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Selective-MS Precoding for Downlink MIMO Transmissions

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1 Introduction

Downlink MIMO techniques for 802.16m systems must capable of providing spatial multiplexing (SM) gain and spatial diversity (SD) gain in the following 802.16m radio environments:

- Macrocell (1-10 km cell radius)
    - Rural Macrocell LOS
    - Urban Macrocell NLOS
    - Suburban Macrocell NLOS
    - Bad Urban Macrocell NLOS
- Microcell (200 m to 1 km 0.3 – 0.5 km cell radius)
    - Urban Microcell LOS/NLOS
    - Bad Urban Microcell
- Picocell (10 to 200 m cell radius)
    - Indoor Hot Spot
    - Indoor Small Office
    - Outdoor to Indoor

The particular downlink MIMO technique that may used depends mainly on BS/MS antenna configurations, MIMO channel conditions within a radio environment, and the availability and accuracy of channel state information (CSI).

In a macrocell BS antennas are typically above rooftops in order to support a large macrocell radius. Transmitted BS downlink signals may not be significantly scattered by channel obstacles since they are mainly below the transmit antennas. At the MS this may result in signals with small angle-or-arrival spreads and low-rank MIMO channel matrices due to high antenna cross correlations (cross-polarized antennas may be used to mitigate this problem). The low-rank channel matrices limit the possibilities for spatial multiplexing and spatial diversity. However, the angle-or-arrival spreads at an MS may allow the BS's transmitted signal energy to be focused to areas where one or more MSs are located (focused signal energy will also mitigate range problems for large cells).

In a macrocell MS mobility may be very high so accurate short-term channel state information (CSI) may not be available at a BS transmitter. Hence an open-loop MIMO technique may be more appropriate. Due to small angular spreads and the possibility of high MS mobility a fixed beamforming approach may also be appropriate for the downlink.

In a microcell BS antennas are typically below rooftops so transmitted BS signals may be significantly scattered and faded by channel obstacles. At the MS this may result in large angle-of-arrival spreads and high or full rank MIMO channel matrices due to low antenna cross correlations. The full-rank channel matrices extend the possibilities for spatial multiplexing and spatial diversity. However, the large angular spreads at MSs make it more difficult to focus a BS's transmitted signal energy to specific areas where one or more MSs are located.

In a microcell MSs may be stationary, slowly moving, or within cars or public transportation vehicles. For stationary or slowly moving MSs accurate short-term CSI may be acquired by a BS in a TDD mode of operation. For stationary or slowly moving MSs a MIMO technique for SM gain or SD gain may be combined with linear precoding operation. Linear precoding may exploit the CSI available at the BS and be used to increase spectral efficiency and also help mitigate multi-user interference at the MSs.

For MSs in vehicles mobility may be high so accurate short-term CSI may not be available at a BS. Hence an open-loop MIMO technique may be more appropriate for MSs within vehicles. Also, due to large angular spreads and the possibility of high MS mobility, a fixed beamforming approach with a set of fixed spatial sub-channels is not appropriate for the downlink.

In a picocell BS antennas are typically low so transmitted BS signals will be significantly scattered and faded by channel obstacles. At the MS this typically results in large angle-of-arrival spreads and high or full rank MIMO channel matrices due to low antenna cross correlations. Similar to the microcell scenario, the full-rank channel matrices extend the possibilities for spatial multiplexing and spatial diversity. The large angular spreads make it more difficult to focus a BS's transmitted signal energy to specific areas where one or more MSs are located.

In a picocell the coverage area is rather small and MSs are stationary or slowly moving. Hence accurate short-term CSI may be acquired by a BS in a TDD mode of operation. The acquired CSI is also valid over larger time intervals. A MIMO technique for SM gain or SD gain may be combined with linear precoding operation to improve performance.

Indeed, the numerous radio environments for 802.16m systems motivates the usage of an adaptive MIMO system. BSs and MSs should be able to adoptively switch between DL MIMO techniques. By switching between DL MIMO techniques an IEEE 802.16m system can dynamically optimize throughput or coverage for a specific radio environment.

To increase spectral efficiency and system throughput this contribution describes a downlink precoding technique that may be used in a microcell or picocell and when MS mobility is low so accurate short-term CSI may be acquired by a BS. The proposed precoding technique is linear and may be combined with any other MIMO technique for SM gain or SD gain.

2 Open-loop Selective-MS Precoding

Space-time coding may be combined with linear precoding in order to increase spectral efficiency and throughput. Figure 1 shows a downlink system model that combines space-time coding and linear precoding. We now describe the signals and the precoding operation shown within Figure 1.

Let N_T denote the number of BS transmit antennas. Let N_R the total number of MS receive antennas and $N_{R,k} \geq 1$ denote the number of receive antennas for the kth MS. Let K_{active} denote the number of active MSs serviced by a BS. Let $N_{S,k} \leq \min(N_T, N_{R,k})$ denote the number of independent spatial streams allocated to the kth active MS. The total number of spatial streams transmitted by the BS is the sum

$$N_S = \sum_{k=1}^{K_{active}} N_{S,k} \tag{1}$$

The composite signal transmitted by the BS is defined as the N_T-by-1 vector

$$\mathbf{x} = \sum_{k=1}^{K_{active}} \mathbf{W}_k \mathbf{P}_k \mathbf{s}_k \tag{2}$$

where $\mathbf{W}_k$ is the N_T-by-$N_{S,k}$ linear precoding matrix, $\mathbf{P}_k$ an $N_{S,k}$-by-$N_{S,k}$ diagonal power loading matrix, and $\mathbf{s}_k$ an $N_{S,k}$-by-1 data symbol vector.

The total number of receive antennas distributed over all K_{active} MSs is

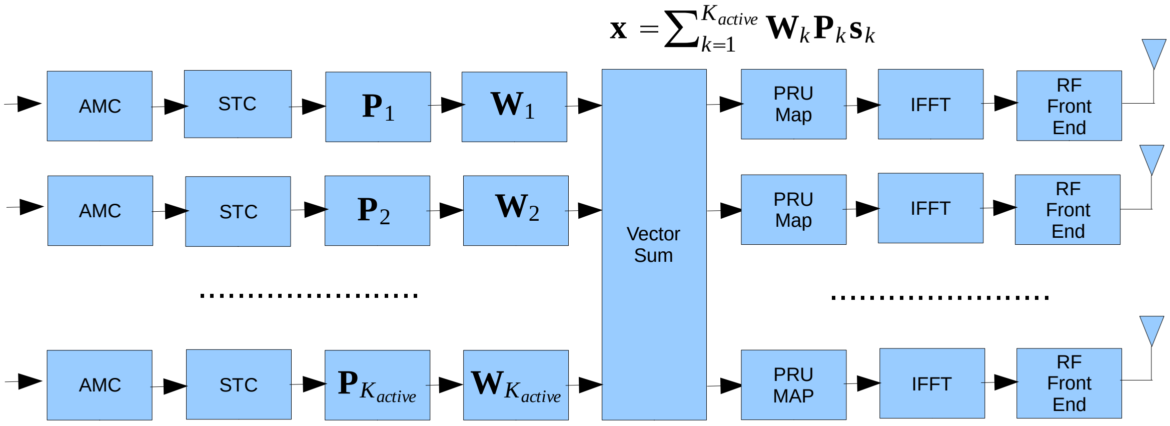

Figure 1: Conceptual block diagram of the precoding operation

$$N_R = \sum_{k=1}^{K_{active}} N_{R,k} \tag{3}$$

The received signal for the kth MS is the $N_{R,k}$-by-1 vector

$$\begin{aligned} \mathbf{y}_k &= \mathbf{H}_k \mathbf{x} + \mathbf{n}_k \\ &= \mathbf{H}_k \mathbf{W}_k \mathbf{P}_k \mathbf{s}_k + \mathbf{H}_k \sum_{j=1, j\neq k}^{K_{active}} \mathbf{W}_j \mathbf{P}_j \mathbf{s}_j + \mathbf{n}_k \end{aligned} \tag{4}$$

where the $N_{R,k}$-by-N_T matrix $\mathbf{H}_k$ denotes the channel matrix for the kth MS. The (i,j)th element of $\mathbf{H}_k$ represents the channel gain and phase associated with the signal path from transmit antenna i to receive antenna j. Elements of $\mathbf{H}_k$ are modeled as independent complex Gaussian random variables with zero mean and unit variance. The first term on the right hand side is the desired signal for the kth MS multiplied by the precoding and the channel matrices. The $N_{R,k}$-by-1 noise vector for the kth user is $\mathbf{n}_k$. The second term is co-channel interference observed by the kth MS.

The co-channel interference term can be eliminated in theory if

$$\mathbf{H}_k \sum_{j=1, j\neq k}^{K_{active}} \mathbf{W}_j = \mathbf{0} \tag{5}$$

Let $M_R = N_R - N_{R,k}$. Using all of the MS's MIMO channel matrices except the kth MS's MIMO channel matrix, we construct the M_R-by-N_T matrix

$$\tilde{\mathbf{H}}_k = \begin{bmatrix} \mathbf{H}_1^T & \dots & \mathbf{H}_{k-1}^T & \mathbf{H}_{k+1}^T & \dots & \mathbf{H}_{K_{active}}^T \end{bmatrix}^T \tag{6}$$

The singular value decomposition of $\tilde{\mathbf{H}}_k$ is defined as

$$\tilde{\mathbf{H}}_k = \tilde{\mathbf{U}}_k \tilde{\mathbf{\Sigma}}_k \tilde{\mathbf{V}}_k^H \tag{7}$$

where superscript H denotes the Hermitian transpose. Matrices $\tilde{\mathbf{U}}_k$ and $\tilde{\mathbf{V}}_k$ are M_R-by-M_R and N_T-by-N_T unitary matrices. Matrix $\tilde{\mathbf{\Sigma}}_k$ is an M_R-by-N_T singular value matrix.

Matrix $\tilde{\mathbf{V}}_k$ can be partitioned as

$$\tilde{\mathbf{V}}_k = \begin{bmatrix} \mathbf{v}_1 & \mathbf{v}_2 & \dots & \mathbf{v}_{r_{\tilde{\mathbf{H}}_k}} & \mathbf{v}_{r_{\tilde{\mathbf{H}}_k}+1} & \dots & \mathbf{v}_{N_T-1} & \mathbf{v}_{N_T} \end{bmatrix}^H \tag{8}$$

where $r_{\tilde{\mathbf{H}}_k} \leq M_R$ denotes the rank of $\tilde{\mathbf{H}}_k$ and each element is an N_T-by-1 vector. Since $\tilde{\mathbf{V}}_k$ is unitary the vectors are orthonormal. Let $\tilde{\mathbf{V}}_k^0$ be defined as the N_T-by-$(N_T - r_{\tilde{\mathbf{H}}_k})$ matrix

$$\tilde{\mathbf{V}}_k^0 = \begin{bmatrix} \mathbf{v}_{r_{\tilde{\mathbf{H}}_k}+1} & \dots & \mathbf{v}_{N_T-1} & \mathbf{v}_{N_T} \end{bmatrix} \tag{9}$$

The orthonormal vectors within $\tilde{\mathbf{V}}_k^0$ form an orthonormal basis for the null space of $\tilde{\mathbf{H}}_k$ hence $\mathbf{H}_j \tilde{\mathbf{V}}_k^0 = \mathbf{0}$ for all $j \neq k$.

The N_T-by-$N_{S,k}$ precoder matrix $\mathbf{W}_k$ for the kth MS may be constructed using a linear combination of the columns of $\tilde{\mathbf{V}}_k^0$. Hence we define the precoder matrix for the kth MS as

$$\mathbf{W}_k = \tilde{\mathbf{V}}_k^0 \tag{10}$$

The precoder matrix will be unitary and the equality $\mathbf{H}_k \mathbf{W}_j = \mathbf{0}$ will be true for all $j \neq k$. Hence the co-channel interference term can be eliminated and the received signal will be

$$\mathbf{y}_k = \mathbf{H}_k \mathbf{W}_k \mathbf{s}_k + \mathbf{n}_k \tag{11}$$

Note that for precoding we cannot use a common pilot for estimating the effective channel $\mathbf{H}_k \mathbf{W}_k$ since each MS uses a different precoding matrix $\mathbf{W}_k$.

If $\tilde{\mathbf{H}}_k$ is of full rank the N_T-by-$(N_T - r_{\tilde{\mathbf{H}}_k})$ matrix $\tilde{\mathbf{V}}_k^0$ is a null matrix so the number of columns is zero and

$$N_T - r_{\tilde{\mathbf{H}}_k} = N_T - N_R + N_{R,k} = 0 \tag{12}$$

To ensure that $N_T - r_{\tilde{\mathbf{H}}_k} > 0$ and there are at least $N_{S,k}$ columns in each $\tilde{\mathbf{V}}_k^0$ a sufficient condition is that

$$N_T \geq N_R = \sum_{k=1}^{K_{active}} N_{R,k} \tag{13}$$

Hence the number of transmit antennas N_T must be as large as the total number of antennas for all MSs. This constrains the number of active MS antennas to be less number of transmit antennas N_T.

The above precoding technique requires that the channels $\mathbf{H}_k$, $k = 1, 2, \dots, K_{active}$, be available at the BS for all active MSs. This can be accomplished using MS-dedicated uplink pilot symbols that the BS may use to estimate $\mathbf{H}_k$. This open-loop approach exploits channel reciprocity and requires a TDD mode of operation. It also requires RF front-end calibration.

2.1 MS Grouping for Selective-MS Precoding

Selective-MS precoding will decrease the number of computations required for precoding and improve performance by better nulling the co-channel interference. In selective-MS precoding only a subset of the K_{active} active MSs is scheduled to receive DL subframe data. The selected subset is called an MS spatial group. We now describe how an MS spatial group may be formed.

A spatial MS grouping of active MSs is a set partition

$$\mathcal{G} = \{G_1, G_2, \dots, G_{N_G}\} \tag{14}$$

where G_i denotes an MS spatial group and N_G the number of groups. Each MS spatial group G_i is a subset of the active MS set

$$\mathcal{M} = \{MS_1, MS_2, \ldots, MS_{K_{active}}\} = \bigcup_{i=1}^{N_G} G_i \tag{15}$$

where MS_i denotes the ith active MS. The total number of MS receive antennas N_R associated with each spatial group G_i must be less than or equal to N_T (i.e. $N_R \leq N_T$). The MS spatial groups are disjoint meaning their set intersections are null.

MSs within a group G_i will have uncorrelated channels $\mathbf{H}_k$. MSs with channel cross correlations $\mathbf{H}_k$ that are below a pre-defined threshold are placed in the same MS group. MSs that have highly correlated channels $\mathbf{H}_k$ are placed into different spatial MS groups. MSs belonging to the same group are assigned different precoding matrices $\mathbf{W}_k$. MSs within a spatial MS group can share the same physical layer resource units within a subframe. The larger an MS group the greater the gain in spectral efficiency and throughput. On the other hand, smaller MS groups allow the BS to transmit with higher average power per MS.

Finding the optimum MS grouping $\mathcal{G}$ requires a comparison between all possible MS groups. This is not practical so reduced complexity algorithms are required to find a sub-optimal MS grouping. Many sub-optimal MS grouping strategies are proposed in the literature.

3 Closed-loop Selective-MS Precoding

The above precoding technique requires that the channels $\mathbf{H}_k$, $k = 1, 2, \ldots, K_{active}$, be available at the BS for all active MSs. This can also be accomplished using a closed-loop approach where each MS-estimated $\mathbf{H}_k$ is provided to the BS via an uplink signal. The closed-loop approach is applicable for both TDD and FDD modes of operation and channel reciprocity is not required.

A closed-loop feedback approach may be categorized as being analog or digital. Analog or unquantized feedback typically refers to the case where non-quantized MS-estimated $\mathbf{H}_k$ are provided to the BS on an uplink signal. Digital or quantized feedback typically refers to the case where the MSs make use of a vector quantization scheme and provide the BS with an encoded and modulated quantization index.

For the digital closed-loop approach it is simple to quantize MS-estimated channels $\mathbf{H}_k$ and transmit the quantized channels back to the BS. Another approach is quantization of the precoder $\mathbf{W}_k$ at the MS as opposed to the channel $\mathbf{H}_k$. However, this approach is can not be used for the above described precoding technique since computation of $\mathbf{W}_k$ requires the kth MS to have channel matrices $\mathbf{H}_j$ for all $j \neq k$. Moreover, even if an MS has these matrices the singular value decomposition described above may be too much for an MS to handle. It is easier for the BS to acquire these matrices and perform the computations for precoding described above.

We now describe how memoryless vector quantization (VQ) may be implemented and used for closed-loop precoding. Other applicable approaches are described in [].

Memoryless vector quantization is a classification or encoding procedure which concerns the non-linear mapping of MS-estimated MIMO channel matrices $\mathbf{H}_k$ into bit vectors $\mathbf{b}_i$. The bit vectors label entries in a MIMO channel codebook $\mathcal{C}$ known by both the MSs and the BS. The bit vectors are transmitted by an MS to a BS. The BS uses the received bit vectors to access a quantized MIMO channel matrix from the codebook $\mathcal{C}$.

More specifically, let

$$Q(\mathbf{H}_k) = \arg\min_{\mathbb{V}\in\mathcal{C}} S(\mathbf{H}_k, \mathbb{H}) = \arg\min_{\mathbb{V}\in\mathcal{C}} \left\| \mathbf{V}_k \mathbf{V}_k^H - \mathbb{V}\mathbb{V}^H \right\|_2 \tag{16}$$

denote an operation that maps $\mathbf{H}_k$ to its quantized representation $Q(\mathbf{H}_k)$. Note that for some matrix $\mathbf{A}$ the matrix norm operator $\|\mathbf{A}\|_2$ produces the largest singular value of $\mathbf{A}$. Some other functions that can be used for $S(\mathbf{H}_k, \mathbb{H})$ are defined within [].

Let the MIMO channel codebook be the K-dimensional set

Quantized MIMO Channel	**3-bit Channel Index**
$\mathbb{H}_1$	$\mathbf{b}_1 = 000$
$\mathbb{H}_2$	$\mathbf{b}_2 = 001$
$\mathbb{H}_3$	$\mathbf{b}_3 = 010$
$\mathbb{H}_4$	$\mathbf{b}_4 = 011$
$\mathbb{H}_5$	$\mathbf{b}_5 = 100$
$\mathbb{H}_6$	$\mathbf{b}_6 = 101$
$\mathbb{H}_7$	$\mathbf{b}_7 = 110$
$\mathbb{H}_8$	$\mathbf{b}_8 = 111$

Table 1: Example codebook entries and indices for closed-loop precoding

$$\mathcal{C} = \left\{ \mathbb{V}_1 \quad \mathbb{V}_2 \quad \ldots \quad \mathbb{V}_K \right\} \tag{17}$$

The channel codebook consists of right singular matrices $\mathbb{V}_i$, $i = 1, 2, \ldots, K$, where each $\mathbb{V}_i$ is computed from the singular value decomposition $\mathbb{H}_i = \mathbb{U}_i \mathbb{D}_{ki} \mathbb{V}_i^H$. Let

$$\mathbf{b}_k = f(Q(\mathbf{H}_k)) \tag{18}$$

denote an operation that maps a quantized MIMO channel $Q(\mathbf{H}_k)$ to length-$\lceil \log_2 K \rceil$ bit vector $\mathbf{b}_k$ within the set

$$\mathcal{B} = \{\mathbf{b}_1, \mathbf{b}_2, \ldots, \mathbf{b}_K\} \tag{19}$$

For example, $f(Q(\mathbf{H}_k))$ may be a simple look-up table operation that outputs a bit vector $\mathbf{b}_k$ given $Q(\mathbf{H}_k)$. When the BS receives $\mathbf{b}_k$ from the kth MS the BS uses $\mathbf{b}_k$ to read the corresponding MIMO channel stored in the codebook $\mathcal{C}$.

Codebook design can be performed in advance by using a set of MIMO channel training matrices obtained via simulations. Codebook design concerns the selection of a set of unitary matrices $\mathbb{V}_i$ that minimizes the mean of a distortion function. Since $\mathbb{V}_i$ is computed from a singular value decomposition of $\mathbb{H}_i$, the distortion function provides a measure of the difference between $\mathbf{H}_k$ and its quantized version $Q(\mathbf{H}_k)$.

Codebook design can be simplified via Grassmannian subspace packing. In this approach the objective is to find a set of unitary matrices $\mathbb{V}_i$, $i = 1, 2, \ldots, K$, such that the minimal subspace distance between the K matrices $\mathbb{V}_i$ is maximized. The following function can be used for this purpose

$$\max_{\mathcal{C}} \min_{\mathbb{V}_i, \mathbb{V}_j \in \mathcal{C}} \left\| \mathbb{V}_i \mathbb{V}_i^H - \mathbb{V}_j \mathbb{V}_j^H \right\|_2, \ i \neq j \tag{20}$$

Besides Grassmannian subspace packing other methods such as the Generalized Lloyd algorithm or a Monte Carlo method may be used. These are described in [].

An example is given in Table 1, where we assume a codebook with $K = 8$ entries.

4 References

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