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Source: Philips, NXP Semiconductors
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1. Introduction

In this contribution we consider some aspects of the calculation of the PMI (precoding matrix indicator) and CQI (channel quality indicator) for multi-user MIMO. As MU-MIMO supports only rank one transmission to a user, hereafter we rename PMI as precoding vector indicator (PVI). We focus on two different assumptions that the UE can make on the nature of the interfering precoding vectors, when calculating the feedback information, and show that the two assumptions lead to the same set of calculations and the values of the feedback, under these assumptions, are equivalent. We therefore propose to include this unified formulation in the specifications describing the nature of the PVI and CQI for MIMO feedback.

2. System model and feedback calculation

Let us consider a codebook for feedback from the terminals, \mathbf{C} , consisting of L unitary matrices of size $M \times M$, M being the number of transmit antennae, $\mathbf{C} = \{\mathbf{C}^{(0)}, \dots, \mathbf{C}^{(L-1)}\}$, such that the overall codebook size in number of vectors is $N_q = LM$. A time sample of a MU-MIMO channel with linear precoding applied at the transmitter is given by

$$\mathbf{y} = \mathbf{H}\mathbf{G}\mathbf{u} + \mathbf{n}, \quad (1)$$

where \mathbf{u} is the vector of independent data symbols transmitted in parallel by the M transmit antennas, such that $E[\mathbf{u}\mathbf{u}^H] = \mathbf{I}$, $\mathbf{y} = (y_0, \dots, y_{K-1})^T$ is the vector of signals individually received by the K users and $\mathbf{n} \sim \mathcal{CN}(\mathbf{0}, \mathbf{I})$ is an i.i.d. proper Gaussian noise vector¹. The matrix $\mathbf{H} = [\mathbf{h}_0^T, \dots, \mathbf{h}_{K-1}^T]^T$ contains the channel coefficients from the M antennas to the K users, where the *row vector* \mathbf{h}_i is the channel of user i . The input is constrained such that

$$\text{trace}(\mathbf{G}^H \mathbf{G}) \leq P. \quad (2)$$

where P denotes the total downlink transmitted power (energy per channel use). We rewrite the precoding matrix in the form

$$\mathbf{G} = \mathbf{F}\text{diag}(\mathbf{p})^{1/2}, \quad (3)$$

¹Here the noise variance, which includes the thermal noise and possibly other interference sources modelled as Gaussian, is normalised to one without loss of generality. In fact, the power constraint P introduced next can be regarded as an SNR (signal-to-noise ratio) value, being normalised by this noise variance.

where the matrix \mathbf{F} is designed without power constraint, and the coefficient p_k is a power adjustment factor for user k . Let \mathcal{S} denote the set of users selected for transmission. We also assume equal power allocation across these users.

Therefore, if the precoding matrix is unitary,

$$\mathbf{F}(\mathcal{S}) \in \{\mathbf{C}^{(0)}(\mathcal{S}), \dots, \mathbf{C}^{(L-1)}(\mathcal{S})\} \quad \text{and} \quad p_k = P/|\mathcal{S}|. \quad (4)$$

If \mathbf{F} is the zero-forcing solution, then

$$\mathbf{F}(\mathcal{S}) = \hat{\mathbf{H}}(\mathcal{S})^H \left(\hat{\mathbf{H}}(\mathcal{S}) \hat{\mathbf{H}}(\mathcal{S})^H \right)^{-1} \quad \text{and} \quad p_k = P/(|\mathcal{S}| \cdot \|\mathbf{f}_k\|^2), \quad (5)$$

where \mathbf{f}_k denotes the k -th column of \mathbf{F} and $\hat{\mathbf{H}}(\mathcal{S}) = [\hat{\mathbf{h}}_1^T, \dots, \hat{\mathbf{h}}_{|\mathcal{S}|}^T]^T$, contains the reported channel vectors of the selected users. We also introduce the cross-talk matrix $[\Phi]_{i,j} = |\mathbf{h}_i \mathbf{f}_j|^2$, such that the received SINR (signal-to-interference plus noise ratio) for user k is given by

$$\gamma_k = \frac{p_k [\Phi]_{k,k}}{1 + \sum_{i \neq k} p_i [\Phi]_{k,i}}. \quad (6)$$

Let us consider two different methods of calculating an index from the codebook, which we refer to equivalently as PVI or channel directional information (CDI), and a real-value quantity signifying an SINR estimate, which represents the channel quality indicator (CQI).

In *method 1*, the following two operations are carried out by each terminal:

- 1) The SINR is computed using (6) for each vector in the codebook taken as own precoding vector. The interfering precoding vectors are assumed to be the other $M - 1$ vectors in the codebook orthogonal to the useful one.
- 2) The largest such SINR is taken as CQI value while the index of the corresponding useful precoding vector represents the PVI.

In *method 2*, the following operations are performed instead

- 1) The measured channel vector² is quantised to the nearest vector in the codebook in terms of chordal distance. The quantisation index represents the CDI.
- 2) The expected SINR is estimated under the following assumptions:
 - a) There are $M - 1$ interfering precoding vectors and they are isotropically distributed on the hyperplane orthogonal to the CDI vector (zero-forcing beamforming assumption).
 - b) The useful precoding vector is in general *slightly offset* from the CDI vector, i.e. the angle between the two vectors is small.

We now show that the operations in method 1 boil down to performing vector quantisation on the channel vector with minimum chordal distance and that the PVI/CQI values for the two methods are totally equivalent.

3. Method 1 performs channel vector quantisation

Let us rewrite the SINR for a generic user k , when the selected precoding matrix is $\mathbf{C}^{(i)} = [\mathbf{c}_0^{(i)}, \dots, \mathbf{c}_{M-1}^{(i)}]$, and the useful precoding vector is $\mathbf{c}_j^{(i)}$. We denote this SINR as $\gamma_k(\mathbf{c}_j^{(i)})$ and define the angle $\theta_{k,j}^{(i)} \in [0, \pi/2]$ between vectors $\tilde{\mathbf{h}}_k^H$ and $\mathbf{c}_j^{(i)}$, where $\tilde{\mathbf{h}}_k = \mathbf{h}_k / \|\mathbf{h}_k\|$, as

$$\cos \theta_{k,j}^{(i)} = | \langle \tilde{\mathbf{h}}_k^H, \mathbf{c}_j^{(i)} \rangle | = | \mathbf{c}_j^{(i)H} \tilde{\mathbf{h}}_k^H |, \quad (7)$$

²For the moment we assume that the terminal has one receiving antenna.

where the angular brackets denote the inner product. As $\mathbf{C}^{(i)}$ is a basis of \mathbb{C}^M , we can decompose $\tilde{\mathbf{h}}_k^H$ as

$$\tilde{\mathbf{h}}_k^H = \alpha_j \mathbf{c}_j^{(i)} + \sum_{l \neq j} \alpha_l \mathbf{c}_l^{(i)}, \quad (8)$$

where $\alpha_l = \langle \tilde{\mathbf{h}}_k^H, \mathbf{c}_l^{(i)} \rangle$ for $l = 0, \dots, M-1$. It follows that $\|\tilde{\mathbf{h}}_k\|^2 = \sum_{l=0}^{M-1} \alpha_l^2 = 1$. Therefore, if the selected precoding matrix is given by $\mathbf{F} = \mathbf{C}^{(i)}$, the power is equally distributed across the M precoding vectors, and user k has reported vector $\mathbf{c}_j^{(i)}$ as PVI, by using (8) into (6) we obtain

$$\gamma_k(\mathbf{c}_j^{(i)}) = \frac{\frac{P}{M} \|\mathbf{h}_k\|^2 \cos^2 \theta_{k,j}}{1 + \frac{P}{M} \|\mathbf{h}_k\|^2 \sin^2 \theta_{k,j}}. \quad (9)$$

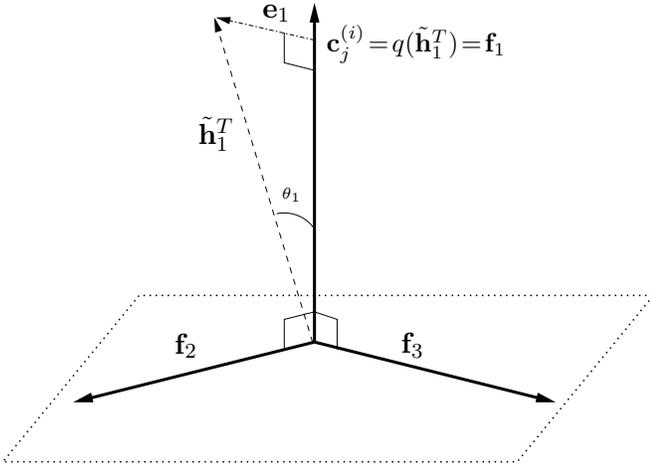


Fig. 1. Representation of precoding vectors, channel vector and PVI vector for method 1. All vectors are defined in \mathbb{R}^3 . $\|\tilde{\mathbf{h}}_1\| = \|\tilde{\mathbf{h}}_1\| = 1$.

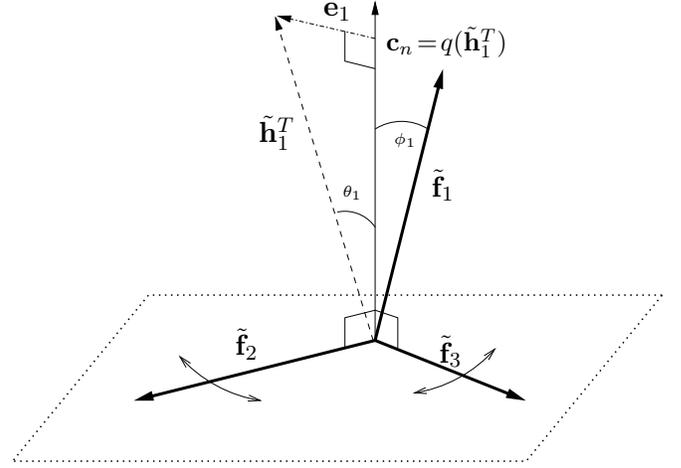


Fig. 2. Representation of precoding vectors, channel vector and CDI vector for method 2. All vectors are defined in \mathbb{R}^3 . $\tilde{\mathbf{f}}_k = \mathbf{f}_k / \|\mathbf{f}_k\|$. $\|\tilde{\mathbf{h}}_1\| = \|\tilde{\mathbf{h}}_1\| = 1$.

We note that the reported PVI can be interpreted as an approximate representation of the conjugate channel vector, $q(\tilde{\mathbf{h}}_k^H)$, where $q(\cdot)$ denotes a vector-quantisation operation. More precisely, the reported PVI corresponds to the vector in the codebook with minimum chordal distance from $\tilde{\mathbf{h}}_k^H$, i.e.

$$q(\tilde{\mathbf{h}}_k^H) = \arg \min_{\mathbf{c}_j^{(i)} \in \mathbf{C}} \sin \theta_{k,j}^{(i)} = \arg \max_{\mathbf{c}_j^{(i)} \in \mathbf{C}} \gamma_k(\mathbf{c}_j^{(i)}). \quad (10)$$

In Fig. 1 we denote with θ_k this minimum quantisation angle for the channel vector of user k . Fig. 1 shows a graphical representation of the channel vector, the selected PVI and the precoding vectors. Note that the precoding vectors $\{\mathbf{f}_1, \mathbf{f}_2, \mathbf{f}_3\}$ are the columns of the selected unitary codebook matrix.

Let us consider method 2. In this case the terminal seeks an accurate representation of the channel vector $\tilde{\mathbf{h}}_k^H$ by one of the vector in the codebook³.

We still use the definition (7) for the angle between vectors $\tilde{\mathbf{h}}_k^H$ and $\mathbf{c}_j^{(i)}$. In this case we see the feedback codebook as a collection of N_q vectors, therefore we introduce the following change of variables for ease of notation: $n = Lj + i$, with $j = 0, \dots, M-1$, $i = 0, \dots, L-1$ and $n = 0, \dots, N_q - 1$. The vector quantisation operation yields the vector in the codebook with minimum chordal distance from $\tilde{\mathbf{h}}_k^H$, which is given by (10), and can be rewritten as

$$q(\tilde{\mathbf{h}}_k^H) = \arg \min_{\mathbf{c}_n \in \mathbf{C}} \sin \theta_{k,n} = \arg \max_{\mathbf{c}_n \in \mathbf{C}} \gamma_k(\mathbf{c}_n). \quad (11)$$

³It is clearly equivalent to represent vector $\tilde{\mathbf{h}}_k^T$ by an element of the codebook \mathbf{C}^* , obtained from \mathbf{C} by taking the conjugate of its vector components. In the case of the DFT-based codebook, depending on whether $\tilde{\mathbf{h}}_k^H$ or $\tilde{\mathbf{h}}_k^T$ is quantised we can use the IDFT and DFT matrix (or the other way round), respectively, as base matrices for the generation of the two codebook, as $\mathbf{C}_{\text{DFT}} = \mathbf{C}_{\text{IDFT}}^*$. The resulting quantisation vectors are one conjugate to the the other.

For the sake of completeness, we recall hereafter the calculation of the SINR in case of channel vector quantisation and zero-forcing precoding, which can be found in [1], and see that it is exactly the same as (9).

When using equal power distribution across the selected users, the SINR (6) can be rewritten as

$$\gamma_k = \frac{\frac{P}{M} \|\mathbf{h}_k\|^2 |\tilde{\mathbf{h}}_k \tilde{\mathbf{f}}_k|^2}{1 + \frac{P}{M} \|\mathbf{h}_k\|^2 \sum_{i \neq k} |\tilde{\mathbf{h}}_k \tilde{\mathbf{f}}_i|^2}. \quad (12)$$

Let us rename the quantisation vector $q(\tilde{\mathbf{h}}_k^H) = \hat{\mathbf{h}}_k^H$ for short, and define the angle, $\phi_k \in [0, \pi/2]$, between $\hat{\mathbf{h}}_k^H$ and the useful precoding vector $\tilde{\mathbf{f}}_k = \mathbf{f}_k / \|\mathbf{f}_k\|$ as

$$\cos \phi_k = | \langle \hat{\mathbf{h}}_k^H, \tilde{\mathbf{f}}_k \rangle | = |\tilde{\mathbf{f}}_k^H \hat{\mathbf{h}}_k^H| = |\hat{\mathbf{h}}_k \tilde{\mathbf{f}}_k|. \quad (13)$$

It is useful to note that, if the precoder is zero-forcing, by construction \mathbf{f}_i is orthogonal to $\hat{\mathbf{h}}_k^H$ for $i \in \mathcal{S} \setminus \{k\}$, i.e. $\hat{\mathbf{h}}_k \mathbf{f}_i = 0$, $i \neq k$, and $|\hat{\mathbf{h}}_k \mathbf{f}_k| = 1$ for every k . Therefore, it follows that

$$\cos \phi_k = |\hat{\mathbf{h}}_k \tilde{\mathbf{f}}_k| = \frac{1}{\|\mathbf{f}_k\|}. \quad (14)$$

We can decompose the channel vector $\tilde{\mathbf{h}}_k^H$ as in (8), by denoting with \mathbf{e}_k the vector component normal to $\hat{\mathbf{h}}_k^H$,

$$\tilde{\mathbf{h}}_k^H = (\hat{\mathbf{h}}_k \tilde{\mathbf{h}}_k^H) \hat{\mathbf{h}}_k^H + \mathbf{e}_k. \quad (15)$$

Clearly \mathbf{e}_k is the quantisation error vector for $\tilde{\mathbf{h}}_k^H$ and $\|\mathbf{e}_k\| = \sin \theta_k$. If ϕ_k is small enough we can approximate $\mathbf{e}_k^H \tilde{\mathbf{f}}_k \approx 0$ and by replacing (15) in (12) and using (14), we obtain

$$\begin{aligned} \gamma_k &\simeq \frac{\frac{P}{M} \|\mathbf{h}_k\|^2 \cos^2 \theta_k |\hat{\mathbf{h}}_k \tilde{\mathbf{f}}_k|^2}{1 + \frac{P}{M} \|\mathbf{h}_k\|^2 \sin^2 \theta_k \sum_{i \neq k} |\tilde{\mathbf{e}}_k^H \tilde{\mathbf{f}}_i|^2} \\ &= \frac{p_k \|\mathbf{h}_k\|^2 \cos^2 \theta_k}{1 + \frac{P}{M} \|\mathbf{h}_k\|^2 \sin^2 \theta_k \sum_{i \neq k} |\tilde{\mathbf{e}}_k^H \tilde{\mathbf{f}}_i|^2} \end{aligned} \quad (16)$$

where $\tilde{\mathbf{e}}_k = \mathbf{e}_k / \|\mathbf{e}_k\|$.

By looking at $\tilde{\mathbf{f}}_i$, $i \neq k$ and $\tilde{\mathbf{e}}_k$ in (16) as random vectors, which are independent and isotropically distributed on the plane represented in Fig. 2, their square inner product $|\tilde{\mathbf{e}}_k^H \tilde{\mathbf{f}}_i|^2$ is a Beta-distributed random variable, $\beta(1, M-2)$, with mean $1/(M-1)$. Therefore, by taking the expectation of (16) w.r.t. the interference terms and using the Jensen's inequality we get the approximate lower-bound

$$E[\gamma_k] \gtrsim \frac{p_k \|\mathbf{h}_k\|^2 \cos^2 \theta_k}{1 + \frac{P}{M} \|\mathbf{h}_k\|^2 \sin^2 \theta_k}. \quad (17)$$

As p_k is known to the Node B but unknown to the terminal, the CQI sent by the terminal k , according to method 2 reads the same as (9), i.e.

$$\text{CQI}_k = \frac{\frac{P}{M} \|\mathbf{h}_k\|^2 \cos^2 \theta_k}{1 + \frac{P}{M} \|\mathbf{h}_k\|^2 \sin^2 \theta_k}, \quad (18)$$

as derived in [1], which is identical to (9). The Node B can then estimate the SINR at user k by simple scaling operation: $\gamma_k = \frac{p_k}{P/M} \text{CQI}_k$. Note that the ratio $\frac{p_k}{P/M} = 1$ for every k if the precoding matrix is unitary, $\frac{p_k}{P/M} \neq 1$ otherwise.

4. Case with multiple receive antennae.

As was discussed e.g. in [2] there are various receive beamformers that can be applied at the terminal to combine the multiple receive antennae into one equivalent antenna. For a given receive beamformer, however, the operations for

calculating the PVI and CQI are the same as outlined in the previous section. We note that the combiner depends, in general, on the codebook vectors being considered. We can identify at least three possible receive beamformer structures:

- 1) *MMSE receiver*. For a given precoding matrix $\mathbf{F} = \mathbf{C}^{(i)}$ and a selected useful precoding vector $\mathbf{c}_j^{(i)}$, the combiner for user k is given by⁴

$$\mathbf{w}_{k,j}^{(i)} = \bar{\mathbf{h}}_{k,j}^{(i)H} \left(\bar{\mathbf{H}}_k^{(i)} \bar{\mathbf{H}}_k^{(i)H} + \mathbf{I}_N \right)^{-1}, \quad (19)$$

where $\bar{\mathbf{H}}_k^{(i)} = \mathbf{H}_k \mathbf{C}^{(i)} = [\bar{\mathbf{h}}_{k,0}^{(i)}, \dots, \bar{\mathbf{h}}_{k,M-1}^{(i)}]$, and \mathbf{H}_k is the $N \times M$ sample channel matrix for user k .

- 2) *Minimum quantisation error receiver*. For a given vector of the codebook, \mathbf{c}_n , the coefficients of the linear combination at the receiver k are given by

$$\mathbf{w}_{k,n} = \mathbf{h}_{k,n}^{(\text{eff})} \mathbf{H}_k^\dagger, \quad (20)$$

where $\mathbf{H}_k^\dagger = \mathbf{H}_k^H (\mathbf{H}_k \mathbf{H}_k^H)^{-1}$ is the right pseudo-inverse of \mathbf{H}_k and $\mathbf{h}_{k,n}^{(\text{eff})}$ is the normalised projection of \mathbf{c}_n^H onto $\text{span}(\mathbf{H}_k)$, i.e. let $\mathbf{Q}_k = [\mathbf{q}_{k,1}^T, \dots, \mathbf{q}_{k,N}^T]^T$ be a generator matrix for $\text{span}(\mathbf{H}_k)$,

$$\mathbf{h}_{k,n}^{(\text{eff})} = \alpha \mathbf{c}_n^H \mathbf{Q}_k^H \mathbf{Q}_k, \quad (21)$$

and α is a normalisation factor, such that, e.g. $\|\mathbf{w}_{k,n}\| = 1$.

- 3) *SVD (singular value decomposition) of the channel*. The receive beamformer in this case depends only on the channel matrix \mathbf{H}_k and not on the feedback codebook. Let $\mathbf{H}_k = \mathbf{U}_k \mathbf{\Lambda}_k \mathbf{T}_k^H$ be the SVD of the channel, with singular values $\lambda_{k,1} \geq \lambda_{k,2} \geq \dots \geq \lambda_{k,N}$ ($N \leq M$), and $\mathbf{U}_k = [\mathbf{u}_{k,1}, \dots, \mathbf{u}_{k,N}]$, $\mathbf{T}_k = [\mathbf{t}_{k,1}, \dots, \mathbf{t}_{k,N}]$, then

$$\mathbf{w}_k = \mathbf{u}_{k,1}^H. \quad (22)$$

No matter what receive beamformer the terminal decides to use, once set the effective one-antenna equivalent channel for a given codebook element n , $\mathbf{h}_{k,n}$ ($n = Lj + i$, as introduced in Sec. 3), where

$$\mathbf{h}_{k,n} \triangleq \begin{cases} \mathbf{w}_{k,j}^{(i)} \mathbf{H}_k & \text{or} \\ \mathbf{w}_{k,n} \mathbf{H}_k & \text{or} \\ \mathbf{w}_k \mathbf{H}_k \end{cases}, \quad (23)$$

the PVI is always given by the equivalent expressions (10) or (11) and the CQI by the equivalent expressions (9) or (18).

5. Numerical results.

We present some numerical results that illustrate the equivalence between method 1 and method 2 of Section 2, when reporting one value of PVI and CQI. We consider two different precoding matrix constructions, unitary precoding (UP for short) where the matrix is taken from the feedback codebook, and zero-forcing (ZF). For the case of multiple receiving antennas we show results with different receive beamformers, namely the MMSE receiver and the minimum-quantisation-error (MQE for short) receiver, presented in the previous section. The main simulation parameters are listed in Table I.

It can be seen that, in the case of 4 tx antennae and 2 rx antennae (Fig. 4), the best performance is obtained by using a receive beamformer derived to minimise the quantisation error (MQE).

⁴Note the the noise variance is normalised to one as it was included in the transmit SNR P .

Number of tx antennae	4
Number of rx antennae	1,2,4
Tx antenna spacing	0.5λ
Type of tx precoding	Unitary (UP), zero-forcing (ZF)
Type of rx beamforming	MMSE, minimum quantisation error (MQE)
Number of users	20
Terminal speed	3 km/h
Transmission bandwidth	5 MHz
Centre frequency	2 GHz
DFT size	512
Feedback codebook type	DFT
Feedback codebook size	4 bits
Channel model	SCM Urban Micro
Number of paths	10
Subframe time	1 ms
Number of subcarriers per RB	12
Feedback granularity	1 per RB
Delay between feedback and data detection	1.5 ms (3 time slots)
Modulation schemes	QPSK, 16QAM, 64QAM
Turbo coding rates	1/3,1/2,2/3,3/4,4/5
Target FER	10%

TABLE I
LINK-LEVEL SIMULATION PARAMETERS.

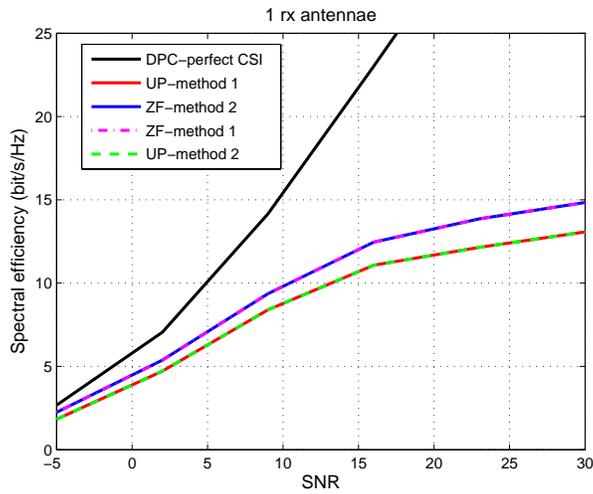


Fig. 3. Unitary precoding and ZF precoding with the two feedback calculation methods described in Sec. 2. The two feedback schemes, as derived in Sec. 3 are equivalent. 4 tx antennae, 1 rx antenna, 20 users.

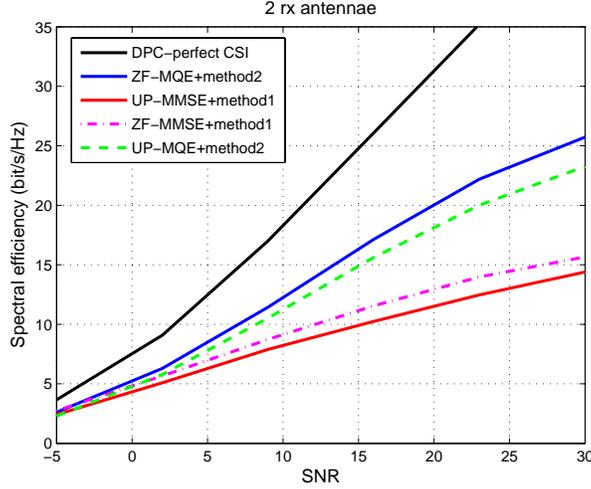


Fig. 4. Unitary precoding and ZF precoding with two different receive beamformers, MMSE and minimum-quantisation-error, as described in Sec. 4 feedback. The calculation of PVI and CQI from the equivalent 1-antenna channel is the same for both types of receivers. 4 tx antennae, 2 rx antenna, 20 users.

6. Conclusion.

In this contribution we have shown the equivalence of the PVI and CQI being fed back by the terminals for either zero-forcing precoding or unitary precoding, when one receiving antenna is present. In case of multiple receiving antennae a combination of these antennae (receive beamformer) is needed at the terminal. The resulting one virtual antenna channel at the output of the receive beamformer can be used for PVI and CQI calculation as for the case of one physical receiving antenna. Therefore, we suggest the following unifying description for the meaning of PVI and CQI in MU-MIMO.

Let $\mathcal{C} = \{c_1, \dots, c_{N_q}\}$ be the codebook available for feedback. When computing the feedback values of PVI and CQI, the UE k derives a linear combination of the receiving antennae, which produces a one virtual antenna channel. Let this channel be represented by a column vector $\mathbf{h}_{k,n}$, which may depend on the codebook entry c_n being considered. Let $\|\mathbf{h}_{k,n}\|$ be the amplitude of such channel and $\tilde{\mathbf{h}}_{k,n} = \mathbf{h}_{k,n}/\|\mathbf{h}_{k,n}\|$ its unit-norm version. The PVI value is the index of the codebook vector with minimum distance from $\tilde{\mathbf{h}}_{k,n}$, where the distance measure is defined as followed

$$d_{k,n}^2 = 1 - |\mathbf{c}_n^H \tilde{\mathbf{h}}_{k,n}|^2.$$

The PVI index for user k , i_k , then reads

$$i_k = \arg \min_{n=1, \dots, N_q} d_{k,n}.$$

The CQI value is given by

$$\text{CQI}_k = \frac{\frac{P}{M} \|\mathbf{h}_{k,i_k}\|^2 (1 - d_{k,i_k}^2)}{1 + \frac{P}{M} \|\mathbf{h}_{k,i_k}\|^2 d_{k,i_k}^2},$$

where P is the transmit SNR (transmit power divided by the noise variance) and M is the number of transmit antennae.

Furthermore, we propose that the following text is adopted as **working assumption** for MU-MIMO operations:

- 1) **The feedback value of PVI (precoding vector indicator) reported by the UE indicates the index of the codebook vector with the smallest chordal distance from the estimated downlink channel vector.**
- 2) **If the UE has more than one receive antenna, the estimated downlink channel vector is a combination of the downlink channel vectors at the individual receiving antennae.**

REFERENCES

- [1] Philips, "Comparison between MU-MIMO codebook-based channel reporting techniques for LTE downlink." R1-062483, Oct 2006. 3GPP TGS RAN WG1 Meeting #46bis, Seoul.
- [2] Philips, "Comparison of MU-MIMO feedback schemes with multiple UE receive antennas." R1-070346, Jan 2007. 3GPP TGS RAN WG1 Meeting #47bis, Sorrento.