SPATIO-TEMPORAL ARRAY PROCESSING FOR FDD/CDMA/SDMA DOWNLINK TRANSMISSION

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ABSTRACT

We address the problem of performing optimum spatio-temporal processing when using adaptive antenna arrays at base stations for multiuser downlink transmission in DS-CDMA systems, using periodic spreading sequences and assuming partial knowledge of the channel parameters of all users. This assumption typically holds in frequency-division duplex (FDD) based mobile communication systems. We consider the SDMA strategy for using antenna arrays to gain system capacity. The channel is assumed to comprise specular multipath, and a per-path argument is pursued to design FIR transmission filters at the base station in order to maximize the signal-to-interference-plus-noise ratio (SINR) at the mobile receivers. Joint optimization of the transmitter and receiver is considered. The per path decorrelating pre-filter is introduced, and it is shown that due to the large number of degrees of freedom available because of the large processing gain (inherent oversampling with respect to the symbol rate in CDMA) and possible multiple antennas/oversampling, the downlink performance can be greatly improved in the FDD problem.

I. INTRODUCTION

The use of adaptive antenna arrays at the base station can increase the capacity of a mobile radio network allowing an increase in the number of users. In the downlink however, the possibility of spatial diversity reception by Multiple Antennas (MA) is limited due to complexity and space limitations. Although transmit antenna diversity does not give the same gains as the receive antenna diversity, pre-processing of some sort at the transmitter, based upon the knowledge of downlink channel parameters can result in improved performance and simplified, low complexity receivers for the mobile stations. The amount and nature of this *a priori* knowledge of the channels depends on the system architecture. In time-division duplex (TDD) based systems the uplink and the downlink channels can be considered to be practically the same (reciprocity), assuming the mobile velocity low enough and the receiver and transmitter appropriately calibrated. Under these circumstances, since the channel is known (or estimated) from the uplink, efficient spatio-temporal processing can be performed at the base station during transmission as well as during reception. We have presented optimal solutions for downlink spatio-temporal processing in the TDD setup for both TDMA and CDMA based systems in [2] and [1] respectively. Contrary to TDD, in the FDD mode, the base station has no direct knowledge of the downlink channel, since it cannot be directly observed and therefore estimated. A solution to this problem consists of providing the base station with feedback from the mobile station about the downlink channel at the cost of reduced spectral efficiency. On the other hand, if such feedback is not provided, the downlink channel characterization can only be based on the estimates of parameters related to the uplink channel, which are relatively frequency independent and whose changing rate is slow with respect to the frame duration. We thus have *partial* channel state information.

The parameters of interest are typically the angle of arrival/departure, the delay, the magnitude and the phase for each path in the multipath propagation. We assume the knowledge of the covariance matrix of the channel impulse response averaged over the path phases and amplitudes (the quantities unknown at the base-station). For the purpose of transmit filter optimization, the specular nature of the paths, and the randomness of the path phases leads to the modeling of the multipath components of a certain mobile user as equivalent to several correlated users, each propagating through a single path. Assuming the individual paths are spatio-temporally resolvable, the averaged covariance matrix can still be built. Delays for paths which are resolvable in space only can be adjusted at transmission to become temporally distinguishable at the receiver.

In the light of the above arguments, we consider here the problem of performing optimal spatio-temporal processing when a FDD/DS-CDMA system is adopted. In typical downlink transmission (e.g., IS-95), the multiuser channel is short (a few chips long), synchronous and the users are assigned orthogonal Walsh-Hadamard sequences. The orthogonality is however destroyed by multipath. When the downlink channels are known, as in the TDD mode, the orthogonality of the codes can however be restored through proper pre-filtering at the base station, which corresponds to Zero-Forcing (ZF) the Inter-User Interference (IUI). When FDD is considered, similar reasoning can be applied, but due to the lack of knowledge of the path phases, the effective number of users is actually given by the sum of all the paths of all the users. If only the spreading (temporal) dimension is exploited, then, in order to restore the orthogonality we need number the total number of paths to be less than the spreading factor. This in turn results in a low loading fraction (number of users over the spreading factor). The loading fraction can be increased by using spatial and other multichannel information in conjunction with the temporal (spreading factor) dimension.

Theoretically, even a number of interfering users larger than the spreading factor may be located in the same cell (interference coming from other cells is neglected, except for the users in soft-handover mode). So zero IUI can be achieved as long as the total number of paths does not exceed the total number of subchannels. The latter can be quite significant is MA and OS is employed. In our treatment, the emphasis is on simple mobile receiver structures (e.g., a correlator or a RAKE receiver) while the optimization criterion consists of maximizing the minimum signal-to-interference and noise ratio (SINR) among the considered users, subject to a total transmit power constraint. Each one of the d mobile receivers is assumed to have one antenna. We introduce the pre-combining like decorrelator filter to decouple the multipath signals [4]. The problem then settles down to the power assignment to signals through these pre-filters, in order to maximize the SINR at the mobile station. We show that the optimal power assignment turns out to correspond to *selection*

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diversity, an approach that has also been followed in [3] based on heuristic reasoning.

II. THE FDD FRAMEWORK AND RECIPROCITY

We consider a specular path channel model that consists of Q_i multipath components for the ith user. The ith's user qth multipath channel component as seen form the base station can be modeled in the continous-time domain as follows

$$\boldsymbol{h}_{iq}^T(\tau,\,t)=\alpha_{iq}(t)\mathbf{a}^T(\theta_{iq})\delta(\tau-\tau_{iq}) \tag{1}$$

where τ_{iq}, θ_{iq}, and $\alpha_{iq}(t)$ denote the delay, the angle and the fading attenuation associated to the qth path of the ith user, respectively, and $\mathbf{a}(\theta)$ represents the array response vector. Assuming a similar multipath channel model for the uplink, the parameters which can be assumed approximately constant between the uplink and the downlink channels are the angles, the delays and the variances of the amplitudes. Since the difference in phase between up- and downlink is random it can be assumed uniformly distributed, whereas the magnitudes for both links are also random but can be assumed to have the same variance. The variances of the path amplitudes can be estimated by non-coherent averaging over a certain time interval. The angles can be estimated if the array manifold at the downlink carrier frequency is known. For particular array geometries and relatively small uplink–downlink frequency shifts, the array response can be transposed from the uplink to the corresponding response in the downlink via a linear transformation [5] without requiring explicit angle estimation. Another approach consists of performing a *beamspace* transformation (namely a spatial DFT) to estimate the beams in which the signal energy is located [6]. The downlink transmission then occurs through the same beams as the uplink reception.

A. The Pathwise Channel-Receiver Cascade

In order to reason in a pathwise manner, we assume that each receiver processes symbol rate data coming from the outputs of a bank of receive (RX) correlators. The number of correlators equals the number of paths for the intended user. For the pulse shaping matched filter at receiver we denote $w_i(\tau)=\sum_{l=0}^{m_c-1}c_{i,\,l}^*\psi(\tau-lT_c)$ as the cascade of the chip-pulse shape matched filter, $\psi(\tau)$, and the ith user correlator $c_i^*(-\tau)=\sum_{l=0}^{m_c-1}c_{i,\,l}^*\delta(\tau-lT_c)$, where T_c is the chip period and m_c the spreading factor. The superscripts *, ^{T} and ^{H} denote complex conjugate, transpose and Hermitian transpose respectively. We assume that $w(\tau)$ is a FIR filter with time duration approximately equal to L_wT. T is the symbol period. Then the following discrete-time channel model at the symbol rate $1/T$, where $T=m_cT_c$, can be described

$$\begin{aligned}\boldsymbol{g}_{iq}^T(k,\,n)&=\alpha_{iq}(n)[\mathbf{a}(\theta_{iq})\otimes\mathbf{w}_{iq}(k)]^T\\ \boldsymbol{G}_{iq}^t(n)&=\alpha_{iq}(n)[\mathbf{a}(\theta_{iq})\otimes\mathbf{W}_i(\tau_{iq})]^t\end{aligned}\quad, \tag{2}$$

where $\otimes$ denotes the Kronecker product and the superscript ^{t} denotes transposition of the blocks in a block matrix, $\mathbf{w}_{iq}(k)=\mathbf{w}_i(t_0+kT-\tau_{iq})$,

$$\mathbf{w}_{iq}(k)=[w(t_0+kT-\tau_q)\;\ldots\;w(t_0+T(k+\frac{m_c-1}{m_c})-\tau_q)]^T$$

and $\mathbf{W}_i(\tau_{iq})=[\mathbf{w}_{iq}(L_w-1)\ldots\mathbf{w}_{iq}(0)]$ [1]. We could also account for OS w.r.t. the chip rate by replacing m_c with m_cm_o in the expression above. We use the notation $\boldsymbol{V}_{iq}=\mathbf{a}(\theta_{iq})\otimes\boldsymbol{W}_i(\tau_{iq})$ in the sequel. One may notice that the $\boldsymbol{V}_{iq}$'s can be built based upon the estimates of the path angles and delays, and the knowledge of the receiver correlator.

We also introduce the spatio-temporal channel covariance matrix associated with $\boldsymbol{G}_{iq}(n)$ averaged over the ith user's qth path phase, given by

$$\boldsymbol{R}_{iq}^{(L)}=\mathrm{E}[\mathcal{T}_L(\boldsymbol{G}_{iq}(n))\mathcal{T}_L^H(\boldsymbol{G}_{iq}(n))]=\sigma_{iq}^2\mathcal{T}_L(\boldsymbol{V}_{iq})\mathcal{T}_L^H(\boldsymbol{V}_{iq}) \tag{3}$$

where $\sigma_{iq}^2=\mathrm{E}[|\alpha_{iq}(n)|^2]$, , $\mathrm{E}[\cdot]$ denotes the expectation operator, and $\mathcal{T}_M(\boldsymbol{A})$ is in general a block Toeplitz matrix with M block rows and $[\boldsymbol{A}\;\;\mathbf{0}_{p\times s(M-1)}]$ as first block row, and $\boldsymbol{A}$ is a matrix with $p\times s$ block entries.

We shall observe that due to the assumption on the receiver structure the delays τ_{iq}'s denote the overall delay between the transmitter antenna(s) and the qth correlator output of the ith receiver. In general, a cost function for the transmit filter optimization should be formulated so as to optimize also each correlator synchronization time, i.e, to optimize the τ_{iq} by properly advancing or retarding the receiver correlator with respect to the base station transmitter clock. For the purpose of the overall channel description and the filter optimization algorithm, we shall assume the delays τ_{iq}'s to be fixed and known at the transmitter.

III. SIGNAL MODEL

Assuming the channels $\boldsymbol{h}_{iq}$ time-invariant for the observation time, the ith user discrete-time received signal, for $i=1\;\ldots,\,d$, is

$$y_{iq}(k)=\mathbf{c}_i^H\mathbf{H}_{iq}^T(\zeta)\sum_{j=1}^{d}\sum_{l=1}^{Q_j}\mathbf{F}_{jl}(\zeta)a_j(k)+v_{iq}(k) \tag{4}$$

where the $a_j(k)$ are the transmitted symbols intended for the jth user, ζ^{-1} is the unit sample delay operator (i.e., $\zeta^{-1}y_i(k)=y_i(k-1)$), $\mathbf{H}_{iq}^T(z)$ is the channel transfer function between the base station and the qth path of the ith user channel, $\mathbf{c}_i^H$ is the ith user correlator, $\mathbf{F}_{jl}(z)=\mathbf{F}'_{jl}(z)\mathbf{c}_j$ is the spatio-temporal filter for the transmitted symbols, accounting for both the actual transmit filter $\mathbf{F}'_{jl}(z)$ to be optimized and the spreading code, $\mathbf{c}_j$, for the jth user, and $v_{iq}(k)$ is the additive noise associated to the qth path of the ith user.

Since we have m_c chips per symbol period, each transmission filter $\mathbf{F}_{iq}(z)$ will perform sampling at least at the chip rate, i.e., it will be at least a $m_c\times 1$ column vector. If no additional OS or MA are provided, the optimization problem for all the $\mathbf{F}_{iq}(z)$'s reduces to one of spreading code optimization at the transmitter in the presence of multiuser multipath channels. Moreover, in general $\mathbf{F}_{jl}(z)$ will be a $m\times 1$ column vector, with $m=m_cm_am_o$, where m_a is the number of MA.

We denote $\mathbf{G}_{iq}^T(\zeta)=\mathbf{c}_i^H\mathbf{H}_{iq}^T(\zeta)$ the overall channel associated with the ith user's qth path as seen from the base station. Note that since the receiver is assumed to sample at the chip rate, $\mathbf{H}_{iq}^T(z)$ is a $m_c\times m$ matrix, $\mathbf{c}_i^H$ is a $1\times m_c$ row vector, so that $\mathbf{G}_{iq}^T(z)$ is a $1\times m$ row vector, and $\mathbf{F}_{jl}(z)$ is a $m\times 1$ column vector. $\mathbf{G}_{iq}(z)$ is the $m\times 1$ qth single path channel in the uplink from the ith user to the m base station channels.

A. Burst Processing Time Domain Signal Model

Consider the I/O transmission chain (see fig. 1) associated to the qth path component of the ith user regardless of the contributions intended for the other paths and other users. The channel $\boldsymbol{g}_{iq}^T(t)=\boldsymbol{c}_i^H\boldsymbol{H}_{iq}^T(t)$ and the transmission filter $\boldsymbol{f}_{iq}(t)=\boldsymbol{F}'_{iq}(t)\boldsymbol{c}_i$ are assumed to be FIR filters with duration $N_{iq}T$ and LT respectively (approximately). In discrete-time representation

[1] The length of L_w may be different for different users, although we shall neglect this issue in this paper

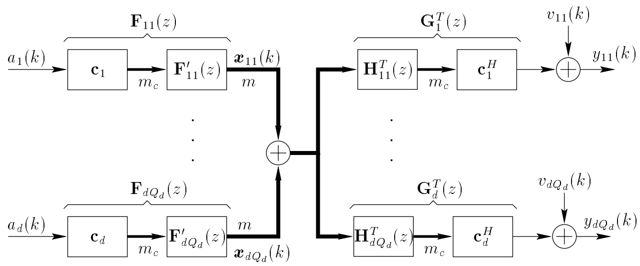

Figure 1. Transmission filters and single-path channels for d users

we have

$$
\begin{aligned}
\boldsymbol{x}_{iq}(k) &= \textstyle\sum_{l=0}^{L-1} \boldsymbol{f}_{iq}(l)a_i(k-l) = \boldsymbol{F}_{iq}A_{i,\,L}(k)\\
y_{iq}(k) &= \textstyle\sum_{n=0}^{L_w-1} \boldsymbol{g}_{iq}^T(n)\boldsymbol{x}_{iq}(k-n) + v_{iq}(k)\\
&= \boldsymbol{G}_{iq}^t \boldsymbol{X}_{iq,\,L_w}(k) + v_{iq}(k)\\
\boldsymbol{G}_{iq}^t &= [\boldsymbol{g}_{iq}^T(L_w-1)\;\ldots\;\boldsymbol{g}_{iq}^T(0)]\\
\boldsymbol{F}_{iq} &= [\boldsymbol{f}_{iq}(L-1)\;\ldots\;\boldsymbol{f}_{iq}(0)]\\
\boldsymbol{X}_{iq,\,L_w}(k) &= [\boldsymbol{x}_{iq}^H(k-L_w+1)\;\ldots\;\boldsymbol{x}_{iq}^H(k)]^H\\
A_{i,\,L}(k) &= [a_i^H(k-L+1)\;\ldots\;a_i^H(k)]^H
\end{aligned}
\tag{5}
$$

If we accumulate M consecutive symbol periods

$$Y_{iq,\,M}(k) = \mathcal{T}_M(\boldsymbol{G}_{iq}^t)\mathcal{T}_{M+L_w-1}(\boldsymbol{F}_{iq})A_{i,\,M+L_w+L-2}(k) + V_{iq,\,M}(k)$$

where, $Y_{iq,\,M}(k) = [y_{iq}^H(k-M+1)\;\ldots\;y_{iq}^H(k)]^H$ and likewise for $V_{iq,\,M}(k)$.
Then, introducing also the contributions of all the other paths and all the other users, for the ith user's qth path component we have

$$
\begin{aligned}
Y_{iq,\,M}(k) = &\sum_{j=1}^{d}\sum_{l=1}^{Q_l} \mathcal{T}_M(\boldsymbol{G}_{iq}^t)\mathcal{T}_{M+L_w-1}(\boldsymbol{F}_{jl})A_{j,\,M+L_w+L-2}(k)\\
&+ V_{iq,\,M}(k)
\end{aligned}
\tag{6}
$$

We observe that $\boldsymbol{G}_{iq} = \alpha_{iq}\boldsymbol{V}_{iq}$.

IV. TRANSMIT FILTER OPTIMIZATION

A major issue in the transmit filter design consists of the power assignment optimization among different paths and different users' pre-filters. In order to find an analytical solution we decople the power assignment optimization problem by considering first the optimization of a set of unit norm transmit filters $\boldsymbol{U}_{iq}$ such that $\boldsymbol{F}_{iq} = \sqrt{p_{iq}}\boldsymbol{U}_{iq}$. Then, once the $\boldsymbol{U}_{iq}$ have been determined the powers p_{iq} will be properly assigned subject to a maximum total transmit power constraint.
For the sake of simplicity in the following developments, we introduce $\boldsymbol{F}_{iq}^t = [\boldsymbol{f}_{iq}^T(L-1)\;\ldots\;\boldsymbol{f}_{iq}^T(0)]$ and the respective unit norm filter $\boldsymbol{U}_{iq}^t$. We also remark that for the convolution of any $\boldsymbol{F}$ and $\boldsymbol{G}^t$, the relation

$$\boldsymbol{F}^t\mathcal{T}_L(\boldsymbol{G}) = \boldsymbol{G}^t\mathcal{T}_N(\boldsymbol{F})$$

holds, where L and N are the durations in symbol periods of $\boldsymbol{F}$ and $\boldsymbol{G}$ respectively.

A. The Per-Path Pre-Decorrelator

A solution for the design of the filters $\boldsymbol{U}_{iq}$'s consists of pre-decorrelating the paths of the user of interst while canceling out the IUI, namely the contributions due to other user's paths. In order to achieve perfect IUI cancellation and IP pre-decorrelation, we shall consider the following set of ZF constraints

$$\max_{\|\boldsymbol{U}_{iq}^t\|_2=1} \|\boldsymbol{U}_{iq}^t\mathcal{T}_L(\boldsymbol{V}_{iq})\|_2^2 \qquad \text{s.t.} \qquad \boldsymbol{U}_{iq}^t\mathcal{T}_L(\boldsymbol{V}_{jl}) = \boldsymbol{0} \tag{7}$$

for any $i,\ j = 1,\ldots,d,\ l = 1,\ldots,Q_j,\ q = 1,\ldots,Q_i$ s.t. $q \neq l$ when $i = j$. Define $\boldsymbol{B}_{iq}$ as $[\mathcal{T}_L(\boldsymbol{V}_{jl})]$ as the matrix accounting for all the paths of all the users but the qth path of the ith user, and $\boldsymbol{A}_{iq} = \mathcal{T}_L(\boldsymbol{V}_{iq})\mathcal{T}_L(\boldsymbol{V}_{iq})$. Then, the solution to problem (7) is $\boldsymbol{U}_{iq}^{tH} = V_{\max}(\mathbf{P}_{B_{iq}}^{\perp}\boldsymbol{A}_{iq}\mathbf{P}_{B_{iq}}^{\perp})$, where $\mathbf{P}_{B_{iq}}^{\perp}$ is the projection matrix onto the null space of the column space of $\boldsymbol{B}_{iq}$. In order for a non-trivial solution to problem (7) to exist we need the length L of the transmit filters $\boldsymbol{U}_{iq}^t$ to be

$$L \geq \frac{(L_w-1)(Q-1)}{m_{\text{eff}}-(Q-1)} \tag{8}$$

where $Q = \sum_i Q_i$ and m_{eff} is defined as the rank of $\boldsymbol{V}_Q = [\boldsymbol{V}_{i1}\;\ldots\;\boldsymbol{V}_{dQ_d}]$. Note that $m_{\text{eff}} = \min\{m,\ L_wQ,\ (L_w-1)Q+\Delta\}$ where $\Delta = \text{rank}([\boldsymbol{v}_{11}(L_w-1)\;\ldots\;\boldsymbol{v}_{dQ_d}(L_w-1)])$. The constraints present in the optimization problem (7) lead to perfect IUI cancellation along with an interpath pre-decorrelation for the user of interest. This is obtained at the expense of increased ISI at the receiver. In order to consider the ISI as well as the IUI rejection in the optimization problem, we rely on the ZF pre-equalization conditions.

B. IP Pre-Decorrelation, ZF Conditions for IUI and ISI Cancellation

In order to ensure ZF conditions for IUI and ISI for the ith user's qth path the set of constraints to be considered is

$$\boldsymbol{U}_{iq}^t\mathcal{T}_L(\boldsymbol{V}_Q) = [0\;\ldots\;0\;\ldots\;\overbrace{|\,0\ldots0\;\alpha\;0\ldots0\,|}^{i\text{th user's } q\text{th path}}\;\ldots\;0\;\ldots\;0] \tag{9}$$

where $\mathcal{T}_L(\boldsymbol{V}_Q) = [\mathcal{T}_L(\boldsymbol{V}_{11})\ldots\mathcal{T}_L(\boldsymbol{V}_{dQ_d})]$, and $\alpha \neq 0$ is an arbitrary constant to be fixed in order to satisfy the constraint on the norm of $\boldsymbol{U}_{iq}^t$. Assuming $m > Q$ and $\mathcal{T}_L(\boldsymbol{V}_Q)$ to be full column rank, to be able to satisfy all the constraints (9) we need to choose the length of each filter $\boldsymbol{U}_{iq}^t$, L, such that the system of equations 9 is exactly or underdetermined. Hence

$$L \geq \underline{L} = \left\lceil \frac{(L_w-1)Q-1}{m_{\text{eff}}-Q} \right\rceil \tag{10}$$

Then assuming $L \geq \underline{L}$ we can consider two limiting set of constraints:

- IUI rejection, no ISI rejection, as in section IV.A.
- both IUI and ISI rejection: in this case, the set of constraints is (9), i.e., we have $L_w + L - 1$ more constraints.

In the absence of IUI (equal to zero due to ZF), the SNR at the output of each correlator is proportional to the energy in the prefilter-single path channel cascade. Then, the SNR decreases if all the energy is constrained in one tap. Hence if no ISI rejection is provided the highest SNR will be achieved, for a specified L, due to the larger number of degrees of freedom. However, in that case, once the strongest path has been selected, the ith receiver needs to equalize a delay spread of up to $L_w + L - 1$ symbol periods, corresponding to the whole delay spread due to the convolution between the single path channel and the selected transmission filter. We may prefer that the introduction of the prefilter does not increase the delay spread, or we may want to limit the delay spread seen by the mobile to limit the complexity for the equalization task in the mobile. In those cases additional constraints in order to obtain at least partial ISI rejection, i.e., limited delay spread, can be added, leading to intermediate solutions between the previous two limiting cases. In general to have complete IUI and partial ISI rejection we add $(L_w + L - 1) - L_{\text{ISI}}$ constraints (coefficients of the prefilter-channel cascade being zero), with $1 \leq L_{\text{ISI}} \leq (L_w + L - 1)$, where L_{ISI} corresponds to the residual delay spread, i.e., residual

ISI. This optimization problem has to be carried out for all possible positions of the nonzero part of length $L_{\rm ISI}$ of the prefilter-channel cascade, and the best position should be chosen. Finally, note that as L increases the SNR increases as well. So, we shall choose the actual length of the transmission filters L according to a trade-off between performance and transmitter complexity.

One might think that by transmitting only through the strongest path per each user the amount of ISI at the receiver is negligible. However, although L_w is in practice very small (2, 3 symbol periods), for high loading fractions, i.e., for a large number of paths, the required L can become relatively large, in order to achieve the above ZF IUI conditions, which in turn results in significant ISI.

Finally, one may note that ZF-pre-decorrelating here corresponds to the design of a bi-orthogonal perfect-reconstruction transmultiplexer in which the $\boldsymbol{F}_{iq}$'s and $\boldsymbol{G}_{iq}$'s are synthesis and analysis filter banks respectively.

C. RX Correlator Positioning / Delay Optimization

The ZF problem in (9) supposes that the delays τ_{iq}, $\forall i = \{1\cdots d\}$, $q = \{1\cdots Q_i\}$, for all users are known at the transmitter. This implies that the correlator at the receiver is also supposed to be located at a known fixed position in time. It is for this overall delay, τ_{iq}, and all others, τ_{jr}, $\forall j = \{1\cdots d\}$, $r = \{1\cdots Q_j\}$, and $r \neq q$ when $j = i$, that the pre-decorrelating conditions are satisfied. In the optimization scheme, due to the presence of the RX correlators in the overall channel, it is taken for granted that the assumed delay would lead to the maximization of the SNR at the output. It would suffice then, that the correlator, in an independent operation mode, searches for the delay by sweeping over the field of interest of the assumed delays. However, since the ZF conditions are being satisfied for a set of discrete delays, the IUI and IPI will have its contributions at all intermediate positions. Furthermore, this may not necessarily be the global SNR maximization delay for the RX correlator. In order to maximize the SNR, let us introduce $\boldsymbol{U}_{iqn}$, as the ZF prefilter for the qth path of the ith user with the correlator placed at a delay of n positions (e.g., chips periods) w.r.t. an arbitrary initial position. This can be seen as an n-shift of the elements in the columns of $\mathcal{T}_L(\boldsymbol{V}_{iq})$, (i.e., the first vector co-efficient now contains n more zeros). The optimization for $\boldsymbol{U}_{iqn}$ is still done at the symbol rate for the new $\mathcal{T}_L(\boldsymbol{V}_{iq})$. The optimization problem still stays the same as (9) and the optimal n is selected to maximize the output SNR: $\max_n \mathrm{SNR}_n$. The RX correlator can still search for the delay. It can be seen, however, that the optimal delay selection is a coupled problem. Its choice, therefore, influences and is influenced by the design of other users' prefilters.

An alternative approach for SNR maximization w.r.t. the correlator delay consists of searching over several transmit filters, $\boldsymbol{U}_{iqn}$ for the one that maximizes the SNR, considering that the RX correlator is fixed. Then, for the optimization problem of section IV.A., assuming $m_a = m_o = 1$,

$$\boldsymbol{U}_{iqn}^{'t} = \begin{bmatrix} \mathbf{0}_{1\times n} & \boldsymbol{U}_{iqn}^{t} & \mathbf{0}_{1\times(m_c-n)} \end{bmatrix},$$

and the $n \in \{0\cdots m_c - 1\}$, in the case where chip-level resolution is sought in the delay optimization. The number of zeros is fixed, and the solution to (7) is still

$$\boldsymbol{U}_{iqn}^{tH} = V_{\max}(\mathbf{P}^{\perp}_{\overline{B}_{iq}} \overline{\boldsymbol{A}}_{iq} \mathbf{P}^{\perp}_{\overline{B}_{iq}}).$$

The matrix $\overline{\boldsymbol{B}}_{iq}$ is built from $\boldsymbol{B}_{iq} = [\mathcal{T}_L(\boldsymbol{V}_{jl})]$ (section IV.A) by appending n zero rows at the top and $(m_c - n)$ zero rows at the bottom. Besides we have $\overline{\boldsymbol{A}}_{iq} = \overline{\mathcal{T}}(\boldsymbol{V}_{iq})\overline{\mathcal{T}}^H(\boldsymbol{V}_{iq})$, where $\overline{\mathcal{T}}(\boldsymbol{V}_{iq})$ is built from $\mathcal{T}_L(\boldsymbol{V}_{iq})$ in a similar fashion as $\overline{\boldsymbol{B}}_{iq}$. We have assumed in the above that the TX filter $\boldsymbol{U}_{iqn}^{t}$ is an integer number of symbols long, since it settles nicely in our framework (see sec. IV.A.). This, however, is not necessary, and the filter length can, for example, be defined in number of chips. The two approaches discussed above lead to similar kinds of delay optimization. Both problems are coupled leading to joint oprimization for all users. Upon solving the joint optimization problem, the optimum delay is determined leading to the maximization of the SNR at the RX correlator output. A simpler, decoupled approach then consists of preselecting (see the following section) the dominant path *a priori*, i.e., before the design of $\boldsymbol{U}_{iq}^{t}$'s, and assuming that the RX correlators for all users are aligned to the delay of the dominant paths. The delay assignment thus assumes that *a priori* and *a posteriori* (after ZF-prefilter design) dominant paths will be the same, a very likely event. The prefilters for all users can now be designed as discussed previously in a decoupled fashion. Fine tuning of TX filter delays as discussed in the previous paragraph can still be applied, subject to the fixed delay constraint for the correlators. We concede that the pre-assigned delays may not, in all cases, be the optimal ones, but this simplifies the optimization problem making it much simpler to implement.

V. TX DIVERSITY AND POWER ASSIGNMENT

We have assumed that each receiver consists of a correlator per multipath component. Assume that the correlator outputs are combined according to the maximum ratio combining (MRC) criterion. The multipath signal components are assumed to be spaced such that the correlator outputs are uncorrelated. The effect if IUI and ISI may be ignored at this point (we have seen that pre-filtering will cancel them). Fig. 2 shows the the TX-channel-RX cascade for the ith user. We assume a constraint on the total transmit power such that $\sum_{q=1}^{Q_i} p_{iq} = p_i$ (with $p_{iq} \geq 0$). The output signal-to-noise ratio (SNR) for the ith user is

$$\mathrm{SNR}_i = \frac{\mathrm{E}[|\sum_{q=1}^{Q_i}|\alpha_{iq}|^2 p_{iq} a_i(k)|^2]}{\sigma_{v_i}^2 \sum_{q=1}^{Q_i}\mathrm{E}|\alpha_{iq}|^2 p_{iq}} = \frac{\sigma_a^2}{\sigma_{v_i}^2}\sum_{q=1}^{Q_i}\mathrm{E}[|\alpha_{iq}|^2]p_{iq}$$

where $\sigma_a^2 = \mathrm{E}[|a_i(k)|^2]$ for all the i's and $\sigma_{v_i}^2$ is the variance of the noise at each correlator output (for the variance of the noise at the correlator output it is assumed that the spreading sequences are sufficiently white). The optimal power assignment among the different paths that maximize the SNR is determined by solving the following problem

$$\max_{p_{iq}}\{\sum_{q=1}^{Q_i}(\mathrm{E}|\alpha_{iq}|^2)p_{iq}\} \qquad \text{s.t. } \sum_{q=1}^{Q_i} p_{iq} = p_i, \tag{11}$$

the solution to which is the well known *selection diversity* which corresponds to assigning the whole transmit power to the path carrying the most power (on the average). Hence, under the conditions above the previous receiver structure collapses into a single pulse shape matched filter and a correlator.

We remark that the strongest multipath component is the one with the maximum energy in the corresponding prefilter-channel.

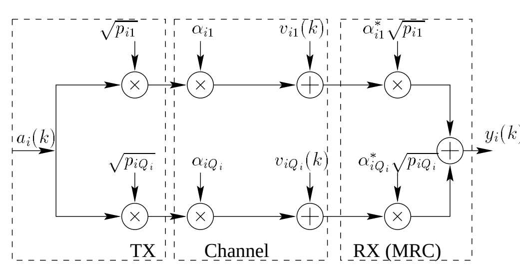

Figure 2. Transmit diversity for the ith user through Q_i diversity branches/paths after pre-decorrelating pre-filtering and ZF IUI

Then in general, when a pathwise pre-filtering is performed at the base station then, strictly speaking, the strongest path selection for a certain user can take place after the pre-filter design for each path. Hence, all paths need to be considered for the pre-filter design.

A. Power Assignment Optimization

Since the transmission strategy consists of exciting one path per user, we refer to $\boldsymbol{U}_i^t$, $\boldsymbol{V}_i$ and σ_i^2 as the filter, the channel (up to the fading coefficient) and the variance of the fading associated with the selected path for the ith user. Once we have designed the normalized transmit filters $\boldsymbol{U}_i^t$ we need to optimize the transmit power assignment among the d users. In the absence of IUI due to ZF, we shall optimize the transmit power assignment in order to make the SNR at the output of each selected path correlator, the same for all the users, subject to a total transmit power constraint. The SNR for the ith user is given by

$$\gamma_i = \frac{\sigma_a^2}{\sigma_{v_i}^2} p_i \sigma_i^2 \|\boldsymbol{U}_i^t \mathcal{T}_L(\boldsymbol{V}_i)\|_2^2 \tag{12}$$

where $\sum_i p_i = p_{\max}$. Since the optimal leads $\gamma_i = \gamma$ for all the users then it is straightforward to derive the following expressions for γ and the optimal p_i's

$$\begin{aligned} \frac{1}{\gamma} &= \frac{1}{p_{\max}\sigma_a^2} \sum_i \frac{\sigma_{v_i}^2}{\sigma_i^2 \|\boldsymbol{U}_i^t \mathcal{T}_L(\boldsymbol{V}_i)\|_2^2} \\ p_i &= \gamma \frac{\sigma_{v_i}^2}{\sigma_a^2} \frac{1}{\sigma_i^2 \|\boldsymbol{U}_i^t \mathcal{T}_L(\boldsymbol{V}_i)\|_2^2} \end{aligned} \tag{13}$$

VI. DISCUSSION

The pre-decorrelating transmit filters designed according to (7) are optimal in the noiseless case. Indeed the limited power constraint does not affect in this case the SINR, which reduces to the Signal to Interference ratio (SIR) at each receiver, and which is infinity for any power assignment when ZF IUI is achieved. However, in the presence of noise at receivers as the number of ZF constraints will increase, a *noise enhancement* phenomenon will arise which might reduce the SINR gain obtained from the IUI cancellation. If the CDMA system under consideration allows a large number of degrees of freedom, namely a large m, compared to the number of paths of all the users, then the noise enhancement phenomenon will be practically negligible compared to the SINR gain yielded by ZF the IUI.
An alternative solution is represented by a pre-RAKE like prefiltering. Due to the lack of knowledge of the path phases (and amplitudes) of the downlink channel, only non-coherent pre-RAKE processing is possible at the base station. However, the result of section V, disagrees with pre-RAKE kind of prefiltering.

VII. SIMULATIONS

We consider an CDMA/SDMA scenario in the presence of $d = 3$ users having $Q_i = 2$ paths each, which receive signals transmitted from a base station. The total power $p_{\max}$ and σ_a^2 are constant, and the noise variances $\sigma_{v_i}^2 = \sigma_v^2$ is assumed and to be the same at all receivers and to be known at the transmitter. The single path delays τ_{iq}'s, the array response vectors $\mathbf{a}(\theta_{iq})$ to build the channels $\boldsymbol{V}_{iq}$'s, and the variances σ_{iq}^2 are estimated from the uplink.
In the first simulation we considered a *saturated* system configuration assuming the $m_c = 8$ and $m_a = m_o = 1$. In this case $m_{\text{eff}} > Q$ ($m_{\text{eff}} = m = 8$) and ZF conditions (9) can be applied, if filter length is $L \geq 4$. We fixed $L = 4$ symbol periods to achieve (9). The resulting performances are plotted in fig. 3(a) in terms of SNR at each receiver versus the residual ISI (L_{ISI} introduced by the pre-filter channel cascade. Due to the high system

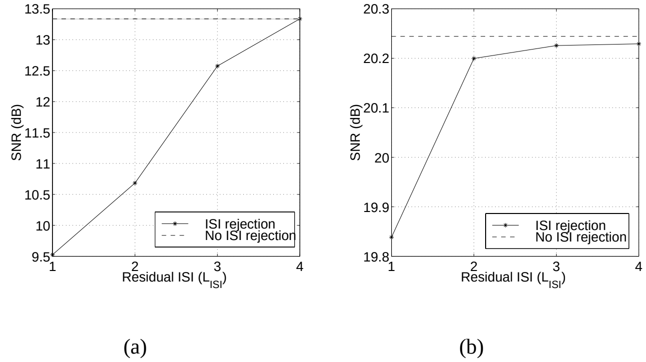

Figure 3. Optimum SNR vs. L_{ISI}, IP Pre-decorrelating ZF solution for $L = 4$, $Q = 6$ paths. $m = m_c = 8$, $m_{\text{eff}} = 8$ in (a), $m_a = 2$, $m_c = 8$, $m_{\text{eff}} = 13$ in (b).

loading significant differences arise for different values of L_{ISI}. In the second simulation we considered the same user scenario as above, $m_c = 8$, but employing $m_a = 2$ antennas at the transmitter. Since $m_{\text{eff}} = 13$ and $Q = 6$ IP pre-decorrelation ZF IUI and ISI conditions (9) can be applied. By setting the length of all the transmit filters equal to $L = 4$ symbol periods (even though $L = 1$ suffices to achieve ZF conditions) we obtain the performances plotted in figure 3(b). Note that in this case due to the large m_{eff}, w.r.t. the number of user paths Q and to the small delay spreads introduced by the path channels the performances are quite insensitive to the residual delay spread L_{ISI}. It can be demonstrated that larger values of L yield improvement of performances, more significant when m_{eff} is not very large compared to Q.

VIII. CONCLUSIONS

The FDD/CDMA downlink problem was addressed. It was shown that due to the partial knowledge of the downlink channel, each path of a particular user could be treated as a seperate user. Pre-decorrelation was applied on the downlink to cancel the IUI and IPI. For the desired user, the path selection diversity scheme was shown to be the best power assignment choice in terms of the SNR optimization. Performance of the receiver *vis-à-vis* the residual ISI was also shown. It was observed that as long as the system has sufficient degrees of freedom (OS/MA factor), IUI can be cancelled by TX pre-filters, leading to low complexity, improved mobile receivers. RX delay optimization was shown to be a coupled problem and a simplified strategy was presented to obtain an individualized framework. We point out that the above framework can easily be extended to include more complex situations, like extracell interference etc.

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