boldsymbol{\Xi}^{\boldsymbol{l}}) = Q(\sqrt{k_l SNR}), \quad l = 0 \div (L-1) \tag{9}$$

3. MODULE STRUCTURE OF SPACE TIME CODE

Let's consider a MIMO system with M transmit and N receive antennas.

At the transmitter in every symbol interval, a set of M $(L+1)$_coefficient vectors

$$\boldsymbol{a_m} = \left(a_m^{(L)}, a_m^{(L-1)}, \dots, a_m^{(0)}\right), \qquad m = 1, \dots, M$$

are modulated into M polynomial phase signals. These M PPS signals are sent out simultaneously from M antennas. The sets of M coefficient vectors are encoding outputs of our *module structure MIMO/Space Time Code*.

Figure 1 shows the block diagram of the module MIMO/STC for PPS modulation. We use the term "*module*" to emphasize the property of independence of each module $STC\_\boldsymbol{a}^{(l)}$ (Space Time Code of coefficient $a^{(l)}$) in the design. The structure consists of L $STC\_\boldsymbol{a}^{(l)}$ modules. The output of each $STC\_\boldsymbol{a}^{(l)}$ module is sequence of $M \times Q$ matrices of coefficient $a^{(l)}$ (assuming block space time code is implemented), with M transmit antennas representing the spatial dimension and Q, the length of the code, representing the temporal dimension. The column elements in each row are grouped consecutively, forming sets of coefficients that are input to the PPS modulators as we can see in the figure 1.

At the receiver, the demodulation and decoding of the $STC\_\boldsymbol{a}^{(l)}$ module are carried out in descending order of coefficient. For example using the Discrete Polynomial Transform algorithm (DPT) described in [11], we decode the module $STC\_\boldsymbol{a}^{(L)}$ first, then $STC\_\boldsymbol{a}^{(L-1)}$ and so on. As a result, the statistical properties of the estimated coefficients are different.

Besides that, different coefficients may have different alphabets.

$$A_i \neq A_j \quad , \text{ with } i \neq j$$

A_l is alphabet of coefficient $a^{(l)}$.

For these reasons, the module STC structure is a suitable choice. We can design and adjust each module $STC\_\boldsymbol{a}^{(l)}$ differently and independently.

The design criteria include type, structure, code rate, diversity, multiplexing of space time code [7, 9, 10, 13].

All design criteria used for conventional modulations in Tarokh's work or others can be modified and applied to a module $STC\_\boldsymbol{a}^{(l)}$.

However, while the symbol of conventional modulations has complex value (2-D space), the coefficients $a^{(l)}$ take real values (1-D space). It makes the design of module STC simpler than conventional STC.

4. PERFORMANCE ANALYSIS AND DESIGN OF MODULE SPACE TIME CODE STRUCTURE

The performance of the module STC varies with the demodulation/decoding algorithms used to estimate the coefficients. In this paper our analysis applies for the Polynomial Transform (PT) algorithm.

Similar to Tarokh's work in [7], we calculate the Pairwise Error Probability (PEP) which is the probability of decoding an error codeword $\boldsymbol{e}$ given the codeword $\boldsymbol{c}$ transmitted. The approximation is good as long as the decoding errors of closest codes are dominant terms in the calculation of error rate.

We will find the PEP for each module $STC\_\boldsymbol{a}^{(l)}$, starting with module $STC\_\boldsymbol{a}^{(L)}$**,** the module of the highest order coefficient. The evaluation is similar to a conventional STC [7, 14, 15] except that the SNR is replaced by $k_L SNR$ at expression (6). We have the average PEP in a Rayleigh fading channel as:

$$P^{ave}\left(\boldsymbol{a}_c^{(L)} \rightarrow \boldsymbol{a}_e^{(L)}\right) \leq \left\{\prod_{m=1}^{M}\left(1 + \left(\lambda_m^{(L)} k_L SNR \text{ div } 4\right)\right)\right\}^{-N}$$

with (10a)

$$k_L{}^{-1} = \{(6L^{2L+1}) \text{ div } (L!^2\, K)\}\binom{2L-2}{L-1}$$

$\boldsymbol{a}_c^{(L)}$: Correct codeword matrix of coefficient $a^{(L)}$.

$\boldsymbol{a}_e^{(L)}$: Incorrect codeword matrix which is the result of a decoding error.

$\lambda_m^{(L)}$ Eigen-values of the Hermitian matrix $\boldsymbol{K}\boldsymbol{a}_{ce}^{(L)}$ which is

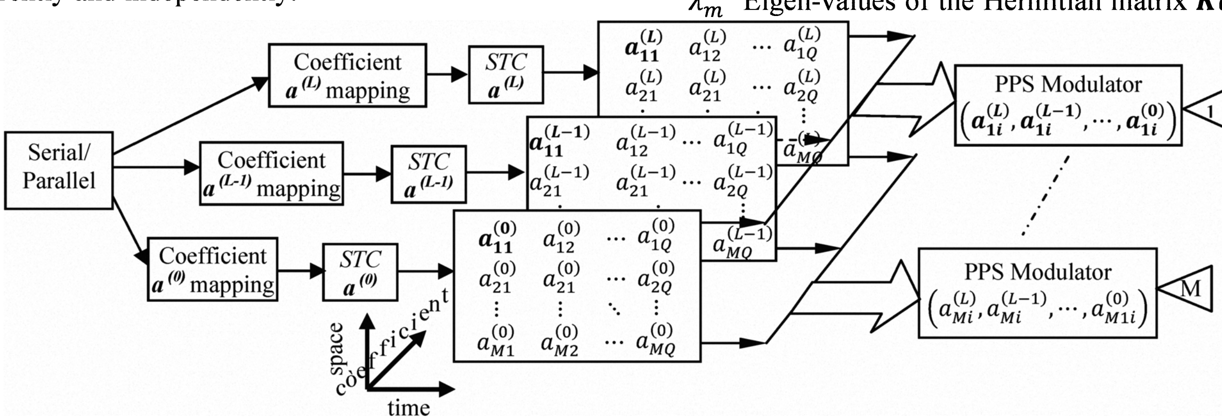


Figure 1. Block diagram of module structure space time code for PPS modulation.

defined as

$$\boldsymbol{Ka}_{ce}^{(L)} = \left(\boldsymbol{a}_c^{(L)} - \boldsymbol{a}_e^{(L)}\right)\left(\boldsymbol{a}_c^{(L)} - \boldsymbol{a}_e^{(L)}\right)^T \quad (10b)$$

The value of $a_{mn}^{(L)}$ in (10b) is scaled-value obtained by dividing the real-value of the coefficient by a constant $\nu^{(L)}$,

$$\nu^{(L)} = \{L! \text{ div } (\pi K L^{L-1})\} \quad (11)$$

which is the result of applying the Discrete Polynomial Transform algorithm for decoding the coefficient $a^{(L)}$ [11]. The real-value of the coefficient $a^{(L)}$ is between the range of $(-\nu^{(L)}, \nu^{(L)})$ [11].

From (10a) we conclude that the rules of the design for the highest-order coefficient $STC\_\boldsymbol{a}^{(L)}$ module are similar to the rules applied for the conventional modulation formats. These are:

- Maximize the rank r of the matrix $\boldsymbol{Ka}_{ce}^{(L)}$.
- Maximize the minimum of the product of all the non-zero eigen-values of the matrix $\boldsymbol{Ka}_{ce}^{(L)}$.

The main difference between the module $STC\_\boldsymbol{a}^{(L)}$ and conventional STC is that the former is designed based on the value of the coefficient $a^{(L)}$ (coefficient distance) while the later based on the Euclidean distance between the signals (signal distance).

Next, we evaluate the performance of modules $STC\_\boldsymbol{a}^{(l)}$ $(l < L)$ with consideration of the error propagation effect caused by sequential descending order decoding principle of the Polynomial Transform (*PT*) algorithm stated in the previous page. First, we compute the conditional PEP and then average the result over the random transmission channel $\boldsymbol{\mathcal{H}}$. We will find the actual PEP of modules $STC\_\boldsymbol{a}^{(l)}$ using the fact that if we mistakenly decode the module $STC\_\boldsymbol{a}^{(l+1)}$, it is unlikely to decode correctly lower-order modules $STC\_\boldsymbol{a}^{(l-k)}$ $(l \geq k \geq 0)$. Interested readers should refer to [11, 16] for more details about *PT* algorithm methods.

Given that $\widehat{\boldsymbol{a}}^{(l+1)}$ is decoded correctly, then using the Gaussian tail function the conditional probability PEP of module $STC\_\boldsymbol{a}^{(l)}$ is upper-bounded by:

$$P\left(\boldsymbol{a}_c^{(l)} \to \boldsymbol{a}_e^{(l)} \mid (\widehat{\boldsymbol{a}}^{(l+1)} = \boldsymbol{a}^{(l+1)})\right) \leq \exp\left\{-d^2(\boldsymbol{a}_c^{(l)}, \boldsymbol{a}_e^{(l)})(k_l SNR \text{ div } 4)\right\} \quad (12)$$

with

$$d^2\left(\boldsymbol{a}_c^{(l)}, \boldsymbol{a}_e^{(l)}\right) = \sum_{n=1}^{N}\sum_{q=1}^{Q}\left|\sum_{m=1}^{M} h_{nm}\left(a_{c,mq}^{(l)} - a_{e,mq}^{(l)}\right)\right|^2 \quad (13)$$

h_{nm}: Multiplicative coefficient of the channel between transmit antenna *m* and receive antenna *n*.

We evaluate the performance of module $STC\_\boldsymbol{a}^{(l)}$ by finding $P_e(\widehat{\boldsymbol{a}}^{(l)})$, the rate of error in decoding the matrix $\widehat{\boldsymbol{a}}^{(l)}$.

$$P_e(\widehat{\boldsymbol{a}}^{(l)}) = P_e\{\widehat{\boldsymbol{a}}^{(l)} | \widehat{\boldsymbol{a}}^{(l+1)} = \boldsymbol{a}^{(l+1)}\}.P\{\widehat{\boldsymbol{a}}^{(l+1)} = \boldsymbol{a}^{(l+1)}\} + P_e\{\widehat{\boldsymbol{a}}^{(l)} | \widehat{\boldsymbol{a}}^{(l+1)} \neq \boldsymbol{a}^{(l+1)}\}.P\{\widehat{\boldsymbol{a}}^{(l+1)} \neq \boldsymbol{a}^{(l+1)}\} \quad (14)$$

Remembering the fact that if $\widehat{\boldsymbol{a}}^{(l+1)}$ is wrongly decoded by $STC\_\boldsymbol{a}^{(l+1)}$ then $STC\_\boldsymbol{a}^{(l)}$ is unlikely to decode $\widehat{\boldsymbol{a}}^{(l)}$ correctly.

$$P_e\{\widehat{\boldsymbol{a}}^{(l)} | \widehat{\boldsymbol{a}}^{(l+1)} \neq \boldsymbol{a}^{(l+1)}\} = \{(V_l - 1) \text{ div } V_l\} \quad (15)$$

V_l: Number of codeword matrices encoded by $STC\_\boldsymbol{a}^{(l)}$.

$$V_l = 2^{QR_l} \quad (16)$$

R_l: Code rate [*bit/symbol*] of $STC\_\boldsymbol{a}^{(l)}$.

The first term on the right hand side of (14) will be replaced by the PEP of the closest codeword matrices. Let's assume a codeword $\boldsymbol{a}_c^{(l)}$ having d_l closest codewords called $\boldsymbol{a}_e^{(l)}$, (d_l depends on the code rate r_l and the code's minimum distance). Substituting (12, 15, 16) into (14), we obtain:

$$P_e(\widehat{\boldsymbol{a}}^{(l)}) \leq d_l \exp\left(-\boldsymbol{d}^2\left(\boldsymbol{a}_c^{(l)}, \boldsymbol{a}_e^{(l)}\right)(k_l SNR \text{ div } 4)\right) (1 - P_e(\widehat{\boldsymbol{a}}^{(l+1)})) + \{(2^{QR_l} - 1) \text{ div } 2^{QR_l}\} P_e(\widehat{\boldsymbol{a}}^{(l+1)}) \quad (17)$$

with

$$k_l^{-1} = \{(6l^{2l+1}) \text{ div } (l!^2 K)\}\binom{2l-2}{l-1}$$

To compute the upper bound on the average probability of error $P_e^{ave}(\widehat{\boldsymbol{a}}^{(l)})$, we simply average $P_e(\widehat{\boldsymbol{a}}^{(l)})$ with respect to the complex Gaussian random variables h_{nm}. From (14), (17) and the assumption of $P_e(\widehat{\boldsymbol{a}}^{(l+1)})$ is much smaller than 1, we get:

$$P_e^{ave}(\widehat{\boldsymbol{a}}^{(l)}) \leq d_l \mathbb{E}_{\boldsymbol{\mathcal{H}}}\left\{\exp\left(-d^2\left(\boldsymbol{a}_c^{(l)}, \boldsymbol{a}_e^{(l)}\right)(k_l SNR \text{ div } 4)\right)\right\} + \{(2^{QR_l} - 1) \text{ div } 2^{QR_l}\}.P_e^{ave}(\widehat{\boldsymbol{a}}^{(l+1)}) \quad (18)$$

The second term in (18) does not affect the design of the module $STC\_\boldsymbol{a}^{(l)}$. Focusing on the first term, the expectation has a format similar to the expression (7) in Tarokh's work [14]. And we can conclude that the rules of design for the modules $STC\_\boldsymbol{a}^{(l)}$, $l = 0 \div (L-1)$, are the same as the rules of design for the module $STC\_\boldsymbol{a}^{(L)}$, or similar to the conventional STC.

Evaluating (18) in a Rayleigh fading channel we get:

$$P_e^{ave}(\widehat{\boldsymbol{a}}^{(l)}) = d_l\left(\prod_{m=1}^{M}\left(1 + \lambda_m^{(l)}(k_l SNR \text{ div } 4)\right)\right)^{-N} + \{(2^{QR_l} - 1) \text{ div } 2^{QR_l}\}.P_e^{ave}(\widehat{\boldsymbol{a}}^{(l+1)}) \quad (19)$$

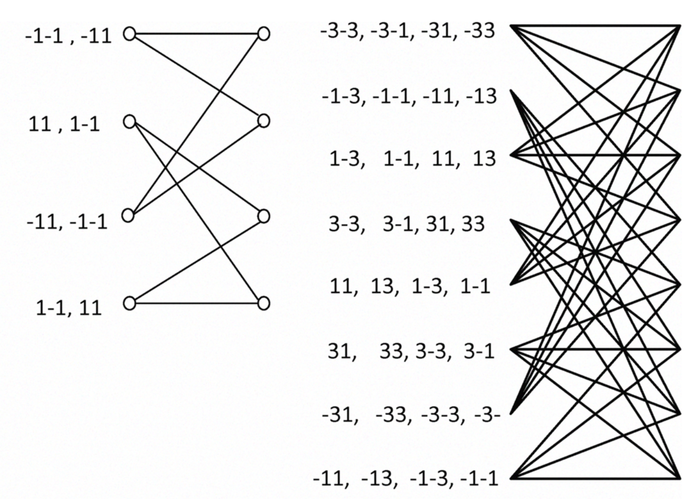
We conclude the performance evaluation section with two remarks.

- The design of all $STC\_\boldsymbol{a}^{(l)}$ modules is based on the criteria of the minimum rank and minimum determinant of the $\boldsymbol{Ka}_{ce}^{(l)}$ matrix, similar to the design of STC for the conventional modulation formats. But differs from conventional STCs whose constructions are based on a 2-D complex signal space. The $STC\_\boldsymbol{a}^{(l)}$ module designs are based on 1-D real-value coefficient space, which means simpler designs and calculations.
- The performance of the systems using STC module for PPS modulation reflects the error propagation from higher order to lower order coefficients.

5. PERFORMANCE RESULTS

We compare the performance of the STC module design to Tarokh's STC code (TSC for short) which is a 8-state space time trellis code (STTC) using 8-PSK modulation and for $M = 2$ and $N = 2$. Our system uses a quadratic PPS signal modulation. The simulations use Frame Error Rate (FER) as the measurement unit. Each frame has a length of 130 transmission symbols. Figure 2 shows two trellis diagrams of space time code that we use for our $STC\_\boldsymbol{a}^{(l)}$ modules.

Our module STC structure is simpler in design and less complexity to decode than conventional STC because of two following reasons:



a) 2-levels, 4-states
1 $bit/symbol$ code rate

b) 4-levels, 8-states
2 $bits/symbol$ code rate

Figure 2. Two trellis diagram for module $STC\_\boldsymbol{a}^{(l)}$

- If two systems have the same performance e.g. same FER rate and are implemented with two different space time codes, one with module STC and the other with conventional STC, the former always has fewer states in trellis than the later. As a result, module STC structure is simpler in design and less complexity in decoding than conventional STC.
- The distance calculations in the design of the module structure space time code $STC\_\boldsymbol{a}^{(l)}$ are carried on the value of the coefficient $a^{(l)}$. Because the coefficient $a^{(l)}$ only takes real value, the designs are carried on 1-D space. On the contrary, the Euclidean distance of conventional space time code is calculated on 2-D space. Again we have simpler design and lower complexity.

The simulation result in figure 3 shows that the module STC structure has good gain compared to the TSC code in the low SNR region. For example at Frame Error Rate (FER) of $10^{-0.9}$, our code structure achieves 1db better than a TSC code. The lower SNR is, the larger gain our codes achieve.

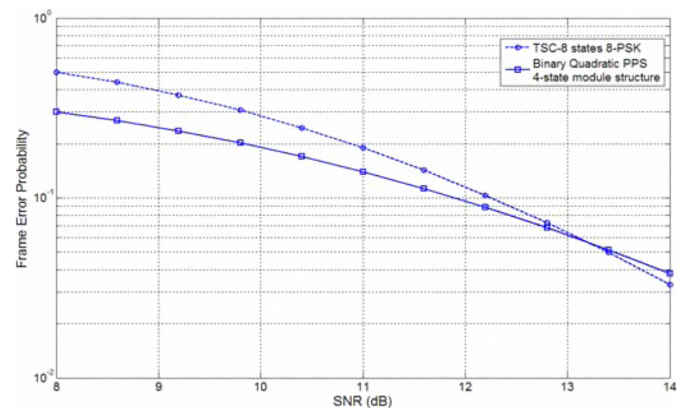


Figure 3. Overall performance of module structure STC

Following are some more simulation results of our module STC structure using PPS signals with different coefficient-alphabet sizes.

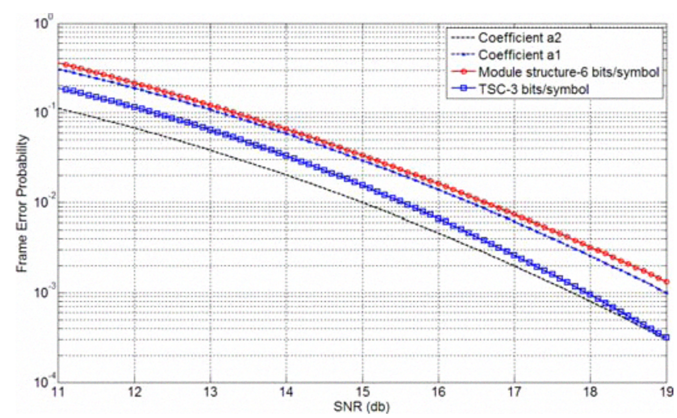


Figure 4. 2 Tx, 2 Rx; quadratic polynomial; $STC\_\boldsymbol{a}^{(2)}, STC\_\boldsymbol{a}^{(1)}, STC\_\boldsymbol{a}^{(0)}$: 4-levels, 8-states; 6 bits/symbol code rate.

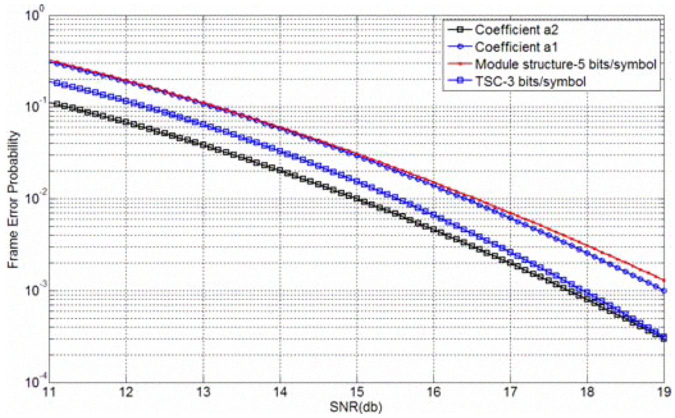


Figure 5. 2 Tx , 2 Rx; quadratic polynomial; $STC\_\boldsymbol{a}^{(2)}$, $STC\_\boldsymbol{a}^{(1)}$: 4-levels, 8-states. $STC\_\boldsymbol{a}^{(0)}$: 2-levels, 4-states; 5 bits/symbol code rate.

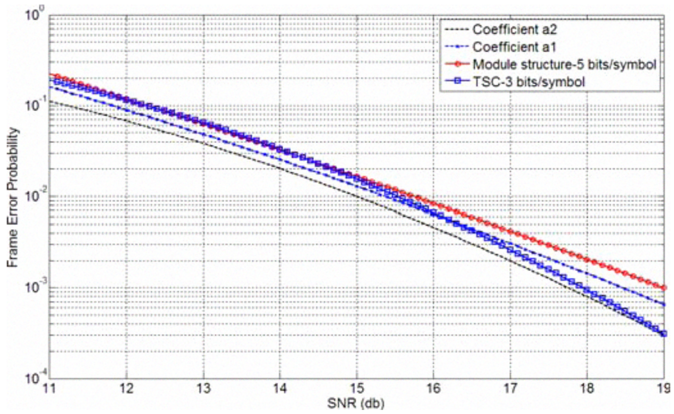


Figure 6. 2 Tx , 2 Rx; quadratic polynomial; $STC\_\boldsymbol{a}^{(2)}$, $STC\_\boldsymbol{a}^{(0)}$: 4-levels, 8-states. $STC\_\boldsymbol{a}^{(1)}$: 2-level, 4-states; 5 bits/symbol code rate.

Notes:

- The simulation results are calculated with Frame Error Rate (FER) and each frame consists of 130 symbols. So the actual Symbol Error Rate (SER) is considerably lower than FER value.
- The FER rate in the figure 3 is impractically high because we use the simple STC in the aim of demonstration. The results are improved with more complex STC design.

6. CONCLUSION

With the understanding of PPS formats and the available tools for the estimation and decoding developed in measurement applications we present a new module structure of Space Time Code (STC) for PPS signal modulation. Beside the well-known benefits of conventional STCs on system's capacity and reliability as shown in the section of performance results, the module STC structure is flexible, adjustable, and easy to implement. It is suitable for adaptive systems.

Our module structure space time code for PPS works well in the low SNR region. The low SNR region is typical for mobile wireless communications because of the limitations in power.

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