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# ADAPTIVE ARRAY PROCESSING FOR TIME-VARYING INTERFERENCE MITIGATION IN IEEE 802.16 SYSTEMS

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## ABSTRACT

In this work, we propose an adaptive technique for interference mitigation based on Minimum Variance Distortionless Response (MVDR) beamforming for the uplink of a WiMAX-compliant system. This method is designed to cope with time-varying interference due to the asynchronous access of users in the neighboring cells. Channel parameters needed for beamforming are obtained by exploiting both the preambles in the transmitted frames and the pilot subcarriers embedded in each information-bearing OFDM symbol. The effectiveness of the proposed technique is shown through numerical simulations of a standard WiMAX uplink over standard multipath channels.

## I. INTRODUCTION

WiMAX (Worldwide Interoperability for Microwave Access) is a standard-based technology that provides fixed last mile broadband wireless access, intended as a cost-effective alternative to existing wired technologies such cable and Digital Subscriber Line (DSL). WiMAX-compliant systems conform to the IEEE 802.16-2004 or the ETSI HiperMAN standards [1] [2]. In the uplink of a cellular WiMAX system, a major source of impairment is the out-of-cell interference. Array processing is a well studied technology for reducing interference from unwanted terminals. In order to make array processing effective, the base station needs to update the spatial filtering based on both the fluctuations of the channel of the desired user and the variations of the spatial features of the interference. In a fixed access scenario, such as the one targeted by the first release of WiMAX [1], the channel coherence time is assumed to be large enough to encompass the entire transmitted frame. However, due to the asynchronicity between the access in different cells, the spatial features of interference may vary within the frame. Therefore, the channel invariance and the noise non-stationarity need to be jointly accounted for when designing spatial processing for interference mitigation.

In this work, we propose a solution to cope with time-varying interference based on Minimum Variance Distortionless Response (MVDR) beamforming. This is coupled with a strategy to estimate both the desired user's channel and the spatial covariance of the interference, assumed to be Gaussian distributed. The method exploits both the preamble within each burst and the pilot subcarriers embedded in each OFDM symbol, working in two steps: 1) Estimate of the desired user's channel and the interference covariance matrix from the measurements of $L$ preambles in the frame; the proposed estimation exploits the stationarity of the channel within the frame and takes into account the possible variations of the interference. 2) Tracking of the interference covariance matrix along OFDM data symbols by using the $K_{\mathrm{p}}$ pilots included in each OFDM data symbol.

The effectiveness of the proposed technique is shown through numerical simulations of a standard WiMAX uplink over conventional multi-path channels.

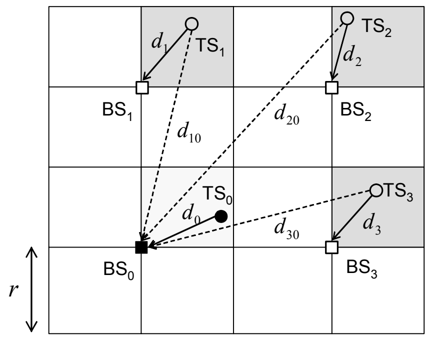


Figure 1: Uplink layout for a wireless cellular system. Shaded cells represent the first ring of interference for reception of user $\mathrm{TS}_0$ by base station $\mathrm{BS}_0$.

## II. SYSTEM AND SIGNAL MODEL

We consider the uplink of a IEEE 802.16-2004 cellular system [1]. Fig. 1 exemplifies the scenario of interest for a squared layout with frequency reuse $F = 4$. In this example, the transmission by the terminal station $\mathrm{TS}_0$ to its own base station $\mathrm{BS}_0$ is impaired by the interference from $N_I = 3$ out-of-cell terminal stations $\{\mathrm{TS}_i\}_{i=1}^{N_I}$ that employ the same carrier frequency. In the figure, $d_i$ denotes the distance of the $i$th terminal from its base station for $i = 0, \cdots, N_I$, while $d_{i0}$ is the distance of the interferer $\mathrm{TS}_i$ (with $i \neq 0$) from the the base station $\mathrm{BS}_0$ of the user of interest. $\mathrm{BS}_0$ is assumed to be equipped with an antenna array of $M$ antennas (covering a 90 degree sector in fig. 1), while TS's have a single omnidirectional antenna.

The signal transmitted by $\mathrm{TS}_0$ is organized into bursts (see fig. 2) and it is received by $\mathrm{BS}_0$ through a multi-path channel. A frame consists of $L$ bursts, with each burst being made of $L_{\mathrm{s}}$ OFDM symbols: the first OFDM symbol (preamble) contains a training sequence for synchronization and channel estimation, whereas the subsequent symbols contain coded data. In addition, each OFDM data symbol includes $K_{\mathrm{p}}$ pilot subcarriers.

Within the $s$th OFDM symbol of the $\ell$th burst ($s = 0$ represents the preamble), the $M \times K$ signal received on the $K$

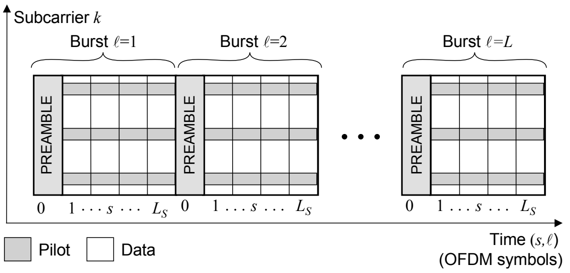


Figure 2: Frame structure for the uplink of a WiMAX-compliant system.

subcarriers can be written as

$$\mathbf{Y}(\ell,s) = \mathbf{H}\mathbf{X}(\ell,s) + \mathbf{N}(\ell,s), \tag{1}$$

where $\mathbf{H} = [\mathbf{h}_1 \cdots \mathbf{h}_K]$ is the $M \times K$ space-frequency channel matrix, whose element $(m,k)$ represents the channel gain for the $m$th receiving antenna on the $k$th subcarrier; the $K \times K$ diagonal matrix $\mathbf{X}(\ell,s) = \mathrm{diag}\{x_1(\ell,s) \cdots x_K(\ell,s)\}$ contains the transmitted symbols (either pilot or data); $\mathbf{N}(\ell,s) = [\mathbf{n}_1(\ell,s) \cdots \mathbf{n}_K(\ell,s)]$ models both the background noise and the out-of-cell interference. The noise is assumed to be zero-mean complex (circularly symmetric) Gaussian, temporally uncorrelated but spatially correlated, with spatial covariance $\mathbf{Q}(\ell,s)$ (equal for all the subcarriers):

$$\mathrm{E}[\mathbf{n}_k(\ell,s)\mathbf{n}_{k+n}^{\mathrm{H}}(\ell+m,s+t)] = \mathbf{Q}(\ell,s)\delta(n)\delta(m)\delta(t).$$

Here $\delta(\cdot)$ denotes the Dirac delta. The channel vector $\mathbf{h}_k$ is assumed to be constant throughout the entire frame, whereas the covariance $\mathbf{Q}(\ell,s)$ may generally vary on each OFDM symbol (i.e., as a function of $s$) due to time-varying interferers. Active interferers may be indeed different in each OFDM symbol, as the access is not synchronized between cells. In fig. 1, for instance, the interferer $\mathrm{TS}_1$ may stop at any given time and a new terminal may become active in the cell, generating an abrupt change in the signal interfering on user $\mathrm{TS}_0$.

### *A. Channel model*

In order to model (and estimate) the space-frequency matrix $\mathbf{H}$, it is useful to write it as $\mathbf{H} = \widetilde{\mathbf{H}}\mathbf{F}^{\mathrm{T}}$, in terms of the $M \times W$ space-time channel matrix $\widetilde{\mathbf{H}}$ that gathers by columns the $W$ taps of the discrete-time channel impulse response in the time-domain. The DFT matrix reads $F_{k,w} = \exp[-j2\pi n_k(w-1)/N]$, with $n_k \in \{0,\ldots,N-1\}$ denoting the frequency index for the $k$th useful subcarrier and $N$ the total number of subcarriers. According to the multipath model [4] for the propagation channel between $\mathrm{TS}_0$ and $\mathrm{BS}_0$, the space-time matrix $\widetilde{\mathbf{H}}$ is assumed to be the superposition of $N_R$ paths' contributions. Each path, say the $r$th, is described by a direction of arrival (DOA) at the receiving array ($\theta_{0,r}$), a delay ($\tau_{0,r}$) and a complex fading amplitude ($\alpha_{0,r}$):

$$\widetilde{\mathbf{H}} = 10^{\frac{P_0^{(\mathrm{R})}}{20}} \sum_{r=1}^{N_R} \alpha_{0,r}\mathbf{a}(\theta_{0,r})\mathbf{g}^{\mathrm{T}}(\tau_{0,r}) = \mathbf{S}\mathbf{A}\mathbf{G}^{\mathrm{T}}. \tag{2}$$

The $M \times 1$ vector $\mathbf{a}(\theta_{0,r})$ denotes the array response to the direction of arrival $\theta_{0,r}$, while the $W \times 1$ vector $\mathbf{g}(\tau_{0,r})$ collects the symbol-spaced samples of the waveform $g(t-\tau_{0,r})$, that is the cascade of transmitter and receiver filters shifted by the delay $\tau_{0,r}$. The fading amplitudes $\{\alpha_{0,r}\}_{r=1}^{N_R}$ are assumed to be uncorrelated and to have normalized power-delay-angle-profile $\Lambda_{0,r} = \mathrm{E}[|\alpha_{0,r}|^2]$ so that $\sum_{r=1}^{N_R} \Lambda_{0,r} = 1$. The matrices $\mathbf{S} = [\mathbf{a}(\theta_{0,1}) \cdots \mathbf{a}(\theta_{0,N_R})]$, $\mathbf{G} = [\mathbf{g}(\tau_{0,1}) \cdots \mathbf{g}(\tau_{0,N_R})]$ and $\mathbf{A} = \mathrm{diag}(\alpha_{0,1},\ldots,\alpha_{0,N_R})$ in (2) gather the channel parameters for the whole multipath set.

The received power $P_0^{(\mathrm{R})}$ [dBm] in (2) is given by

$$P_0^{(\mathrm{R})} = P_0^{(\mathrm{T})} + G - L(d_0) + S_0, \tag{3}$$

and it depends on: the transmitted power $P_0^{(\mathrm{T})}$ [dBm]; the transmitter-receiver antenna gain $G = G^{(\mathrm{T})} + G^{(\mathrm{R})}$ [dB]; the power loss $L(d_0)$ [dB] experienced over the distance $d_0$ between $\mathrm{TS}_0$ and $\mathrm{BS}_0$; the random fluctuations $S_0 \sim \mathcal{N}(0,\sigma_s)$ due to shadowing. As recommended in [1], the path-loss is herein modelled according to the Hata-Okamura model [4]. Notice also that $P_0^{(\mathrm{T})}$ is limited by the maximum power available at the TS's, i.e. $P_0^{(\mathrm{T})} \leq P_{\max}^{(\mathrm{T})}$.

### *B. Interference model*

As previously explained, due to the asynchronicity of the access in different cells, at any given time instant (here assumed to be a multiple of the OFDM symbol time), the position and therefore the power of the terminals interfering from neighboring cells may change. As a consequence, the non-stationary process vector $\mathbf{n}_k(\ell,s)$ has time-varying covariance $\mathbf{Q}(\ell,s) = \mathbf{Q}_n + \mathbf{Q}_I(\ell,s)$, sum of the background noise matrix $\mathbf{Q}_n = \sigma_n^2\mathbf{I}_M$ and the contribution $\mathbf{Q}_I(\ell,s)$ from the $N_I$ out-of-cell active interferers.

We assume that the signal from each interferer $\mathrm{TS}_i$, $i = 1,\ldots,N_I$, is received by $\mathrm{BS}_0$ through a multipath channel with the same characteristics as in (2). It follows that the $i$th interferer spatial covariance (averaged with respect to the fast fading) depends on the DOA's $\{\theta_{i,r}(\ell,s)\}_{r=1}^{N_R}$, the normalized power-angle-profile $\{\Lambda_{i,r}(\ell,s)\}_{r=1}^{N_R}$ and the received power $P_{i0}^{(\mathrm{R})}(\ell,s)$ [dBm], according to:

$$\mathbf{Q}_I(\ell,s) = \sum_{i=1}^{N_I} 10^{\frac{P_{i0}^{(\mathrm{R})}(\ell,s)}{10}} \sum_{r=1}^{N_R} \Lambda_{i,r}(\ell,s)\mathbf{a}(\theta_{i,r}(\ell,s))\mathbf{a}^{\mathrm{H}}(\theta_{i,r}(\ell,s)). \tag{4}$$

As in (3), the received power is obtained from the power $P_i^{(\mathrm{T})}(\ell,s)$ transmitted by $\mathrm{TS}_i$, taking into account the path-loss over the distance $d_{i0}(\ell,s)$ and the shadowing effect $S_{i0}(\ell,s) \sim \mathcal{N}(0,\sigma_s)$ over the link $\mathrm{TS}_i$-$\mathrm{BS}_0$ (see fig. 1)

$$P_{i0}^{(\mathrm{R})}(\ell,s) = P_i^{(\mathrm{T})}(\ell,s) + G - L(d_{i0}(\ell,s)) + S_{i0}(\ell,s). \tag{5}$$

Some further comment is in order about the transmitted power. Since adaptive modulation and coding is adopted to satisfy a fixed bit error rate (BER$= 10^{-6}$), the modulation/coding scheme selected (among the seven possible transmission modes listed in [1]) by the $i$th user ($i \neq 0$) and the corresponding transmitted power will be functions of the path loss (over the distance $d_i(\ell,s)$) and the shadowing (over the link $\mathrm{TS}_i$-$\mathrm{BS}_i$).

## III. Interference mitigation through array processing

On the signal (1), the base station $\mathrm{BS}_0$ performs the MVDR [5] spatial filtering $\hat{x}_k(\ell,s) = \mathbf{w}_k^{\mathrm{H}}(\ell)\mathbf{y}_k(\ell,s)$ with:

$$\mathbf{w}_k(\ell,s) = \mathbf{Q}^{-1}(\ell,s)\mathbf{h}_k\left(\mathbf{h}_k^{\mathrm{H}}\mathbf{Q}^{-1}(\ell,s)\mathbf{h}_k\right)^{-1}. \quad (6)$$

Implementation of such a beamforming requires an estimate of the channel on each subcarrier, i.e. of the whole stationary matrix $\mathbf{H}$ and the current interference covariance matrix $\mathbf{Q}(\ell,s)$. In this Section, two techniques suited for the estimation of such parameters are proposed: a first one estimates the parameters from the preambles (Sec. III-A) and a second one tracks the variations of the interference covariance matrix along the data symbols (Sec. III-B).

As a preliminary observation, notice that from (1) the received signal can be written in terms of the space-time channel matrix $\tilde{\mathbf{H}}$ as

$$\mathbf{Y}(\ell,s) = \tilde{\mathbf{H}}\tilde{\mathbf{X}}(\ell,s) + \mathbf{N}(\ell,s), \quad (7)$$

where $\tilde{\mathbf{X}}(\ell,s) = \mathbf{F}^{\mathrm{T}}\mathbf{X}(\ell,s)$ is the $W \times K$ convolution matrix with the transmitted signal in the time-domain. The alternative signal model (7) is useful for deriving the channel estimator as discussed in the following.

### A. Multi-preamble estimation in time-varying noise

Given the signal received on the preamble ($s=0$) of any burst, the conventional approach for the estimation of $\mathbf{H}(\ell,0)$ is the Least Squares (LS) technique [6]:

$$\mathbf{H}_{\mathrm{LS}}(\ell,0) = [\mathbf{Y}(\ell,0)\tilde{\mathbf{X}}(\ell,0)^{\dagger}]\cdot\mathbf{F}^{\mathrm{T}} = \tilde{\mathbf{H}}_{\mathrm{LS}}(\ell,0)\cdot\mathbf{F}^{\mathrm{T}}, \quad (8)$$

where $\tilde{\mathbf{H}}_{\mathrm{LS}}(\ell,0) = \mathbf{Y}(\ell,0)\tilde{\mathbf{X}}(\ell,0)^{\dagger}$ is the LS estimate of the space-time channel $\tilde{\mathbf{H}}(\ell,0)$, and $(\cdot)^{\dagger}$ denotes the pseudoinverse operator. Moreover, the estimate of the covariance $\mathbf{Q}(\ell,0)$ can be obtained from $\mathbf{N}_{\mathrm{LS}}(\ell,0) = \mathbf{Y}(\ell,0) - \mathbf{H}_{\mathrm{LS}}(\ell,0)\mathbf{X}(\ell,0)$ as

$$\mathbf{Q}_{\mathrm{LS}}(\ell,0) = \frac{1}{K}\mathbf{N}_{\mathrm{LS}}(\ell,0)\mathbf{N}_{\mathrm{LS}}^{\mathrm{H}}(\ell,0). \quad (9)$$

The long coherence time of the channel $\mathbf{H}$, not considered in the preamble-by-preamble estimation above, can be exploited by simply averaging the LS estimates over the preambles (i.e., over $\ell$). This approach will be referred to as the multi-preamble LS estimate (MLS):

$$\mathbf{H}_{\mathrm{MLS}} = \frac{1}{L}\sum_{\ell=1}^{L}\mathbf{H}_{\mathrm{LS}}(\ell,0) \quad (10)$$

$$\mathbf{Q}_{\mathrm{MLS}}(\ell,0) = \frac{1}{K}\mathbf{N}_{\mathrm{MLS}}(\ell,0)\mathbf{N}_{\mathrm{MLS}}^{\mathrm{H}}(\ell,0), \quad (11)$$

with $\mathbf{N}_{\mathrm{MLS}}(\ell,0) = \mathbf{Y}(\ell,0) - \mathbf{H}_{\mathrm{MLS}}\mathbf{X}(\ell,0)$.

Even though the MLS estimate (10) is consistent (the estimate error goes to zero for $L\to\infty$, due to the independence of the $L$ measures), it is suboptimal as it does not account for the non-stationarity of the noise. A weighting should be introduced in the average (10) to account for time-varying second-order statistics of noise. To this aim, let us perform a spatial pre-whitening before channel estimation:

$$\mathbf{Y}_w(\ell,0) = \mathbf{Q}_{\mathrm{LS}}^{-\mathrm{H}/2}(\ell,0)\mathbf{Y}(\ell,0).$$

This, for $K\to\infty$ and thus $\mathbf{Q}_{\mathrm{LS}}(\ell,0)\to\mathbf{Q}(\ell,0)$, makes the noise $\mathbf{N}_w(\ell,0) = \mathbf{Q}_{\mathrm{LS}}^{-\mathrm{H}/2}(\ell,0)\mathbf{N}(\ell,0)$ be stationary over the preambles (i.e., over $\ell$). However, it has to be noticed that the whitened channel $\tilde{\mathbf{H}}_w(\ell,0) = \mathbf{Q}_{\mathrm{LS}}^{-\mathrm{H}/2}(\ell,0)\tilde{\mathbf{H}}(\ell,0)$ is now time-varying. More specifically, only the spatial component has been modified, from (1), into $\mathbf{S}_w(\ell,0) = \mathbf{Q}_{\mathrm{LS}}^{-\mathrm{H}/2}(\ell,0)\mathbf{S}$, and it is now varying from preamble to preamble. The temporal component is still constant over the whole frame. The optimal estimate for such a channel structure, characterized by a non-stationary spatial component and a constant temporal component, can be derived following the maximum likelihood approach [6]. We refer to the resulting estimate as multi-preamble space-time estimate (MST) given by

$$\mathbf{H}_{\mathrm{MST}}(\ell,0) = \left[\tilde{\mathbf{H}}_{\mathrm{LS}}(\ell,0)\mathbf{R}_{\tilde{x}\tilde{x}}^{-\mathrm{H}/2}\mathbf{P}\,\mathbf{R}_{\tilde{x}\tilde{x}}^{-\mathrm{H}/2}\right]\cdot\mathbf{F}^{\mathrm{T}} \quad (12)$$

where $\mathbf{R}_{\tilde{x}\tilde{x}} = \tilde{\mathbf{X}}(\ell,0)\tilde{\mathbf{X}}^{\mathrm{H}}(\ell,0)$, and $\mathbf{P}$ is the projector onto the $r_0$ dominating eigenvectors of the temporal correlation

$$\mathbf{R} = \frac{1}{L}\sum_{\ell=1}^{L}\mathbf{R}_{\tilde{x}\tilde{x}}^{-1/2}\tilde{\mathbf{H}}_{\mathrm{LS},w}^{\mathrm{H}}(\ell,0)\mathbf{Q}_{\mathrm{LS}}^{-1}(\ell,0)\tilde{\mathbf{H}}_{\mathrm{LS},w}(\ell,0)\mathbf{R}_{\tilde{x}\tilde{x}}^{-\mathrm{H}/2} \quad (13)$$

Notice that the estimation of the temporal part of the channel is obtained from (13) based on a multi-preamble observation, while the estimate of the spatial part is updated within each preamble. The noise covariance estimate is again obtained as $\mathbf{Q}_{\mathrm{MST}}(\ell,0) = \frac{1}{K}\mathbf{N}_{\mathrm{MTS}}(\ell,0)\mathbf{N}_{\mathrm{MST}}^{\mathrm{H}}(\ell,0)$, with $\mathbf{N}_{\mathrm{MST}}(\ell,0) = \mathbf{Y}(\ell,0) - \mathbf{H}_{\mathrm{MST}}\mathbf{X}(\ell,0)$.

### B. Tracking of the interference covariance from data symbols

The discussion above covered the computation of the channel and interference covariance matrices on the preamble of each burst. However, these estimates cannot be used for evaluating the MVDR beamformer (6) within the data burst since the interference covariance matrix $\mathbf{Q}(\ell,s)$ may also vary within the burst (i.e., along $s$). Therefore, a technique should be devised in order to track the variations of $\mathbf{Q}(\ell,s)$ by using the $K_{\mathrm{p}}$ pilots included in each data OFDM symbol. Given any channel estimate $\hat{\mathbf{H}}$ and labeling by the subscript p the signals on the $K_{\mathrm{p}}$ pilot subcarriers, a preliminary estimate of $\mathbf{Q}(\ell,s)$ can be obtained as

$$\mathbf{Q}_{\mathrm{p}}(\ell,s) = \frac{1}{K_{\mathrm{p}}}\mathbf{N}_{\mathrm{p}}(\ell,s)\mathbf{N}_{\mathrm{p}}^{\mathrm{H}}(\ell,s), \quad (14)$$

from $\mathbf{N}_{\mathrm{p}}(\ell,s) = \mathbf{Y}_{\mathrm{p}}(\ell,s) - \hat{\mathbf{H}}\mathbf{X}_{\mathrm{p}}(\ell,s)$. The estimate (14) can be compared with the estimate in the previous OFDM symbol in order to decide whether the interference has changed or not. This operation is herein performed by computing the correlation between the noise covariance matrix at two successive instants as:

$$\rho(\ell, s) = \frac{\mathrm{tr}[\mathbf{Q}_{\mathrm{p}}(\ell, s)\mathbf{Q}_{\mathrm{p}}(\ell, s-1)]}{||\mathbf{Q}_{\mathrm{p}}(\ell, s)|| \cdot ||\mathbf{Q}_{\mathrm{p}}(\ell, s-1)||}$$

where $||\cdot||$ denotes the Frobenius norm of the argument matrix. If the correlation $\rho(\ell, s)$ is larger than a given threshold $\bar{\rho}$ (to be determined experimentally), the interference covariance estimate can be refined by a sample average, otherwise it needs to be re-initialized according to the new estimate value (14) as

$$\hat{\mathbf{Q}}(\ell, s) = \begin{cases} [\mathbf{Q}_{\mathrm{p}}(\ell, s) + (T-1)\hat{\mathbf{Q}}(\ell, s-1)]/T, & \rho \geq \bar{\rho} \\ \mathbf{Q}_{\mathrm{p}}(\ell, s), & \rho < \bar{\rho} \end{cases}$$

where $T \leq s+1$ is the number of averaged matrices at the $s$th OFDM symbol.

## IV. Numerical results

In this Section, the uplink of a IEEE 802.16-2004 compliant system [1] is considered with a cellular layout as in fig. 1 and cell side $r = 1$km. The main system parameters used for simulations are listed in Table 1. A uniform linear array (ULA) of $M = 4$ elements is adopted by $\mathrm{BS}_0$ with inter-element spacing of $d = 1.8\lambda$ [3]. The receiver at $\mathrm{BS}_0$ consists of MVDR filtering, soft demodulation, Log-MAP convolutional decoding and Reed-Solomon decoding. The user $\mathrm{TS}_0$ transmits at maximum power $P_0^{(T)} = P_{\max}^{(T)}$ with transmission mode QPSK-$\frac{1}{2}$ (QPSK modulation and coding rate $1/2$ [1]). Interferers $\{\mathrm{TS}_0\}_{i=1}^{3}$ are uniformly distributed in their cells. Their power and transmission mode are adaptively selected based on the spatial position and the shadowing effects as described in Sec. II-B.

Delays and amplitudes of the multipath channel (2) are selected according to the SUI-3 model. Directions of arrival of both user and interferers are drawn from a Gaussian distribution $\theta_{i,r} \sim \mathcal{N}(\theta_i, \sigma_\theta)$ with mean $\theta_i$ uniformly distributed in the 90deg sector and standard deviation $\sigma_\theta = 5\,\mathrm{deg}$. For

Table 1: System parameters

| | |
|---|---|
| Carrier frequency $f_c$ | 3.5GHz |
| Channel bandwidth | 4MHz |
| N. of subcarriers $N$ | 256 |
| N. of useful subcarriers $K$ | 200 |
| N. of pilot symbols per OFDM symbol $K_p$ | 8 |
| TS maximum power $P_{\max}^{(\mathrm{T})}$ | 27dBm |
| TS omnidirection antenna gain $G^{(\mathrm{T})}$ | 2dBi |
| BS directional antenna gain (broadside) $G^{(\mathrm{R})}$ | 16dBi |
| Path-loss exponent $\gamma$ | 4 |
| Reference path-loss distance $d_{\mathrm{ref}}$ | 100m |
| Shadowing standard deviation $\sigma_s$ | 8dB |
| N. of paths for each interfer $N_R$ | 3 |
| Temporal channel support $W$ | 32 |
| Cyclic prefix length | 32 |

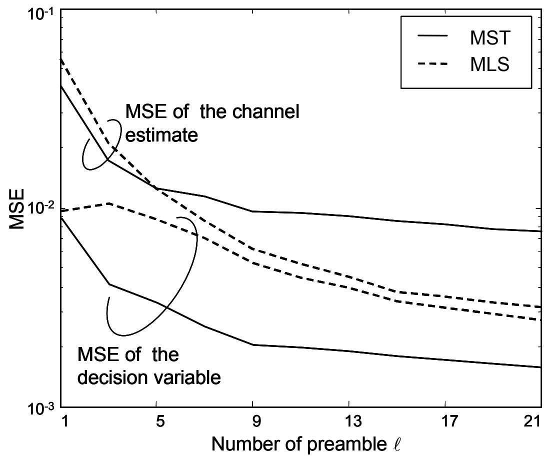


Figure 3: Normalized MSE for the channel estimate and the decision variable, for MLS and MST, versus the number of preamble $\ell$.

each interferer a uniform power-angle delay profile is adopted ($\Lambda_{i,r} = 1/N_R$, for $i \neq 0$).

The system performance is first evaluated for the case where the interference covariance matrix varies at the beginning of each burst but it is constant within each burst: $\mathbf{Q}(\ell, s) = \mathbf{Q}(\ell)$, $\forall s$ (symbol index is dropped). The user $\mathrm{TS}_0$ is placed at distance $d_0 = 0.8$km from $\mathrm{BS}_0$ with DOA $\theta_0 = 0$deg. We compare the mean square error (MSE) on the channel estimate $\mathrm{MSE}_h(\ell) = \mathrm{E}[||\hat{\mathbf{H}}(\ell) - \mathbf{H}||^2]$ and the MSE on the decision variable $\mathrm{MSE}_x(\ell) = \mathrm{E}[|\hat{x}_k(\ell, s) - x_k(\ell, s)|^2]$ versus the preamble number $\ell$ for the different estimation techniques. The error on the decision variable clearly depends on both the interference and the channel estimate accuracy. This is shown briefly in the following. Denoting by $\Delta\mathbf{h}_k(\ell) = \hat{\mathbf{h}}_k(\ell) - \mathbf{h}_k$ the channel estimate error on the $k$th subcarrier, from (1) and (6) it is

$$\mathrm{MSE}_x(\ell) = \underbrace{\mathbf{w}_k^{\mathrm{H}}(\ell)\mathbf{Q}(\ell)\mathbf{w}_k(\ell)}_{\mathrm{MSE}_{x,1}(\ell)} + \underbrace{\mathbf{w}_k^{\mathrm{H}}(\ell)\,\mathrm{Cov}(\Delta\mathbf{h}_k(\ell))\mathbf{w}_k(\ell)}_{\mathrm{MSE}_{x,2}(\ell)} \tag{15}$$

where $\mathbf{w}_k(\ell)$ is the MVDR filter calculated as in (6) from the channel estimate $\hat{\mathbf{h}}_k(\ell)$ and for known spatial covariance $\hat{\mathbf{Q}}(\ell) = \mathbf{Q}(\ell)$ (as for for $K \to \infty$). To simplify, we have assumed uncorrelation between the channel estimate error $\Delta\mathbf{h}_k(\ell)$ and interference $\mathbf{n}_k(\ell, s)$, and also $\mathrm{E}[|x_k(\ell, s)|^2] = 1$. We notice that the first term in (15) depends on the interference only, while the second one is also affected by the channel estimate covariance $\mathrm{Cov}(\Delta\mathbf{h}_k(\ell))$. Fig. 3 compares the two squared errors $\mathrm{MSE}_h(\ell)$ (top figure) and $\mathrm{MSE}_{x,2}(\ell)$ (bottom figure) for MLS and MST. It can be seen that, even though the MLS estimate is more convenient than MST in terms of $\mathrm{MSE}_h(\ell)$, its error on the decision variable $\mathrm{MSE}_{x,2}(\ell)$ is significantly larger than MST. We can thus conclude that MST is better suited to be used for MVDR beamforming (6).

In fig. 4-top we compare the estimation techniques with the ideal case of known channel in terms of average BER (after channel decoding) versus the angular position of the user placed at a distance $d = 0.8$km from the BS. The probability

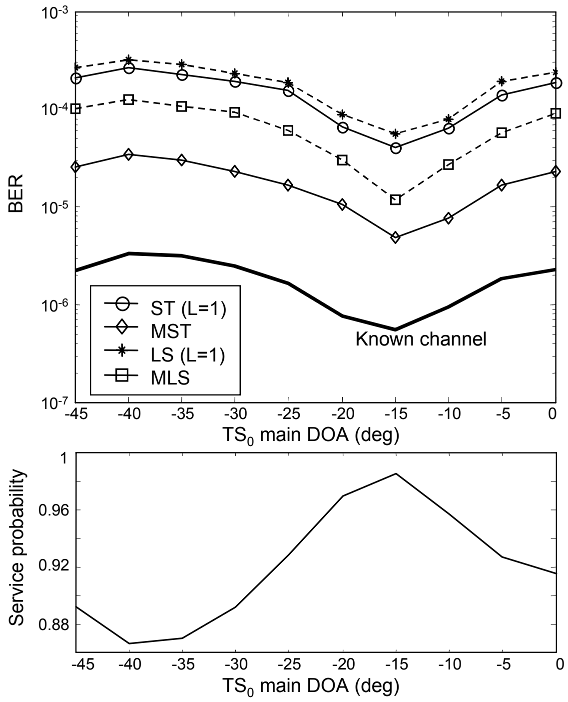


Figure 4: Average BER (with minimal allowed BER= $10^{-3}$) for MVDR receiver with different parameter estimation techniques (top); probability of service (bottom).

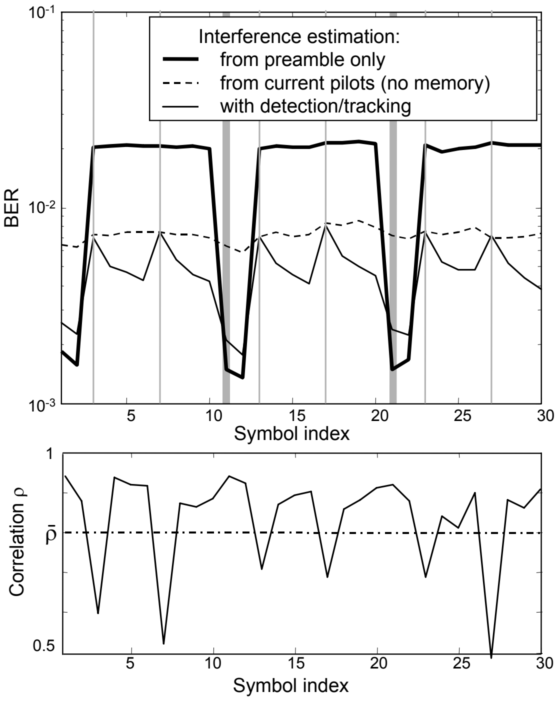


Figure 5: BER (top) and correlation value $\rho$ (bottom) as a function of the time index over the frame.

of service, calculated as $1 - P_{out}$ from the outage probability $P_{out}$, is also shown on the bottom of the figure. The outage probability is here defined as $P_{out} = \Pr\left\{P_b \geq \bar{P}_b\right\}$ where the minimum BER is set to $\bar{P}_b = 10^{-3}$ and the reference BER $P_b$ is computed for a MVDR receiver with known channel. The average BER is obtained by averaging only over the channel instances that satisfy $P_b < \bar{P}_b$. The results show that MST can gain a decade in terms of BER with respect to MLS.

We now let the interference covariance matrix vary asynchronously within each burst. In particular, we consider $L = 3$ bursts of $L_s = 10$ symbols and the user $TS_0$ placed in broadside at a distance $d = 0.8$km from $BS_0$. The interference scenario changes at the third and seventh symbol of each burst, with positions of the three interferers selected uniformly within their cell. The threshold is set to $\bar{\rho} = 0.8$. Fig. 5 shows the BER (top) and the interference correlation $\rho(\ell, s)$ (bottom) over the OFDM symbols. The estimation of the interference matrix $\mathbf{Q}(\ell, s)$ is obtained as in Sec. III using three different approaches: estimation only from the preamble of the current burst (thick line); re-estimation within each OFDM symbol without tracking (dashed line); tracking in each OFDM symbol by the method in Sec. III with change detection (thin line). The BER results confirm that the proposed tracking method is an effective approach for time-varying interference mitigation.

## V. Conclusion

In this work, an adaptive technique based on MVDR beamforming that copes with out-of-cell asynchronous interference in the uplink of a WiMAX-compliant system has been proposed. The method exploits both the preambles and the pilot subcarriers embedded in each data OFDM symbol in order to estimate the time-invariant wireless channel of the desired use and track the variations of spatial characteristics of interference. Performance of the discussed technique has been validated through numerical results of a multi-cell system in a standard multipath propagation environment.

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