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Full Length Research Paper

Full-diversity high-rate space-time block-coded systems using estimated channel state information for symbol detection

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In this paper, the frame-error-rate (FER) performance of full-diversity high-rate space-time block-coded (STBC) system is presented for quadrature phase-shift keying (QPSK) digital modulation technique, which is using the numeric-variable-forgetting-factor (NVFF) least-squares (LS) algorithm based channel estimator. The STBC transmission data-rates $5/4$ with two-transmitter antennas and $9/8$ with four-transmitter antennas are achieved by maximizing the coding gain and by minimizing the peak-to-minimum-power-ratio while using the selective power scaling of the information symbols, which also guarantees full-diversity in case of the two-transmitter antennas. The imperfect channel estimation leads to degradation in the FER performance of the low complexity maximum likelihood decoding algorithms, though the optimum power scaling factor is incorporated along with the constellation rotation scheme. Simulation results are presented to demonstrate that the FER performance of high-rate STBC system under idealistic conditions is quite different from its performance under realistic wireless environment due to the channel estimation errors. Therefore, the tracking performance of the pilot channel estimator is found to play significant role in the overall performance evaluation of the proposed wireless systems under slowly time-varying channels also in case of the general M-ary phase-shift keying (M-PSK) constellations.

Key words: Space-time block-code, coding gain, peak-to-minimum-power-ratio, numeric-variable-forgetting-factor, variable-forgetting-factor, least-squares algorithm.

INTRODUCTION

A new class of full-diversity high-rate space-time block-codes has appeared as a breakthrough in the emerging field of multiple-antenna wireless communication engineering (Das et al., 2006), which exploits the innate algebraic structure in the existing orthogonal designs based on the quaternions for two-transmitter (TXR) antennas (Alamouti, 1998) and quasi-orthogonal designs for four-transmitter antennas (Jafarkhani, 2001). The achievable transmission rate $R_{STBC,symbol} = K/P$ can be more than unity in the case of the space-time signals by using the concept of signal constellation rotation (Su and Xia, 2004) and selective power scaling; where P time slots are used to transmit K information symbols. Das et al. (2006) demonstrated that the incorporation of optimum selective power scaling factor in the high-rate space-time block-coded (STBC) signal constellation leads to the maximization of coding gain while minimizing the transmission signal peak-to-minimum-power-ratio (PMPR). Consequently, it is apparent from the simulation results that the effective throughput of STBC system significantly increases under high signal-to-noise-ratio (SNR) conditions, which also supersedes the rate-1 Alamouti STBC system (Alamouti, 1998).

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The simplest space-time block-code with complex orthogonal design is the 2×2 code, which achieves $R_{STBC,symbol}=1$ at the full-diversity, which uses the low computationally complex maximum likelihood (ML) decoding algorithms. The two different configurations of 2×2 STBCs are as follows:

$$X_{TXR}(2)=J1\{x_1,x_2\}\quad for\ \{d_0=0\}$$
$$X_{TXR}(2)=\begin{bmatrix}+x_1 & +x_2\\ -x_2^* & +x_1^*\end{bmatrix}=X_T \qquad (1)$$

$$X_{COSET}(2)=\begin{bmatrix}1 & 0\\ 0 & -1\end{bmatrix}X_{TXR}(2)=\begin{bmatrix}+x_1 & +x_2\\ +x_2^* & -x_1^*\end{bmatrix} \qquad (2)$$
$$X_{COSET}(2)=J2\{x_1,x_2\}$$

$$X_C=J2\left\{x_{c1}/Ks,\ x_{c2}/Ks\right\}\quad for\ \{d_0=1\}$$
$$X_C=\begin{bmatrix}+x_{c1}/Ks & +x_{c2}/Ks\\ +x_{c2}^*/Ks & -x_{c1}^*/Ks\end{bmatrix} \qquad (3)$$

where $(\ )^*$ denotes the complex conjugate operator and $X_{COSET}(2)=J2\{x_1,x_2\}$ (Das et al., 2006) is the coset of $X_{TXR}(2)=J1\{x_1,x_2\}$ (Alamouti, 1998) in Equations 1 and 2, which form the basis of two different STBCs X_T *and* X_C in Equations 1 and 3 with the power scaling factor $1/Ks$.

The low complexity ML decoding procedure for STBC system requires the knowledge of perfect channel-state-information (CSI) at the receiver, which seems to be impractical under the time-varying channel. Under such Rayleigh fading environment, the linear-least-squares algorithm (linear polynomial model-based approach (Borah and Hart, 1999)) using variable-forgetting-factor (LSn-VFF) is developed for the channel estimation (Song et al., 2002), which can be used for the pilot channel estimation under the nonstationary environment. The LSn-VFF algorithm is reported to perform well at high SNRs at the cost of increased computational complexity (Song et al., 2000). However, Kohli et al. (2011) presented a computationally efficient channel estimation method using the linear-least-squares algorithm in combination with the numeric-variable-forgetting-factor (LSn-NVFF), which is based on the extended estimation error criterion. In this correspondence, we propose the utilization of LSn-NVFF algorithm based estimated complex channel coefficient matrix for the information symbol detection in high-rate STBC wireless systems akin to that of Grover and Kohli (2011), which is expected to improve the bit-error-rate (BER) performance of the high-rate quasi-orthogonal STBCs (QSTBCs) even at the low SNRs.

This paper is organized as follows. We first describe the proposed high date-rate STBC system model working under the time-varying environment. Then, we give details about the simulation results to compare the frame-error-rate (FER) performance of proposed system using different polynomial based variable forgetting factor LS algorithms along with ML decoding algorithm. The main focus/emphasis is on the usage/incorporation of channel matrix obtained by using the LSn-NVFF (first-order polynomial based approach) and LSn2-NVFF (second-order polynomial based approach) pilot channel estimation algorithms (Grover and Kohli, 2011), which is the estimated CSI. The value of NVFF increases under stationary conditions for the accurate channel estimation and its value decreases under nonstationary conditions to reduce the lag noise (Kohli et al., 2011). Finally, conclusions and future scope are given for the proposed high data-rate STBC, that is, $R_{STBC,data}>1$ wireless systems with quaternions and M-PSK constellations (Su and Xia, 2004).

PARADIGM FOR HIGH DATA-RATE STBC SYSTEM

We first consider a wireless system equipped with two-transmitter antennas and two-receiver antennas to explore the transmitter-diversity as well as the receiver-diversity benefits as shown in Figure 1. The received signals (X_T configuration for 2×2 STBCs) can be arranged in the following matrix form using Equation 1 as:

$$\underbrace{\begin{bmatrix}+y_1 & +y_2\\ -y_2^* & +y_1^*\end{bmatrix}}_{\hat{Y}_T}=\underbrace{\begin{bmatrix}+c_1 & +c_2\\ -c_2^* & +c_1^*\end{bmatrix}}_{C}\underbrace{\begin{bmatrix}+x_1 & +x_2\\ -x_2^* & +x_1^*\end{bmatrix}}_{X_T}+\underbrace{\begin{bmatrix}+\eta_1 & +\eta_2\\ -\eta_2^* & +\eta_1^*\end{bmatrix}}_{Z_T} \qquad (4)$$

where $\hat{Y}_T$ is the received symbol signal matrix, Z_T denotes the additive white Gaussian noise (AWGN) sample matrix with zero-mean and variance σ_η^2, X_T is the transmitted symbol signal matrix with zero-mean and variance σ_x^2, the symbol duration is T_s and C is the channel coefficient matrix. The channel coefficients are assumed to be flat-fading slowly time-varying, which follow the first-order autoregressive process, because the realistic channel variations can be well approximated by using the first-order Markov process (Kohli, 2011). In the block diagram shown in Figure 1, the symbols x_1 and x_2 from quadrature phase-shift keying (QPSK) constellation are chosen for the data bit sets $\{d_1d_2\}$ and $\{d_3d_4\}$, respectively if $d_0=0$. Similarly, the symbols x_{c1} and

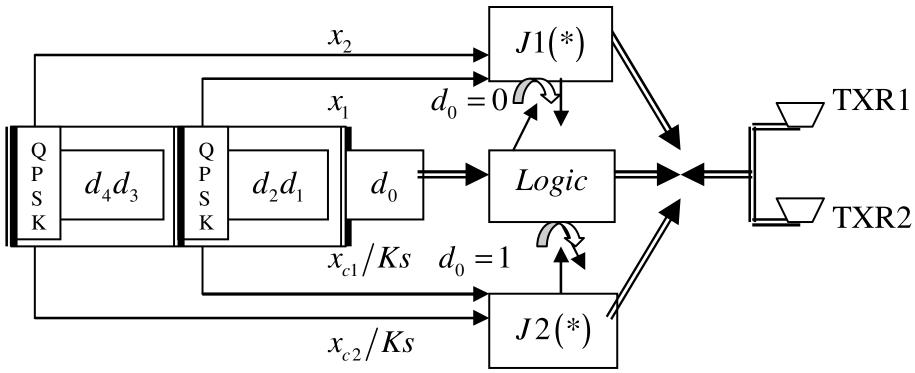

Figure 1. The block diagram of data rate – 5/4 STBC with QPSK modulation for the proposed system.

x_{c2} are chosen from QPSK constellation as shown in Figure 2 for the data bit sets $\{d_1 d_2\}$ and $\{d_3 d_4\}$, respectively in case $d_0 = 1$.

The channel coefficients are assumed to be constant for the period of $2T_s \; with \; P = 2$. To decode the actual transmitted symbol signal matrix X_T from the processed received signal matrix $\hat{Y}_T$, we form the decision matrix $\hat{X}_T$ as:

$$\hat{X}_T = \hat{C}^H \hat{Y}_T = \hat{C}^H C X_T + \hat{C}^H Z_T$$
$$\tilde{X}_T = ML\left(\hat{X}_T\right) \tag{5}$$

where $(.)^T$, $(.)^H$, $ML(.)$ and $\tilde{X}_T$ are the simple transpose operator, the Hermitian transpose operator, the maximum likelihood operator and the finally detected symbol matrix, respectively. For the symbol decoding in Equation 5, we propose the usage of the estimated channel coefficient matrix $\hat{C}$. If the estimated channel coefficient matrix approach the actual channel conditions, that is, $\hat{C} \to C$, then only $\left(\hat{c}_2^* c_1 - \hat{c}_1 c_2^*\right) \to 0$, $\left(\hat{c}_1^* c_2 - c_1^* \hat{c}_2\right) \to 0$ and $SNR_{Avg} \to \left[|c_1|^2 + |c_2|^2\right] \sigma_x^2 / \sigma_\eta^2$ are at each element of the decision matrix $\hat{X}_T$. As the two-branch transmitter-diversity scheme with one receiver is equivalent to that of the two-branch maximal-ratio receive combining (MRRC) only under the perfect channel estimation conditions (Alamouti, 1998), therefore, the 2×2 STBC system doubles the benefit of diversity gain by improving SNR at the receiver. For two equal power uncorrelated transmitters, while keeping the total transmitted power constant (Gesbert et al., 2003) that is, $P_{TXR} = P_{TXR}/2 + P_{TXR}/2$, the capacity of such multi-input multi-output system is given as:

$$capacity_{STBC2\times 2} = \log_2\left[\det\left\{I_2 + \frac{\sigma_x^2}{\sigma_\eta^2}\frac{C^H C}{2}\right\}\right] \; bits/s/Hz \tag{6}$$

(under ideal conditions with known channel coefficient matrix)

$$\hat{c}apacity_{STBC2\times 2} = \log_2\left[\det\left\{I_2 + \frac{\sigma_x^2}{\sigma_\eta^2}\frac{E\{\hat{C}^H C\}}{2}\right\}\right] \; bits/s/Hz \; \leq capacity_{STBC2\times 2} \tag{7}$$

(under realistic conditions with estimated channel coefficient matrix) where, $E\{\;\}$ is the expectation operator and $\det\{\;\}$ is the determinant operator for matrices. Similarly, the received signals (X_C configuration for 2×2 STBCs) can be arranged in the following matrix form using Equation 3 as:

$$\underbrace{\begin{bmatrix} +y_{c1} & +y_{c2} \\ +y_{c2}^* & -y_{c1}^* \end{bmatrix}}_{\hat{Y}_C} = \underbrace{\begin{bmatrix} +c_1 & +c_2 \\ -c_2^* & +c_1^* \end{bmatrix}}_{C} \underbrace{\begin{bmatrix} +x_{c1}/Ks & +x_{c2}/Ks \\ +x_{c2}^*/Ks & -x_{c1}^*/Ks \end{bmatrix}}_{X_C} + \underbrace{\begin{bmatrix} +\eta_1 & +\eta_2 \\ +\eta_2^* & -\eta_1^* \end{bmatrix}}_{Z_C} \tag{8}$$

where $\hat{Y}_C$ is the received symbol signal matrix, Z_C denotes AWGN sample matrix with zero-mean and variance σ_η^2, X_C is the transmitted symbol signal matrix with zero-mean and variance σ_{xc}^2/Ks^2 with real scalar $Ks > 1$. The symbols x_{c1} and x_{c2} are scaled by the factor $1/Ks$ for the selective power scaling to guarantee the full diversity for the proposed high-rate STBC system with two-transmitter antennas, which is equivalent to the radius of inner QPSK constellation circle, whereas the QPSK symbols x_1 and x_2 on the outer constellation circle are normalized to unity, which leads to the transmitted signal peak-to-minimum-power-ratio

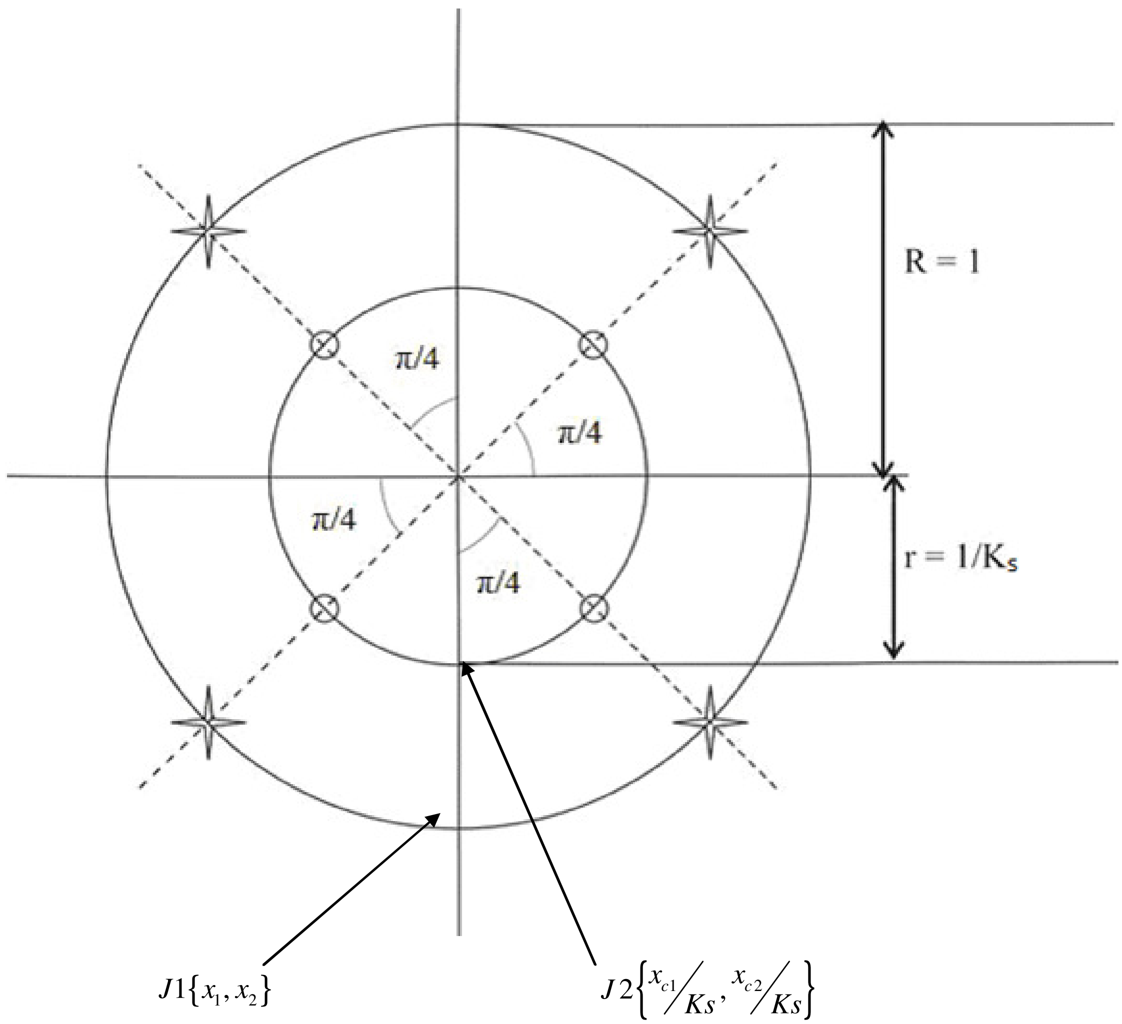

Figure 2. Signal constellation depiction for data rate - $5/4$ STBC with QPSK modulation.

(PMPR) equal to $Ks^2 > 1$. To decode the actual transmitted symbol matrix X_C from the processed received symbol signal matrix $\hat{Y}_C$, we form the decision matrix $\hat{X}_C$ as:

$$\begin{aligned} \hat{X}_C &= \hat{C}^H \hat{Y}_C \\ \tilde{X}_C &= ML\left(\hat{X}_C\right) \end{aligned} \tag{9}$$

where $\tilde{X}_C$ is the finally detected symbol matrix. For symbol decoding in Equation 9, we propose the usage of the estimated channel coefficient matrix $\hat{C}$ like in Equation 5. Subsequently, it can be demonstrated that:

$$\hat{C}^H C = \begin{bmatrix} \left(\hat{c}_1^* c_1 + \hat{c}_2 c_2^*\right) & \left(\hat{c}_1^* c_2 - \hat{c}_2 c_1^*\right) \\ \left(\hat{c}_2^* c_1 - \hat{c}_1 c_2^*\right) & \left(c_1^* \hat{c}_1 + c_2 \hat{c}_2^*\right) \end{bmatrix} \tag{10}$$

However,

$$C^H C = \begin{bmatrix} \left(|c_1|^2 + |c_2|^2\right) & 0 \\ 0 & \left(|c_1|^2 + |c_2|^2\right) \end{bmatrix} \text{ for } \hat{C} \to C \tag{11}$$

which results in the ideal decoding of transmitted symbol signal matrix. After deciding about $\tilde{X}_T$ or $\tilde{X}_C$ using ML scheme, the final decoding of d_0 is as follows:

$$\begin{aligned} decoded\left\{\tilde{X}_T\right\} &\to d_0 = 0 \\ decoded\left\{\tilde{X}_C\right\} &\to d_0 = 1 \end{aligned} \tag{12}$$

To obtain the optimum value of power scaling factor Ks, we need to maximize the cost function given by

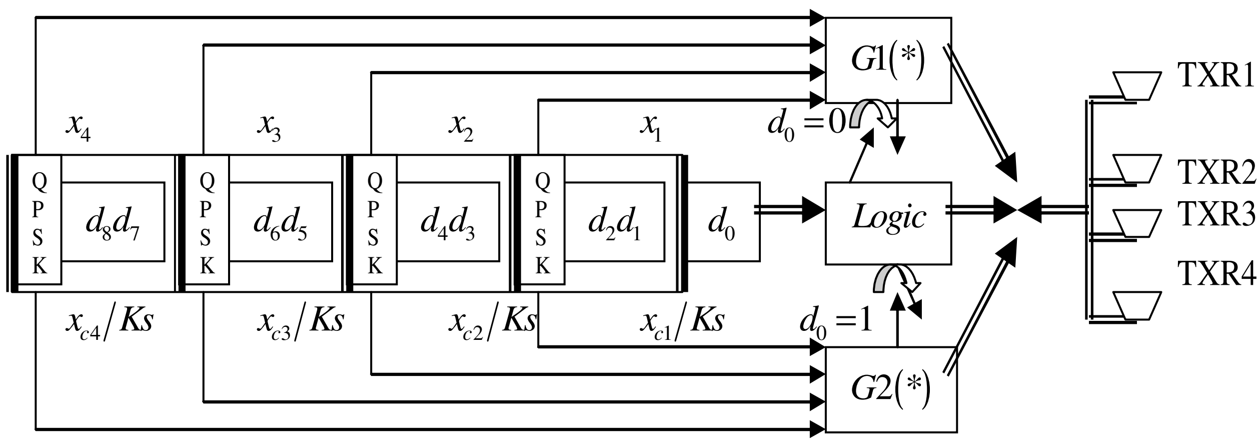

Figure 3. The block diagram of data rate – $9/8$ STBC with QPSK modulation for the proposed system.

$$C_F = \arg \max_{Ks>1} \left\{ \frac{Coding\ Gain}{PMPR} \right\} \quad (13)$$

$$C_F = \arg \max_{Ks>1} \left\{ \frac{\min\{\det\{D_{TC} D_{TC}^H\}\}}{Ks^2} \right\} \quad \text{with}$$

$$D_{TC} = X_T - X_C \quad (14)$$

By simple mathematical simplification, it can be shown that the optimum value of Ks is $\sqrt{3}$ (Appendix) and the achievable maximum transmission rate is $R_{STBC,data} = 5/4 > 1$ for non-zero positive value of Ks. Thus, the frame of five data bits may be decoded.

Further, the aforementioned scheme can be extended to achieve the maximum transmission rate of $R_{STBC,data} = 9/8 > 1$ by using the four-transmitter antennas in combination with the concept of constellation rotation or rotation of information symbols (Su and Xia, 2004; Jafarkhani and Seshadri, 2003) as shown in Figures 3 and 4. For the four-transmitter antennas case, the rate-1 complex quasi-orthogonal space-time block-code (QSTBC) configurations are presented (Jafarkhani, 2001; Su and Xia, 2004; Grover and Kohli, 2011). It follows that:

$$X_{TXR}(4) = \left(\begin{array}{cc|cc} x_1 & x_2 & x_3 & x_4 \\ -x_2^* & x_1^* & -x_4^* & x_3^* \\ \hline -x_3^* & -x_4^* & x_1^* & x_2^* \\ x_4 & -x_3 & -x_2 & x_1 \end{array} \right) = X_{T4} \quad (15)$$

However, this QSTBC does not exhibit the full-diversity, as the diversity gain is sacrificed to boost the rate of this QSTBC design (Su and Xia, 2004). Here, the channel coefficients are assumed to remain unchanged for the duration of $4T_s\ with\ P = 4$. Similar to the two-transmitter antennas based STBC systems, it can be shown that:

$$X_{TXR}(4) = G1\{x_1, x_2, x_3, x_4\} \quad for\ \{d_0 = 0\} \quad (16)$$

$$X_{TXR}(4) = \begin{pmatrix} +J1\{x_1, x_2\} & +J1\{x_3, x_4\} \\ -\bar{J}1\{x_3, x_4\} & +\bar{J}1\{x_1, x_2\} \end{pmatrix} \quad (17)$$

$$\underbrace{X_{C4}}_{COSET} = G2\left\{ x_{c1}/Ks, x_{c2}/Ks, x_{c3}/Ks, x_{c4}/Ks \right\} \quad for\ \{d_0 = 1\} \quad (18)$$

$$X_{C4} = \begin{pmatrix} +J2\left\{ x_{c1}/Ks, x_{c2}/Ks \right\} & +J2\left\{ x_{c3}/Ks, x_{c4}/Ks \right\} \\ +\bar{J}2\left\{ x_{c3}/Ks, x_{c4}/Ks \right\} & -\bar{J}2\left\{ x_{c1}/Ks, x_{c2}/Ks \right\} \end{pmatrix} \quad (19)$$

where

$$\bar{J}1\{x_1, x_2\} = conjugate\left[J1\{x_1, x_2\} \right] \quad (20)$$

$$\bar{J}2\left\{ x_{c1}/Ks, x_{c2}/Ks \right\} = conjugate\left[J2\left\{ x_{c1}/Ks, x_{c2}/Ks \right\} \right] \quad (21)$$

It is noteworthy that the following information symbols belong to M-PSK constellation with the rotation angle ϕ in the signal constellation (Su and Xia, 2004).

$$x_3 = x_1' e^{j\phi}\ and\ x_4 = x_2' e^{j\phi} \quad (22)$$

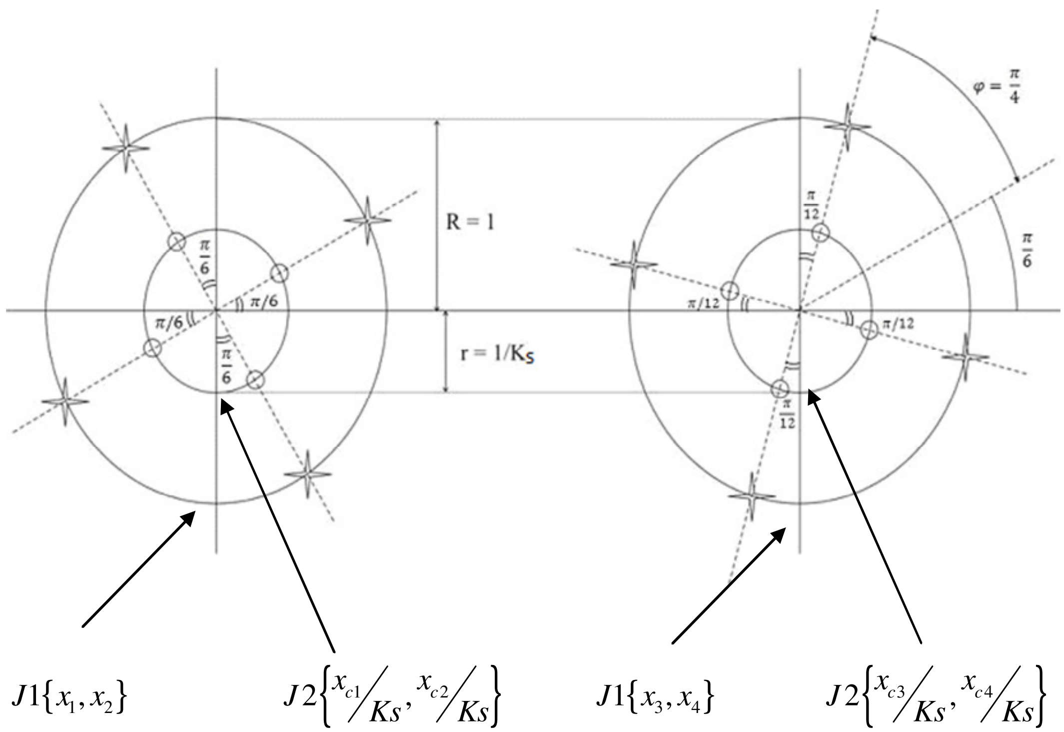

Figure 4. Signal constellation depiction for data rate – $9/8$ STBC with QPSK modulation.

where $J1\{x_1, x_2\}$ and $J1\{x_1', x_2'\}$ belong to the same constellation points.

$$x_{c3} = x_{c1}' e^{j\phi} \ , \ and \ \ x_{c4} = x_{c2}' e^{j\phi} \tag{23}$$

where $J2\{x_{c1}/Ks, x_{c2}/Ks\}$ and $J2\{x_{c1}'/Ks, x_{c2}'/Ks\}$ belong to the same constellation points. For simulation results presented subsequently, the optimum value of rotation angle is considered to be $\phi = \pi/4$. In the block diagram as shown in Figure 3, the symbols x_{c1} , x_{c2} , x_{c3} and x_{c4} from the QPSK constellation are chosen for the data bit sets $\{d_1 d_2\}$, $\{d_3 d_4\}$, $\{d_5 d_6\}$ and $\{d_7 d_8\}$, respectively if $d_0 = 1$. The symbols x_{c1} , x_{c2} , x_{c3} and x_{c4} are scaled by the factor $1/Ks$ for the selective power scaling to guarantee high diversity gain for the proposed high-rate STBC system, which is equivalent to the radius of inner QPSK constellation circle as shown in Figure 4, whereas the QPSK symbols x_1 , x_2 , x_3 and x_4 on the outer constellation circle are normalized to unity, which leads to the transmitted signal PMPR equal to $Ks^2 > 1$. After deciding about $\tilde{X}_{T4}$ or $\tilde{X}_{C4}$, the final decoding of d_0 is as follows:

$$\begin{aligned} decoded\{\tilde{X}_{T4}\} &\rightarrow d_0 = 0 \\ decoded\{\tilde{X}_{C4}\} &\rightarrow d_0 = 1 \end{aligned} \tag{24}$$

Thus, the frame of nine data bits may be decoded to provide $R_{STBC,data} = 9/8 > 1$, in which the optimum value of $Ks = \sqrt{2}$ is considered for MATLAB simulations in the simulation results.

SIMULATION RESULTS

For simulations, we consider a single antenna at the receiver along with maximum likelihood decoder. The performance index is effective throughput in terms of the FER, which is defined as:

$$Effective\ Throughput = \ (1 - FER) \times R_{STBC,data} \times \log_2 M \tag{25}$$

In the case of $R_{STBC,data} = 5/4 > 1$ with QPSK modulation, the value of M = 4, and therefore, the maximum value of effective throughput is 2.5. However, the maximum value of effective throughput is 2.25 in case of $R_{STBC,data} = 9/8 > 1$ with FER = 0. To obtain simulation results for the proposed full-diversity high-rate STBC systems using the estimated CSI at the receiver, we considered the pilot channel tracking using LSn-VFF (Song et al., 2002), LSn2-VFF (Song et al., 2000), LSn-NVFF (Kohli et al., 2011) and LSn2-NVFF (Grover and Kohli, 2011) adaptive algorithms. For smoothly time-varying

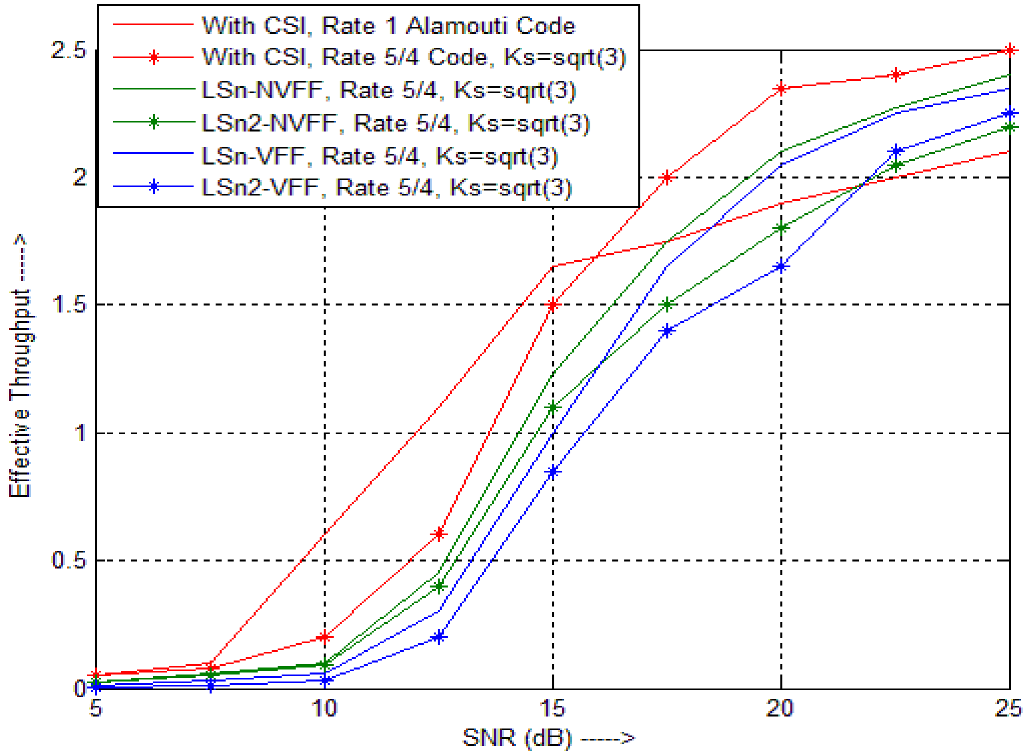

Figure 5. Effective throughput comparison for rate – $5/4$ STBC with power-scaling.

varying fading channels, the LSn-NVFF algorithm reduces the tracking weight error more in comparison to LSn-VFF algorithm.

For the throughput performance evaluation of the proposed high-rate STBC communication systems, the time-varying wireless fading channels with fading rate $f_d(PT_s) \le 0.01$ are assumed, where f_d is the maximum Doppler frequency under nonstationary environment. The presented results are based on the ensemble average of 250 independent simulation runs for the different channel realizations; such that $FER_{avg} = \sum_{m=1}^{250} \frac{FER(m)}{250}$. For known CSI, rate-1 Alamouti STBC systems are compared with $R_{STBC,data} = 5/4$ STBCs, $Ks = \sqrt{3}$ and with $R_{STBC,data} = 5/4$ STBCs, $x_{c1}e^{j\pi/4}$, $x_{c2}e^{j\pi/4}$, $Ks = 1$ (ordinary M-PSK case) in Figures 5 and 6, respectively; which depict that the usage of optimum power scaling factor results in higher effective throughput than the optimum rotation angle based strategy under similar conditions for all the values of SNR. At high value of SNR, we can observe cross-over in the effective throughput performance because the optimum power scaling factor and optimum rotation angle based schemes outperform the Alamouti rate-1 STBC systems.

For high-rate $R_{STBC,data} = 5/4 > 1$ and full spatial diversity (2×2) orthogonal STBC wireless systems, it may be inferred from Figures 5 and 6 that at the lower values of SNRs, $FER_{LSn-NVFF} < FER_{LSn2-NVFF} < FER_{LSn-VFF} < FER_{LSn2-VFF}$ and at the higher values of SNRs, $FER_{LSn-NVFF} < FER_{LSn-VFF} < FER_{LSn2-VFF} < FER_{LSn2-NVFF}$ Similarly, for the high-rate $R_{STBC,data} = 9/8 > 1$ and partial spatial diversity (4×4) quasi-orthogonal STBC wireless systems with $Ks = \sqrt{2}$, it is apparent from Figure 7 that at the lower values of SNRs, $FER_{LSn-NVFF} < FER_{LSn2-NVFF} < FER_{LSn-VFF} < FER_{LSn2-VFF}$ and at the higher values of SNRs, $FER_{LSn-VFF} < FER_{LSn-NVFF} < FER_{LSn2-VFF} < FER_{LSn2-NVFF}$ However, at high value of SNR, we can notice a clear cross-over in the effective throughput performance as shown in Figure 7, as the optimum power scaling factor and the optimum rotation angle based schemes outperform the rate-1 QSTBC systems. Slight overall performance degradation is evident in the high-rate QSTBC system due to the loss of diversity-gain caused by quasi-orthogonal designs. Therefore, if we know the operating SNR level at the transmitter, we can switch between different pilot channel estimators to control the

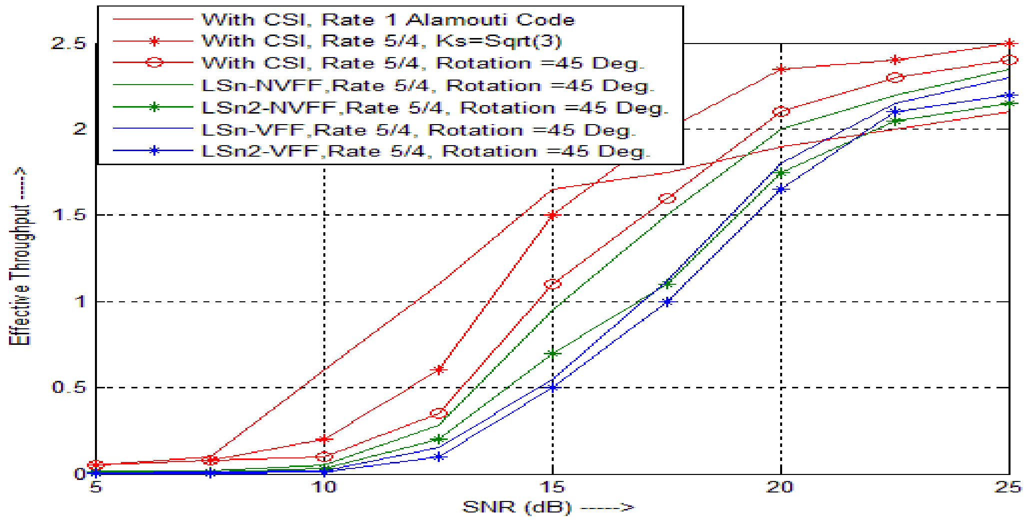

Figure 6. Effective throughput comparison for rate – $5/4$ STBC with phase-rotation (M-PSK).

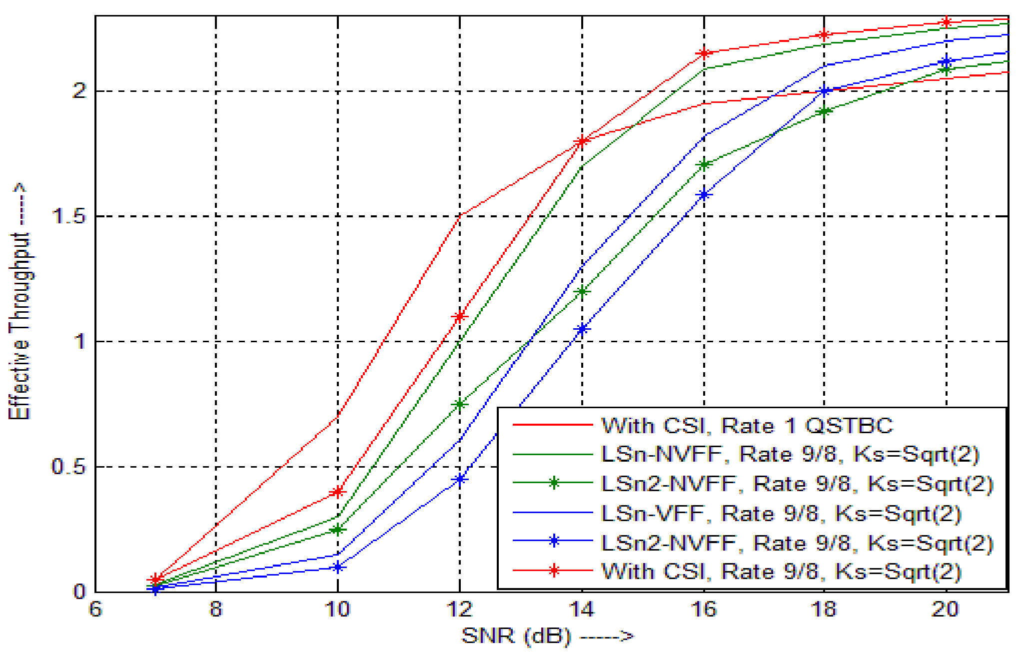

Figure 7. Effective throughput comparison for rate – $9/8$ STBC with phase-rotation and power-scaling.

effective throughput of the high-rate proposed STBC systems.

Conclusion

In this correspondence, we proposed high-rate ($R_{STBC,data} = 5/4 > 1$) with full-diversity gain and high-rate ($R_{STBC,data} = 9/8 > 1$) with high partial-diversity gain space-time block-coded systems, while using estimated channel state information for symbol decoding at the receiver, which provide the maximum value of effective throughput 2.5 and 2.25, respectively with zero frame error rate. Under realistic wireless conditions, the QPSK and M-PSK digital modulation techniques have been incorporated in combination with the selective optimum power scaling and the optimum angle rotation for signal constellation. It may be inferred from the simulation results that the optimum power scaling factor provides an efficient solution to maximize the coding gain and to reduce PMPR of the transmitted signal, which results in better effective throughput of $R_{STBC,data} = 5/4 > 1$ STBC systems than the $R_{STBC,data} = 5/4 > 1$ STBC systems using pure angular rotations without power scaling. Similar throughput improvement is observed in the extended design of high-rate $R_{STBC,data} = 9/8 > 1$ STBC systems, though, there is diversity-gain loss due to the quasi-orthogonal design configurations. The cross-over points in the throughput performance versus SNR graphs give us an opportunity to choose appropriate pilot channel estimator for the proposed high-rate moderate-diversity STBC systems. The proposed scheme can be extended using the eight-transmitter antennas with QSTBC (Su and Xia, 2004), but the throughput performance advantage is marginal because of diversity-gain loss in final ML decoding using CSI.

Future scope includes the merger of the proposed high-rate space-time block-code scheme with orthogonal-frequency-division-multiplexing communication systems in 4G for the enhancement of date transmission rates.

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APPENDIX

To calculate the optimum value of power scaling factor, the maximization of cost function in Equation 14 leads to

$$D_{TC} = \begin{bmatrix} +\left(Ksx_1 - x_{c1}\right) & +\left(Ksx_2 - x_{c2}\right) \\ -\left(Ksx_2^* + x_{c2}^*\right) & +\left(Ksx_1^* + x_{c1}^*\right) \end{bmatrix}\left(1/Ks\right) \quad (A.1)$$

It is a well known exposition that:

$$\det\left\{D_{TC}D_{TC}^H\right\} = \det\left\{D_{TC}\right\}\det\left\{D_{TC}^H\right\} \quad (A.2)$$

Using Equation A.1, it can be demonstrated that

$$\det\left\{D_{TC}\right\} = \alpha_{TC} - Ks\beta_{TC} \quad (A.3)$$

where

$$\alpha_{TC} = \left[\left(\left|x_1\right|^2 + \left|x_2\right|^2\right)Ks^2 - \left(\left|x_{c1}\right|^2 + \left|x_{c2}\right|^2\right)\right]\left(1/Ks^2\right) \quad (A.4)$$

$$\beta_{TC} = \left[\left(x_1^* x_{c1} + x_2^* x_{c2}\right)Ks^2 - \left(x_1 x_{c1}^* + x_2 x_{c2}^*\right)\right]\left(1/Ks^2\right) \quad (A.5)$$

Similarly,

$$\det\left\{D_{TC}^H\right\} = \alpha_{TC} + Ks\beta_{TC} \quad (A.6)$$

Equation A.2 can be solved by using Equations A.3 and A.6 to give:

$$\det\left\{D_{TC}D_{TC}^H\right\} = \alpha_{TC}^2 - Ks^2\beta_{TC}^2 \quad (A.7)$$

The cost function in Equation 14 may be maximized under the following condition,

$$\begin{aligned} x_1 x_{c1}^* + x_2 x_{c2}^* &= 0 \\ x_1^* x_{c1} + x_2^* x_{c2} &= 0 \end{aligned} \quad (A.8)$$

which is only possible, if

$$x_1 = x_{c1} \quad and \quad x_2 = -x_{c2} \quad (A.9)$$

The substitution of the earlier condition of Equation A.9 in Equations A.4, A.5 and 14 results in

$$\alpha_{TC} = \left[\left(Ks^2 - 1\right)/Ks^2\right]\left(\left|x_1\right|^2 + \left|x_2\right|^2\right) \quad (A.10)$$

$$C_F = \arg \max_{Ks>1}\left\{\frac{\alpha_{TC}^2}{Ks^2}\right\} \quad (A.11)$$

$$C_F = \arg \max_{Ks>1}\left\{\left(Ks^2 - 1\right)\left(\left|x_1\right|^2 + \left|x_2\right|^2\right)/Ks^3\right\}^2 \quad (A.12)$$

The maximization of cost function is only possible for the optimum value of power scaling factor, which maximizes the coding gain and minimizes the PMPR, such that:

$$\frac{dC_F}{dKs} = 0 = \frac{d}{dKs}\left\{\left(Ks^2 - 1\right)/Ks^3\right\} \quad (A.13)$$

Consequently, $Ks = \sqrt{3}$ for the proposed 2×2 STBC system with maximum transmission rate of $R_{STBC} = 5/4 > 1$.