ADAPTIVE MIMO-DIVERSITY SELECTION WITH CLOSED-LOOP POWER CONTROL OVER WIRELESS CDMA RAYLEIGH-FADING CHANNELS

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ABSTRACT

This contribution considers adaptive (multiple input multiple output) MIMO-diversity selection at the receiver jointly with closed-loop power control (PC) to efficiently combat fading with only few transmit (Tx) and receive (Rx) antennas. Hence we avoid resorting to antenna selection among large MIMO-arrays without PC which would be otherwise required to keep complexity low. With low Doppler closed-loop PC significantly increases the transmission capacity and reduces the MIMO-array size.

1. INTRODUCTION

Use of multiple antennas at both the transmitter and the receiver (*i.e.*, MIMO) increases diversity and enables significant gains in capacity over fading channels [1],[2]. Their exploitation in a low-cost implementation is, however, still challenging.

Early work on diversity-selection combining [3] considered a pre-determined subset of the strongest paths among those resolvable in order to reduce complexity without significant loss in performance. For the same purpose, more recent works extended selection combining to switched MIMO diversity between $M_t < M_T$ and $M_r < M_R$ elements among larger antenna-arrays of M_T and M_R elements at the transmitter and the receiver, respectively, for both the single- and multiple-stream signaling approaches (see [4] and references therein).

Closed-loop PC can be viewed as an additional means to efficiently combat fading with only few Tx and Rx antennas without resorting to antenna selection among larger arrays to keep complexity low. Jointly, we fully exploit the same smaller set of constantly power-controlled MIMO diversity components by capturing their power fractions whenever detectable at the receiver which, otherwise, would contribute to interference leakage. Simulations with low Doppler suggest that adaptive MIMO-diversity selection with closed-loop PC significantly increases the transmission capacity and reduces the MIMO-array size.

2. DELAY TX DIVERSITY AND DATA MODEL

We consider a MIMO CDMA link between M_t Tx and M_r Rx antennas (applicable to either uplink or downlink). At the transmitter the interleaved binary output of the channel encoder is differentially coded as a DBPSK data sequence $b(t)$ with rate $1/T$, spread with a long spreading sequence $c(t)$ at the chip rate $1/T_c = L/T$ where L is the processing gain, then possibly amplified by a PC gain $a(t)$ (see section 3), resulting in a data stream $d(t) = a(t)b(t)c(t)$. Multiplexing the output of a space-time channel encoder instead is an option that results in multiple data streams $d_i(t) = a_i(t)b_i(t)c_i(t)$ with additional benefits from the coding gain of space-time-codes (STC) [1],[2]. We leave such an extension to a future work. We are mainly interested here in isolating the MIMO diversity gains that could be achieved by increasing the set of independent diversity components of the same data stream $d(t)$ (and therefore of each stream $d_i(t)$ with STC or multiple-stream signaling).

For simplicity, we assume a non-selective flat Rayleigh-fading channel. We also assume that no channel state information is available at the transmitter, except for a possible PC command from the receiver (see section 3). We consider the simple delay Tx diversity scheme [1] that replicates delayed versions of the data stream $d(t)$ over the Tx antennas with uniformly distributed powers, *i.e.*, $d(t-D_p)/\sqrt{M_t}$ for $p = 1, \ldots, M_t$ where the transmission delays D_p are selected multiples of T_c. Delay Tx diversity requires a single spreading code per data stream $d(t)$ for CDMA and hence avoids expensive despreading of multiple codes otherwise necessary to resolving the fade components of the Tx antennas. Indeed, delay Tx diversity transforms a flat fading channel into virtually a frequency selective channel with M_t resolvable equal-power paths that could be processed by regular CDMA receivers. Hence the following readily applies to the selective-channel case.

At time t, the M_r Rx antennas collect the following $M_r \times 1$ observation vector:

$$X(t) = \sum_{p=1}^{M_t} \mathcal{G}_p(t) d(t - \tau_0 - D_p)/\sqrt{M_t} + N(t) , \quad (1)$$

where τ_0 denotes the propagation delay between the transmitter and the receiver and $\mathcal{G}_p(t)$ the $M_r \times 1$ vector of Rayleigh-fading components[1] between the p-th Tx antenna and the M_r Rx antennas. Spacing between antennas in both

[1] They include large-scale variations such as path-loss and shadowing.

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transmitter and receiver is assumed large enough to result into $M_t \times M_r$ i.i.d. Rayleigh-fade components in $\mathcal{G}_p(t)$ for $p = 1, \ldots M_t$ without, however, exacting resolvable propagation delays across the Tx or Rx antennas. The $M_r \times 1$ noise vector $N(t)$ denotes the total contribution from all interference sources in the network. For simplicity, it is assumed Gaussian and uncorrelated both in space and time.

After appropriate transformations, we can rewrite $X(t)$ in Eq. (1) as follows:

$$\begin{aligned} X(t) &= \psi(t) \sum_{p=1}^{M_t} \varepsilon_p(t) G_p(t) \delta(t-\tau_p) \otimes c(t) b(t) + N(t) \\ &= \psi(t) H(t) \otimes c(t) b(t) + N(t) , \end{aligned} \tag{2}$$

where $\tau_p = \tau_0 + D_p$, $G_p(t) = \sqrt{M_r}\mathcal{G}_p(t)/\|\mathcal{G}_p(t)\|$ denotes $\mathcal{G}_p(t)$ normalized, $\varepsilon_p(t)^2 = \|\mathcal{G}_p(t)\|^2 / \sum_{k=1}^{M_t} \|\mathcal{G}_k(t)\|^2$ the normalized fraction of the total received power $\psi(t)^2 = a(t)^2 \sum_{p=1}^{M_t} \|\mathcal{G}_p(t)\|^2/(M_t \times M_r)$ collected from the p-th Tx antenna, and $H(t)$ is the spatio-temporal channel.

The new expression for the observation $X(t)$ in Eq. (2) is that of the wideband CDMA channel model presented in [5] from which we infer the post-correlation model (PCM) of the $M_r \times L$ despread data block for the n-th DBPSK symbol as follows [we use the notation $x_n = x(nT)$]:

$$\mathbf{Z}_n = \psi_n b_n \mathbf{G}_n \mathbf{\Upsilon}_n \mathbf{D}_n^T + \mathbf{N}_n = s_n \mathbf{H}_n + \mathbf{N}_n , \tag{3}$$

where $s_n = \psi_n b_n$ denotes the signal component, $\mathbf{H}_n = \mathbf{G}_n \mathbf{\Upsilon}_n \mathbf{D}_n^T$ is the $M_r \times L$ spatio-temporal channel normalized to $\sqrt{M_r}$, $\mathbf{G}_n = [G_{1,n}, \ldots, G_{M_t,n}]$ is the $M_r \times M_t$ spatial channel matrix, and $\mathbf{\Upsilon}_n = \text{diag}[\varepsilon_{1,n}, \ldots, \varepsilon_{M_t,n}]$. $\mathbf{D}_n = [D_{1,n}, \ldots, D_{M_t,n}]$ is the $L \times M_t$ temporal channel matrix with p-th column for $p = 1, \ldots, M_t$ given by:

$$D_{p,n} = [\rho_c(-\tau_p), \rho_c(T_c - \tau_p), \ldots, \rho_c((L-1)T_c - \tau_p)]^T, \tag{4}$$

where $\rho_c(t)$ is a truncated raised-cosine pulse [5].

For the sake of clarity, we exploit the PCM model of Eq. (3) to present the adaptive MIMO-diversity-selection combining scheme implemented by the spatio-temporal array-receiver (STAR) over the despread data block $\mathbf{Z}_n$, although combining of the data before despreading was shown to reduce complexity [5].

3. MIMO DIVERSITY WITH CLOSED-LOOP PC

The receiver STAR implements matched-channel maximum ratio combining (MRC) as follows [5]:

$$\hat{s}_n = \text{Real}\left\{ \underline{\hat{\mathbf{H}}}_n^H \underline{\mathbf{Z}}_n / M_r \right\} , \tag{5}$$

where $\underline{\mathbf{V}}_n$ denotes the $M_r \times L$ matrix $\mathbf{V}_n$ reshaped as an $M_r L \times 1$ column vector, before passing $\hat{s}_n \hat{s}_{n-1}$ to the soft Viterbi decoder after deinterleaving. The spatio-temporal channel estimate in Eq. (5) is given by [5]:

$$\hat{\mathbf{H}}_n = \hat{\mathbf{G}}_n \hat{\mathbf{\Upsilon}}_n \hat{\mathbf{D}}_n^T , \tag{6}$$

where $\hat{\mathbf{G}}_n = \left[\hat{G}_{1,n}, \ldots, \hat{G}_{\hat{P},n}\right]$, $\hat{\mathbf{\Upsilon}}_n = \text{diag}\left[\hat{\varepsilon}_{1,n}, \ldots, \hat{\varepsilon}_{\hat{P},n}\right]$, $\hat{\mathbf{D}}_n = \left[\hat{D}_{1,n}, \ldots, \hat{D}_{\hat{P},n}\right]$, and $\hat{P}$ is the number of detected paths being tracked. The synchronization module constantly monitors the vanishing and appearing paths by comparing their power estimates (*i.e.*, $\hat{\psi}_n^2 \hat{\varepsilon}_{p,n}^2$) to the same detection threshold and accordingly adjusts the size of the matrices in the spatio-temporal reconstruction of $\hat{\mathbf{H}}_n$ in Eq. (6) (please refer to [5] for more details about channel synchronization).

In essence, STAR exploits adaptive space-time processing to implement a very robust and fast adaptive MIMO-diversity selection combining scheme (well adjusted to the channel time-variations) where the selection criterion used is that paths exceed a given detection threshold (around -12 dB) in order to be selected for combining. This selection-combining feature has never been underlined as such in previous works. We exploit it jointly with closed-loop PC.

Indeed, STAR estimates the total power collected at the receiver from all the MIMO diversity paths as follows [5]:

$$\hat{\psi}_{n+1}^2 = (1-\alpha)\, \hat{\psi}_n^2 + \alpha\, \hat{s}_n^2 , \tag{7}$$

where $0 < \alpha \ll 1$ is a smoothing factor, compares $\hat{\psi}_n^2$ to the target power (usually 1) and accordingly instructs the transmitter to increase/decrease the amplification factor $a(t)$ (see section 2) at a given PC update rate.

In contrast, previous contributions focussed only on increasing the diversity gains with more Tx and/or Rx antennas to combat fading and hence proposed new implementations and/or analyzed the performance of MIMO transceivers (or STC schemes) in the absence of PC (*i.e.*, $a(t) = 1$). Hence, they recently resorted to antenna selection among large MIMO-arrays to keep complexity low (see [4] and references therein). Diversity is indeed a powerful means for combating fades. With (M_t, M_r) antennas, it achieves a diversity gain of $M_t \times M_r$ (plus a coding gain in STC schemes) that translates into the exponential order by which the BER decays asymptotically versus the SNR [2].

Power control, however, steepens the rate of descent of the BER curve faster and results in a log-normal distribution of ψ_n^2 that quickly approaches the AWGN-channel case (*i.e.*, perfect PC). Indeed, it attempts to bind ψ_n^2 around a constant power by compensating the instantaneous average power of the fade components $\sum_{p=1}^{M_t} \|\mathcal{G}_p(t)\|^2/(M_t \times M_r)$ (*i.e.*, ψ_n^2 without PC[2]) with its inverse profile in $a^2(t)$; the equalization burden being made easier with the help of a moderate diversity order only. Hence, PC trades the high diversity order otherwise required to reduce the large power variations at the receiver for distributed yet overall weaker power variations at both the transmitter and the receiver, each with few antennas only (see section 4).

Stege *et al* [6] recently studied the impact of PC on a $(2,1)$-MIMO link. They did not, however, recognize the

[2]It also approaches the AWGN-channel case when $M_t \times M_r \to \infty$.

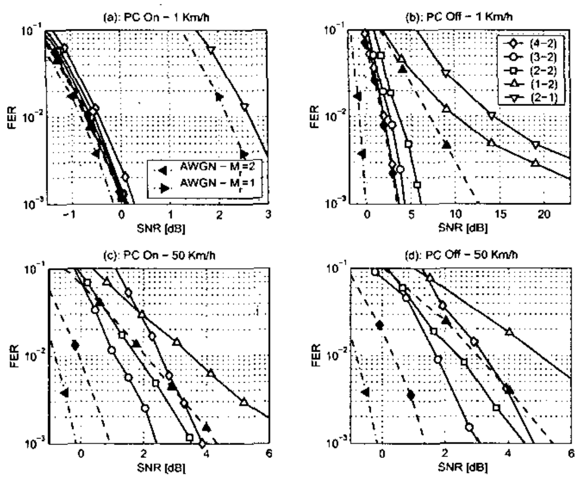

Figure 1: FER vs. SNR in AWGN-channel (semi-dashed), perfect ID (dashed), and imperfect ID (solid) cases.

(M_t, M_r)	PC On			PC Off		
	$\frac{E_b}{N_0}$	C	$\mathcal{E}$	$\frac{E_b}{N_0}$	C	$\mathcal{E}$
Speed of 1 Km/h (*i.e.*, $f_D \simeq 2$ Hz)						
Active Channel Synchronization						
(2,1)	2.6	2	0.031	14.3	0	0
(1,2)	-0.7	5	0.039	10.1	0	0
(2,2)	-0.6	11	0.086	3.6	3	0.023
(3,2)	-0.5	13	0.102	2.5	4	0.031
(4,2)	-0.4	12	0.09	1.9	6	0.047
Perfect Channel Synchronization						
(1,2)	-0.7	6	0.047	7.1	0	0
(4,2)	-0.7	13	0.102	1.7	6	0.047
Speed of 50 Km/h (*i.e.*, $f_D \simeq 90$ Hz)						
Active Channel Synchronization						
(1,2)	3.5	1	0.008	5.0	1	0.008
(2,2)	1.8	5	0.039	2.4	4	0.031
(3,2)	1.1	7	0.055	1.8	6	0.047
(4,2)	2.5	4	0.031	3.2	3	0.023
Perfect Channel Synchronization						
(1,2)	2.1	3	0.023	2.9	3	0.023
(4,2)	-0.1	10	0.078	0.4	10	0.078

Table 1: Required SNR value $\frac{E_b}{N_0}$ in dB, resulting capacity C in number of users per cell and spectrum efficiency per Rx antenna $\mathcal{E}$ in b/s/Hz/M_r for a target FER of 10^{-2}.

above advantages of combining low-complexity closed-loop PC and low-order diversity without resorting to antenna selection. Additionally, they implemented a common PC command for the two streams generated by a space-time block code [6]. As explained earlier in section 2, *ad hoc* extension of this work to the multiple-stream signaling approach implements independent PC $a_i(t)$ for each data steam $d_i(t)$. Simulations not shown for lack of space confirm that independent PC of two Tx streams with half the feedback rate outperforms common PC. Other interesting findings and simulation results are reported and analyzed next.

4. SIMULATION RESULTS

We consider data transmissions of 19.2 Kb/s over a CDMA system operating at a carrier of 1.9 GHz with a chip rate $1/T_c = 1.2288$ Mcps. Information bits are convolutionally-coded at rate-1/2 with constraint length of 9, grouped into 10-ms frames of 384 symbols after block interleaving with size 24×16, modulated DBPSK at $1/T = 38.4$ Kb/s, then upconverted at $1/T_c$ with a spreading factor $L = 32$. When active, closed-loop PC instructs ± 0.25 dB increments at 1.6 KHz with a Tx delay of 0.625 ms and a 5% feedback BER.

Simulation results suggest the following new findings:

• When PC is active, the FER curves in Fig. 1 (as well as the BER curves before/after FEC not shown for lack of space) indicate that a MIMO-array nominally (*i.e.*, perfect ID) approaches the AWGN-channel case (*i.e.*, perfect PC) much more rapidly with increasing diversity, suggesting that lower order of diversity may suffice. This allows reducing the number of antennas.

• In practical systems, where channel identification is active, there is a diversity limit beyond which no gain is achieved. On one hand, increased diversity increases the slope of descent of the BER at the link level. At the system-level, it reduces the variances of both the Rx and Tx (when PC is active) powers as well as the outage probability due to reduced variations of the incell and outcell interferences. As shown in Tab. 2, when PC is active in the low Doppler case, σ_{a^2} reduces from 3.2 with (1,2) antennas, to 0.7 with (2,2) antennas, to 0.4 with (3,2) antennas resulting in capacity gains[3] in Tab. 1 of 120 (5 to 11) and 20% (11 to 13), respectively. On the other hand, increased diversity increases the amount of uncaptured power that leaks to interference through increasingly weaker and less detectable paths due to power "fractioning" among Tx antennas, more so when PC is active, because PC further shrinks the range of multipath amplitude variations, or at high Doppler, because tracking becomes more unstable (see P_{M_t} and P_{M_t-1} in Tab. 2). Overall, best tradeoff in capacity versus diversity in Tab. 1 is with (3,2) antennas in all but the low Doppler case without PC, where losses due to "fractioning" are still negligible compared to the diversity gains and beyond the reversal point (to be reached with larger MIMO arrays).

• A good indicator of the achievable capacity in Tab. 1 is the level of received-power std σ_{ψ^2} in Tab. 2 as long as diversity gains dominate losses due to "fractioning". Without PC, ψ^2 is χ-distributed with an std $\sigma_{\psi^2} = 1/\sqrt{M_t M_r}$ that reflects

[3]For qualitative illustration of the results, we used the system-level capacity evaluation tool in [5] with an uplink setup for simplicity.

(M_t, M_r)	PC On									PC Off				
	P_{M_t} [%]	P_{M_t-1} [%]	β^2 [dB]	$\overline{\psi^2}$	σ_{ψ^2}	$\overline{\psi^2_{\text{dB}}}$	$\sigma_{\psi^2_{\text{dB}}}$	$\overline{a^2}$	σ_{a^2}	P_{M_t} [%]	P_{M_t-1} [%]	β^2 [dB]	$\overline{\psi^2}$	σ_{ψ^2}
Speed of 1 Km/h (*i.e.*, $f_D \simeq 1.8$ Hz) - Delay Drift of 0.046 ppm														
(2,1)	73.3	26.6	-23.8	0.8	0.2	-1.0	0.9	1.5	3.5	95.8	3.6	-19.6	1.0	0.7
(1,2)	100	0	-28.0	0.8	0.1	-1.1	0.7	1.4	3.2	100	0	-20.4	1.0	0.7
(2,2)	98.4	1.5	-28.8	0.8	0.1	-1.1	0.7	1.0	0.7	99.6	0.3	-30.5	1.0	0.5
(3,2)	92.2	7.6	-28.2	0.8	0.1	-1.1	0.7	0.9	0.4	96.7	3.2	-29.8	1.0	0.4
(4,2)	80.6	18.7	-27.4	0.8	0.1	-1.1	0.7	0.9	0.4	88.3	11.3	-29.4	1.0	0.3
Speed of 50 Km/h (*i.e.*, $f_D \simeq 90$ Hz) - Delay Drift of 2 ppm														
(1,2)	100	0	-14.2	1.0	0.9	-0.9	3.3	1.1	0.6	100	0	-14.4	1.0	0.7
(2,2)	100	0	-17.1	1.0	0.5	-0.7	2.2	1.0	0.4	100	0	-17.8	1.0	0.5
(3,2)	99.8	0.1	-16.9	1.0	0.4	-0.5	1.8	1.0	0.3	99.8	0.1	-17.8	1.0	0.4
(4,2)	0	99.8	-12.5	1.3	0.5	0.7	1.7	1.3	0.4	0	99.9	-13.0	1.0	0.3

Table 2: ID & PC statistics at $\frac{E_b}{N_0}$ values in Tab. 1: $P_p = \text{Prob}[\hat{P} = p]$, $\beta^2 = E[\|\hat{\underline{H}}_n - \underline{H}_n\|^2]/(M_r L)$, $\psi^2_{\text{dB}} = 10\ \log_{10}(\psi^2)$.

well the saturation in performance gains due to increased diversity. Although ψ^2 has log-normal distribution with PC (*i.e.*, ψ^2_{dB} is Gaussian, see mean and variance in Tab. 2), capacities achieved with or without PC are about the same (6-7, see Tab. 1) at similar std levels (0.3-0.4, see Tab. 2).

• PC reduces the required order of diversity and the size of the MIMO array. In Tab. 2, while measured σ_{ψ^2} decays in $1/\sqrt{M_t M_r}$ without PC, it reduces fast to around 0.1 with PC at low Doppler. Based on these std observations, we expect to obtain the same capacity with PC and (3,2) antennas as would be obtained without PC with (30,2) antennas. Practical limits on path tracking and "fractioning" may force one to reduce this array-size using Tx-antenna selection. Without antenna-selection feedback, however, closed-loop PC enables a small-size MIMO-array of (3,2) antennas to serve 13 users/cell versus 4 without PC, resulting in a capacity gain of 225% (see Tab. 1).

• At high Doppler, closed-loop PC is too slow to achieve the same gains obtained at low Doppler (see Fig. 1 and Tabs. 1 and 2).

• With closed-loop PC over non-selective channels, results at either Doppler rate suggest that $M_t = 3$ Tx antennas per data stream provides the best performance. Over selective channels with 3 equal-power paths, only $M_t = 1$ Tx antenna per data stream is therefore needed. In a multiple-stream signaling approach, more Tx antennas could be necessary to create a sufficient number of distinct channels for spatial multiplexing.

• These results further motivate the study of MIMO capacity with PC where the decision variable has a log-Normal instead of a χ-distribution.

5. CONCLUSIONS

We implemented adaptive MIMO-diversity selection at the receiver jointly with closed-loop PC to efficiently combat fading with only few Tx and Rx antennas. Hence we avoided resorting to antenna selection among large MIMO-arrays without PC which would have been otherwise required to keep complexity low. With low Doppler, closed-loop PC significantly increases the transmission capacity and reduces the size of the required MIMO-array.

Ongoing work assesses STC with MIMO-diversity selection and closed-loop PC. We will report the results in a future publication.

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