Investigating the Performance of Smart Antenna Systems at the Mobile and Base Stations in the Down and Uplinks

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Abstract— Future wireless communication systems are expected to support a wide range of services which will include video, data and voice. To achieve the necessary capacity for these systems, the use of adaptive antennas has been proposed at the basestation. Further increases in capacity may be possible by including adaptive antennas at both the base and mobile stations in the up and down links. In this paper, we investigate the performance of such a system.

I. INTRODUCTION

Smart or adaptive antennas promise to provide significant increases in system capacity and reduce the average error probability in wireless communication systems. However, to date, smart antennas have generally only been utilized at the basestation (BS). In this paper we investigate the performance of using adaptive antennas at both the BS and mobile station (MS) in the up and down links. Potential advantages include further reductions in average error probability and increases in capacity [1].

The practical implementation of such a system would require antenna arrays at both the BS and MS. For this reason, only systems where the MS is larger than the size of several antennas are feasible targets for our system. Possible examples include wireless LAN systems that operate at 2.4GHz (or higher) in which the MS is typically a notebook computer [2], [3], [4].

To analyse the performance of the adaptive base and mobile (SBM) system, we develop a formulation to determine the antenna weights for the base and mobile antennas that give the optimum SINR in both flat and frequency selective fading channels. Using these weights, we then employ computer simulations for various configurations of antennas at both the MS and BS to determine the average bit error rate (BER) performance. Throughout, we utilize QPSK modulation in an environment with ISI, AWGN and CCI in a similar way to that described in [5].

This paper is organized as follows. In Section II, we introduce the system model. Section III provides analytical expressions for optimal antenna weight vectors for SBM. Section IV proposes a SBM practical implementation and comparative simulation results are presented in Section V. Finally, Section VI concludes our work.

II. PRELIMINARIES

The configuration of the SBM system is shown in Figure 1 where K antennas are located at the BS and L

antennas are located at the MS. The signal transmitted by the k th antenna in the n th symbol interval is multiplied by a controllable complex weight which we write as $w_B^k(n)$. The signal received by the ℓ th element is multiplied by another controllable complex weight $w_M^\ell(n)$ and the weighted signals from all L antennas are summed to form a scalar output $s_o(n)$. For convenience, we write the antenna weights at the BS and MS as vectors, so that

$$\mathbf{w}_B(n) = [w_B^1(n) \ w_B^2(n) \ \cdots \ w_B^K(n)]^T \quad (1)$$

$$\mathbf{w}_M(n) = [w_M^1(n) \ w_M^2(n) \ \cdots \ w_M^L(n)]^T \quad (2)$$

where the superscript T denotes the transpose operation.

The antenna elements transmit or receive information through a wireless communication channel which is here characterised by a multipath fading model. For a particular channel, the multipath model is represented by its channel impulse response using a m-ray model defined as

$$c_{k,\ell}(t) = \sum_{m=0}^{M-1} \beta_{k,\ell}^m \delta(t - \tau_{k,\ell}^m) \quad (3)$$

where the subscripts k, ℓ refer to the channel between the k th and ℓ th antenna at the BS and MS, respectively and M is the total number of paths [6]. Likewise, $\beta_{k,\ell}^m$ and $\tau_{k,\ell}^m$ are, respectively, the complex gain and time delay for the m th path of the diversity channel.

To determine $\beta_{k,\ell}^m$, we use a statistical approach, and assume that paths with different delays are uncorrelated (i.e. uncorrelated scattering). In addition, we assume that the paths are uncorrelated for each antenna branch so as to provide perfect spatial diversity. We model $\beta_{k,\ell}^m$ statistically by zero mean, complex Gaussian random variables, with their power following the exponential delay profile given by

$$E[|c_{k,\ell}(t)|^2] = \begin{cases} \frac{1}{D} \exp(-\frac{t}{D}) & \text{for } t \geq 0 \\ 0 & \text{elsewhere.} \end{cases} \quad (4)$$

in which $D \triangleq \tau/T$ where τ and T are the *rms* delay spread and symbol period, respectively. Moreover, we assume a constant increment delay, so $\tau_{k,\ell}^m = \frac{5D}{M}m$.

We let $\mathbf{H}(m)$ represent the channel matrix for the m th

symbol, so that

$$\mathbf{H}(m) = \begin{bmatrix} h_{1,1}(mT) & h_{1,2}(mT) & \cdots & h_{1,L}(mT) \\ h_{2,1}(mT) & h_{2,2}(mT) & \cdots & h_{2,L}(mT) \\ \vdots & \vdots & \cdots & \vdots \\ h_{K,1}(mT) & \cdots & \cdots & h_{K,L}(mT) \end{bmatrix} \quad (5)$$

where $h_{k,\ell}(t)$ is equal to $(g \otimes c_{k,\ell})(t)$ in which $c_{k,\ell}(t)$ is defined in (3), and $g(t)$ is a raised cosine rolloff pulse shaping function, and $\otimes$ denotes the convolution operator between two continuous time signals.

Interference is also considered and P interferers are assumed. For now, we consider the down link only. However, the uplink will be similar. The interfering channel to the ℓ th antenna of the desired mobile also has multipath and is defined similar to (3). Hence, the channel vector from the p th interferer is

$$\tilde{\mathbf{h}}_p(m) = [\tilde{h}_p^1(mT) \ \tilde{h}_p^2(mT) \ \cdots \ \tilde{h}_p^L(mT)]^T \quad (6)$$

where $\tilde{h}_p^\ell(\cdot)$ is the composite channel impulse response representing the p th interference to the ℓ th antenna of the desired MS link.

In the context of a QPSK signaling scheme, the transmitted baseband signal is

$$s(t) = \sum_n z(n)g(t - nT) \quad (7)$$

where $\{z(n)\}$ is the complex symbol sequence having the values of $\pm 1 \pm j$ and $g(t)$ is the pulse shaping function described earlier.

In this paper, a TDMA based transmission system is assumed. In particular, data are transmitted in packets over the radio channel and these packets may include a training sequence for equalization and/or synchronization purposes along with information data. We assume the channel is quasi-stationary so that the channel is time-invariant over a packet.

III. OPTIMAL WEIGHTS

Using the system model described in Section II, our objective is to find the optimum BS and MS weight vectors $(\mathbf{w}_B, \mathbf{w}_M)$ which maximize the overall SINR with the constraint of a fixed transmitter power P_T . In the down link, this is expressed as

$$(\mathbf{w}_B, \mathbf{w}_M)_{\text{opt}} = \arg \max_{\mathbf{w}_M, \|\mathbf{w}_B\|^2 = P_T} \{\text{SINR}\} \quad (8)$$

From this solution, we can determine the overall BER performance of SBM and compare it to a conventional adaptive antenna system at the BS only. In our analysis, we assume that the desired channel matrix $\mathbf{H}(\cdot)$ is known. In Section IV, we consider the case where $\mathbf{H}(\cdot)$ is not known.

A. Flat Fading

In a flat fading radio environment, ISI is negligible, so that only $\mathbf{H}(0)$ in (5) and $\tilde{\mathbf{h}}_p(0)$ in (6) are significant. As

such, the received signal can be written as

$$s_o(n) = z(n)\mathbf{w}_M(n)^T \underline{\mathbf{H}}^T \mathbf{w}_B(n) + \sum_{p=1}^P \tilde{z}_p(n)\mathbf{w}_M(n)^T \tilde{\mathbf{h}}_p + \mathbf{w}_M(n)^T \mathbf{x}_n(n) \quad (9)$$

where $\underline{\mathbf{H}}$, $\tilde{\mathbf{h}}_p$ and $\mathbf{x}_n(n)$ are used to represent $\mathbf{H}(0)$, $\tilde{\mathbf{h}}_p(0)$ and AWGN noise, respectively (See Section II).

Given $\underline{\mathbf{H}}$ and $\tilde{\mathbf{h}}_p$, the average SINR can be written as

$$\text{SINR} = \frac{\sigma_z^2 \mathbf{w}_M(n)^H \left(\underline{\mathbf{H}}^T \mathbf{w}_B(n) \right)^* \left(\underline{\mathbf{H}}^T \mathbf{w}_B(n) \right)^T \mathbf{w}_M(n)}{\mathbf{w}_M(n)^H \left[\sum_{p=1}^P \sigma_{\tilde{z}_p}^2 \tilde{\mathbf{h}}_p \tilde{\mathbf{h}}_p^* + \sigma_n^2 \mathbf{I} \right] \mathbf{w}_M(n)} \quad (10)$$

where $*$ and H denote the complex conjugate and complex hermitian operators, respectively, and $\mathbf{I}$ is the identity matrix. In the above equation, $\sigma_z^2 \triangleq \mathbb{E}[|z(n)|^2]$ and $\sigma_{\tilde{z}_p}^2 \triangleq \mathbb{E}[|\tilde{z}_p(n)|^2]$ are defined as the digital symbol power of the desired user and interfering users, respectively. AWGN with variance σ_n^2 is also assumed.

By utilizing a standard result in [7], the weight vector, $\mathbf{w}_M(n)$, which maximizes the output SINR is

$$\mathbf{w}_M(n)|_{\text{SINR}_{\text{max}}} = \mu \Phi_u^{-1} \left(\underline{\mathbf{H}}^T \mathbf{w}_B(n) \right)^* \quad (11)$$

where $\Phi_u \triangleq \sum_{p=1}^P \sigma_{\tilde{z}_p}^2 \tilde{\mathbf{h}}_p \tilde{\mathbf{h}}_p^* + \sigma_n^2 \mathbf{I}$ is defined as the correlation matrix of the undesired signal, and μ is an arbitrary real constant.

In other words, we can find the optimum weights $\mathbf{w}_M(n)$ given that $\mathbf{w}_B(n)$ is known. The difficulty, however, is how to determine the optimum $\mathbf{w}_B(n)$ and the remainder of this section discusses this.

In general, Φ_u is a hermitian and invertible matrix. Thus, $(\Phi_u^{-1})^*$ is also a hermitian matrix as is $\mathbf{A} \triangleq \underline{\mathbf{H}}^* (\Phi_u^{-1})^* \underline{\mathbf{H}}^T$. As a result, it follows that $\mathbf{A}$ can be decomposed into

$$\mathbf{A} = \Psi \Lambda \Psi^H \quad (12)$$

where Ψ is a unitary matrix whose columns are the eigenvectors of $\mathbf{A}$ and Λ is a diagonal matrix which contains the corresponding eigenvalues. Consequently,

$$\begin{aligned} \text{SINR} &= \sigma_z^2 \left(\Psi^H \mathbf{w}_B(n) \right)^H \Lambda \left(\Psi^H \mathbf{w}_B(n) \right) \\ &= \sigma_z^2 \mathbf{w}'_B(n)^H \Lambda \mathbf{w}'_B(n) \\ &= \sigma_z^2 (\lambda_1 |\dot{w}_B^1(n)|^2 + \cdots + \lambda_K |\dot{w}_B^K(n)|^2) \end{aligned} \quad (13)$$

where $\mathbf{w}'_B(n)$ is equal to $\Psi^T \mathbf{w}_B(n)$ and $\{\lambda_k\}_{k=1}^K$ are the eigenvalues of $\mathbf{A}$. Our constant transmitted power constraint, $\|\mathbf{w}_B(n)\|^2 = P_T$ then becomes $\|\mathbf{w}'_B(n)\|^2 = P_T$. Maximizing the SINR (13) can then be performed by using standard theory [8] so that the optimum transmission weight vector is

$$\mathbf{w}_B(n)|_{\text{SINR}_{\text{max}}} = \sqrt{P_T} \mathbf{v}(\mathbf{A}) \quad (14)$$

where $\mathbf{v}(\mathbf{A})$ is the eigenvector of the matrix $\mathbf{A}$ which corresponds to the largest eigenvalue. Throughout, we will refer to $\mathbf{v}(\mathbf{A})$ as the principle eigenvector of the matrix $\mathbf{A}$. The corresponding maximum SINR is

$$\text{SINR}|_{max} = \sigma_z^2 P_T \cdot \max_k \{\lambda_k\} \quad (15)$$

Hence, the maximum SINR we can achieve with the constraint $\|\mathbf{w}_B(n)\|^2 = P_T$ is given by (15) and is achieved by using weight vectors given by (11) and (14).

B. Frequency Selective Fading

In frequency selective fading channels, the received signal includes multiple versions of the transmitted waveform which are attenuated and delayed in time. In general, the optimization problem (8) cannot be expected to have a closed form solution for the weight vectors as in the case of flat fading. Fortunately, approximate solutions can be provided. Specifically, assuming that the transmission weights $\mathbf{w}_B(n)$ are fixed, then the reception weight vector which maximizes the SINR is

$$\mathbf{w}_M(n)|_{\text{SINR}_{max}, \mathbf{w}_B(n)} = \mu_1 \Phi_1^{-1} \left(\mathbf{H}(0)^T \mathbf{w}_B(n) \right)^* \quad (16)$$

where μ_1 is a real constant and Φ_1 is given by

$$\begin{aligned} \Phi_1 = & \sigma_z^2 \sum_{\substack{m=-\infty \\ m \neq 0}}^{\infty} \left(\mathbf{H}(m)^T \mathbf{w}_B(n) \right)^* \left(\mathbf{H}(m)^T \mathbf{w}_B(n) \right)^T \\ & + \sum_{p=1}^P \sum_{m=-\infty}^{\infty} \sigma_{z_p}^2 \tilde{\mathbf{h}}_p(m)^* \tilde{\mathbf{h}}_p(m)^T + \sigma_n^2 \mathbf{I}. \end{aligned} \quad (17)$$

Similarly, for a fixed reception weight vector $\mathbf{w}_M(n)$, the optimal transmission weight vector is

$$\mathbf{w}_B(n)|_{\text{SINR}_{max}, \mathbf{w}_M(n)} = \mu_2 \Phi_2^{-1} \left(\mathbf{H}(0) \mathbf{w}_M(n) \right)^* \quad (18)$$

where μ_2 is a real constant and Φ_2 is defined as

$$\begin{aligned} \Phi_2 = & \sigma_z^2 \sum_{\substack{m=-\infty \\ m \neq 0}}^{\infty} \left(\mathbf{H}(m) \mathbf{w}_M(n) \right)^* \left(\mathbf{H}(m) \mathbf{w}_M(n) \right)^T \\ & + \frac{\zeta}{P_T} \mathbf{I} \end{aligned} \quad (19)$$

with $\zeta = \mathbf{w}_M(n)^H [\sigma_n^2 \mathbf{I} + \sum_{p=1}^P \sum_{m=-\infty}^{\infty} \sigma_{z_p}^2 \tilde{\mathbf{h}}_p(m)^* \tilde{\mathbf{h}}_p(m)^T] \mathbf{w}_M(n)$. Using (16) and (18), the optimal weight vectors for transmission and reception in frequency selective fading radio channels can be estimated iteratively. That is by guessing an initial $\mathbf{w}_B(n)$, we can find a $\mathbf{w}_M(n)$ from (16). Using this $\mathbf{w}_M(n)$, we can then update our initial guess for $\mathbf{w}_B(n)$ using (18), and so on iteratively.

In order to have fast convergence, a good choice for an initial guess of $\mathbf{w}_B^{(0)}(n)$ is important. A logical guess is

$$\mathbf{w}_B^{(0)}(n) = [1 \ 1 \ \dots \ 1]^T \quad (20)$$

although better choices are available.

IV. SYSTEM CONFIGURATION

In the previous section, the optimal solution for SBM system in flat and frequency selective channels are given in (11), (14), (16) and (18) respectively. However, perfect knowledge of the channels and noise statistics have been assumed. Therefore, the results are only useful in evaluating the optimum performance of SBM system. In practice, channel estimation is needed. In the following, we suggest two ways to achieve SBM in conjunction with channel estimation. We concentrate on Time Division Duplex (TDD) systems, so the uplink signal can be used to estimate the down link channels and vice versa, thereby making channel estimation easier. However, modifications for FDD operation are possible.

A. SBM

Here, we suggest the use of pilot tones [9] to perform channel estimation. In TDD systems, pilot tones can be used on the down link to estimate the uplink since the channels are reciprocal. The down link can be estimated in a similar way. For FDD systems, feedback of the channel estimations would be necessary.

An important assumption in both approaches is that the channel is stationary over two packets. Having the knowledge of channel responses, the adaptive array processor at BS determines the weight vectors $\mathbf{w}_B(n)$ and $\mathbf{w}_M(n)$ using a Sample Matrix Inversion (SMI) processor during the reference sequence reception period [10]. We refer to this pilot tone approach as *SBM*.

B. Adaptive SBM

The problem with the pilot tone approach is the overhead associated with the tones. To keep the system implementation as simple as possible, we have developed another algorithm for determining the antenna weights that does not require the use of tones in a TDD system.

The adaptation scheme proposed relies on the channel being invariant for several packets. In initialization, the BS weights are set equal (so the antenna pattern is omnidirectional), a transmission from BS to the MS occurs, and SMI is used to find the optimum weights for reception at the MS. These weights are then used to form an antenna pattern for a transmission from the MS to BS. Once this transmission is received at the BS, SMI is used to determine the optimal weights for reception at the BS. These weights are used to form an updated pattern for transmission from BS to MS, and the process continues until convergence.

To describe this procedure exactly, we define some terminology. Any transmission occurring from BS to MS is regarded as a down link transmission and the corresponding SMI update to the weights $\mathbf{w}_M(n)$ is known as a down link adaptation. Similarly, a transmission from MS to BS is termed an uplink transmission and the update to the weights $\mathbf{w}_B(n)$ an uplink adaptation. The exact process to find the weights can then be described as

- *Step 1* : BS transmits omnidirectionally to MS and the antenna weights at the MS are calculated using

SMI (First Down link adaptation).

- *Step 2* : The MS antenna weights are then used for transmission back to the BS on the uplink.
- *Step 3* : The antenna weights at BS are calculated based on SMI so that the uplink channel effectively contains smart antennas at both the BS and MS (First Uplink adaptation).
- *Step 4* : The BS antenna weights are then used for transmission so that the system has multiple smart down link channels.
- *Step 5* : Based on SMI, the antenna weights at MS are then updated (Second Down link adaptation).

The performance of the system can be further enhanced if more adaptations are utilized. However, more adaptations require more packets transmissions, thereby, implying a larger processing delay. Once the adaptation process is finished, more adaptation is only required to adapt to the time-varying channel. We will refer to this method as *Adaptive SBM*.

V. SIMULATION RESULTS

The proposed SBM (Section IV.A) and Adaptive SBM (Section IV.B) systems are investigated for a TDMA/TDD based wireless communication system by computer simulation. The configuration considered is for 2 antenna elements at the both the BS and MS. For each technique, average error probability (P_e) are provided for various CCI (Signal-to-interference ratio (SIR)) and AWGN (SNR). The results are compared to a conventional smart antenna system with 2 antennas at the BS only and also standard QPSK modulation without smart antennas.

For each P_e simulation, more than 10,000 data packets are transmitted with independent channels are simulated [5]. A data packet consists of 1,000 data symbols for each transmission. For the QPSK signal, a raised cosine pulse shaping filter with rolloff factor of 0.3 is used. In the simulation, the channel of each link contains 10 random multipath components ($M = 10$ in (3)) and one interferer.

A. Flat Fading

In Figure 2, results are provided for a flat Rayleigh fading channel in the absence of interference. By comparing the conventional adaptive antenna at the BS only and our SBM systems, we can observe that there is up to a 100 times decrease in P_e for our system. This reveals that significant advantages can be achieved by using smart antennas at both the BS and MS. Moreover, the results show that the error probability performance of the Adaptive SBM with 3 adaptations has only 0.5dB difference compared to the optimum SBM system. Thereby, implying that the proposed Adaptive SBM approach is an effective way in implementing adaptive antennas at the BS and MS in flat fading environments.

In Figure 3, results are provided for a flat Rayleigh fading channel with CCI at SIR = 15dB. These results demonstrate that SBM and adaptive SBM provide a 100

and a 10 fold decrease in P_e as compared to a conventional adaptive antenna at BS only. The large difference in performance of SBM and adaptive SBM can be explained by realizing that the mechanism of the Adaptive SBM system takes advantages of the symmetry of the duplex channels. Under interference conditions, the down link and uplink channels are no longer identical. Thus, unless the interfering channels are estimated, the asymmetry between the down and uplink channels will limit the performance of the Adaptive SBM antenna system.

B. Frequency Selective Fading

In Figures 4 and 5, we provide the same results as in V.B, but with a frequency selective fading channel which can be characterized by the normalized delay spread, D (See Section II). In our simulations, we set $D = 0.5$.

Close observation of Figures 4 and 5 reveals that more than a 10 fold reduction in P_e is possible for the Adaptive SBM system with 3 adaptations compared with the conventional adaptive antenna system. Likewise, the optimum SBM system has a 100 fold decrease in P_e performance compared to a smart antenna at the BS only. Note that with and without CCI, SBM performs significantly better than Adaptive SBM. For SBM system, perfect ISI channel information are used to achieve the best transmit antenna weights. In contrast, for Adaptive SBM system, the smart transmit antenna is obtained by SMI which indirectly measures the channel state information. As a result, the system performance of SBM should always be better than that of Adaptive SBM. Therefore, it is recommended that SBM should be used for large values of D .

VI. CONCLUSIONS

The investigation presented here has demonstrated the feasibility of utilizing smart antennas at the base and mobile stations in the up and down links.

Simulation results reveal that our system reduces the average error probability by up to 100 times for channels with flat fading and no interference. When interference and ISI are included, a reduction in average error probability of up to 100 times is possible with SBM and a 10 times reduction with adaptive SBM. The reason for the significant improvement in system performance is that on transmission, the smart antenna focuses the radiated power into the best directions. The implementation of the system requires that multiple antennas be incorporated into the MS and consequently, we suggest its suitability for wireless LAN applications in combination with recent advances in antenna design.

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Fig. 1. System configuration in which smart antennas are included at both BS and MS in the down and uplinks.

Fig. 2. Average error probability performance of SBM under conditions of no interference and flat fading.

Fig. 3. Average error performance of SBM in flat fading conditions with a single dominant CCI at SIR=15dB.

Fig. 4. Average error performance of SBM in frequency selective fading channels ($D = 0.5$) with no CCI.

Fig. 5. Average error performance of SBM in frequency selective fading channels ($D = 0.5$) and CCI at SIR=15dB.

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