Supporting Integrated Services in Wireless Networks with Space-Time Block-Coded Transmissions*

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Abstract—**Integrated Services in Wireless Networks call for a physical layer which can support multirate transmissions and guarantee symbol recovery at the receiver. Multirate transmissions are necessary for data applications with diverse needs in throughput/delay, whereas symbol recovery leads to Quality of Service guarantees in terms of bit error rate. In this work, we develop a multiuser physical layer framework, which can be used in a wireless integrated services architecture. Capitalizing on space-time coding and block transmissions, we show how space-time transmit antenna diversity can provide transmission rates of arbitrarily fine resolution while guaranteeing symbol recovery irrespective of the FIR wireless channel.**

Keywords—**space-time coding, block transmissions, wireless networks**

I. INTRODUCTION

Broadband wireless networks are envisioned to provide high data rates and support Integrated Services. To make them possible, the physical layer should i) cope with the variable channel capacity due to multipath fading; and ii) facilitate multiuser/multirate transmissions. Time, frequency, and spatial diversity have been employed to combat channel fading and the high bit-error-rates (BERs) it induces (see e.g., [10, 11] and references therein). Among the three, spatial diversity especially with transmit-antennas has drawn a lot of attention recently as it promises to increase the achievable transmission rates significantly.

In this paper we develop a novel multirate/multiuser scheme relying on space-time (ST) coded Generalized MultiCarrier (GMC) CDMA that exploits transmit antenna diversity. The resulting transceivers rely on block-spreading to guarantee symbol recovery with multipath-diversity gains regardless of the (possibly unknown) frequency-selective FIR channels and multiuser interference (MUI). Our approach is based on a three-level user code design: the outer code handles MUI, the middle code results in ST diversity gains, and the inner code mitigates inter-symbol interference (ISI).

The merits of block-spreading with an inner/outer code design have been introduced in [6, 7, 16–18] for block transmission systems. The inner-code can be realized either through channel coding or redundant linear precoding. The latter introduces significantly less redundancy than channel coding and may be more effective in mitigating the effects of FIR channels (see [6] and references therein). The outer code (which could take the form of time slots in TDMA or user codes in CDMA) eliminates MUI and converts multiuser environments to single-user ones. Based on subcarrier allocation, the outer code of [6, 7, 16–18] offers flexible rate allocation and deterministic MUI elimination regardless of the underlying FIR channels, while adhering to a low-implementation cost (most of operations are FFT-based). However, the framework of [6, 7, 16–18] does *not* incorporate ST coding.

ST coding capitalizes on multiple transmit antennas to increase considerably the channel capacity [5]. Though it is well known that array processing and receive antennae improve BER performance, the size–power limitations of handheld devices make the deployment of multiple transmit antennas at the basestation an attractive alternative. ST coding can achieve spatial diversity gains with just a single receive antenna; multiple receive antennas are optional (if the receiver is equipped with more than one receive antenna, further performance improvement can be achieved [15]). Systems equipped with ST coding are multi-input multi-output (MIMO) systems where serial information symbols are coded across both "space" (as the same symbols are transmitted by multiple transmit-antennas) and "time" (as the same symbols are repeated at different time slots). The received symbols are jointly decoded to recover the information symbols with diversity and coding gains. Tutorials on ST coding can be found in [10, 11].

In broadband wireless networks, ST coding must take into consideration channel frequency-selectivity [10]. ST decoding schemes based on maximum-likelihood estimators may not be attractive, especially given the limited processing resources of handheld devices and the fact that their computational complexity grows exponentially with the transmission rate. Decoupling ST decoding from channel equalization appears to be a computationally attractive alternative, and it has already been proposed in the form of ST coded OFDM [1]. However, the inner/outer code designs of [16, 17] exhibit improved BER performance with respect to OFDM, which in part motivated the development of *single-rate* ST block codes in [9] with *a single*

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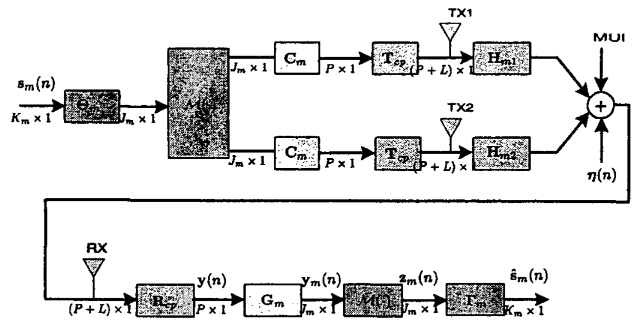

Fig. 1. Physical layer

receive-antenna.

Herein we combine the strengths of [17] (multirate services, deterministic MUI/ISI suppression) with those of [9] (ST diversity gains in the presence of frequency-selective fading channels), while augmenting both schemes. In Section II, we incorporate the ST code of [9] as the *middle code* into the inner/outer code framework of [17]. The merits of our system are verified through simulations in Section III.

II. Physical Layer

Our model, shown in Fig. 1, represents signals, codes and channels by chip-rate samples of their complex envelopes. The information symbols $s_m(k)$ of user m are grouped together in blocks $\mathbf{s}_m(n) := \left(s_m(nK) \cdots s_m(nK+K-1)\right)^T$ of size K_m, which are mapped to blocks $\tilde{\mathbf{s}}_m(n) = \mathbf{\Theta}_m \mathbf{s}_m(n)$ of length J_m through the tall $J_m \times K_m$ matrix $\mathbf{\Theta}_m$ (the redundancy introduced by $\mathbf{\Theta}_m$ facilitates ISI elimination and symbol recovery regardless of the physical channel [9, 17]). In practice, the precoder $\mathbf{\Theta}_m$ can be implemented either by a DSP processor and matrix operations, or by a filterbank with K_m FIR filters $\theta_{m,k}(q)$, each of length J_m: $\theta_{m,k}(q) := [\mathbf{\Theta}_m]_{q,k}$. Then, the blocks $\tilde{\mathbf{s}}_m(n)$ are fed into the ST mapper $\mathcal{M}(\cdot)$, which takes two consecutive precoded blocks, $\tilde{\mathbf{s}}_m(2n)$ and $\tilde{\mathbf{s}}_m(2n+1)$, to output the following $2J_m \times 2$ code matrix ($*$ denotes conjugation):

$$\begin{aligned} \mathbf{S}(n) &:= \begin{pmatrix} \bar{\mathbf{s}}_{m1}(2n) & \bar{\mathbf{s}}_{m1}(2n+1) \\ \bar{\mathbf{s}}_{m2}(2n) & \bar{\mathbf{s}}_{m2}(2n+1) \end{pmatrix} \\ &= \begin{pmatrix} \tilde{\mathbf{s}}_m(2n) & -\tilde{\mathbf{s}}_m^*(2n+1) \\ \tilde{\mathbf{s}}_m(2n+1) & \tilde{\mathbf{s}}_m^*(2n) \end{pmatrix} \begin{matrix} \rightarrow & \text{time} \\ \downarrow & \text{space} \end{matrix} \end{aligned} \tag{1}$$

Each column corresponds to a transmission slot ("time"), whereas each row holds the symbols to be transmitted by each antenna ("space"). Note that without blocking (i.e., $J_m = 1$) the code matrix in (1) reduces to the ST block code in [2]. For notational convenience, we denote by $\bar{\mathbf{s}}_{mi}(n)$, $i = 1, 2$, the $J_m \times 1$ block transmitted through the ith transmit-antenna at the time slot n, and observe from (1) that e.g., $\bar{\mathbf{s}}_{m1}(2n) = \tilde{\mathbf{s}}_m(2n)$, and $\bar{\mathbf{s}}_{m2}(2n+1) = \tilde{\mathbf{s}}_m^*(2n)$.

Instrumental in MUI elimination is the user-specific block-spreading code, which is represented by the $P \times J_m$ matrix $\mathbf{C}_m$. The user code is applied to $\bar{\mathbf{s}}_{mi}(n)$ to produce the block $\mathbf{u}_{mi}(n) := \mathbf{C}_m \bar{\mathbf{s}}_{mi}(n)$ of length P, which is transmitted by the m-th user's i-th antenna through the L-th order FIR channel $\{h_{mi}(l)\}_{l=0}^{L}$. In order to eliminate the IBI caused by FIR channels, similar to OFDM, we rely on inserting a CP of length L at the beginning of $\mathbf{u}_{mi}(n)$; at the receiver, the corresponding first L received symbols are discarded. As explained in [17] and shown in Fig. 1, the CP insertion can be described by the transmit-matrix $\mathbf{T}_{cp} := \left(\mathbf{I}_{cp}^T \ \mathbf{I}_P^T\right)^T$, where $\mathbf{I}_{cp}$ is formed by the last L rows of the $P \times P$ identity matrix $\mathbf{I}_P$. Correspondingly, the receiver operation of discarding the first L received symbols is described by the receive-matrix $\mathbf{R}_{cp} := \left(\mathbf{0}_{P\times L} \ \mathbf{I}_P\right)$. The effect of the FIR channel $h_{mi}(l)$ on the transmitted data is described by the $(P+L)\times(P+L)$ Toeplitz (convolutional) matrix $\mathbf{H}_{mi}$ with first row $\left(h_{mi}(0)\mathbf{0}_{1\times(P+L-1)}\right)$, and first column $\left(h_{mi}(0)\ldots h_{mi}(L)\ \mathbf{0}_{1\times(P-1)}\right)^T$. With $\tilde{\mathbf{H}}_{mi} := \mathbf{R}_{cp}\mathbf{H}_{mi}\mathbf{T}_{cp}$, the $P \times 1$ IBI-free received symbol block $\mathbf{y}(n)$ is given by (see Fig. 1):

$$\mathbf{y}(n) = \mathbf{x}_m(n) + \sum_{\mu=0,\ \mu\neq m}^{M-1} \mathbf{x}_\mu(n) + \mathbf{R}_{cp}\boldsymbol{\eta}(n), \tag{2}$$

where: $\mathbf{x}_m(n) := \tilde{\mathbf{H}}_{m1}\mathbf{C}_m\bar{\mathbf{s}}_{m1}(n) + \tilde{\mathbf{H}}_{m2}\mathbf{C}_m\bar{\mathbf{s}}_{m2}(n)$ denotes the received symbol block from the two transmit-antennas of the mth user, and $\boldsymbol{\eta}(n)$ is the $(P+L)\times 1$ additive white Gaussian noise vector.

The received symbol block in (2) contains three terms: the first term is the useful received block $\mathbf{x}_m(n)$ from the desired mth user; the second term constitutes the MUI caused by simultaneous transmissions of the other $M-1$ users; the third term is the additive noise. Next, we present our code designs, which guarantee symbol recovery with diversity gains for each user regardless of the MUI and underlying FIR channels. As a prelude to Section II-B, we note that the flexibility in assigning different information block length K_m's to different users enables our physical layer to support packet-level multirate services.

A. Code Design

To achieve improved BER performance at the physical layer, our system employees redundancy at 3 different layers: *block-of-symbols* level, where the tall precoder matrix $\mathbf{\Theta}_m$ combats ISI and guarantees symbol recovery regardless of the FIR channels (this operation can be thought of as the *inner-code* in our transmission scheme); *transmit-antennae* level, where the ST mapper is designed such that diversity gain at the receiver is guaranteed regardless of the channel (this operation constitutes the *middle-code* of our framework); *user* spreading level, where the code matrix $\mathbf{C}_m$ guarantees MUI elimination regardless of the FIR channels and implements the *outer-code*. In the following, we describe our code design procedure starting from the outside ("outer-code") and moving to the inside ("inner-code").

Outer Code: MUI Elimination In order to design the code matrix $\mathbf{C}_m$ and the receive matrix $\mathbf{G}_m$, we will borrow the techniques of [17] to ensure that MUI is deterministically eliminated while still guaranteeing symbol recovery for user m. The designs in [17] are for single-transmit antenna and capitalize on block-spreading and the core concept of subcarrier allocation: similar to OFDMA, every user m is allocated a distinct set of J_m subcarriers to transmit information, but unlike OFDMA, the all-digital framework of [17] guarantees symbol recovery irrespective of channel nulls. One of the fundamental insights of [17] is that an L-th order FIR channel manifests itself as at most L deep fades on the FFT grid. Hence, when the channel is not known to the transmitter, even in the noiseless case, source symbols transmitted on L (out of the possible J_m) subcarriers may be lost; this potential loss is taken care of by our inner code. Herein, we allocate J_m subcarriers to every user m; these subcarriers are used by both transmit antennas to produce two interfering signals at the receiver end (something that [17] did not have to cope with). For the time being we focus on eliminating MUI, leaving it up to our ST coding to take care of the two interfering signals generated by user m. Note also that both antennas transmit at the same time.

To express the aforementioned idea of subcarrier allocation mathematically, let $P = \sum_{\mu=0}^{M-1} J_\mu$ be the total number of subcarriers available to M users, and define the index set $\mathcal{I} = \{0, 1, \ldots, P-1\}$ as the collection of all subcarrier indices. Each user μ is allocated a set $\mathcal{I}_\mu = \{j_{m,0}, \cdots, j_{m,J_m-1}\}$ of J_μ subcarriers $\{\rho_{m,j_m}\}_{j_m=0}^{J_m-1}$; in our discrete-time equivalent baseband model, the j_m-th subcarrier is represented as $\rho_{m,j_m} = \exp\{j2\pi j_m/P\}$. Users have non-intersecting subcarrier sets:

$$\bigcup_{\mu=0}^{M-1} \mathcal{I}_\mu = \mathcal{I}, \quad \mathcal{I}_\mu \bigcap \mathcal{I}_m = \emptyset, \ \forall \mu \neq m, \tag{3}$$

where $\emptyset$ denotes the empty set. Note that this modeling does not preclude the use of some subcarriers as guards bands as in, e.g., HiperLan 2 [3]. In our model, these subcarriers could be assigned to a "system user". The set of subcarriers is used to define for the μth user the subcarrier selector matrix $\mathbf{\Phi}_\mu = \mathbf{I}_P(:, \mathcal{I}_\mu)$ where $\mathbf{I}_P(:, \mathcal{I}_\mu)$ is a $P \times J_\mu$ matrix built from the $\mathcal{I}_\mu$ columns of $\mathbf{I}_P$. Using (3), it is readily proved that the subcarrier selector matrices are orthogonal to each other, i.e.,

$$\mathbf{\Phi}_\mu^T \mathbf{\Phi}_m = \delta(\mu - m)\mathbf{I}_{J_\mu \times J_m}\,. \tag{4}$$

Denoting by $\mathbf{F}$ the $P \times P$ FFT matrix $\mathbf{F}$ with its (k, p) entry $[\mathbf{F}]_{k,p} = P^{-\frac{1}{2}} \exp\{(-j2\pi kp)/P\}$, the subcarrier selector $\mathbf{\Phi}_\mu$ is used to design the user coder $\mathbf{C}_\mu$ and the receive-filter $\mathbf{G}_\mu$ as:

$$\mathbf{C}_\mu = \mathbf{F}^{\mathcal{H}} \mathbf{\Phi}_\mu, \quad \mathbf{G}_\mu = \mathbf{\Phi}_\mu^T \mathbf{F}, \quad \mu = 0, \ldots, M-1\,. \tag{5}$$

In other words, the matrix $\mathbf{\Phi}_\mu$ selects the subcarriers allocated to user μ; at the receiver end, only these subcarriers are taken into account for the recovery of the transmitted symbols. We obtain from (2) and (5) that the receive-filter output $\mathbf{y}_m(n) := \mathbf{G}_m \mathbf{y}(n)$ can be written as:

$$\mathbf{y}_m(n) = \sum_{\mu=0}^{M-1} \mathbf{\Phi}_m^T \mathbf{F}[\tilde{\mathbf{H}}_{\mu 1} \mathbf{F}^{\mathcal{H}} \mathbf{\Phi}_\mu \bar{\mathbf{s}}_{\mu 1}(n) + \tilde{\mathbf{H}}_{\mu 2} \mathbf{F}^{\mathcal{H}} \mathbf{\Phi}_\mu \bar{\mathbf{s}}_{\mu 2}(n)] + \mathbf{G}_m \mathbf{R}_{cp} \boldsymbol{\eta}(n). \tag{6}$$

To show that indeed MUI has been eliminated from (6), we use the following three facts [17]:

Fact 1: The equivalent channel matrix $\tilde{\mathbf{H}}_{\mu i} = \mathbf{R}_{cp} \mathbf{H}_{\mu i} \mathbf{T}_{cp}$ is a $P \times P$ circulant matrix.

Fact 2: The circulant $\tilde{\mathbf{H}}_{\mu i}$ is diagonalized by the IFFT/FFT at the transmitter/receiver, i.e.,

$$\mathbf{F}\tilde{\mathbf{H}}_{\mu i}\mathbf{F}^{\mathcal{H}} = \mathbf{D}_{H_{\mu i}} := \mathrm{diag}[H_{\mu i}(e^{j0}), \ldots, H_{\mu i}(e^{j2\pi(P-1)/P})],$$

where $H_{\mu i}(\rho) = \sum_{l=0}^{L} h_{\mu i}(l) \rho^{-l}$ is the frequency response of the channel $h_{\mu i}(l)$ at the subcarrier ρ.

Fact 3: With $\mathcal{D}_{\mu i} := \mathbf{\Phi}_\mu^T \mathbf{D}_{H_{\mu i}} \mathbf{\Phi}_\mu$ denoting a $J_\mu \times J_\mu$ diagonal matrix holding in its diagonal the frequency responses of the FIR channel $h_{\mu i}(l)$ at ρ_{μ, j_μ}'s, i.e.,

$$\mathcal{D}_{\mu i} = \mathbf{\Phi}_\mu^T \mathbf{F} \tilde{\mathbf{H}}_{\mu i} \mathbf{F}^{\mathcal{H}} \mathbf{\Phi}_\mu = \mathrm{diag}\big(H_{\mu i}(\rho_{\mu,0}), \ldots, H_{\mu i}(\rho_{\mu, J_\mu - 1})\big) \tag{7}$$

it holds that $\mathbf{D}_{H_{\mu i}} \mathbf{\Phi}_\mu = \mathbf{\Phi}_\mu \mathcal{D}_{\mu i}$. Note that $\mathbf{D}_{H_{\mu i}}$ is $P \times P$ whereas $\mathcal{D}_{\mu i}$ is $J_\mu \times J_\mu$.

Exploiting Facts 1)-3), the receive-filter output $\mathbf{y}_m(n)$ is given by [17]:

$$\mathbf{y}_m(n) = \mathcal{D}_{m1} \bar{\mathbf{s}}_{m1}(n) + \mathcal{D}_{m2} \bar{\mathbf{s}}_{m2}(n) + \mathbf{G}_m \mathbf{R}_{cp} \boldsymbol{\eta}(n), \tag{8}$$

where in deriving the third equality we have used the orthogonality in (4). It is clear from (8) that the MUI has been eliminated through the joint designs of $\mathbf{C}_\mu$ and $\mathbf{G}_\mu$ in (5). Note also that the MUI elimination is achieved without requiring the knowledge of the channel impulse response (CIR) at the receiver. Having eliminated the MUI, we exploit next the ST coding in (1) to recover the transmitted symbols $\mathbf{s}_m(n)$ with guaranteed diversity gains.

Middle-Code: Diversity Gain The objective of the middle-code is to collect diversity gains by combining constructively the transmitted signals from the two antennas. Knowing $\mathcal{D}_{m1}$ and $\mathcal{D}_{m2}$, we consider two consecutive MUI-free blocks: $\mathbf{y}_m(2n)$ and $\mathbf{y}_m(2n+1)$. Recall that $\bar{\mathbf{s}}_m(n)$ is mapped to $\bar{\mathbf{s}}_{mi}(n)$ as:

$$\begin{array}{ll} \bar{\mathbf{s}}_{m1}(2n) \longleftrightarrow \bar{\mathbf{s}}_m(2n) \quad , & \bar{\mathbf{s}}_{m2}(2n) \longleftrightarrow \bar{\mathbf{s}}_m(2n+1) \\ \bar{\mathbf{s}}_{m1}(2n+1) \longleftrightarrow -\bar{\mathbf{s}}_m^*(2n+1)\,, & \bar{\mathbf{s}}_{m2}(2n+1) \longleftrightarrow \bar{\mathbf{s}}_m^*(2n)\,. \end{array} \tag{9}$$

Plugging the mapping (9) into (8), we express $\mathbf{y}_m(2n)$ and $\mathbf{y}_m(2n+1)$ as:

$$\mathbf{y}_m(2n) = \mathcal{D}_{m1}\tilde{\mathbf{s}}_m(2n) + \mathcal{D}_{m2}\tilde{\mathbf{s}}_m(2n+1) + \mathbf{G}_m\mathbf{R}_{cp}\boldsymbol{\eta}(2n) \tag{10a}$$

$$\mathbf{y}_m(2n+1) = -\mathcal{D}_{m1}\tilde{\mathbf{s}}^*_m(2n+1) + \mathcal{D}_{m2}\tilde{\mathbf{s}}^*_m(2n) + \mathbf{G}_m\mathbf{R}_{cp}\boldsymbol{\eta}(2n+1). \tag{10b}$$

In order to exploit the embedded diversity of (10a) and (10b), we design the ST decoder $\bar{\mathcal{M}}(\cdot)$ (see Fig. 1) by forming its two consecutive output blocks, $\mathbf{z}_m(2n)$ and $\mathbf{z}_m(2n+1)$, as:

$$\begin{pmatrix} \mathbf{z}_m(2n) \\ \mathbf{z}_m(2n+1) \end{pmatrix} = \begin{pmatrix} \mathcal{D}^*_{m1} & \mathcal{D}_{m2} \\ \mathcal{D}^*_{m2} & -\mathcal{D}_{m1} \end{pmatrix} \begin{pmatrix} \mathbf{y}_m(2n) \\ \mathbf{y}^*_m(2n+1) \end{pmatrix}. \tag{11}$$

Substituting (10a) and (10b) into (11), we arrive at:

$$\mathbf{z}_m(n) = [\mathcal{D}^*_{m1}\mathcal{D}_{m1} + \mathcal{D}^*_{m2}\mathcal{D}_{m2}]\tilde{\mathbf{s}}_m(n) + \tilde{\boldsymbol{\eta}}_m(n), \tag{12}$$

where the additive noise vector $\tilde{\boldsymbol{\eta}}_m(n) := [\boldsymbol{\eta}^T_m(2n) \;\; \boldsymbol{\eta}^T_m(2n+1)]^T$ is given by:

$$\tilde{\boldsymbol{\eta}}_m(n) := \begin{pmatrix} \mathcal{D}^*_{m1} & \mathcal{D}_{m2} \\ \mathcal{D}^*_{m2} & -\mathcal{D}_{m1} \end{pmatrix} \begin{pmatrix} \mathbf{G}_m\mathbf{R}_{cp}\boldsymbol{\eta}(2n) \\ \mathbf{G}^*_m\mathbf{R}_{cp}\boldsymbol{\eta}^*(2n+1) \end{pmatrix}. \tag{13}$$

If the receiver noise is temporally white, then $\boldsymbol{\eta}(n)$ is a white Gaussian random process with covariance matrix $\sigma^2_\eta\, \mathbf{I}_{P+L}$. It readily follows from (5) and Fact 3) that the covariance matrix of $\tilde{\boldsymbol{\eta}}_m(n)$ is diagonal, and the blocks $\mathbf{z}_m(n)$ can be processed individually without loss in performance. To show how the transmit-diversity gains are achieved, we plug (7) into (12), to obtain:

$$\begin{aligned}\mathbf{z}_m(n) &= \mathcal{D}_m\tilde{\mathbf{s}}_m(n) + \boldsymbol{\eta}_m(n) \\ &= \mathcal{D}_m\boldsymbol{\Theta}_m\mathbf{s}_m(n) + \boldsymbol{\eta}_m(n),\end{aligned} \tag{14}$$

where $\mathcal{D}_m := \mathcal{D}^*_{m1}\mathcal{D}_{m1} + \mathcal{D}^*_{m2}\mathcal{D}_{m2}$ $= \text{diag}[\sum_{i=1}^2 |H_{mi}(\rho_{m,1})|^2, \ldots, \sum_{i=1}^2 |H_{mi}(\rho_{m,J_m})|^2]$.

As each channel can have at most L nulls, (14) implies that transmit-diversity gains are achieved in at least $J_m - 2L$ subcarriers. Moreover, $J_m - L$ subcarriers are guaranteed to be non-zero and can be used for symbol recovery (regardless of the underlying channels). The latter observation brings us to the third component of our (pre)coding, namely, the inner code.

Inner-Code: Guaranteed Symbol Recovery We deduce from (14) that the recovery of $\mathbf{s}_m(n)$ from $\mathbf{z}_m(n)$ requires $\mathcal{D}_m\boldsymbol{\Theta}_m$ to be full column rank. As shown in [7,9,17,18], the full column rank is guaranteed by selecting: i) $J_m = K_m + L$, and ii) any K_m rows of $\boldsymbol{\Theta}_m$ to be linearly independent.

B. Supporting Integrated Services

The fact that the m-th user's transmission rate is $K_m/(P+L)$ reveals another attractive feature of our designs: namely, multirate allocation with arbitrarily fine resolution (as we increase P). In an integrated services framework, the flexibility in assigning values to J_m allows the physical layer to implement bandwidth allocation as specified by the network layer scheduler[1]. To appreciate the importance of a multirate-transparent physical layer, we should take into consideration that nowadays, a large class of QoS network schedulers are based on the Generalized Processor Sharing (GPS) [13] scheduling policy. According to GPS, every user m should be allocated bandwidth proportional to a user-specific positive weight ϕ_m. As shown in [14], multirate transmissions at the physical layer enable fair bandwidth allocation at the network layer without resorting to TDMA. The basic idea is that in every transmission round each user m is allocated a portion BW_m of the total bandwidth BW according to $\text{BW}_m/\text{BW} = \phi_m/\sum_{\mu\in\mathcal{A}}\phi_\mu$, where $\mathcal{A}$ is the set of active users. By setting

$$J_m = \lfloor \frac{\text{BW}_m}{\text{BW}}(P-L) \rfloor, \quad K_m = J_m - L,$$

we ensure that the bandwidth assignments as dictated by the network layer are implemented exactly at the physical layer. If we take into account that in quasi-synchronous CDMA systems the order of the channel is $L \leq 7$, and that P is a multiple of, say, 424 bits (the size of an ATM cell), we can see that $K_m/P \approx \text{BW}_m/\text{BW}$. Hence, our code assignment procedure could also play a very important role in translating BER improvements at the physical layer to throughput improvements at the network layer. This is because in each transmission round, every mobile user is assigned its fair share of the bandwidth. As a direct result, the mobile user can immediately exploit any BER gains at the physical layer.

At this point, we would like to re-iterate that if one wants to dispense with TDMA and still provide integrated services, then the physical layer should possess multirate/multiuser capabilities. The subcarrier-allocation scheme of [7,9,17,18] is not the only physical layer framework capable of providing such services. Multicode schemes (based on orthogonal variable-spreading codes or multiple pseudonoise codes) can also support different transmission rates and multiple users [8, 12]. However, pseudonoise codes do not guarantee deterministic MUI elimination, and orthogonal codes (at the transmitter) may not be orthogonal at the receiver (because of the effect of the convolutive FIR channels [17]). Thus, in broadband wireless networks, where unknown frequency-selective multipath naturally arises, the multi-carrier framework of [7,9,17,18] is well-motivated for improved BER performance.

III. SIMULATIONS

When all mobile users have been assigned the same rates, [9] provides simulations which show the benefits of ST coding. Herein we are interested in the multirate aspects of our

[1] The network layer scheduler has a profound effect on the Quality of Service (QoS) guarantees that can be provided to users, since it is the scheduler the one who determines the transmission order of packets in the network.

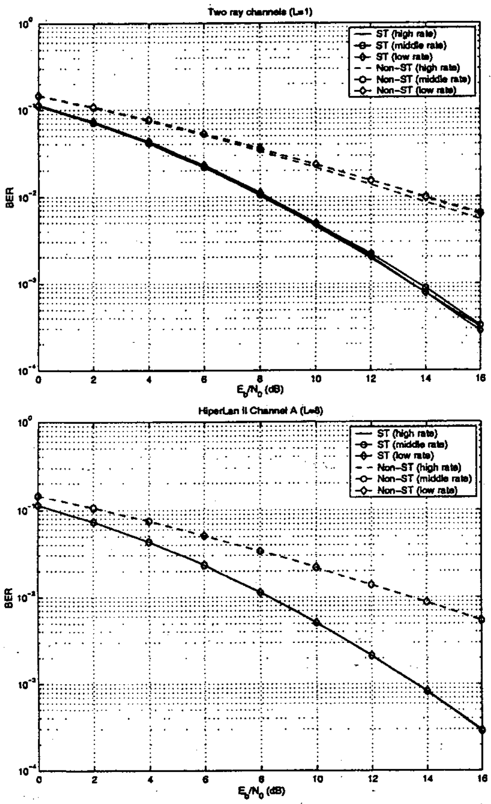

Fig. 2. Space-time gains for $L = 2$ (up) and $L = 8$ (down)

framework, and how our designs improve upon the single-transmit/single-receive antenna scenario (which amounts to no spatial diversity). We study a system with $M = 8$ users, where each user is equipped with two transmit-antennas; two high-rate users have been assigned the weight $\phi_{hr} \approx 1/4$, two middle-rate users have $\phi_{mr} \approx 1/8$, whereas four low-rate users have been assigned the weight $\phi_{lr} \approx 1/16$. In all simulations, the BER is chosen as the figure of merit and is averaged over 600 random channel and noise realizations for various E_b/N_0 points. The random channels are generated based on two different channel models:

Two-ray channel ($L = 1$): This corresponds to the case where the multipath consists of two equal-power dominant rays, and one ray is delayed with respect to the other by a chip duration.

HiperLan 2 channel ($L = 8$): The random channels are based on the HiperLan 2 channel model A, which corresponds to a typical office environment [4].

We compare our design with $N_t = 2$ transmit-antennas and $N_r = 1$ receive-antenna to that with a single transmit- and receive-antenna. The results are depicted in Fig. 2 for the aforementioned two channel models. Observing that the BER gains achieved by ST coding for all users is almost 5.8 dB at BER $= 10^{-2}$, it is deduced that our ST block-spreading code designs result in BER improvements even in the presence of frequency-selective FIR channels.

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