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Source:	**LSI Corporation (formerly Agere Systems)**
Title:	**Further results on QO-SFBC as a TxD scheme for 4 transmit antennas**
Agenda Item:	**7.10.3**
Document for:	**Discussion and Decision**

I INTRODUCTION

In our earlier contributions [1] and [2], we presented a Rate-1[1] Quasi-Orthogonal Space Frequency Block code with constellation rotation (QO-SFBC-CR) for 4 transmit antennas that achieves a full diversity order[2] of 4 with a maximum likelihood (ML) receiver. From a performance perspective, it is desirable for the open loop transmit diversity (TxD) scheme to achieve maximal diversity gain and maximal coding gain. However, from a complexity perspective, it is not attractive that a maximum likelihood receiver has to be implemented to realize the full diversity and coding gain.

In this contribution, we present a slight variation to the QO-SFBC-CR code we presented in [1] such that maximal diversity and coding gain is achieved with a reduced complexity receiver. In our analysis, we compare the performance of our proposed QO-SFBC-CR code with SFBC-FSTD for both an MMSE receiver and a reduced complexity ML receiver that supports matrix partitioning and exploits spherical decoding techniques.

II CODE CONSTRUCTION FOR 4 TRANSMIT ANTENNAS

II.a Design Criteria

Let the transmitted 4x4 space-frequency code matrix be:

$$\mathbf{X} = \begin{bmatrix} x(1,1) & x(1,2) & x(1,3) & x(1,4) \\ x(2,1) & x(2,2) & x(2,3) & x(2,4) \\ x(3,1) & x(3,2) & x(3,3) & x(3,4) \\ x(4,1) & x(4,2) & x(4,3) & x(4,4) \end{bmatrix} \tag{1}$$

where $x(i, j)$ is the symbol transmitted from antenna j on frequency index i. Let the code matrix detected at the receiver be defined as:

$$\hat{\mathbf{X}} = \begin{bmatrix} \hat{x}(1,1) & \hat{x}(1,2) & \hat{x}(1,3) & \hat{x}(1,4) \\ \hat{x}(2,1) & \hat{x}(2,2) & \hat{x}(2,3) & \hat{x}(2,4) \\ \hat{x}(3,1) & \hat{x}(3,2) & \hat{x}(3,3) & \hat{x}(3,4) \\ \hat{x}(4,1) & \hat{x}(4,2) & \hat{x}(4,3) & \hat{x}(4,4) \end{bmatrix} \tag{2}$$

[1] Rate = 1 corresponds to 1 symbol transmitted per channel use. Note, with N_{TX} transmit antennas, the maximum rate that can be achieved is N_{TX} symbols transmitted per channel use.
[2] Diversity order is defined as the inverse slope of BER (or FER) versus SNR. The steeper the slope, the faster the error rate decreases with increasing SNR.

We define the codeword difference matrix as $\mathbf{B} = \mathbf{X} - \hat{\mathbf{X}}$. The codeword distance matrix is defined as $\mathbf{A} = \mathbf{B}^H \cdot \mathbf{B}$. To minimize the pairwise error probability, i.e. the probability that $\mathbf{X}$ was sent but $\hat{\mathbf{X}}$ was detected, several criteria have been proposed in literature [3],[4]:

- Rank criterion – maximize the minimum rank of **B** across all pairs of distinct codewords. Larger rank of **B** implies higher diversity gain.
- Determinant criteria – impacts the coding gain. Two metrics have been suggested:
  - Maximize the minimum determinant of **A** across all pairs of distinct codewords.
  - Minimize the average of 1/det(**A**) across all pairs of distinct codewords.
- Trace criteria – impacts the coding gain. Objective is to maximize the minimum trace of **A** across all distinct codewords.

II.b Code Orthogonality

Consider the 2x2 Space-Frequency Alamouti Code [5], whose rate = 1:

$$\mathbf{S} = \begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{bmatrix} \tag{3}$$

Since the column vectors of the Alamouti code matrix are orthogonal, **S** is a full rank matrix with rank 2. Now consider the transmission of the space-frequency Alamouti code through a flat fading channel to a single antenna receiver:

$$\begin{aligned} r_1(f_1) &= h_1(f_1)\cdot s_1(f_1) + h_2(f_1)\cdot s_2(f_1) + n_1 \\ r_2(f_2) &= -h_1(f_2)\cdot s_2^*(f_2) + h_2(f_2)\cdot s_1^*(f_2) + n_2 \end{aligned} \tag{4}$$

We can rewrite (4) as:

$$\begin{bmatrix} r_1(f_1) \\ r_2^*(f_2) \end{bmatrix} = \begin{bmatrix} h_1(f_1) & h_2(f_1) \\ h_2^*(f_2) & -h_1^*(f_2) \end{bmatrix} \cdot \begin{pmatrix} s_1 \\ s_2 \end{pmatrix} + \begin{pmatrix} n_1 \\ n_2 \end{pmatrix} \tag{5}$$

If we assume that the channel is invariant from frequency f_1 to f_2, i.e. $h_1(f_1) \simeq h_1(f_2)$ and $h_2(f_1) \simeq h_2(f_2)$, then we can simplify (5) as:

$$\begin{gathered} \begin{bmatrix} r_1(f_1) \\ r_2^*(f_2) \end{bmatrix} = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix} \cdot \begin{pmatrix} s_1 \\ s_2 \end{pmatrix} + \begin{pmatrix} n_1 \\ n_2 \end{pmatrix} \\ \vec{r} = \mathbf{H}\cdot\vec{s} + \vec{n} \end{gathered} \tag{6}$$

Maximum diversity gain is achieved with a maximum likelihood receiver. Assuming AWGN, we get:

$$\arg\max_s p(\vec{r} \mid \vec{s}) \Rightarrow \vec{s}_{est} = \arg\min_s \left\| \vec{r} - \mathbf{H}\cdot\vec{s} \right\|_2^2 \tag{7}$$

Since **H** is a unitary matrix, $\mathbf{H}^{\mathbf{H}}\mathbf{H} = \gamma\mathbf{I}$, where $\gamma = h_1 h_1^* + h_2 h_2^*$. If $E\{\vec{n}\vec{n}^H\} = \sigma^2\mathbf{I}_2$, then $E\{\mathbf{H}^H\vec{n}\vec{n}^H\mathbf{H}\} = \mathbf{H}^H\mathbf{H}\sigma^2\mathbf{I}_2 = \gamma\sigma^2\mathbf{I}_2$. Hence, $\mathbf{H}^H\vec{n}$ is also white noise. Therefore, we can recast (7) as:

$$\vec{s}_{est} = \arg\min_s \left\| \mathbf{H}^H\vec{r} - \gamma\mathbf{I}_2\vec{s} \right\|_2^2 = \arg\min_s \left\| \mathbf{H}^H\vec{r} - \gamma\vec{s} \right\|_2^2 \tag{8}$$

From (8), we can readily see that the maximum likelihood estimate is given by:

$$\vec{s}_{ML,est} = \frac{1}{\gamma}\mathbf{H}^H \vec{r} \tag{9}$$

The formulation in (9) corresponds to a linear zero-forcing receiver. Therefore, we can state two important attributes of an orthogonal space-frequency block code:

- Since the code matrix is of full rank, the code will achieve full diversity with an ML receiver
- Since the channel transfer matrix is orthogonal, the linear zero-forcing receiver is equivalent to the maximum likelihood receiver.

II.c Construction of Quasi-Orthogonal Codes with full diversity and reduced decoding complexity

It is shown in [6] that for $n > 2$, no complex orthogonal (i.e. unitary) code exists. Hence, our first task is to construct a 4x4 code matrix with full rank, despite the fact that the code matrix will not be orthogonal. We start by constructing a quasi-orthogonal code matrix, based on the 2x2 orthogonal Alamouti sub-matrix, as proposed in [7]:

$$\mathbf{S} = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & -s_3 & -s_2 & s_1 \end{bmatrix} \tag{10}$$

The code matrix in (10) is called quasi-orthogonal because 2 out of the 6 pairs of columns are not orthogonal. The 4 column pairs that are orthogonal to each other are:

- (1,2), (1,3), (2,3), and (3,4)

The remaining 2 columns of the code matrix in (10) are not orthogonal:

- (1,4) and (2,4)

In [1], we showed that by introducing constellation rotation for some of the symbols in (10), we can achieve full diversity[3] (i.e. diversity order = 4) with a maximum likelihood receiver. Now, we state that by rotating each symbol in the code matrix (10) by a certain amount, we can partition the order-4 ML receiver into two order-2 ML receivers. The code matrix that permits partitioning is given by:

$$\mathbf{S} = \begin{bmatrix} c_1 e^{j\theta_1} & c_2 e^{j\theta_2} & c_3 e^{j\theta_3} & c_4 e^{j\theta_4} \\ -c_2^* e^{j\theta_2} & c_1^* e^{j\theta_1} & -c_4^* e^{j\theta_4} & c_3^* e^{j\theta_3} \\ -c_3^* e^{j\theta_3} & -c_4^* e^{j\theta_4} & c_1^* e^{j\theta_1} & c_2^* e^{j\theta_2} \\ c_4 e^{j\theta_4} & -c_3 e^{j\theta_3} & -c_2 e^{j\theta_2} & c_1 e^{j\theta_1} \end{bmatrix} \tag{11}$$

where, $s_i = c_i e^{j\theta_i}$, $c_i \in \{QPSK, 16QAM, 64QAM\}$, θ_i is the rotation applied to constellation symbol c_i, and $i = \{1,2,3,4\}$. The corresponding channel matrix for the code matrix in (11) is given by:

[3] Note that full diversity with constellation rotation is only achieved with square constellations.

$$\mathbf{H} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & h_4^* & -h_1^* & -h_2^* \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix} \tag{12}$$

where we've assumed that the channel is invariant from frequency f_1 to f_4. Now we can express the 4x1 vector at one receiver antenna as:

$$\begin{bmatrix} r_1 \\ r_2^* \\ r_3^* \\ r_4 \end{bmatrix} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & h_4^* & -h_1^* & -h_2^* \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix} \cdot \begin{bmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \\ n_3 \\ n_4 \end{bmatrix} \tag{13}$$

Applying $\mathbf{H}^{\mathrm{H}}$ to the received signal, we get:

$$\begin{bmatrix} \tilde{r}_1 \\ \tilde{r}_4 \\ \tilde{r}_2 \\ \tilde{r}_3 \end{bmatrix} = \begin{bmatrix} \mathbf{H_1} & \mathbf{0} \\ \mathbf{0} & \mathbf{H_2} \end{bmatrix} \cdot \begin{bmatrix} s_1 \\ s_4 \\ s_2 \\ s_3 \end{bmatrix} + \mathbf{H}^H \begin{bmatrix} n_1 \\ n_4 \\ n_2 \\ n_3 \end{bmatrix} \tag{14}$$

where,

$$\begin{bmatrix} \tilde{r}_1 \\ \tilde{r}_2 \\ \tilde{r}_3 \\ \tilde{r}_4 \end{bmatrix} = \mathbf{H}^H \begin{bmatrix} r_1 \\ r_2^* \\ r_3^* \\ r_4 \end{bmatrix} \tag{15}$$

$$\mathbf{H}_1 = \begin{bmatrix} \sum_{i=1}^{4} |h_i|^2 & 2\Re\left(h_1 h_4^* - h_2 h_3^*\right) \\ 2\Re\left(h_1 h_4^* - h_2 h_3^*\right) & \sum_{i=1}^{4} |h_i|^2 \end{bmatrix} \tag{16}$$

and,

$$\mathbf{H}_2 = \begin{bmatrix} \sum_{i=1}^{4} |h_i|^2 & 2\Re\left(h_2 h_3^* - h_1 h_4^*\right) \\ 2\Re\left(h_2 h_3^* - h_1 h_4^*\right) & \sum_{i=1}^{4} |h_i|^2 \end{bmatrix} \tag{17}$$

Note that $\mathbf{H}_1$ and $\mathbf{H}_2$ are real symmetric matrices.

Noise gets correlated:

$$E\left\{\mathbf{H}^H \vec{n}\vec{n}^H \mathbf{H}\right\} = \sigma^2 \begin{bmatrix} \mathbf{H_1} & \mathbf{0} \\ \mathbf{0} & \mathbf{H_2} \end{bmatrix} \tag{18}$$

Now we can set up two order-2 ML decoders. One ML decoder will extract (s_1, s_4), and the second ML decoder will extract (s_2, s_3). First, we decorrelate the noise:

$$\mathbf{z}_1 = \mathbf{H_1}^{-1/2} \begin{bmatrix} \tilde{r}_1 \\ \tilde{r}_4 \end{bmatrix} \tag{19}$$

and,

$$\mathbf{z}_2 = \mathbf{H}_2^{-1/2} \cdot \begin{bmatrix} \tilde{r}_2 \\ \tilde{r}_3 \end{bmatrix} \tag{20}$$

Hence, as a result of partitioning (13), the two ML decoder metrics are given by:

$$\left(s_{est,1}, s_{est,4}\right) = \arg\min_{s1,s4} \left\| \mathbf{z}_1 - \mathbf{H}_1^{1/2} \cdot \begin{pmatrix} s_1 \\ s_4 \end{pmatrix} \right\|_2^2 \tag{21}$$

$$\left(s_{est,2}, s_{est,3}\right) = \arg\min_{s2,s3} \left\| \mathbf{z}_2 - \mathbf{H}_2^{1/2} \cdot \begin{pmatrix} s_2 \\ s_3 \end{pmatrix} \right\|_2^2 \tag{22}$$

Compared to full-ML implementation, the complexity reduction with partitioned ML is substantial:

- 8x for QPSK
- 128x for 16-QAM
- 2048x for 64-QAM

The decoding complexity of each of the order-2 ML decoders can be further reduced without any loss in performance by exploiting spherical decoding techniques, as shown in [8]. Computation of the ML soft bit values is given by:

$$L_i = \min_{\substack{(s1,s4) \\ b_i=0}} \left\| \mathbf{z}_1 - \mathbf{H}_1^{1/2} \cdot \begin{pmatrix} s_1 \\ s_4 \end{pmatrix} \right\|_2^2 - \min_{\substack{(s1,s4) \\ b_i=1}} \left\| \mathbf{z}_1 - \mathbf{H}_1^{1/2} \cdot \begin{pmatrix} s_1 \\ s_4 \end{pmatrix} \right\|_2^2 \tag{23}$$

where, for example for 16-QAM, we would need to compute 4 soft bit values for each symbol s_i. Thus, the computation of the soft bit values is a two-step process:

1. For all possible (s_1, s_4) pairs (256 for order-2 16-QAM), compute the Euclidean distances.
2. Find the minimum distances corresponding to $b_i = 0$ and $b_i = 1$.

In Figure 1, we show the complexity reduction achieved by using spherical decoding for calculating the Euclidean distance in (23) for 16-QAM in a 4x2 antenna configuration (i.e. 4 Tx antennas at NodeB and 2 Rx antennas at the UE). We observe that for SNRs of interest, the complexity reduction is ~ 4x.

In Figure 2, we show the complexity reduction achieved by using spherical decoding for determining the minimum distances in (23) for 16-QAM in a 4x2 antenna configuration. We again observe that for SNRs of interest, the complexity reduction is ~ 4x.

We can now state that using spherical decoding techniques for a partitioned ML decoder, we are able to achieve substantial reductions in complexity, which makes the implementation of such decoders viable. In the case of 16-QAM, the complexity reduction is ~ 512x relative to a full-ML decoder.

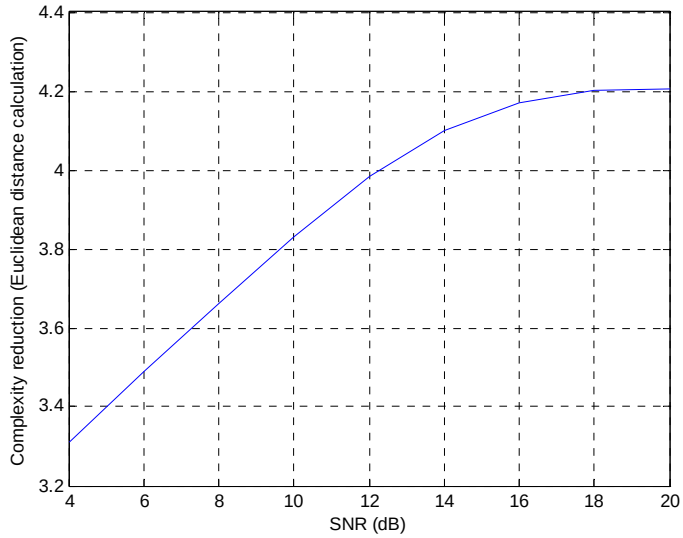

Figure 1: Complexity reduction using spherical Decoding for Euclidean distance calculation for 4x2 16-QAM

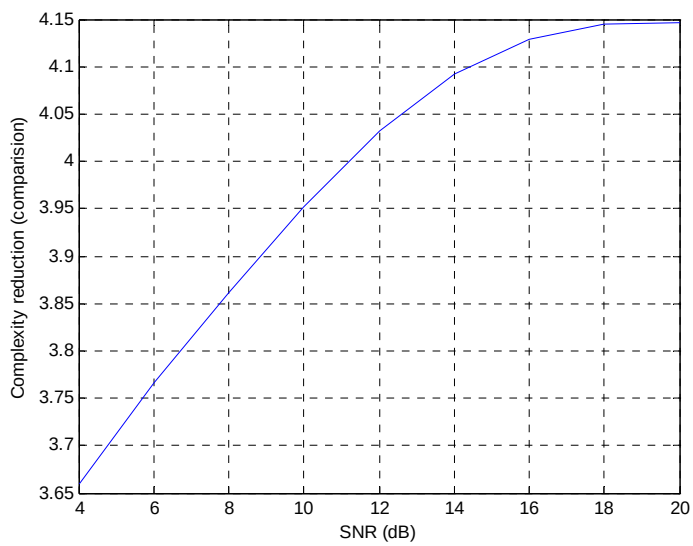

Figure 2: Complexity reduction using spherical decoding for comparison calculation for 4x2 16-QAM

II.d Optimization of Code parameters

For the QO-SFBC-CR code presented in (11), our next task is to determine the optimal rotation angles θ_i to achieve full diversity order and maximize coding gain. By virtue of constellation rotation, we showed in [1] that with a maximum likelihood receiver, the codeword difference matrix **B** is well-behaved, and we are able to realize full diversity performance. Hence, we now focus on maximizing the coding gain, for which we use the max min det(A) and the min average(1/det(A)) criteria (as discussed in II.a). We should note that since the minimum of trace(A) is invariant to rotation angles, we do not consider it in our optimization process.

We start with an examination of the determinant of **A**:

$$\det(\mathbf{A}) = \left[\left(\left|\Delta_1 + \Delta_4\right|^2 + \left|\Delta_2 - \Delta_3\right|^2\right) \times \left(\left|\Delta_1 - \Delta_4\right|^2 + \left|\Delta_2 + \Delta_3\right|^2\right)\right] \tag{24}$$

where $\Delta_i = S_i - \hat{S}_i$. We note that det(**A**) will be minimum when $\Delta_1 = \Delta_4$ and $\Delta_2 = \Delta_3$. Hence, we only need to consider the relative angle between Δ_1 and Δ_4, and the relative angle between Δ_2 and Δ_3. Therefore, without loss of generality, we set $\theta_1 = \theta_2 = 0$. To further simply the design, we set $\theta_3 = \theta_4$. We computed det(A) and the average of 1/det(A) as a function of rotation angles from 0 to π/4. The results for QPSK are shown in Figure 3 and Figure 4. The results for 16-QAM are shown in Figure 5 and Figure 6.

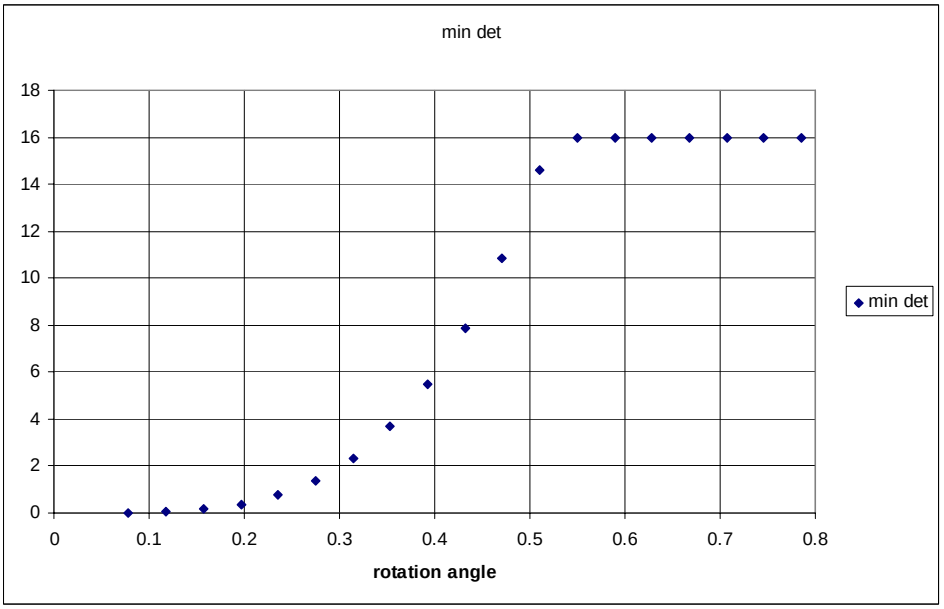

Figure 3: QPSK

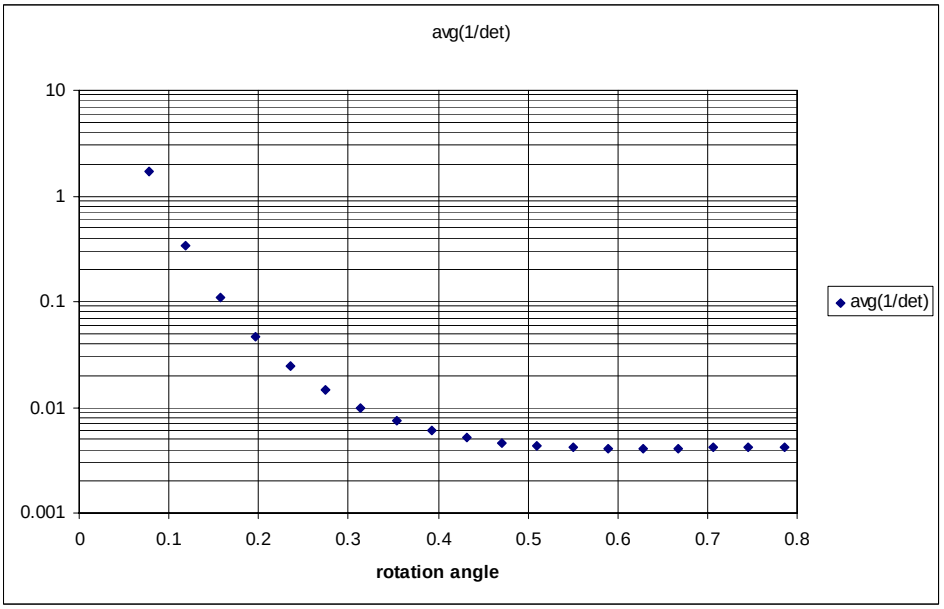

Figure 4: QPSK

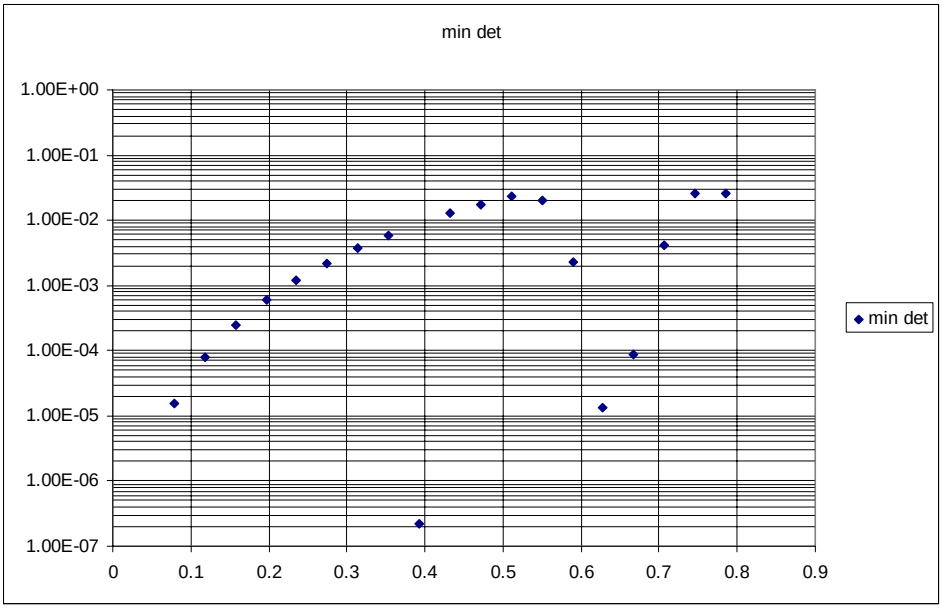

Figure 5: 16-QAM

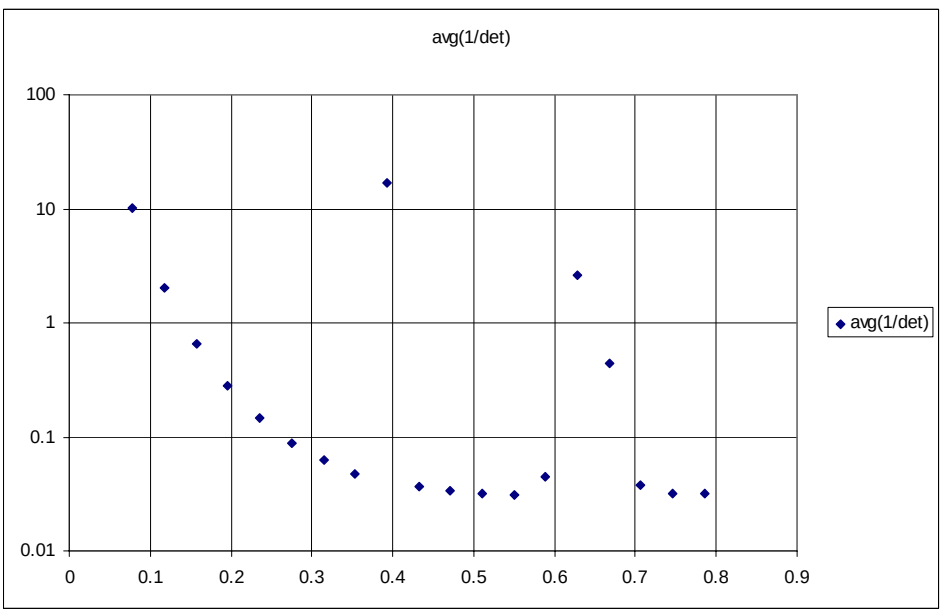

Figure 6: 16-QAM

Based on these results, we see that for QPSK, any rotation angle between 0.5236 radians (π/6) and 0.7854 radians (π/4) is optimal. For 16-QAM, the optimal interval is split into two segments: $0.44 < \theta < 0.56$ and $0.71 < \theta < 0.7854$. For both constellations, we observe that π/4 is an optimal rotation angle. From an implementation point of view, π/4 is a good choice. Hence, we recommend the following rotation angles for the QO-SFBC-CR code in (11) for both QPSK and 16-QAM constellations:

$$\begin{aligned} \theta_1 = \theta_2 &= 0 \\ \theta_3 = \theta_4 &= \frac{\pi}{4} \end{aligned} \tag{25}$$

III PERFORMANCE RESULTS

In this section, we present performance results for QO-SFBC-CR and SFBC-FSTD. The code for QO-SFBC-CR is given in (11), and the corresponding CR (constellation rotation) parameters are given in (25). The code for SFBC-FSTD is given by:

$$\mathbf{S}_{SFBC-FSTD} = \begin{bmatrix} s_1 & s_2 & 0 & 0 \\ -s_2^* & s_1^* & 0 & 0 \\ 0 & 0 & s_3 & s_4 \\ 0 & 0 & -s_4^* & s_3^* \end{bmatrix} \tag{26}$$

We compare results with two receiver types: MMSE and MLD-RC. MLD-RC[4] is the partitioned ML receiver (given in (21) and (22)) with spherical decoding [8]. We should mention that the MLD-RC receiver for SFBC-FSTD is simply the linear Alamouti receiver, as discussed in II.b. The MMSE receiver is given by:

$$s_{MMSE,est} = \left(\mathbf{H}^H\mathbf{H} + \sigma_n^2\mathbf{I}\right)^{-1}\mathbf{H}^H\vec{r} \tag{27}$$

We first present results for an uncoded 4x1 system. The motivation for studying uncoded 4x1 is that the actual artifacts of the space-frequency code and the impact of different receiver architectures can be better studied and analyzed. Results for QPSK are shown in Figure 7. We make several observations from Figure 7:

- The diversity gain of the QO-SFBC-CR code, as measured by the slope of the FER-SNR curve, with an MLD receiver is 3.3. The theoretically maximum diversity gain with 4 antennas is 4. The loss in diversity from the theoretical maximum can be attributed to antenna correlation [9], which is present in the SCM-C channel model.
- There is a small loss in performance for the QO-SFBC-CR code with the partitioned MLD receiver (MLD-RC). This is due to the lack of channel invariance across 4 frequency tones, an assumption made by the partitioned ML receiver. Note that spherical decoding does not results in any loss of performance.
- There is loss of both diversity gain and coding gain for the QO-SFBC-CR code with an MMSE receiver. The diversity gain drops sharply from 3.3 to 2.0.
- The diversity gain of SFBC-FSTD is 1.62, irrespective of the receiver type, i.e. performance of SFBC-FSTD is similar with both MLD and MMSE receivers.
- Hence, we can state that SFBC-FSTD offers the smallest diversity gain. And with an MLD-RC receiver, the diversity gain offered by QO-SFBC-CR is 2x that of SFBC-FSTD.
- Finally, SFBC-FSTD offers a much lower coding gain compared to QO-SFBC-CR, irrespective of the receiver type.

[4] MLD-RC stands for Maximum Likelihood Decoding with Reduced Complexity

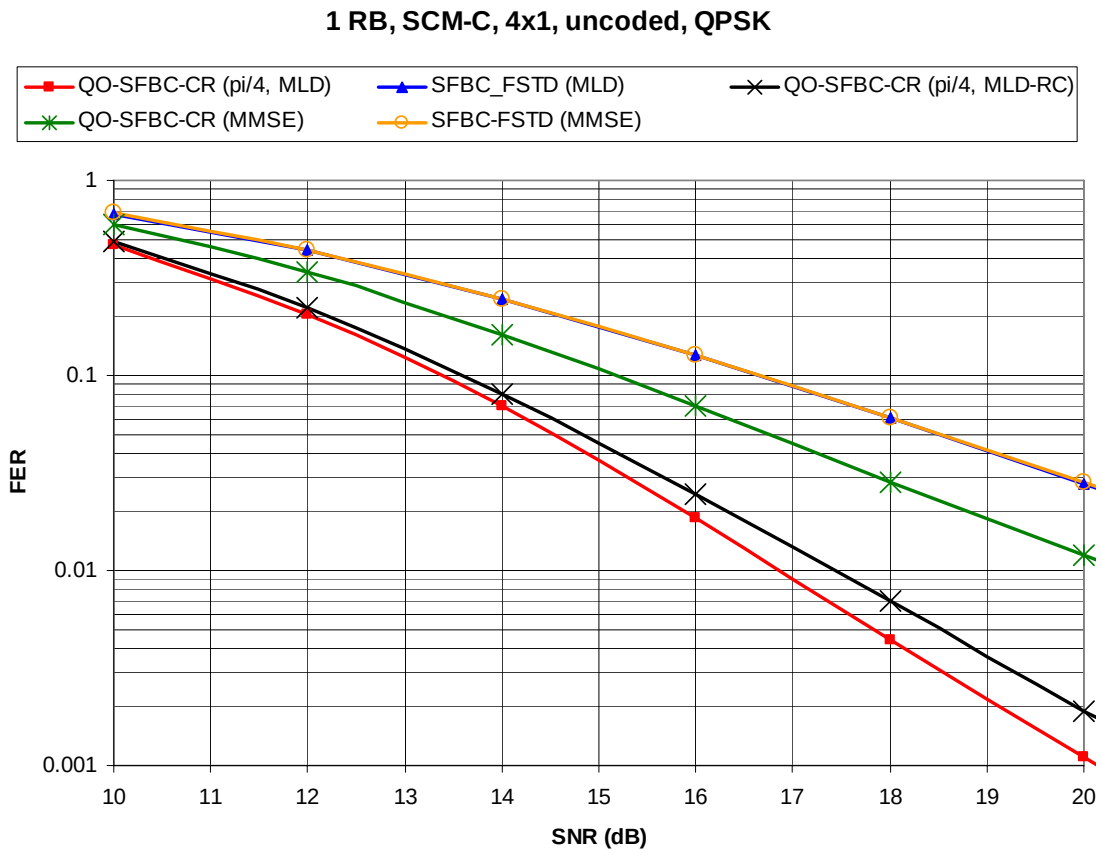

Figure 7: Uncoded 4x1 QPSK FER results for QO-SFBC-CR and SFBC-FSTD

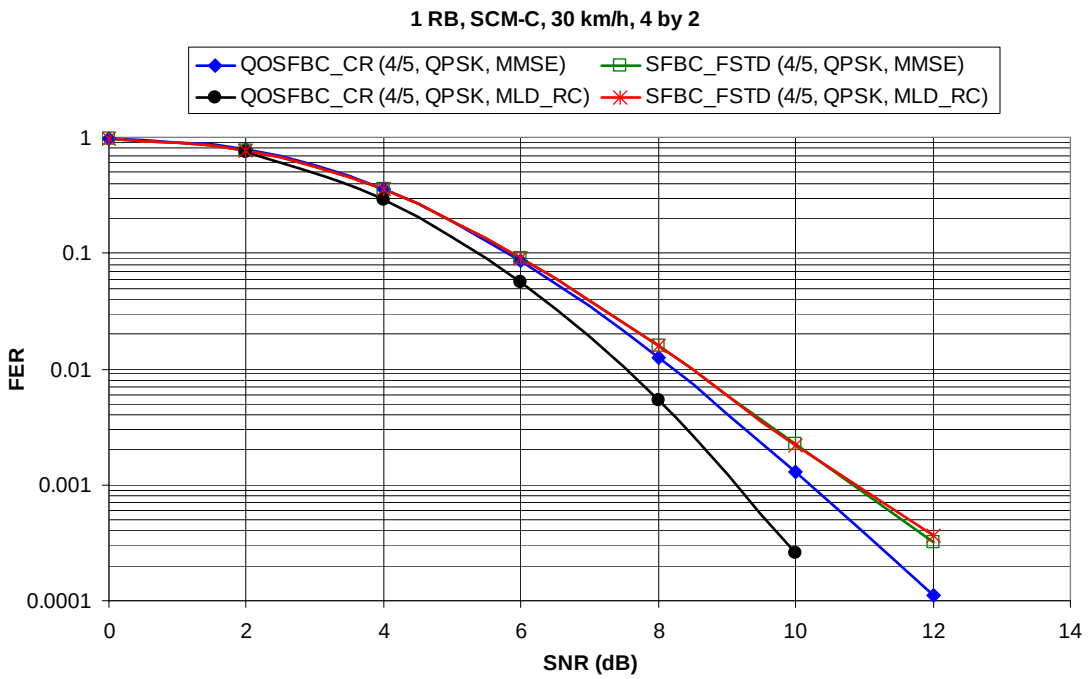

Figure 8: Performance comparison between SFBC-FSTD and QO-SFBC-CR for QPSK

In Figure 8, we present FER results for QPSK constellation with a 4x2 antenna configuration, scheduling across 1RB with 30km/hr of Doppler in an SCM-C channel with coding rate = 4/5. We make several observations:

- Best performance is obtained for QO-SFBC-CR with an MLD-RC receiver.
- For an MMSE receiver, performance of QO-SFBC-CR is better than SFBC-FSTD.

In Figure 9, we present FER results for 16-QAM constellation with a 4x2 antenna configuration, scheduling across 1RB with 30km/hr of Doppler in an SCM-C channel with coding rate = 4/5. We make the following observations:

- Best performance is obtained for QO-SFBC-CR with an MLD-RC receiver.
- With an MMSE receiver, the performance gap between QO-SFBC-CR and SFBC-FSTD reduces.

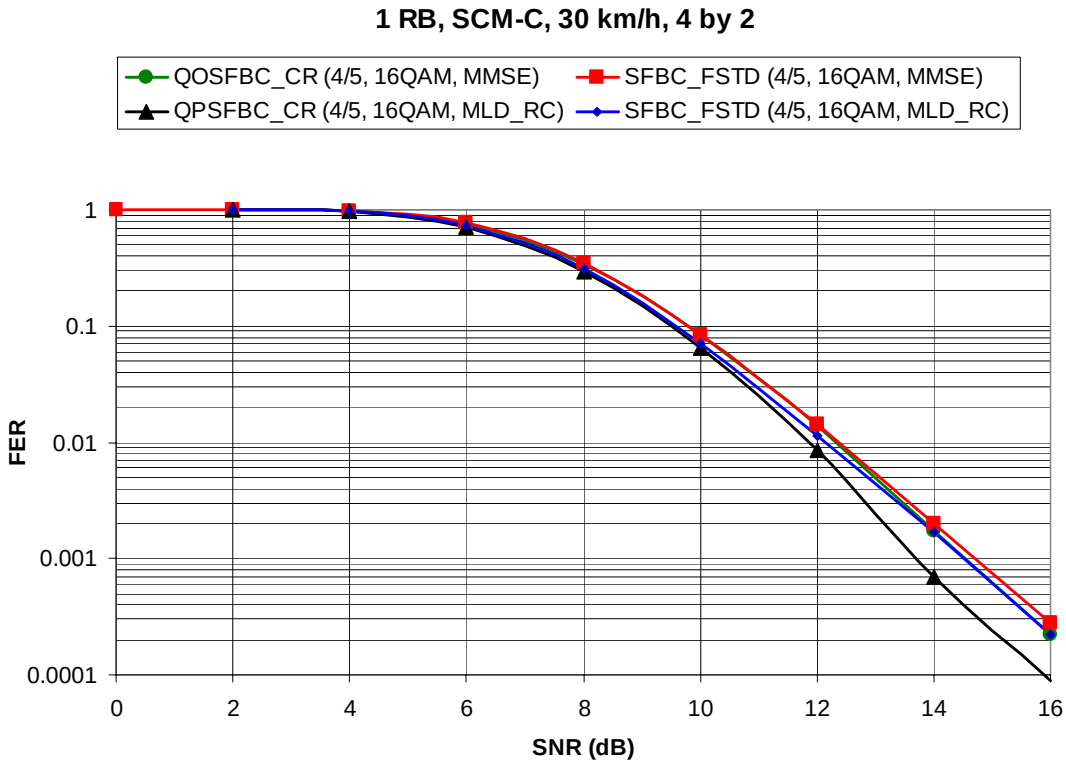

Figure 9: Performance comparison between SFBC-FSTD and QO-SFBC-CR for 16-QAM

Hence, we can state that with an MLD-RC receiver, QO-SFBC-CR performs better than SFBC-FSTD, irrespective of constellation. With an MMSE receiver, QO-SFBC-CR performs better than SFBC-FSTD for QPSK constellations. And, for 16-QAM with an MMSE receiver, the performance of both codes is similar. Therefore, QO-SFBC-CR is a more attractive code than SFBC-FSTD when both performance and complexity considerations are taken into account.

IV CONCLUSION

In this contribution, we've provided details on code construction of high performance QO-SFBC-CR codes. These codes are an extension of the 2-antenna SFBC, and they

exploit constellation rotation (CR) to maximize diversity gain and coding gain. We've also provided details on how to implement a substantially complexity reduced ML decoder (MLD-RC) for QO-SFBC-CR codes. We achieve complexity reduction from two aspects: (1) the proposed QO-SFBC-CR code can be partitioned, and (2) we can leverage well known spherical decoding techniques to further reduce complexity.

We presented performance results comparing QO-SFBC-CR with SFBC-FSTD. With an MLD-RC receiver, we show that QO-SFBC-CR offers superior performance compared to SFBC-FSTD, regardless of constellation size. For QPSK with an MMSE receiver, we show that QO-SFBC-CR performs better than SFBC-FSTD. For 16-QAM with an MMSE receiver, the performance of the two codes is comparable.

Therefore, our recommendation is to adopt the QO-SFBC-CR code as a technique for open loop transmit diversity with 4 transmit antennas.

V REFERENCES

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