3GPP TSG-Ran WG1 #46 **R1-062122**
Tallinn, Estonia
Aug 28 - Sep 1, 2006

Agenda Item: 8.5
Source: NEC Group
Title: Link Analysis of Multi-rank Beamforming
Document for: Discussion and Decision

---

1 Introduction

In this contribution, we first describe our proposed multi-rank beamforming scheme in more detail. This contribution also presents link-level simulation results on the achievable data rate by employing 2×2 and 4×2 MIMO systems in OFDM downlink for E-UTRA. We also present optimal precoding bounds that serve as upper bounds on any precoding scheme using finite rate feedback.

2 Overview

Precoding was discussed in Shanghai meeting where it was agreed to further study only linear precoding. We state our views on the following issues that were identified for further discussion. Supporting arguments are provided in the sequel.

1. Whether to use unitary or non-unitary precoding matrix.

   -We support unitary precoding for SU-MIMO.

2. Codebook based versus non-codebook based precoding.

   -We support codebook based precoding

3. Number of codewords, i.e., SCW versus MCW.

   -This issue needs further study. The need for a higher throughput decoder in SCW versus MCW is also one of important issues that should be considered.

4. Upper bounds on the performance of precoding schemes and the optimal precoding codebook size.

   -Three different upper bounds for MIMO OFDM have been derived and the performance of multi-rank beamforming and some other schemes, e.g., antenna cycling have been compared with the derived upper bounds. The derived upper bounds hold for any unitary or non-unitary precoding techniques and any codebook based or non-codebook based approach. Based on the simulation results, we also provide some views on the codebook size to achieve certain performance in different 2×2 and 4×2 DL MIMO scenarios. The feedback overhead should be considered along with the increase in downlink capacity.

The results in the following indicate the effectiveness of unitary precoding and codebook-based scheme in approaching the precoding bounds. We believe unitary linear precoding is an essential and important part of the SU-MIMO performance improvement technique. To this end, we believe appropriate precoding structure is necessary to simultaneously consider the trade-off between several important factors: higher throughput gain, lower feedback overhead, lower memory requirement, and lower computational complexity. For a given feedback overhead, a rank specific codebook is essential to achieve the highest throughput. As presented in Seoul meeting [2], the structured codebook and successive precoding substantially reduces the computational complexity and the memory requirement for storing the codebooks. In this contribution, we describe the codebook structure in more detail.

3 Codebook Structure

Multi-rank beamforming uses rank specific codebooks as described in [2, 3] . Therefore, we need to construct codebooks for all the cases of 4×4, 4×3, 4×2, and 4×1 depicted in Figure 1 for 4 transmit antennas at node B and 2×2, 2×1 cases for two transmit antennas at node B. However, by using successive beamforming described in the next

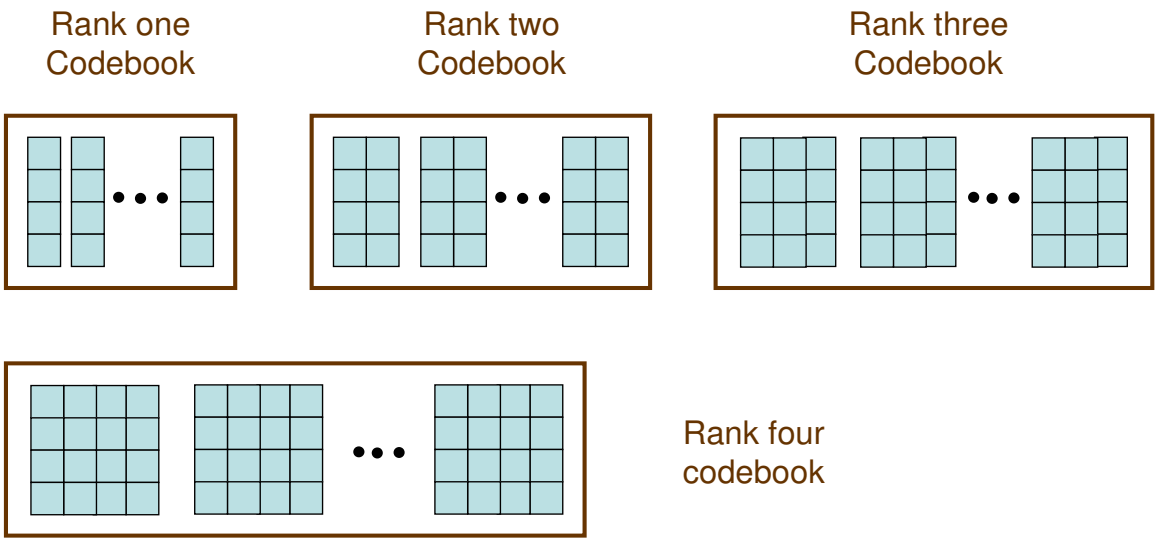

Figure 1: Representation of rank specific codebook for 4 transmit antennas at node B

section, all the codebooks are derived from 3 vector codebooks C_4, C_3, and C_2 consisting of 2^{B_4} (4×1)-dimensional vectors, 2^{B_3} (3×1)-dimensional vectors, and 2^{B_2} (2×1)-dimensional vectors, respectively (please see Figure 2 and 3). Each vector codebook is optimally designed by maximizing the minimum distance (e.g, chordal distance, Fubini-Study distance, p-norm distance) between each pair of the vectors in the codebook.

For rank one transmission the vectors from the codebooks C_4 and C_2 are used directly, and the FBI requirement is B_4 and B_2 bits plus bits required to specify the rank, respectively. The 2×2 codebook is simply constructed by adding an orthonormal vector to each of the vectors in the codebook C_2 to construct 2^{B_2} matrices of size 2×2 each. For rank two transmission from Node B with 4 antennas, the 4×3 codebook is constructed by considering the cartesian product $C_4 \times C_3$. Therefore, the FBI feedback requirement is $B_4 + B_3 + rank$ bits. For rank three transmission, only a subset C_2' of size $2^{B_2'}$ from the codebook C_2 is considered and the 4×3 codebook is constructed as the cartesian product $C_4 \times C_3 \times C_2'$ and the FBI feedback requirement is $B_4 + B_3 + B_2' + rank$ bits. By adding the fourth orthonormal vector to each of the matrix in 4×3 codebook, we obtain the rank 4 codebook, in this case.

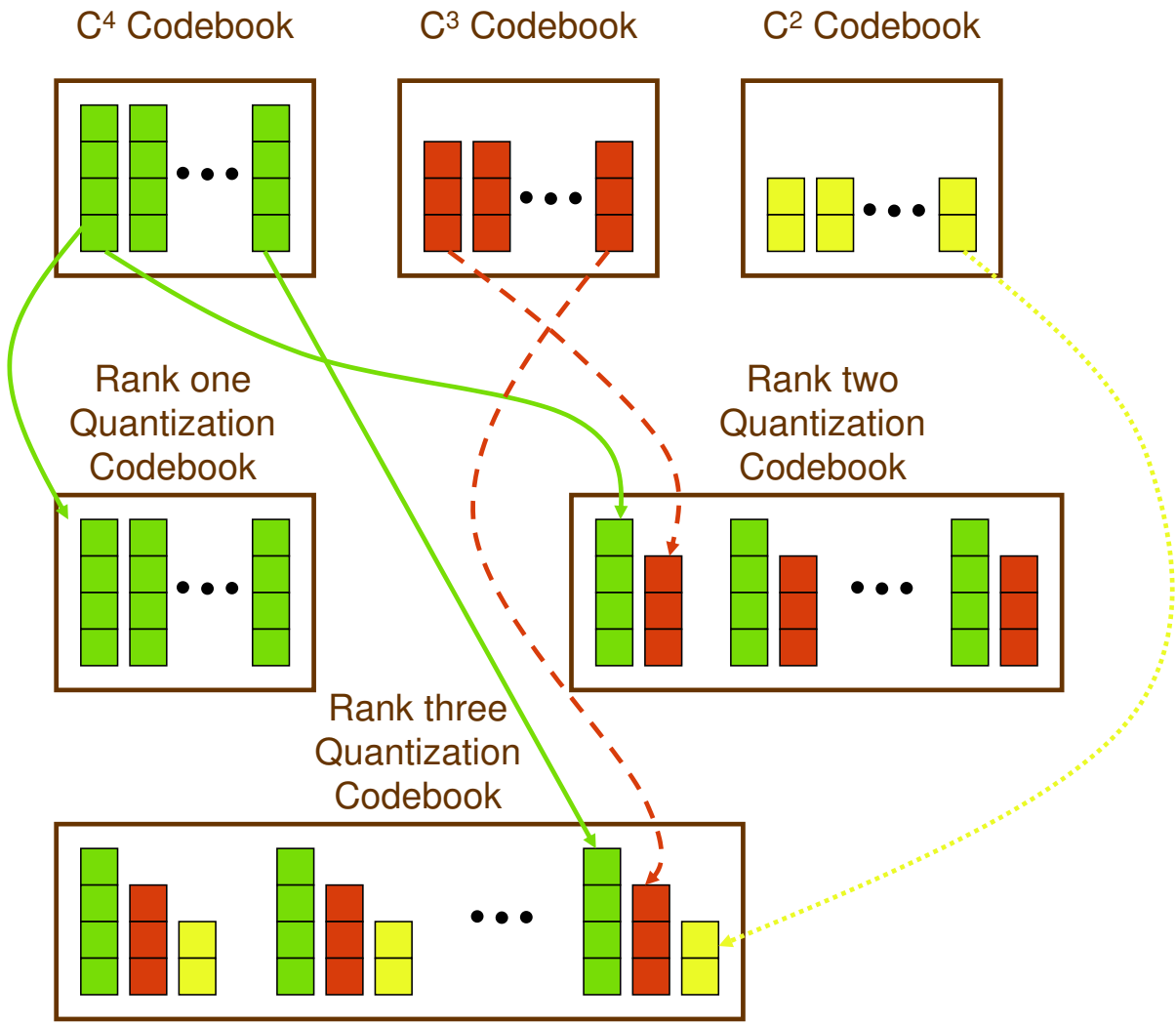

Figure 2: Representation of rank specific codebook derived from 3 vector codebook of 4, 3, and 2-dimensional vectors.

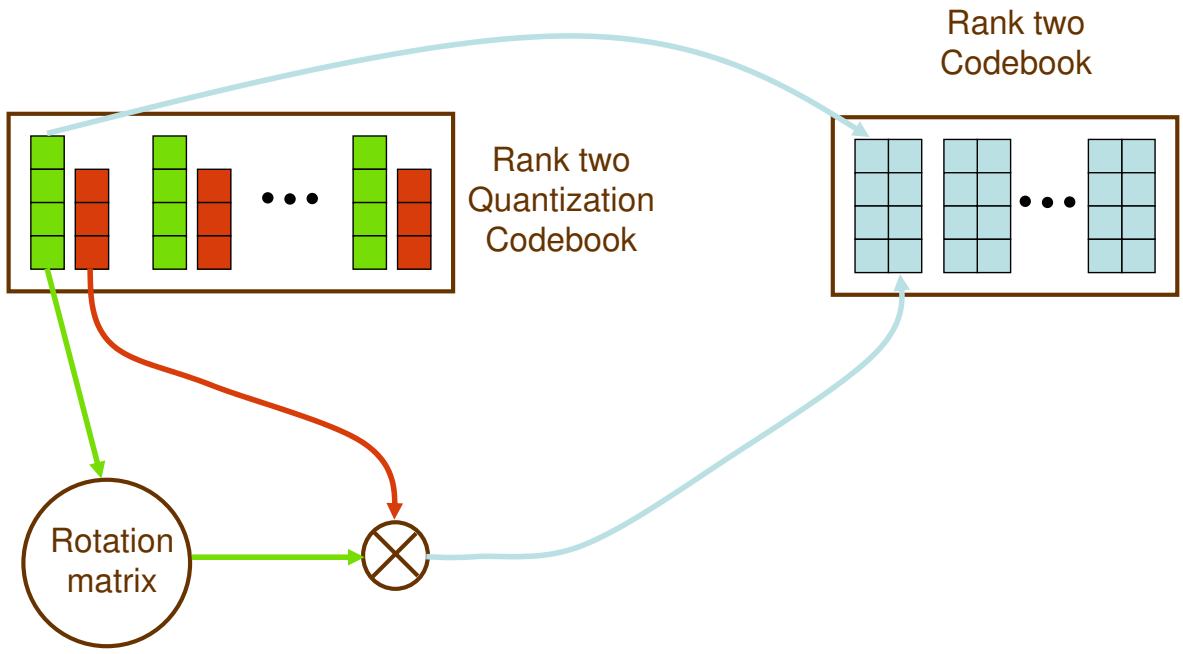

Figure 3: Example of Regenerating transmission codebook from quantization codebook for rank two.

While the bits for the FBI feedback for both rank one and rank two transmission from a Node B equipped with 2 antennas are the same, the above derivation of the codebook for the base station with 4 antennas presents a variable feedback for different ranks, i.e., $B_4 + rank < B_4 + B_3 + rank < B_4 + B_3 + B_2' + rank = B_{\max}$. Clearly, if the FBI feedback bits for all the cases are equal, the above construction would incur a loss from the possible optimal rank specific codebook design. One possibility, for example, is to construct larger codebook C_4' of size $2^{B_{\max}-rank}$ with C_4 as embedded codebook.

To avoid the extra memory requirement for keeping such large codebook, we propose a recursive beamforming also described in the following sections. The recursive beamforming again uses the same vectors from C_4, C_3 and C_2' to construct a $2^{B_4+B_3+B_2'}$ sized vector codebook of (4×1)-dimensional vectors. Such an extension of C_4 to a larger codebook does not require extra memory, while in practice the performance degradation w.r.t. the optimal design is negligible.

4 Successive Beamforming

The optimal precoding index can be found through wide-band precoding over the set of OFDM tones in each sub-band, i.e, multiple adjacent chunks over which the FBI feedback is reported. To avoid the computational complexity of finding the optimal wide-band precoder over large cluster sizes, e.g., where $T_2 - T_1 = 75$ OFDM tones $= 3$ chunks, by calculating the SINRs for all the entries of the codebook, we first compute a channel representative for the whole sub-band from T_1 to T_2 as follows

$$H_{Rep} = \left\{ \sum_{i=T_1}^{T_2} H_i^* H_i \right\}^{1/2} \tag{1}$$

where $\{.\}^{1/2}$ denotes the matrix square root operation and H_i is the estimate of the channel for the ith OFDM tone. Then, we select the transmission rank, denoted by N, based on a modified capacity measure [3] and by using our proposed successive beamforming, we quantize the first N dominant right eigen-modes of H_{Rep} using SVD decomposition.

Let SVD decomposition of the channel estimate $\mathbf{H_k} = \mathbf{H_{Rep}}$ for k^{th} user channel in the sub-band T_1 to T_2 be given as

$$\mathbf{H_k} = \mathbf{U_k} D_k \mathbf{V_k^*} \tag{2}$$

where the columns of the unitary matrix $\mathbf{V_k} = [\mathbf{v_1}\ \mathbf{v_2}\ \ldots\ \mathbf{v}_{N_k}]$ represent different eigen-vectors of the channel. For a given rank N, the aim of quantized beamforming is to find the best quantization of the first N dominant eigen-vectors of the channel. If B_i bits is used for the quantization of ith eigen vector $\mathbf{v_i}$, the quantized vector $\hat{\mathbf{v}}_\mathbf{i}$ is given by

$$\begin{aligned} \hat{\mathbf{v}}_i = & \qquad \arg\max \qquad |\mathbf{v}_i^*\ w| \\ w \in \mathcal{C}_i = & \{w_1, w_2, \ldots, w_{2^{B_i}}\} \end{aligned} \tag{3}$$

where $\mathcal{C}_i$ is the corresponding quantization codebook.

The successive beamforming provides a more efficient quantization of the eigen-vectors by using the fact that the column vectors $\mathbf{v_1}\ \mathbf{v_2}\ \ldots\ \mathbf{v}_N$ are orthogonal, therefore, the quantization of each vector can be successively performed in a lower dimensional sub-space using the knowledge of previously quantized vectors. The successive quantization is performed as follows.

Step 1: First, $\mathbf{v_1}$ is quantized. Since, $\mathbf{v_1}$ is an M dimensional vector, the quantization codebook $\mathcal{C}^{(M)} = \{w_1, w_2, \ldots, w_{2^{B_1}}\}$ consisting of the M-dimensional vectors in complex space $\mathbb{C}^M$ will be used which is of size 2^{B_1} using B_1 bits of feedback for quantization of $\mathbf{v_1}$. The quantized vector $\hat{\mathbf{v}}_1 \in \mathcal{C}^{(M)}$ is chosen to maximize $| < \mathbf{v_1}, \hat{\mathbf{v}}_1 > | = |\mathbf{v_1^*}\hat{\mathbf{v}}_1|$. The index of $\hat{\mathbf{v}}_1$ then constitutes the quantized feedback information for the first vector.

Step 2: Then, we find a rotation matrix $\phi(\hat{\mathbf{v}}_1)$ such that

$$\phi(\hat{\mathbf{v}}_1)\hat{\mathbf{v}}_1 = \mathbf{e}_1 = [1;\ 0;\ 0;\ \ldots;\ 0] \tag{4}$$

Step 3: Next, we rotate all the vectors $\mathbf{v_1}$ $\mathbf{v_2}$ ... $\mathbf{v}_N$ by the rotation matrix $\phi(\hat{\mathbf{v}}_1)$. We have

$$V' = [\mathbf{v}_1'\ \mathbf{v}_2'\ \ldots\ \mathbf{v}_n'] = \phi(\hat{\mathbf{v}}_1)V = [\phi(\hat{\mathbf{v}}_1)\mathbf{v}_1\ \phi(\hat{\mathbf{v}}_1)\mathbf{v}_2\ \ldots\ \phi(\hat{\mathbf{v}}_1)\mathbf{v}_N]. \tag{5}$$

If $\hat{\mathbf{v}}_1 = \mathbf{v_1}$ the first vector $\mathbf{v_1'}$ becomes $\mathbf{e_1}$ and all the first elements of $\mathbf{v_2'}$ $\mathbf{v_3'}$ ... $\mathbf{v_N'}$ are zero due to the fact that V is a unitary matrix and all of its column are orthogonal. Thus, if the quantization codebook is fine enough such that $\hat{\mathbf{v}}_1$ closely approximates $\mathbf{v_1}$, then we can ignore the first element of the rotated eigen-vectors $\mathbf{v_2'}$ $\mathbf{v_3'}$ ... $\mathbf{v_N'}$ and quantize the next eigen-vector in a lower dimensional subspace.

Step 4: We consider the new matrix $\tilde{V} = V'(2:M, 2:M)$ and perform the same beamforming procedure described in Step 1 through Step 3 above on the new matrix $\tilde{V}$. This step will successively be performed until all the vectors are quantized. It should be pointed out that since the length of the vectors are reduced by one, the codebook used in Step 1 is now one dimension lower than the codebook used in previous iteration.

The above successive beamforming technique generates different amount of feedback for different rank. Therefore, for a fixed number of feedback bits per block of transmission, we can generate different size codebooks in order to use up all available bits for different rank beamforming. For example, consider a case where the possible ranks of the channel are 1 or 2, and 8 bits of feedback is available. We can dedicate 1 bit for rank selection and use the other 7 bits for description of the eigenmodes. For rank 2, we can use 4 bits for the first eigenmode and 3 bits for the second eigenmodes. Therefore, we have

$2^4 = 16$ M-vectors to quantize the first eigenmode and $2^3 = 8$ (M-1)-vector to quantize the second eigenmode. For the rank 1, we then use all 7 bits and therefore generate $2^7 = 128$ M-vectors to be used for the quantization of the dominant eigenmode of the channel.

Due to the higher computational complexity and memory requirement of such beamforming strategy, we present an alternative beamforming strategy which uses the same idea of successive beamforming described above to quantize the vectors for the lower rank, e.g. rank 1 case in the above example. The following section describes the recursive beamforming strategy which illustrates how in our example the same codebook of 2^4 M-vectors plus 2^3 (M-1)-vectors used for rank 2 beamforming can also be used for the rank 1 beamforming.

5 Recursive beamforming of lower rank vectors

In the recursive beamforming, we use the idea of successive beamforming presented in the previous section in order to quantize the vectors in a lower rank transmission. In successive beamforming we quantize a set of orthogonal vectors $\mathbf{v_1}, \mathbf{v_2}, \ldots, \mathbf{v_k}$, however, in recursive beamforming, we recursively generate this set for only one vector $\mathbf{v_1}$. After quantizing $\mathbf{v_1}$ to $\hat{\mathbf{v}}_\mathbf{1}$ in the first step, we define $\mathbf{v_2}$ as the residual of $\mathbf{v_1}$ on the orthogonal space $span\{\hat{\mathbf{v}}_1^\perp\}$, and so on. Another difference between the successive beamforming strategy on the set of orthogonal vectors $\mathbf{v_1}, \mathbf{v_2}, \ldots, \mathbf{v_k}$ and the recursive beamforming of $\mathbf{v_1}$ and its residuals $\mathbf{v_2}, \mathbf{v_3}, \ldots, \mathbf{v_k}$ is that we need to quantize the ratio of the projection of $\mathbf{v_1}$ in each successive space for the reconstruction for the recursive beamforming. For example, for $k = 2$ we need to quantize the value $\alpha = \mathbf{v}_1^* \hat{\mathbf{v}}_1$. Since the recursive beamforming is used when the chosen rank k by UE is lower than the maximum possible rank $N_{\max}$, the reserved feedback bits for quantization of other eigenmodes can be used for quantizing the projection values. As a result, the recursive beamforming can be performed without increasing the overall feedback overhead for the system.

Assume that the vector $\mathbf{v_1}$ is to be quantized using k codebooks of M-vectors, (M-

1)-vectors, up to (M-k+1)-vectors, respectively.

Step 1: Since $\mathbf{v_1}$ is an M dimensional vector, it is quantized using the codebook $\mathcal{C}^{(M)}$ of the M-vectors in $\mathbb{C}^M$. The quantized vector $\hat{\mathbf{v}}_1 \in \mathcal{C}^{(M)}$ is chosen to maximize $| < \mathbf{v}_1, \hat{\mathbf{v}}_1 > | = |\mathbf{v}_1^* \hat{\mathbf{v}}_1|$.

Step 2: To find a finer description of $\mathbf{v_1}$, we find the residual part of $\mathbf{v_1}$ that lies in the orthogonal space defined by $span\{\hat{\mathbf{v}}_1^{\perp}\}$. We define

$$\mathbf{v'}_2 = \mathbf{v}_1 - (\mathbf{v}_1^* \hat{\mathbf{v}}_1)\hat{\mathbf{v}}_1 \tag{6}$$

and normalize the vector, i.e., $\mathbf{v_2} = \mathbf{v'_2}/|\mathbf{v'_2}|$.

Step 3: Then, we find a rotation matrix $\phi(\hat{\mathbf{v}}_1)$ such that

$$\phi(\hat{\mathbf{v}}_1)\hat{\mathbf{v}}_1 = \mathbf{e}_1 = [1;\ 0;\ 0;\ \ldots;\ 0] \tag{7}$$

Step 4: Next, we rotate the vector $\mathbf{v_2}$ by the rotation matrix $\phi(\hat{\mathbf{v}}_1)$. We have

$$\mathbf{v}_2'' = \phi(\hat{\mathbf{v}}_1)\mathbf{v}_2 \tag{8}$$

where the first elements of $\mathbf{v}_2''$ is zero due to the fact that $\mathbf{v}_2 \in span\{\hat{\mathbf{v}}_1^{\perp}\}$, i.e., $< \mathbf{v}_2, \hat{\mathbf{v}}_1 >= 0$.

Step 5: Therefore, we consider the new vector $\tilde{\mathbf{v}}_2 = \mathbf{v}_2''(2 : M)$ and perform the same beamforming procedure described in Step 1 through Step 3 on the new vector $\tilde{\mathbf{v_2}}$. This step will successively be performed until all the codebooks of M-vectors, (M-1)-vectors, up to (M-k+1)-vectors are used.

We explain the reconstruction of the vector $\mathbf{v_1}$ with the following example. Assume only two codebooks are used and we successively obtained $\hat{\mathbf{v}}_1$ and $\hat{\mathbf{v}}_2$ as the quantized

vectors. We have

$$\begin{aligned}
\mathbf{v}_1 &= (\mathbf{v}_1^*\hat{\mathbf{v}}_1)\hat{\mathbf{v}}_1 + \mathbf{v}'_2 && (9)\\
&= (\mathbf{v}_1^*\hat{\mathbf{v}}_1)\hat{\mathbf{v}}_1 + |\mathbf{v}'_2|\mathbf{v}_2 && (10)\\
&= |\mathbf{v}'_2|(\frac{\mathbf{v}_1^*\hat{\mathbf{v}}_1}{|\mathbf{v}'_2|}\hat{\mathbf{v}}_1 + \mathbf{v}_2) && (11)\\
&= |\mathbf{v}'_2|(\frac{\mathbf{v}_1^*\hat{\mathbf{v}}_1}{|\mathbf{v}'_2|}\hat{\mathbf{v}}_1 + \phi(\hat{\mathbf{v}}_1)^{-1}\mathbf{v}''_2) && (12)\\
&= |\mathbf{v}'_2|(\frac{\mathbf{v}_1^*\hat{\mathbf{v}}_1}{|\mathbf{v}'_2|}\hat{\mathbf{v}}_1 + \phi(\hat{\mathbf{v}}_1)^{-1}\begin{bmatrix} 0 \\ \tilde{\mathbf{v}}_2 \end{bmatrix}) && (13)\\
&\approx \sqrt{1-|\hat{\alpha}|^2}(\frac{\hat{\alpha}}{\sqrt{1-|\hat{\alpha}|^2}}\hat{\mathbf{v}}_1 + \phi(\hat{\mathbf{v}}_1)^{-1}\begin{bmatrix} 0 \\ \hat{\mathbf{v}}_2 \end{bmatrix}) && (14)
\end{aligned}$$

where in (14) all the quantities are known and we have $|\mathbf{v}'_2| = \sqrt{1-|\hat{\alpha}|^2}$ which is obtained through the fact that the beamforming vector $\mathbf{v}_1$ is unit norm, i.e., $|\mathbf{v}_1| = 1$.

6 CQI reporting

In order to avoid capacity loss and achieve better rates for rank r transmission, one may either use SCW transmission by encoding over all r streams or MCW by encoding one codeword per stream and perform SIC decoding. Using SCW versus MCW has its pros and cons. SCW requires only one CQI feedback but MCW requires multiple CQI feedback for all the streams. On the other hand, SCW needs a higher throughput decoder, e.g., SCW would a single decoder with the throughput of 100 MBPS at the anticipated peak rate of E-UTRA for 2×2 scenario, whereas the peak rate of each stream at MCW is half, i.e., 50 MPBS. The CQI feedback requirement of multiple codewords can be reduced by layer permutation. However, the use of MMSE-SIC for downlink is not recommended as it imposes much higher computational burden for the UE decoding operation and increased decoding delay, therefore we concentrate on the scheme with LMMSE or QR-MLD demodulators for which there are no extra decoding delays. In order to reduce the throughput requirement for the decoding of each codeword and benefit from MCW operation but not to increase the CQI reporting requirement, we use pseudo-random scrambling of each transmission codeword into all layers. As a result, only

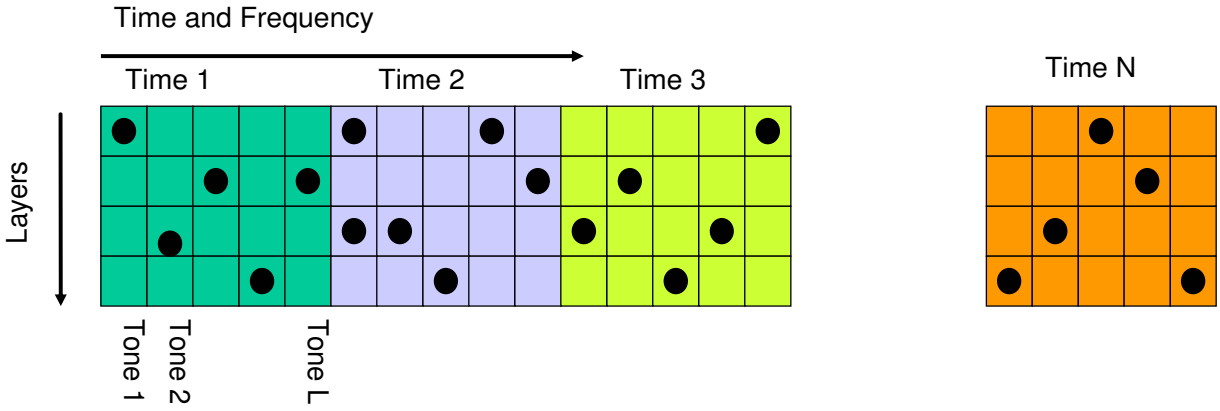

Figure 4: Representation of stream scrambling over multiple layers

one CQI will be reported that is used by Node B to choose the rate for the encoding of all codewords. Such random scrambling of the codewords into the layers symmetrizes the system while in combination with LMMSE or better QR-MLD decoding, the performance loss is negligible. The Pseudo-random number generator that is used for such layering is chosen in static way and is known to all UEs and Node B in advance. An example of such scrambling is depicted in Figure 4

The pseudo-random scrambling reduces the CQI reporting requirement for the MCW so that with the same feedback requirement MCW performs as well as SCW with much lower decoder throughput requirement, because of dividing the throughput between the streams. However, if multiple CQI reporting is available, SCW scheme can also benefit by using different rates and power control over each layer (although only one streams is encoded over multiple layers) to improve the performance.

7 Simulation setup

Simulation parameters are the same as in [3] except that we included link-level SCME channel models proposed in [1] and we considered both 2 and 4 transmit antennas at Node B. We evaluated MIMO schemes in the SCME Urban Macro-cell for 2×2 scenario and SCM-B Urban Macro-cell model with 4×2 systems. The simulation parameters are

Parameter	Assumption
Access	OFDM
RF carrier frequency	2.0 GHz
Bandwidth	5.0 MHz
Number of paths (Multi-path model)	6
Sub-carrier spacing	15.0 kHz
Sampling frequency	7.68 MHz
Number of occupied sub-carriers	300
Number of OFDM symbols per TTI	12 (6 x 2)
Number of data symbols per TTI	3600 (1800 x 2)
Symbol rate	7.2 M/s
CP length	4.82 micro second
FFT point	512
Number of antennas at BS	4 and 2
Number of antennas at MS	2
Number of feedback bits	Various cases (beween 2 to 8 bits/sub-band)
CQI description	5 bits
Channel model	SCME Urban Macro-cell: 3kmph
Channel estimation	Ideal channel estimation
Tx and Rx antenna correlation	Without spatial correlation

Table 1: Simulation parameters

listed in Table 1. We use LMMSE demodulation. We have considered sub-band precoding and evaluated the effect of different cluster sizes (the number of adjacent OFDM chunks in a scheduled sub-band) and various FBI feedback reporting per sub-band. We consider 5MHz bandwidth and the mobile speed of 3 km/h (corresponding to the fading maximum doppler frequency of 5.55 Hz with 2.0GHz carrier frequency).

8 Simulation Results

We evaluated the performance of multirank beamforming (MRBF) scheme. Two precoding bounds have been derived and evaluated. On is a per tone precoding with rate

control (denoted in the figures by PT/RC), $C_{PT/RC}$, and the other one is per cluster precoding (denoted in the figures by PC). The PC precoding bound is in turn evaluated under two different conditions, where the rank control is performed (denoted by RC in the figures), $C_{PC/RC}$, and where only rank 2 precoder is used (denoted by R=2 in the figures), $C_{PC/R=2}$. In general, we have $C_{PT/RC} > C_{PC/RC} > C_{PC/R=2}$. The performance is also compared with two base-line scheme with minimal or no FBI feedback requirement, i.e., antenna cycling (no FBI feedback) and antenna cycling with RC (only 1 bit of feedback). The optimal bounds with perfect channel knowledge at both UE and Node B and without channel state information at Node B are also evaluated as references (denoted by CSIR and CSIRT, respectively).

Although precoding for the full rank cases (2×2 with two streams and 4×4 with 4 streams) does not provide a gain in terms of capacity if MMSE-SIC is used, it has been observed that precoding provides a notable gain in these cases when LMMSE demodulation is used. Intuitively, if perfect precoding is performed so that the streams in different layers become orthogonal the performance of LMMSE is equivalent to MMSE-SIC. The precoding with quantized feedback also benefits from such orthogonalization and notably improves the performance.

One of the goal of this contribution is to evaluate the trade-off between cluster and chunk size and throughput. If the cluster size increases, beyond a certain point, even the optimal precoder with perfect feedback might not show much gain, i.e., there is no gain in increasing the number of feedback bits. This observation has also been made in some earlier contributions that if a single precoding is used for the whole band the gain is marginal. However, this should not conclude that precoding has marginal gain if sub-band scheduling is performed. Moreover, the increase in cluster size can also result in a loss in multi-user diversity via sub-band scheduling. On the other hand, if the cluster and chunk size decrease when the overall feedback overhead is kept constant, there would be a loss in throughput because of the reduction in the precoding codebook size.

Figures 5 and 6 illustrate the performance of 4×2 system with various cluster sizes

of 1, 2, 3, and 12 chunks (corresponding to 25, 50, and 75, and 300 OFDM tones) for 6 bits and 8 bits of feedback per cluster (or sub-band), respectively. Figures 7 and 8 illustrate the performance of 2×2 system for the same set of cluster sizes for 4 and 5 bits of feedback per sub-band, respectively.

The simulation results show that MRBF even with LMMSE decoding and finite codebook generally outperforms $C_{PC/R=2}$ precoding bound because of exploiting rank control. Moreover, MBRF can come fairly close to the optimal precoding bound $C_{PC/RC}$ with 5 bits of feedback per cluster (of size 1) for the case of 2×2. However, a considerable gap remains between the performance of MRBF and the precoding bound $C_{PC/RC}$ in the case of 4×2 even with 8 feedback bits per cluster.

For both cases of 2×2 and 4×2, the precoding bound $C_{PT/RC}$ is very close to optimal CSIRT bound C_{CSIRT} which means that most of the gain in power control can be captured by just rank control.

In general, the memory requirement and complexity of the UE action grows exponentially with the size (in bits) of the precoding codebook. Therefore, for a given (total) number of feedback bits it is more desirable to have a smaller cluster size. However, a very small number of bits per cluster (in other words a small precoding codebook) may provide only a marginal precoding gain. This aspect should also be considered in the tradeoff between the feedback rate and throughput via using the same precoder for a cluster of multiple adjacent chunks.

Figure 7 and 8 illustrate that increasing the cluster size for 2×2 case, in general, the performance of precoding reduces (please see the precoding bounds). It is also shown that a finer quantization (i.e. larger precoding codebook size) is more effective only with smaller cluster size. In fact, for larger cluster sizes, increasing quantization bits and feedback overhead is not much helpful which is in line with a number of previous contributions where a precoding is done for the whole band.

More precisely, the optimal precoding bound $C_{PC/RC}$ for cluster size 12, i.e., the whole band, is very close to the capacity C_{CSIR}, therefore, precoding provides very marginal

performance gain even if MCW and SIC are allowed. For cluster of size 3 chunks, one may also note that, the performance of MCW with SIC is almost equal to LMMSE with 2 bit precoding. To implement MCW with SIC, we need to feedback two CQIs which results in a higher overall feedback requirement versus LMMSE with precoding that requires only one CQI feedback.

Current evaluation denotes three possible feedback scenarios for 2x2 case: (1) a high feedback load Scenario where 5 bits of feedback per cluster of size 1 chunk is used (Figure 9), (2) a moderate feedback load Scenario where 2 bits of feedback for cluster of size 3 chunks is used(Figure 10), and (3) a low feedback load Scenario where 2 bits of feedback for the cluster of size 12 chunks is used.

Compared to the gain offered by CSIRT over baseline antenna cycling that requires no CSIT, the above choices of partial CSIT thorough quantized feedback and MRBF achieve about $80\%, 50\%, 20\%$ percent of this gain gap, respectively.

Similarly, for the 4×2 case two feedback scenario are recognized: (1) a high feedback load scenario where 8 bits feedback per cluster of size 1 chunk is used ((Figure 11), and (2) a low feedback load scenario where 6 bits feedback per cluster of size 12 chunk is used (Figure 12). Comparing the relative gain of the high and low feedback load scenarios, using 8 bits of feedback per cluster of size 3 offers a good tradeoff between performance gain, feedback overhead, and complexity.

9 Conclusion

We have seen that unconstrained precoding bound (bound valid for all unitay and non-unitary precoding schemes and all codebook based or non-codebook based) can be approached with am appropriate unitary precoding, i.e., multi-rank beamforming which is a codebook-based precoding scheme. Therefore, we believe that unitary linear precoding is an essential and important part of the SU-MIMO performance improvement technique.

For a given feedback overhead, a rank specific codebook is essential to achieve the highest throughput. With the goal of achieving the best tradeoff between higher through-

put gains, lower feedback overhead, lower memory requirement, and lower computational complexity, we believe EUTRA DL MIMO should benefit from an structured precoding codebook. We also provided some views on the required feedback bits to achieve certain performance in different 2×2 and 4×2 DL MIMO scenarios.

References

[1] R4-060334, Ericsson, Elektrobit, Nokia, Motorola, Siemens, "LTE channel models and simulations"

[2] R1-051455, NEC Group, "MIMO techniques for Downlink E-UTRA" *TSG-RAN WG1 Meeting #43*, Seoul, Republic of Korea November 7th - 11th 2005

[3] R1-060362, NEC Group, "MIMO techniques for Downlink E-UTRA: Detailed description and Simulation results" *TSG-RAN WG1 Meeting #44*, Denver, Colorado, USA February 13th - 17th 2006

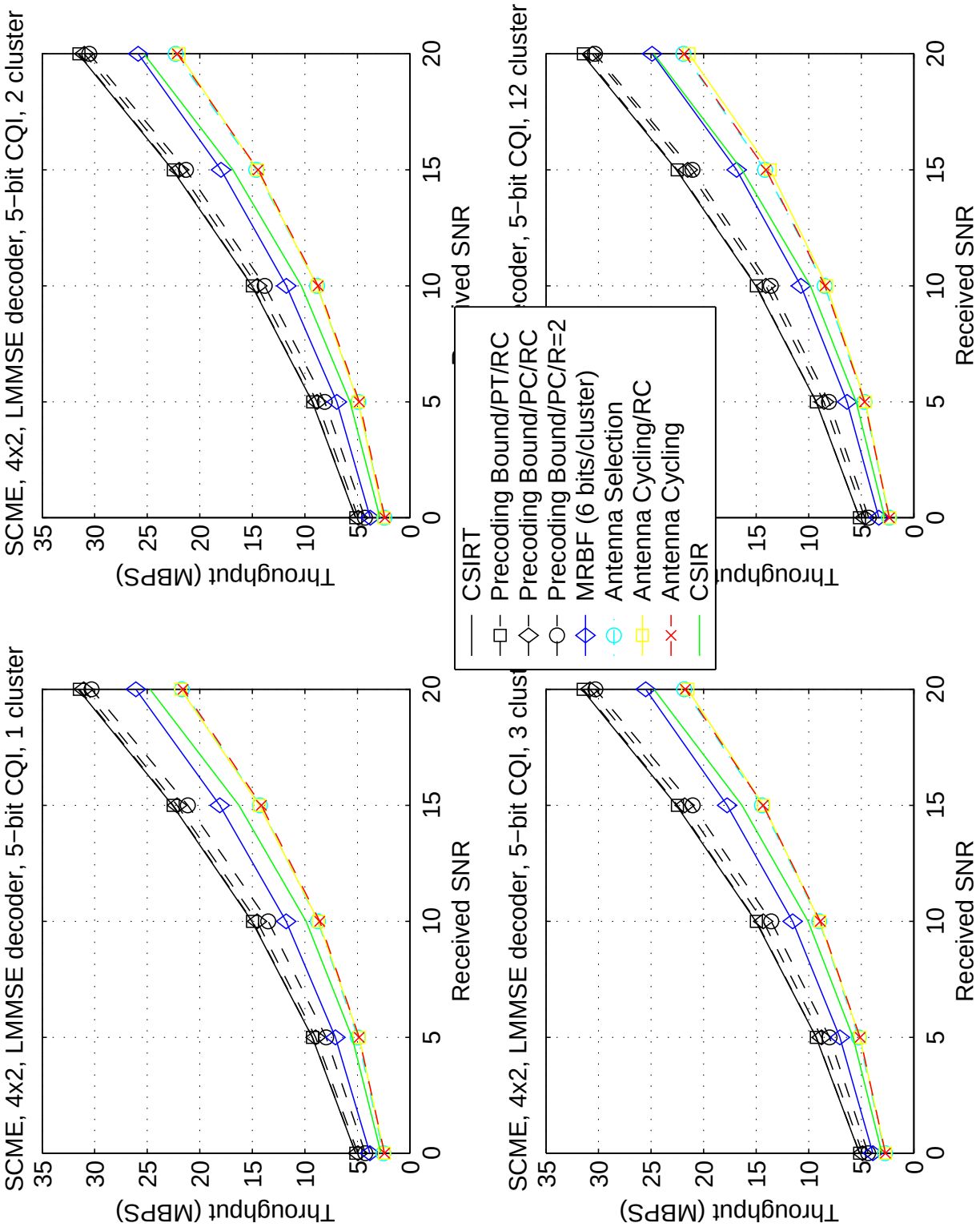

Figure 5: Effect of Various cluster sizes and sub-band scheduling for 4×2 system with 6 bits FBI feedback per sub-band.

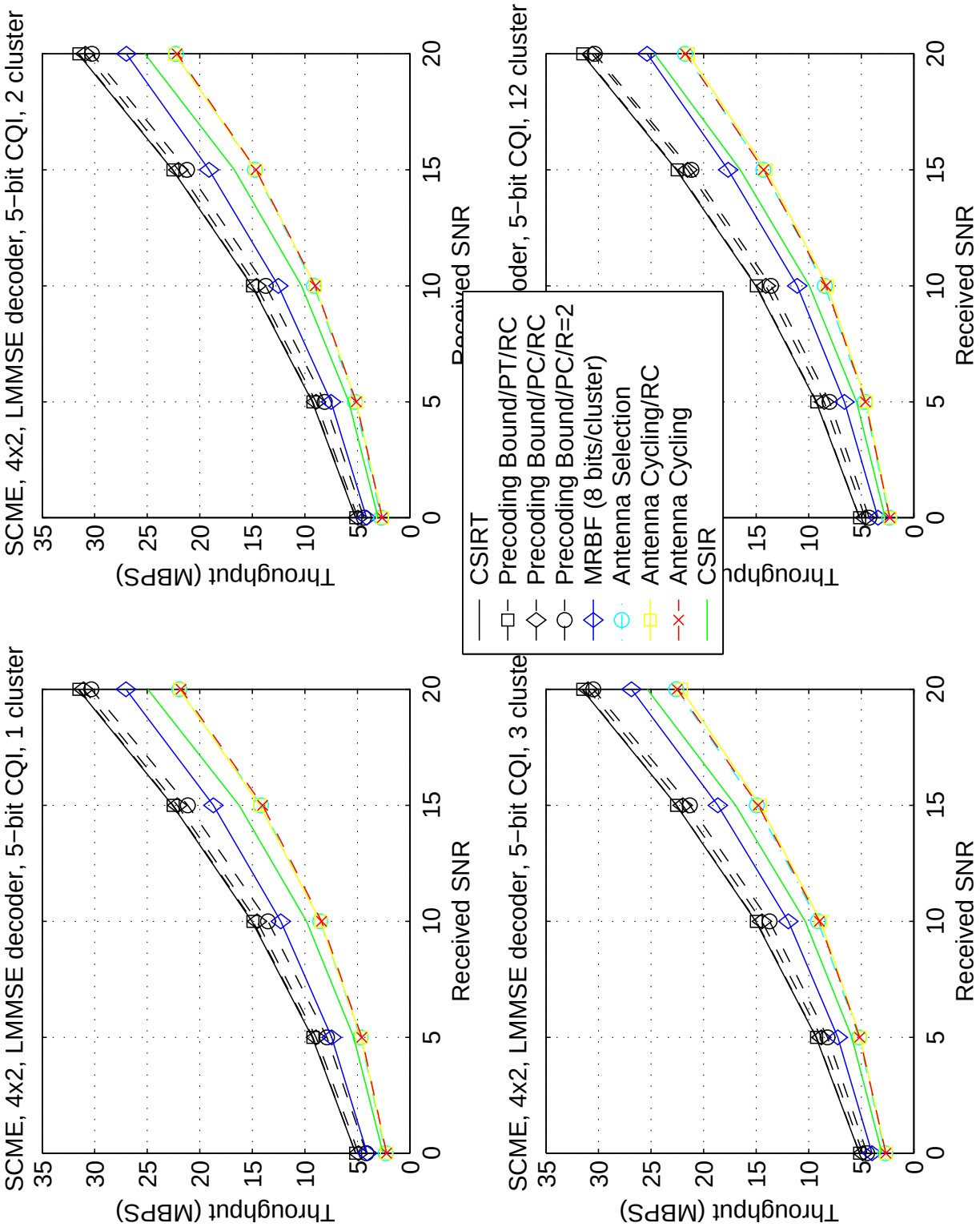

Figure 6: Effect of Various cluster sizes and sub-band scheduling for 4×2 system with 8bits FBI feedback per sub-band.

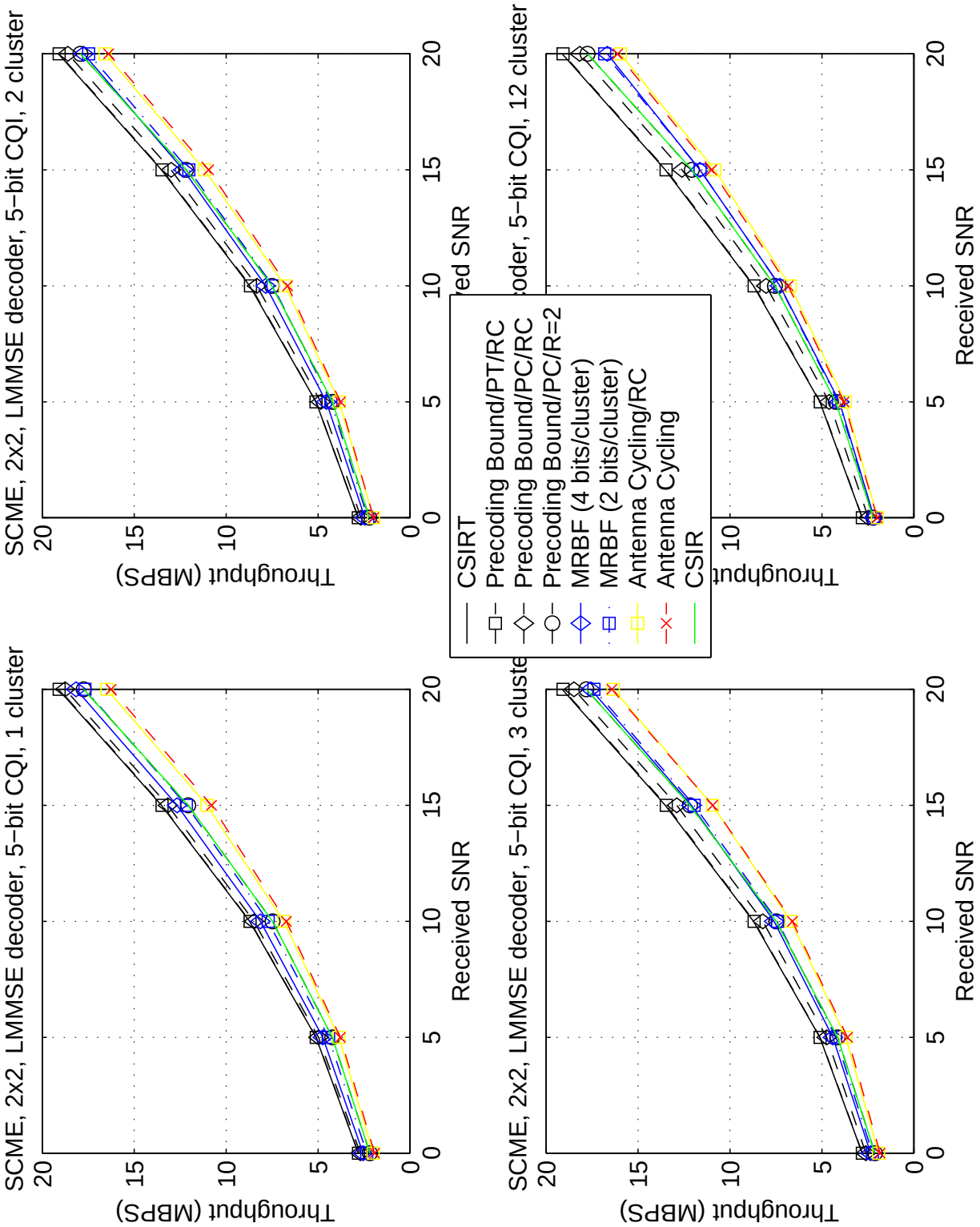

Figure 7: Effect of Various cluster sizes and sub-band scheduling for 2×2 system with 2 and 4 bits FBI feedback per sub-band.

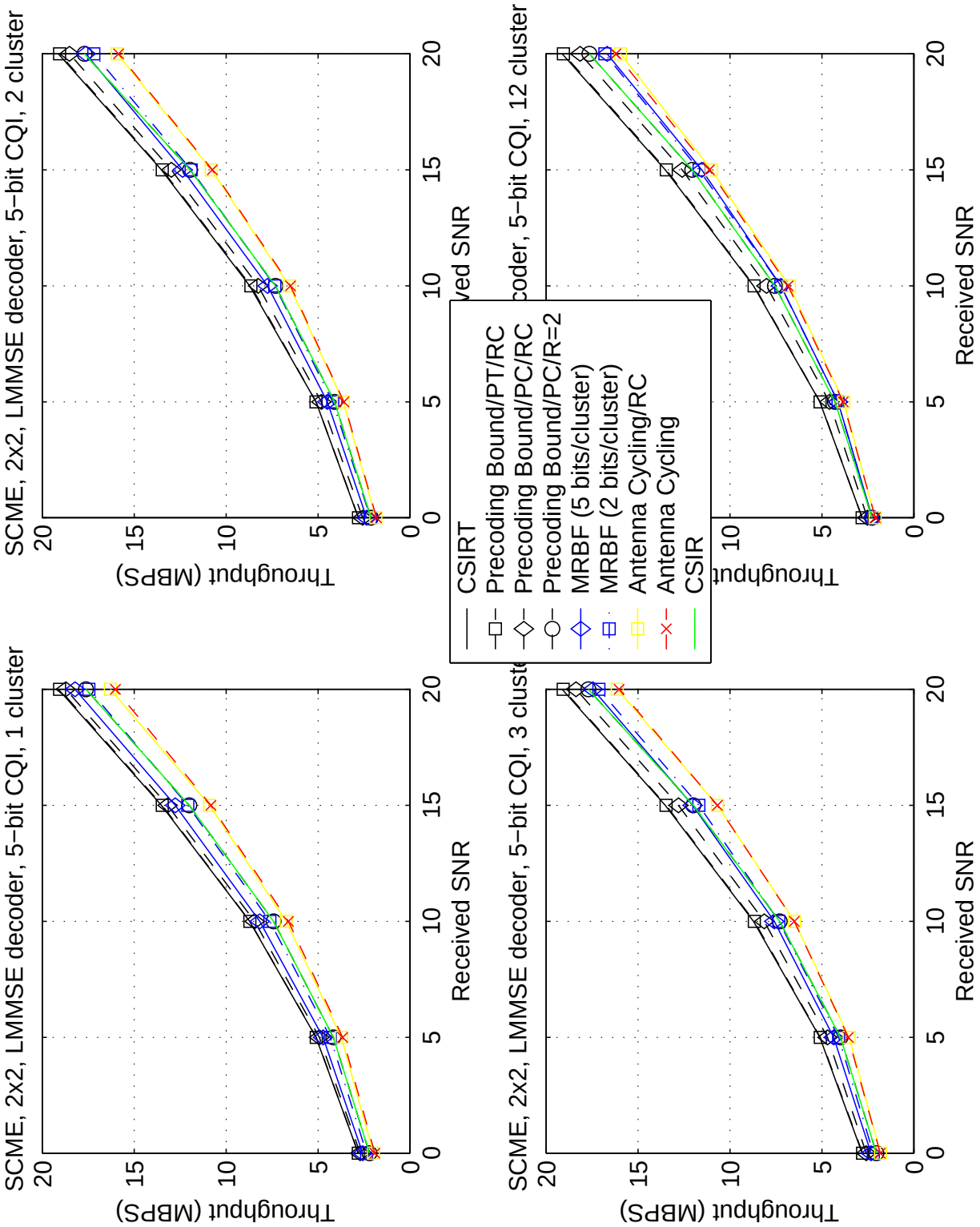

Figure 8: Effect of Various cluster sizes and sub-band scheduling for 2×2 system with 2 and 5 bits FBI feedback per sub-band.

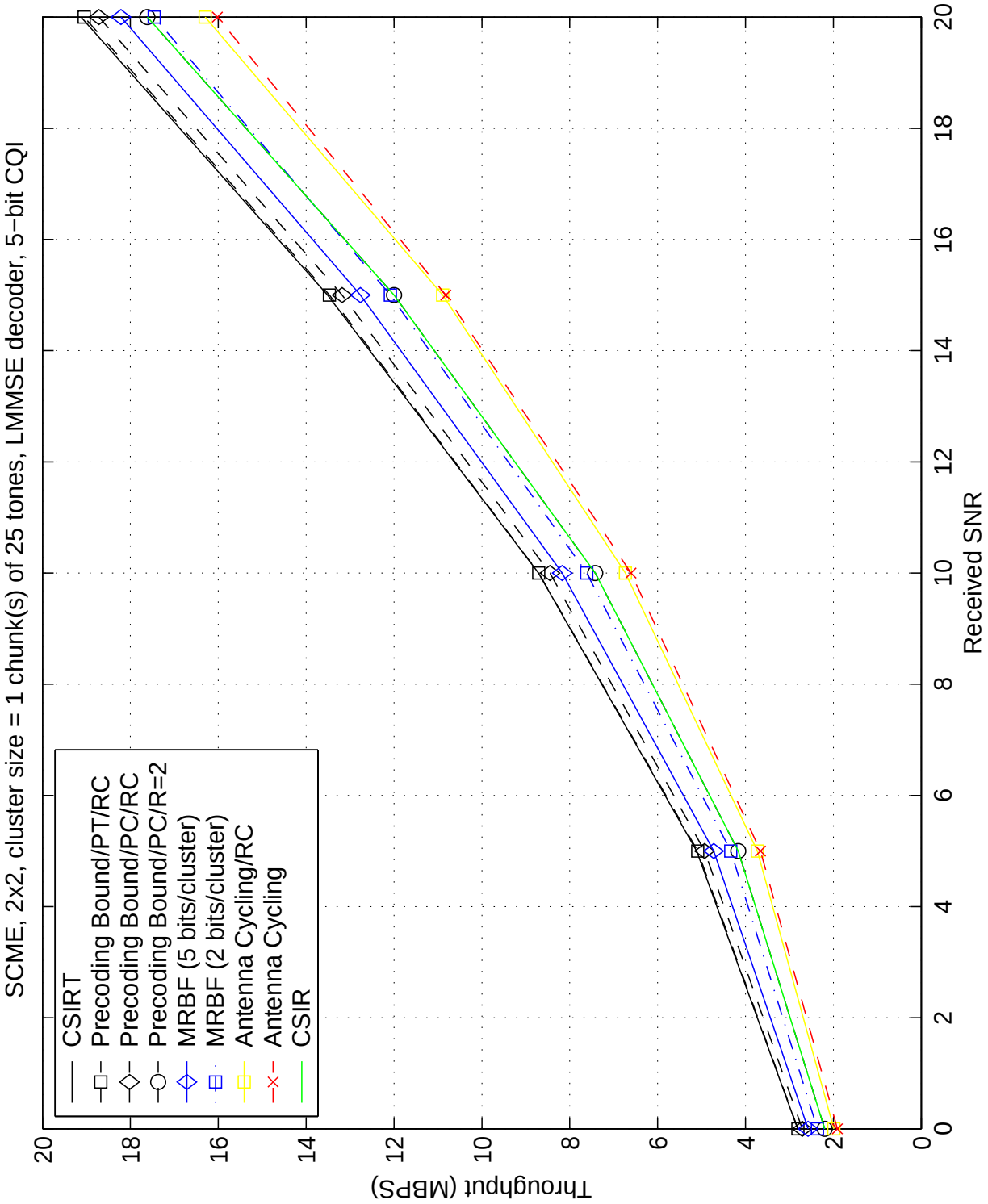

Figure 9: High rate feedback scenario for 2×2 system.

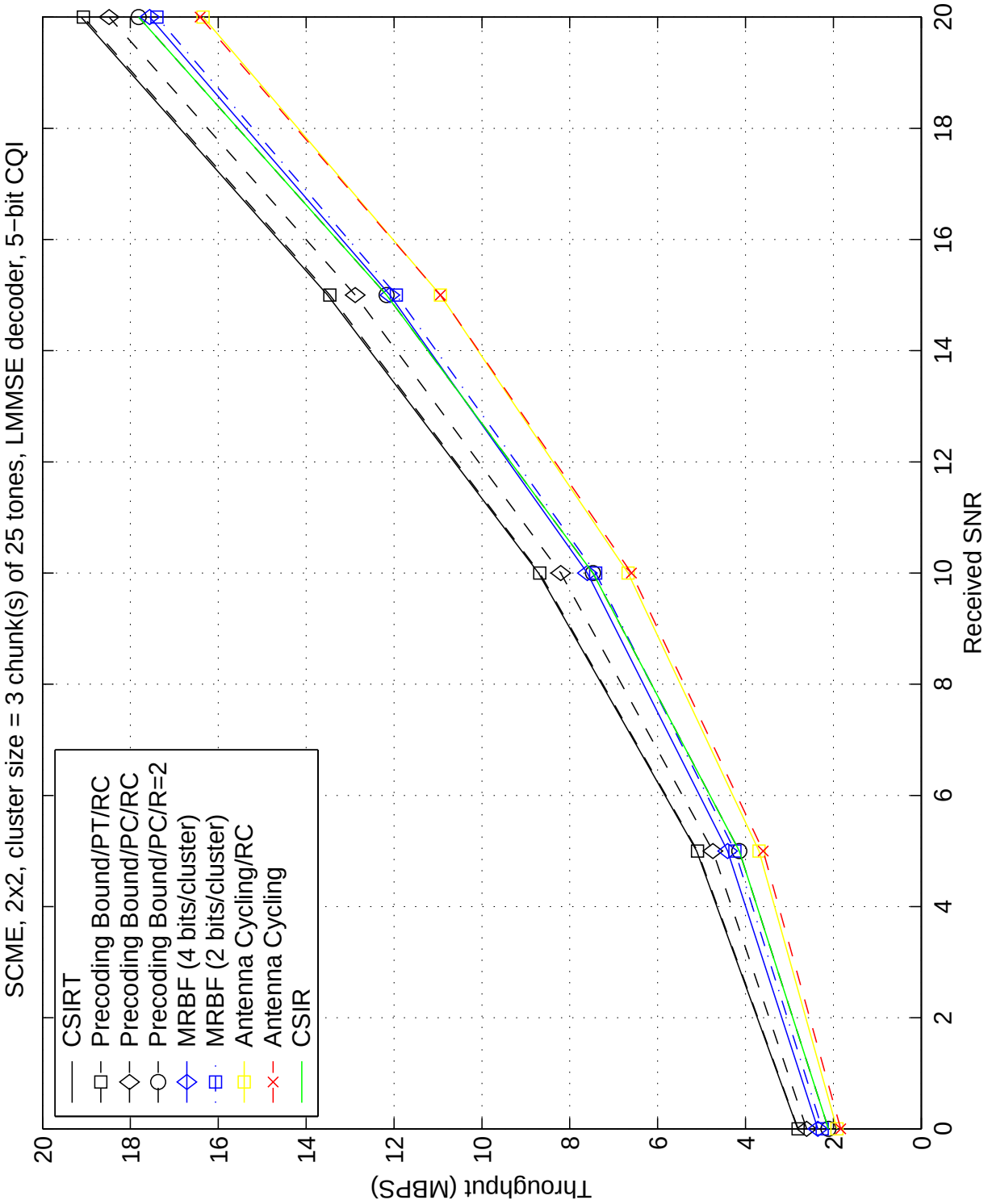

Figure 10: Low rate feedback scenario for 2×2 system.

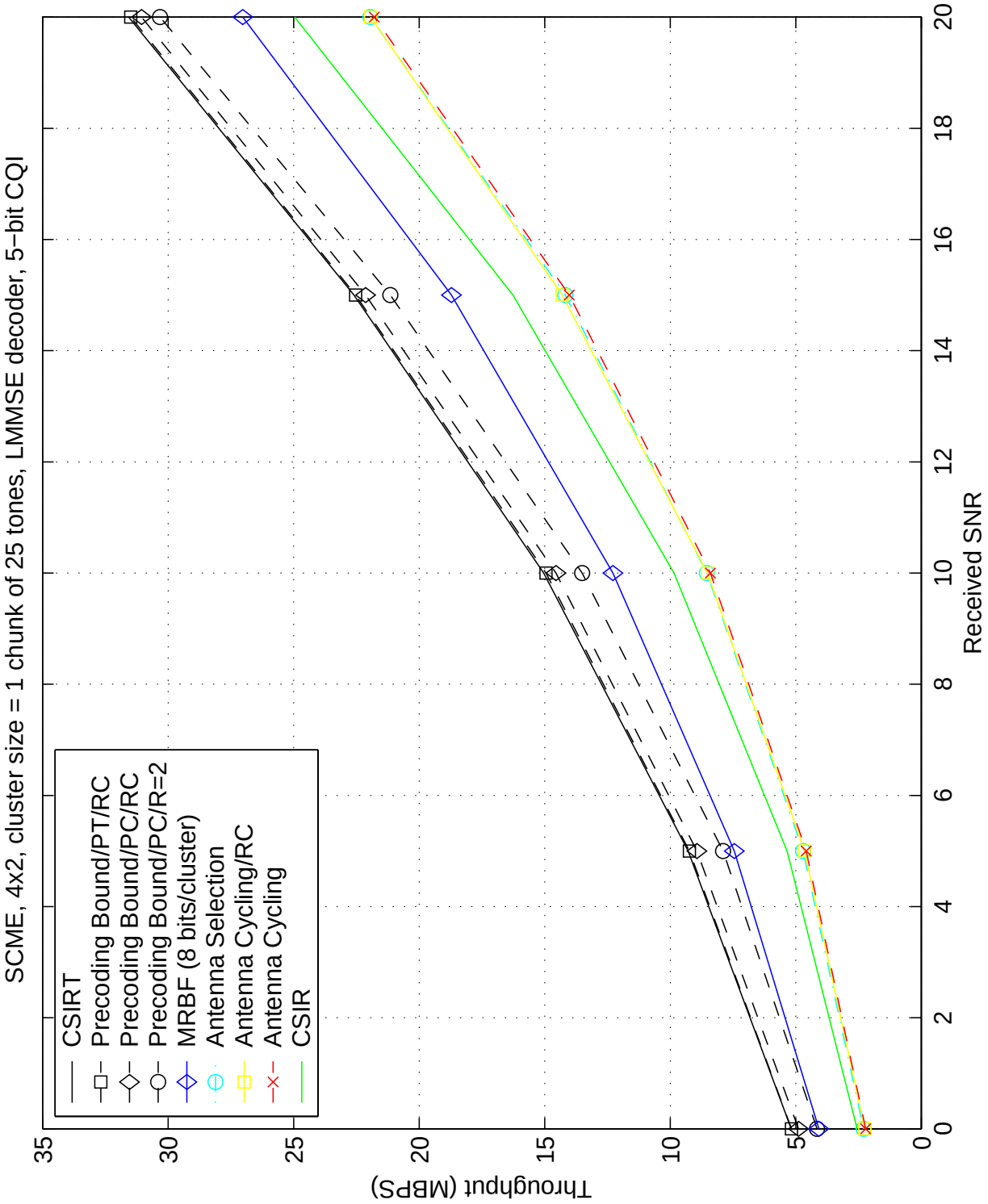

Figure 11: High rate feedback scenario for 4×2 system.

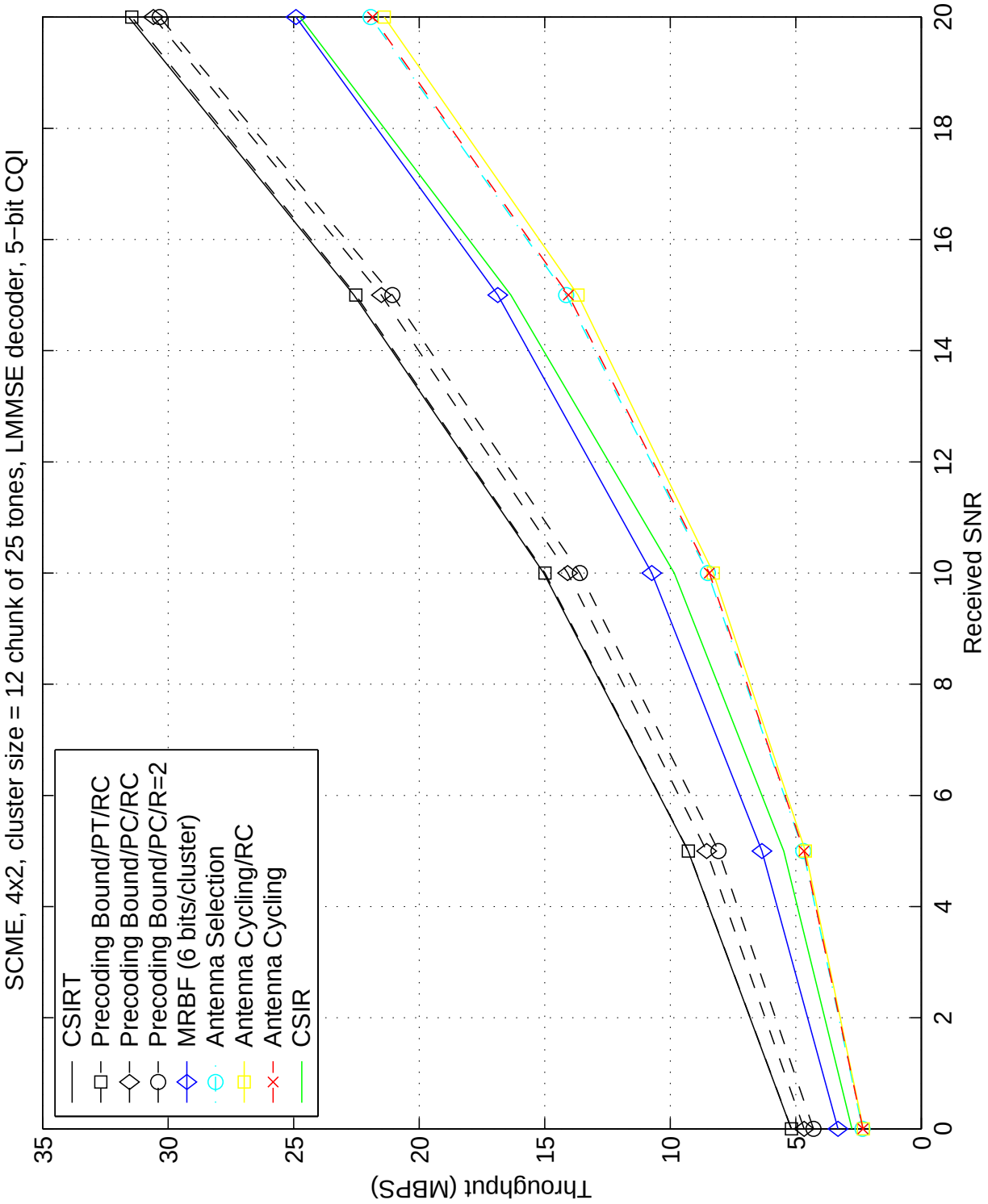

Figure 12: Low rate feedback scenario for 4×2 system.