Simplified Receiver Design for STBC Binary Continuous Phase Modulation

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***Abstract*— Existing space-time codes have focused on multiple-antenna systems with linear modulation schemes such as phase-shift keying and quadrature amplitude modulation. Continuous phase modulation (CPM) is an attractive scheme for digital transmission because of its constant envelope which is needed for power efficient transmitters. Recent research has shown that space-time coded CPM can achieve transmit diversity to improve performance while maintaining the compact spectrum of CPM signals. However, these efforts mainly combine space-time coding (STC) with CPM to achieve spatial diversity at the cost of a high decoding complexity. In this paper, we design space-time block codes (STBC) for binary CPM with modulation index $h = 1/2$ and derive low-complexity receivers for these systems. The proposed scheme has a much lower decoding complexity than STC CPM with the Viterbi decoder and still achieves near-optimum error performances.**

***Index Terms*— Continuous phase modulation, wireless communications, space-time block codes, low-complexity decoding.**

I. Introduction

CONTINUOUS phase modulation (CPM) [1] is a very attractive scheme for wireless communications because of its constant envelope, compact spectrum, and flexible bandwidth-performance tradeoffs. Binary CPM (BCPM) with a modulation index $1/2$ (for brevity, we call it BCPM0.5 in this paper) is widely used in wireless communication systems. For example, Gaussian minimum shift keying (GMSK) has been used in the global system for mobile communications (GSM). The duration of the impulse response of the pre-modulation filter L is one of the parameters that control the signal spectrum; increasing L results in a more compact spectrum at the expense of a higher bit-error rate (BER) under the same bit-energy-to-noise-density ratio $E_b/\mathcal{N}_0$ due to the increased level of inter-symbol interference (ISI). These properties make BCPM0.5 an attractive scheme to use, especially in power-constrained applications. Appropriately designed space-time codes for BCPM0.5 add diversity while having constant envelope properties and low receiver complexity.

Space-time code design criteria for general CPM are developed by Zhang and Fitz [2], and a reduced-complexity receiver for multiantenna layered space-time systems with binary CPM is described by Zhao and Giannakis [3]. Space-time coded MSK is analyzed by Cavers [4]. In particular, the relationship between offset and non-offset modulation formats and the effect of pulse shape are explored. Orthogonal space-time coding with CPM for systems with two transmit antennas is introduced in [5], [6] to reduce decoding complexity. In [7], concatenation of convolutional code and MSK-STBC is studied. Later, the Alamouti code [8] is applied to orthogonal frequency shift keying in [9], and blind detection scheme of MSK with the Alamouti code in fast fading channels is proposed in [10]. Viterbi decoders are used to recover the transmitted symbols in these schemes and decoding complexity is still relatively high.

In this paper, we design orthogonal space-time block codes (OSTBC) [8], [11] for BCPM with a modulation index 0.5. The orthogonal code design is based on Laurent decomposition of BCPM signals combined with differential precoding. We then derive a simplified decoder with a linear finite impulse response (FIR) filter to reduce ISI inherent in BCPM0.5 signals with two transmit antennas. The proposed scheme significantly improves the error performance. For STBC BCPM0.5 with more than two transmit antennas, decoding based on FIR filtering becomes inefficient. Therefore, we derive a soft decision feedback decoding scheme to simplify the receiver while maintaining a satisfactory performance. The STBC BCPM0.5 designed together with the proposed decoding schemes has a much lower complexity than STC BCPM0.5 with Viterbi decoding, especially when L of the pre-modulation filter is large, while their performances are similar.

The design of STBC with BCPM0.5 based on Laurent decomposition is briefly discussed in Section II. Two low-complexity, ISI-resistant decoding schemes for STBC BCPM0.5 are presented in Section III. Section IV provides numerical results to assess the diversity and error performance of the proposed code over quasi-static channels as well as time-varying fading channels and the effect of various system parameters and receiver designs.

II. Code Design Based on Laurent Decomposition

Consider a system with M transmit antennas and one receive antenna. The received signal is expressed as

$$r(t) = \sqrt{\frac{1}{M}}\boldsymbol{h}^T\boldsymbol{s}(t) + n(t) \qquad (1)$$

where $(\cdot)^T$ denotes transpose, $n(t)$ is a complex Gaussian noise with power spectral density $\mathcal{N}_0$, $\boldsymbol{h} = [h_1, h_2, \cdots, h_M]^T$ is the channel coefficient vector with h_m being the coefficient from transmit antenna m to the receive antenna, and $\boldsymbol{s}(t) =$

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$[s_1(t), s_2(t), \cdots, s_M(t)]^T$ is the transmitted signal vector with $s_m(t)$ being the CPM signal from transmit antenna m.

The complex baseband binary CPM signal $s(t)$ can be written as the sum of $K = 2^{L-1}$ pulse-amplitude modulation (PAM) signals as [12]–[14]

$$s(t) = \sqrt{\frac{E_b}{T}} \sum_{k=0}^{K-1} \sum_{n=0}^{N-1} e^{j\pi h a_{k,n}} c_k(t - nT), \ t \in [LT, NT] \tag{2}$$

where T is the bit interval, N is the number of consecutive bits, h is the modulation index, which equals 0.5 for the signaling scheme of interest in this paper. The pseudo-symbols $\{a_{k,n}\}$ can be derived from the information bits α_n, and $c_k(t)$ is the expression for the kth PAM pulse.

Among the K terms of PAM signals in Eq. (2), the first term $c_0(t)$ usually contains the bulk of the total signal energy [13], and the length of $c_0(t)$ is $(L+1)T$ [14]. Therefore, considering only the first term will significantly reduce the decoding complexity at the expense of a small performance loss. By keeping only the major term, the binary CPM signal is approximated as

$$s(t) \approx \sqrt{\frac{E_b}{T}} \sum_{n=0}^{N-1} e^{j\pi h a_{0,n}} c_0(t - nT). \tag{3}$$

For example, for MSK, a special case of GMSK with $L = 1$, $K = 2^{L-1} = 1$ and the Laurent decomposition consists of only the $c_0(t)$ term, which is expressed as

$$c_0(t) = \begin{cases} \sin(\frac{\pi t}{2T}), & t \in [0, 2T) \\ 0, & \text{otherwise.} \end{cases} \tag{4}$$

This also leads to the well-known interpretation of MSK as offset-QPSK in which the pulse shape is a half-cycle sinusoid with period $4T$ [12].

Information bits $\hat{\alpha}_n$ can be decoded by differential decoding of the estimated pseudo-symbols $\hat{a}_{0,n}$ and $\hat{a}_{0,n-1}$. Note that $\hat{a}_{0,n} = \sum_{i=0}^{n} \hat{\alpha}_i$ [12], [13]. In this case, the Viterbi algorithm can be applied because of the memory structure of CPM signals. However, there is a performance loss due to differential detection. This loss can be eliminated by a data precoding algorithm applied to the non-return-to-zero (NRZ) source data symbols prior to BCPM0.5 modulation [15]. The precoding scheme is described briefly as follows.

Let d_k denote the equally probable source data bits at time $t = kT$. The input to the BCPM0.5 modulator is formed as $\alpha_k = (-1)^k d_k d_{k-1}$ with $d_{-1} = 1$. Since symbols d_k and α_k have identical statistics, BCPM0.5 signals with and without precoding have the same power spectrum. When precoding is applied to the source data symbols d_k, we have

$$e^{j\pi h a_{0,n}} = \begin{cases} jd_n, & n = 0, 2, 4, \cdots \\ d_n, & n = 1, 3, 5, \cdots \end{cases} \tag{5}$$

With precoding, the memory in BCPM0.5 is eliminated; thus we can use a linear receiver, rather than a Viterbi decoder, to decode the precoded BCPM0.5 signals. Now, we can readily apply OSTBC for BCPM0.5 with precoding.

A modified Alamouti code obtained by taking the transpose of the original Alamouti code can be used for 2 transmit antennas with BCPM0.5. The modified code allows us to keep the transmitted signal from the first antenna the same as the transmitted signal in a single-input single-output (SISO) antenna system. From the first transmit antenna, we transmit $d_0, d_1, d_2, d_3, \cdots$; for the second transmit antenna, we transmit $-d_1, d_0, -d_3, d_2, \cdots$, as illustrated in Fig. 1. Obviously, signals from the two transmit antennas have the same spectrum.

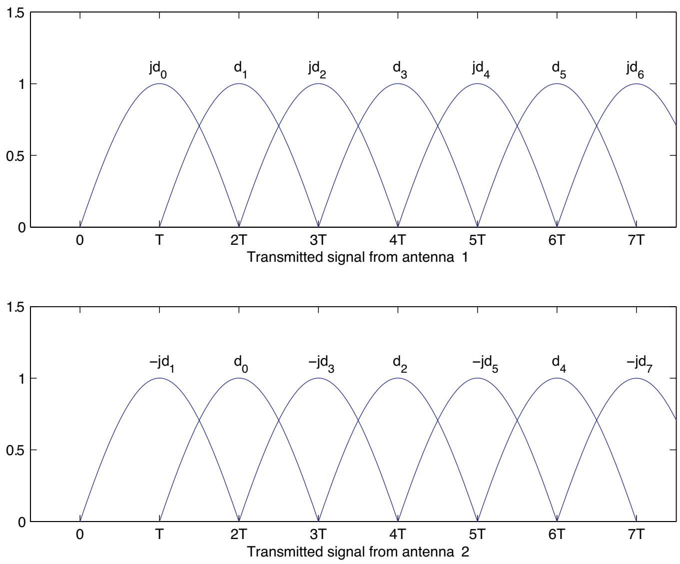

Fig. 1. The modified Alamouti scheme ($M = 2$) for MSK with precoding based on Laurent decomposition.

III. Low-Complexity Decoding

A. Receiver without an FIR filter

After passing through a real-valued matched filter $c_0(-t)$, the received signal is sampled at time $t = kT$. The output of the sampler is expressed as $r_i = \int_{iT}^{(i+L+1)T} r(t) c_0(t - iT) dt$. Space-time decoding for linear modulations can also be applied on r_i.

Neglecting ISI in detecting BCPM0.5 signals leads to the simplest receiver whose decoding complexity is the same as that of linear modulations. However, ISI increases as the number of transmit antennas and L increase, which will result in performance degradation.

Let us consider MSK with the Alamouti space-time coding scheme as an example. We use the outputs of the sampler after the matched filter (i.e., r_{2n} and r_{2n+1}) to decode symbols d_{2n} and d_{2n+1}. From Fig. 1, we have

$$\begin{aligned} r_{2n} = &\sqrt{\frac{E_b}{MT}} \left[h_1 \left(j d_{2n} c_{\text{full}} + (d_{2n-1} + d_{2n+1}) c_{\text{half}} \right) \right. \\ &\left. + h_2 \left(-j d_{2n+1} c_{\text{full}} + (d_{2n-2} + d_{2n}) c_{\text{half}} \right) \right] + n_{2n} \end{aligned} \tag{6a}$$

$$\begin{aligned} r_{2n+1} = &\sqrt{\frac{E_b}{MT}} \left[h_1 \left(d_{2n+1} c_{\text{full}} + j(d_{2n} + d_{2n+2}) c_{\text{half}} \right) \right. \\ &\left. + h_2 \left(d_{2n} c_{\text{full}} - j(d_{2n+1} + d_{2n+3}) c_{\text{half}} \right) \right] + n_{2n+1} \end{aligned} \tag{6b}$$

where

$$c_{\text{full}} = \int_0^{(L+1)T} c_0^2(t)dt = T \tag{7a}$$

$$c_{\text{half}} = \int_0^{(L+1)T} c_0(t)c_0(t-T)dt \tag{7b}$$

$$n_{2n} = \int_{2nT}^{(2n+L+1)T} n(t)c_0(t-2nT)dt \tag{7c}$$

$$n_{2n+1} = \int_{(2n+1)T}^{(2n+L+2)T} n(t)c_0(t-(2n+1)T)dt. \tag{7d}$$

Applying the space-time decoding algorithm described by Alamouti [8], we have

$$\begin{aligned} d'_{2n} &= \Re\left\{h_1^* r_{2n}/j + h_2 r_{2n+1}^*\right\} \\ &= \Re\left\{\sqrt{\frac{E_b}{MT}}\left[(|h_1|^2+|h_2|^2)d_{2n}c_{\text{full}} - jh_1^*h_2c_{\text{half}}\right.\right. \\ &\quad \left.\left.(2d_{2n}+d_{2n-2}+d_{2n+2})\right] - jh_1^*n_{2n} + h_2n_{2n+1}^*\right\} \end{aligned} \tag{8a}$$

$$\begin{aligned} d'_{2n+1} &= \Re\left\{-h_2^* r_{2n}/j + h_1 r_{2n+1}^*\right\} \\ &= \Re\left\{\sqrt{\frac{E_b}{MT}}\left[(|h_1|^2+|h_2|^2)d_{2n+1}c_{\text{full}} + jh_1h_2^*c_{\text{half}}\right.\right. \\ &\quad \left.\left.(2d_{2n+1}+d_{2n-1}+d_{2n+3})\right] + jh_2^*n_{2n} + h_1n_{2n+1}^*\right\} \end{aligned} \tag{8b}$$

where $\Re\{\cdot\}$ denotes the real part.

The decision variables can be written as $\hat{d}_n = \text{sgn}\{d'_n\}$, where $\text{sgn}\{\cdot\}$ is the sign function. Note that the nonzero terms $2d_{2n}+d_{2n-2}+d_{2n+2}$ and $2d_{2n+1}+d_{2n-1}+d_{2n+3}$ in Eqs. (8a) and (8b) cause ISI. If ISI is completely canceled, space-time coded MSK has the same error performance as space-time coded BPSK.

B. Receiver with an FIR filter

Examining Eqs. (8a) and (8b) and noting that $\Re\{-jh_1^*h_2\} = \Re\{jh_1h_2^*\}$, we found that d'_{2n} and d'_{2n+1} are equivalent to the outputs when the input information sequence $\boldsymbol{d} = [d_0, d_1, d_2, d_3, \cdots]^T$ is passed through a pseudo-channel modeled as a 5-tap symmetric FIR filter with an impulse response

$$\begin{aligned} \boldsymbol{h}_{\text{imp}} &= \sqrt{\frac{E_b}{MT}}\Re\left\{\left[-jh_1^*h_2c_{\text{half}}, 0, (|h_1|^2+|h_2|^2)c_{\text{full}} - \right.\right. \\ &\quad \left.\left. 2jh_1^*h_2c_{\text{half}}, 0, -jh_1^*h_2c_{\text{half}}\right]^T\right\}. \end{aligned} \tag{9}$$

The output of the pseudo-channel is further corrupted by additive Gaussian noise.

The optimum FIR filter in the sense of minimum mean-square error (MMSE) $\boldsymbol{c}_{\text{mmse}}$ is an effective way to recover the information bits [16, ch. 6]. This filter is obtained as $\boldsymbol{c}_{\text{mmse}} = \boldsymbol{R}^{-1}\boldsymbol{z}$, where $\boldsymbol{R} = E\{\boldsymbol{d}'\boldsymbol{d}'^T\}$ ($E\{\cdot\}$ denotes expectation) is the autocorrelation matrix of the output of the pseudo-channel, $\boldsymbol{d}' = [d'_0, d'_1, \cdots, d'_{\tau-1}]^T$, τ is the length of $\boldsymbol{h}_{\text{imp}}$, and $\boldsymbol{z}$ is the cross-correlation vector between the input and the output of the pseudo-channel. Note that, $\boldsymbol{h}_{\text{imp}}$ always has a major tap (the third one) that is significantly larger than other taps. As a result of this major tap, the convolution of $\boldsymbol{c}_{\text{mmse}}$ and $\boldsymbol{h}_{\text{imp}}$ always approaches the delta function.

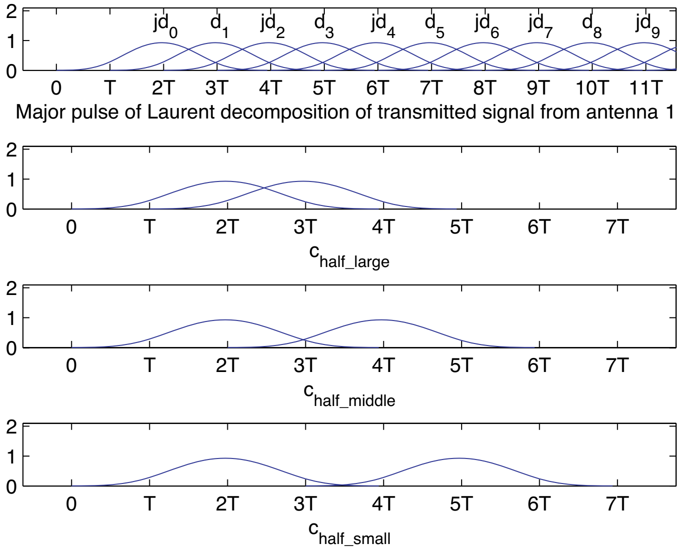

Fig. 2. GMSK with $BT = 0.3$, $L = 3$.

Ignoring some small terms of $\boldsymbol{h}_{\text{imp}}$ could further reduce the order of $\boldsymbol{c}_{\text{mmse}}$, which consequently reduces the decoding complexity for BCPM0.5 with $L > 1$. Although a higher order of $\boldsymbol{c}_{\text{mmse}}$ usually results in a better performance, the minimum and necessary order should be chosen to achieve the desired performance-complexity tradeoff. In fact, $\boldsymbol{c}_{\text{mmse}}$ is essentially the inverse filter of $\boldsymbol{h}_{\text{imp}}$ with the noise effect taken into consideration. In practice, if noise power is unknown, we can design a zero-forcing filter $\boldsymbol{c}_{\text{zf}}$ based on similar procedures.

The Levinson-Durbin recursion algorithm could be applied to efficiently calculate $\boldsymbol{R}^{-1}$. Details of such algorithms can be found in [17]. Moreover, the inversion can be simplified further by considering the zero elements in $\boldsymbol{R}$. Overall, the proposed linear receiver has a much lower computational complexity than existing schemes that employ the Viterbi decoder.

The filter impulse response $\boldsymbol{h}_{\text{imp}}$ given by Eq. (9) for BCPM0.5 with the Alamouti code always has a symmetrical structure, resulting in a relatively low complexity of the MMSE filter. Partial response BCPM0.5 ($L > 1$) may have higher order of the filter $\boldsymbol{h}_{\text{imp}}$. For example, let B be the 3-dB bandwidth of the Gaussian pre-modulation filter. Then $\boldsymbol{h}_{\text{imp}}$ for GMSK with $BT = 0.5$ and $L = 2$ has five taps, three of which are nonzero; $\boldsymbol{h}_{\text{imp}}$ for GMSK with $BT = 0.3$ and $L = 3$ has nine taps, five of which are nonzero. Among the five nonzero taps, two small ones can be ignored to lower the order to five as followings.

Recall that the length of $c_0(t)$ is $L+1$. Therefore, there are L different overlapping terms that cause ISI. For instance, there are three overlapping terms among $c_0(t)$ for BCPM0.5 with $L = 3$ and its time-shifted copies as illustrated in Fig. 2:

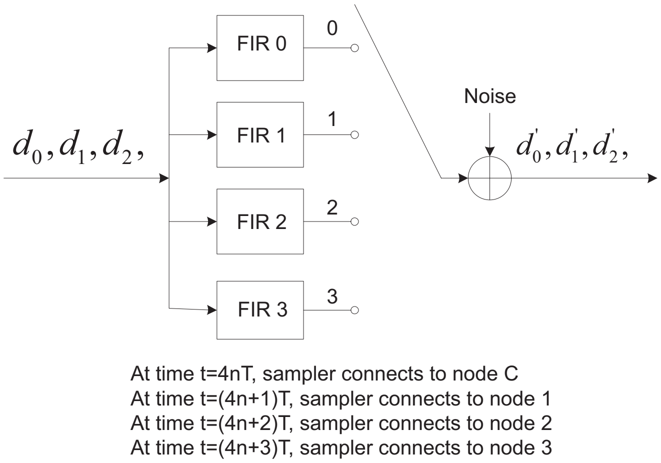

Fig. 3. Equivalent system model for space-time block coded GMSK with 4 antennas.

$$c_{\text{half\_large}} = \int_0^{(L+1)T} c_0(t)c_0(t-T) \approx 0.52T \quad (10a)$$
$$c_{\text{half\_middle}} = \int_0^{(L+1)T} c_0(t)c_0(t-2T) \approx 0.06T \quad (10b)$$
$$c_{\text{half\_small}} = \int_0^{(L+1)T} c_0(t)c_0(t-3T) \approx 0.0008T. \quad (10c)$$

It found that $c_{\text{half\_small}}$ accounts for only about 0.1% of the total interference energy; thus we may ignore its contribution to lower decoding complexity. Since two of the five nonzero taps are the contribution from $c_{\text{half\_small}}$ only, we can thus reduce the order of $\boldsymbol{h}_{\text{imp}}$ from nine to five. In fact, one could even keep $c_{\text{half\_large}}$ only by ignoring both $c_{\text{half\_middle}}$ and $c_{\text{half\_small}}$. Although the resulting filter order is still five for this case, the computation needed to determine each filter tap is reduced.

With these strategies, GMSK with $L = 3$ has a comparable decoding complexity as that with $L = 1$ and 2. Similarly, for BCPM0.5 with $L > 3$, keeping the pulses that contribute the largest amount of ISI could significantly decrease the decoding complexity at the expense of a slight error performance loss.

C. Receiver with decision feedback

When the number of transmit antennas is greater than two, the symmetry property of $\boldsymbol{h}_{\text{imp}}$ in Eq. (9) does not hold anymore. As an example, let us consider 4 transmit antennas. We apply the transpose of the code matrix (4) given in [11] as our transmission matrix. By applying the decoding algorithm for linear modulations, we obtain similar results as given by Eq. (8) for $d'_{4n}, d'_{4n+1}, d'_{4n+2},$ and d'_{4n+3}. If we treat information bits $d_0, d_1, d_2, d_3, \cdots$ as the system input and $d'_0, d'_1, d'_2, d'_3, \cdots$ as the output, the equivalent model of this system is illustrated in Fig. 3. The equivalent system model includes a sampler and four FIR filters with different coefficients, i.e., with $BT = 0.3$ and $L = 3$, FIR0 and FIR3 in Fig. 3 have eleven taps while FIR1 and FIR2 have twelve taps. For four antennas, ignoring $c_{\text{half\_middle}}$ and $c_{\text{half\_small}}$ cannot reduce the order of any of the four filters. However, the number of computations needed to calculate the filter taps will be reduced significantly, and the four filters share common values for most taps.

The overall system is no longer linear, and linear receivers will suffer from an irreducible error floor, especially when the value of L is large. However, soft decision-feedback receiver works effectively.

Generally speaking, d'_n is a linear combination of $d_{n-i}, \cdots, d_n, \cdots, d_{n+j}$ corrupted by noise, where $i + j + 1$ is the number of taps of the corresponding FIR for d_n. For instance, $i + j + 1$ equals eleven or twelve for GMSK with $BT = 0.3$. At each iteration of the decision-feedback process, soft decisions are used to cancel the interference caused by $d_{n-i}, \cdots, d_{n-1}, d_{n+1}, \cdots, d_{n+j}$.

A common method is to use the $\tanh$ function [18] to derive soft decisions; however, in practice, $\tanh$ function is difficult to realize because of its nonlinearity. We will apply a linear function to obtain optimum soft decisions in the sense of minimum mean-square error.

The output of the space-time decoder can be expressed as

$$d'_n = \sum_{k=n-i}^{n+j} \delta_k d_k + \mathcal{N} \quad (11)$$

where δ_k is the power of d_k and the power of noise $\mathcal{N}$ is P_n. The optimal scaler x used to approximate d_n should minimize the mean-square error as

$$E\left\{\left(\frac{d'_n}{x} - d_n\right)^2\right\} = E\left\{\left(\frac{\delta_n d_n + \sum_{k=n-i, k\neq n}^{n+j} \delta_k d_k + \mathcal{N}}{x} - d_n\right)^2\right\} = 0. \quad (12)$$

We can solve the equation to obtain

$$x = \frac{\sum_{k=n-i}^{n+j} \delta_k^2 + P_n}{\delta_n}. \quad (13)$$

Strictly speaking, x in Eq. (13) is optimal only for the first iteration, as the signal-to-interference-plus-noise ratio (SINR) will be slightly different after the first iteration. In our simulation, we apply αx in the second and the third iterations, where the optimal values of α $(\alpha > 1)$ are found via simulation. The magnitude of d'_n/x should be further bounded for binary CPM, i.e., soft decisions for d_n should satisfy $\text{sgn}\{d'_n/x\}$ if $|d'_n/x| > \eta$ [19]. We set $\eta = 1$ for any number of transmit antennas and any pulse length L. Compared with the optimal η values, which can be found via exhaustive search, applying $\eta = 1$ will slightly degrade the error performance.

To obtain x, we must calculate the power of the FIR filter in the equivalent system model, which implies that we need to compute the power of the four FIR filters for all the four transmit antennas as shown in Fig. 3. Fortunately, the major taps δ_n of these four filters are very close to one another while the weight magnitudes for the minor taps δ_k $(k = n - i, \cdots, n-1, n+1, \cdots, n+j)$ are from the same set but with different permutations. Therefore, we can use any of the four filters to compute x, rather than calculating x four times.

The complexity of the decision-feedback receiver with soft decisions depends on the number of iterations and the number

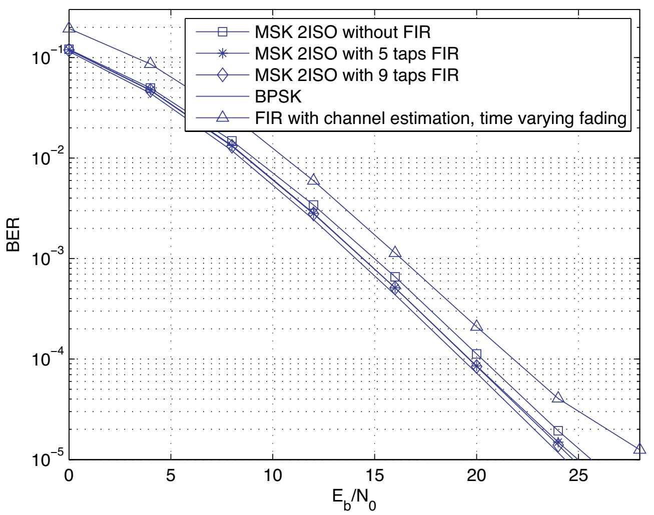

Fig. 4. Performance of space-time block coded GMSK ($BT = \infty$, $L = 1$, $M = 2$).

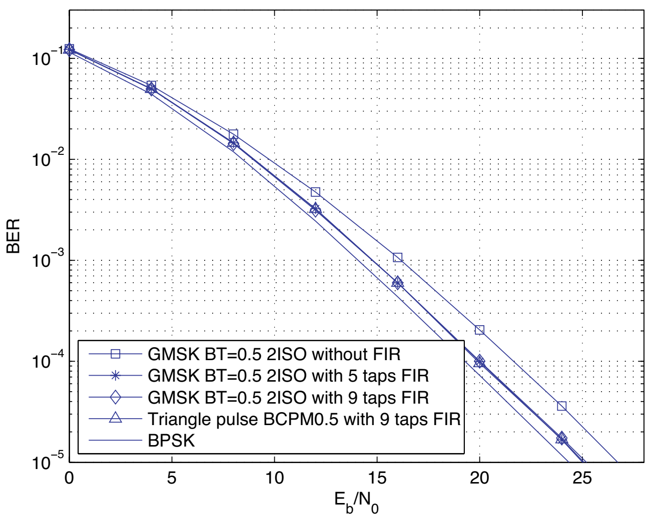

Fig. 5. Performance of space-time block coded GMSK ($BT = 0.5$, $L = 2$, $M = 2$).

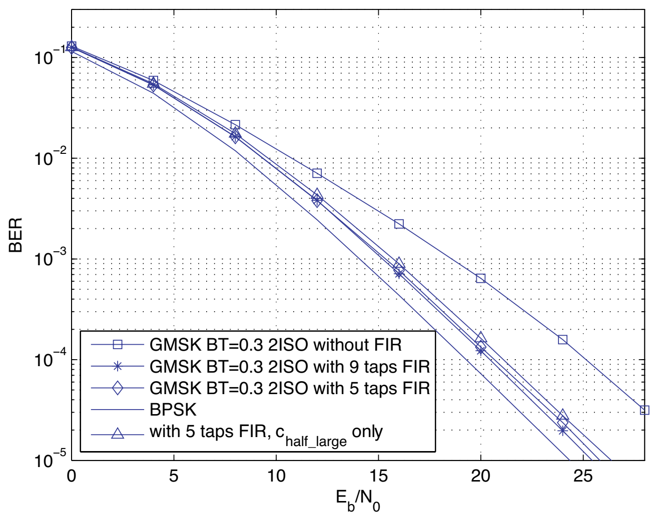

Fig. 6. Performance of space-time block coded GMSK ($BT = 0.3$, $L = 3$, $M = 2$).

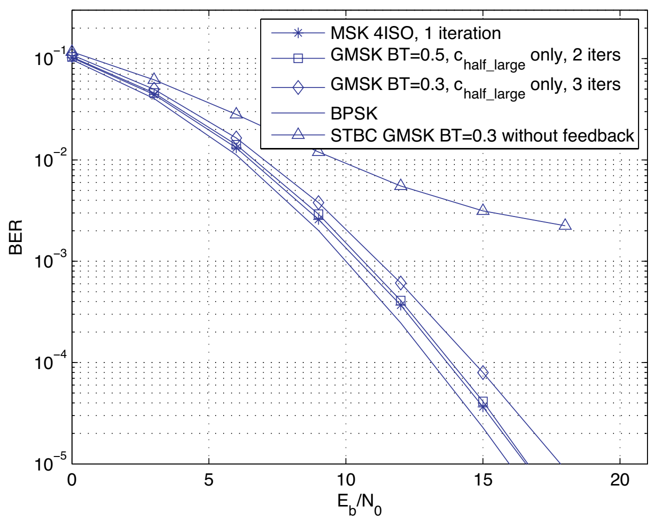

Fig. 7. Performance of space-time block coded GMSK with 4 transmit antennas.

of taps of the FIR filters. The number of iterations is dependent upon mainly L. However, the number of taps of the FIR filters depends on L and the number of antennas.

IV. Simulation Results

Error performances of orthogonal space-time block coded BCPM0.5 with two transmit antennas over frequency-flat Rayleigh fading channels are shown in Figs. 4, 5, and 6, which correspond to GMSK with $BT = \infty$ $(L = 1, \text{MSK})$, $BT = 0.5$ $(L = 2)$, and $BT = 0.3$ $(L = 3)$, respectively. The error performance of systems employing a triangle frequency pulse with $L = 2$ is also provided to show that proposed scheme is applicable for all BCPM0.5. Signal waveforms over one bit interval T are represented by 16 samples in the waveform-based simulation. BER curves of BPSK systems with OSTBC is used as the baseline performance. The exact error probability of BPSK with full diversity employing the Alamouti code can be found in [20].

All simulations assume a quasi-static channel with perfect channel estimation except in Fig. 4, which includes the case of MSK with pilot-aided channel estimation in fast time-varying fading environments. For quasi-static channels, the size of a frame over which the channel coefficients remain constant is 200 data bits. For time-varying fading, channel estimation for the proposed design is very simple and can be developed based on Eq. (6); appropriate pilot sequences could be chosen to cancel c_{half} from the received signal as $d_{2n-1} + d_{2n+1} = 0, d_{2n-2} + d_{2n} = 0, d_{2n} + d_{2n+2} = 0, d_{2n+1} + d_{2n+3} = 0$ and transmitted periodically. For instance, for MSK a minimum of 6 pilot bits are required and $[1, 1, -1, -1, 1, 1]$ is a good training sequence. To improve estimation quality, we add two more pilot bits to form an 8-bit sequence. This allows the receiver to obtain two estimates of the same set of coefficients: the first estimate is obtained using pilot bits 1 to 6, the second using pilot bits 3 to 8, and the final estimate is the average of the two. For GMSK with $BT = 0.3, L = 3$, a pilot sequence of minimum of 10 bits is required. In obtaining the error performance for time-varying fading channels with pilot-aided channel estimates in Fig. 4, we assume a system operating at a date rate of 270 kbps and carrier frequency of 1.8 GHz with a vehicle speed of 60 mph. The channel is generated by using the Jakes' model. A pilot sequence is inserted at the middle of every data 50 bits and used to estimate the coefficients for these 50 bits without interpolation over pilot periods. Thus channel estimation complexity for the proposed design is comparable to that of STBC BPSK. It is observed from Fig. 4 that with actual channel estimates the performance degradation compared with the ideal case of perfect channel estimates is within about 2 dB at a BER of 10^{-4}.

It is found from the simulation results that the linear receiver with an MMSE FIR filter is very robust to ISI. By examining the slopes of the BER curves, we found that the proposed STBC BCPM0.5 system with linear receivers achieves nearly the same diversity as STBC BPSK. From Figs. 4 and 5, we find that increasing the order of $\boldsymbol{c}_{\text{mmse}}$ results in only negligible performance improvement; thus, keeping the filter order to minimum is highly recommended. For the scenario simulated in Fig. 6, it is found that keeping $c_{\text{half\_large}}$ only still achieves almost the same diversity and the performance loss is within 1dB when compared with the receiver with a nine-tap FIR.

Performance of systems with four antennas is shown in Fig. 7. Soft decision feedback by considering $c_{\text{half\_large}}$ only is found achieving good performance too. Without decision feedback, GMSK with $BT = 0.3$ for 4 antennas suffers from an irreducible error floor due to large ISI.

Comparing our results with the results in [4], we find that for MSK signals the proposed simplified receiver achieves almost the same performance as an ML receiver. This is expected since the decomposition for MSK is exact even though the decomposition for BCPM0.5 with $L > 1$ is approximate. Since the various approximations to achieve the minimum complexity for STBC BCPM0.5 with $L > 1$ results in only a negligible performance loss, the proposed STBC scheme and simplified receivers could have good practical values.

V. CONCLUSION

We have proposed a space-time block code for BCPM0.5 signals. This scheme is based on Laurent decomposition combined with data precoding, which allows us to apply the orthogonal code structure. We have also derived a linear MMSE receiver for the proposed space-time block coded BCPM0.5 with two transmit antennas, and a nonlinear decision-feedback receiver for systems with more than two antennas. The simplified receiver is very robust to ISI inherent in BCPM0.5 signals. The combination of the proposed orthogonal code and the simplified receiver achieves good performance for BCPM0.5 systems, and the decoding complexity is much lower than existing schemes. For the scenarios where the time-bandwidth product is not smaller than 0.5, the performance gap between the proposed STBC BCPM0.5 and the baseline system – STBC BPSK over frequency-flat Rayleigh fading channels – is typically within about half a dB at a BER of 10^{-4}, and is slight different depending on the frequency pulse length L and the number of transmit antennas.

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