Subspace-based methods for channel estimation in the TD-SCDMA system

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In this paper we propose a subspace method to estimate time-varying channels by taking into account the different rates of variation between delays/directions of arrival (slowly-varying) and faded amplitudes (fast-varying) of the multipath propagation. The stationarity of angles and delays across several time-slots is exploited to perform a subspace estimation that avoids the explicit computation of the path parameters. Numerical analysis for a TD-SCDMA system shows the relevant advantage of the multi-slot approach compared to single-slot techniques.

INTRODUCTION

In mobile communication systems the channel is time-varying due to the movement of the terminal. Each path of the multipath propagation channel is characterized by an angle (or direction of arrival), a delay and an amplitude. Since the angles and delays are stationary compared to the amplitudes (even for fast moving mobile terminals), we propose a method that exploits these different varying rates to improve the accuracy in the estimate of the space-time channel matrix.

For slow fading channels, the accuracy of the channel estimate can be increased by simply averaging the information from successive slots so as to extend the effective training data length [2]. However, multi-slot averaging is not effective for the estimation of fast varying channels. In this case only the slowly varying features of the propagation channel can be obtained from long-term observations: the subspaces spanned by the spatial/temporal channel signatures can be estimated by exploiting the training data of successive slots, while the amplitudes have to be calculated on a slot-by-slot basis.

As an application, in this paper we focus on the uplink of a TD-SCDMA system [1] with antenna array at the receiver. The angle (or delay) stationarity has to be compared to the spatial (or temporal) resolution of the array (or the waveform). For instance, consider a base station with an array of $M = 8$ half wavelength spaced elements and a mobile terminal at a distance of 500 m, the space-time channel can be considered as stationary within $L = 10$ frames (100 ms) as far as the velocity of the mobile terminal is below 1000 km/h (in this case, the temporal resolution is more restrictive compared to spatial resolution).

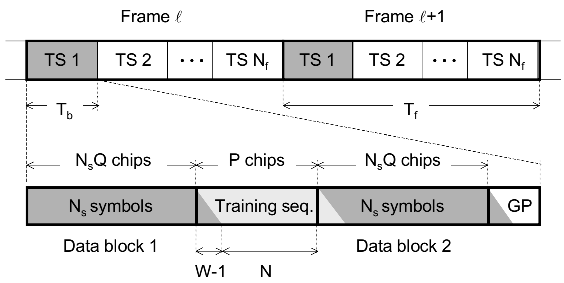

Figure 1: Frame structure of a TD-CDMA system.

HYBRID TD-CDMA SIGNAL MODEL

The frame of a time slotted CDMA system (see Figure 1) contains N_f time-slots, each can be allocated to either the uplink or the downlink. We will concentrate on the uplink and assume that the base station is equipped with a linear antenna array of M half wavelength spaced elements. Within each frame a time-slot is considered in which K users are transmitting data at the same carrier frequency but using different spreading codes. During the time-slot, each user transmits a burst that consists of two data fields of N_S information symbols (spread by a code of length Q), a midamble of P chips and a guard period. Within the frame only one time-slot is allocated to the K users of interest, so that the temporal interval between two successive bursts equals the frame duration T_f. Under these conditions the multi-slot approach can be carried out on a frame-by-frame basis as multi-frame processing.

The midamble of the kth user ($k = 1, .., K$) contains a training sequence of N chips, $\{x_k(i)\}_{i=0}^{N-1}$, that is known at the receiver and is used for channel estimation. The last $W-1$ chips of the sequence are repeated at the beginning of the midamble as a cyclic prefix ($P = N + W - 1$).

The multipath channel from the kth user to the receiving antennas is assumed to be quasi-static within the burst interval (or at least within the training period) but varying from frame to frame. It is modelled by the $M \times 1$ impulse response vector $\mathbf{h}_k(t;\ell)$ (ℓ denotes the frame index) whose temporal support is $t \in [0, WT)$ (T is the chip interval). This implies that the training sequences are received free from the interference of data symbols.

At each antenna the baseband received signal is passed through a filter matched to the transmitted pulse waveform and sampled at the symbol rate $1/T$. The discrete-time impulse response for the channel of the kth user is represented by the $M \times W$ space-time matrix $\mathbf{H}_k(\ell) = [\mathbf{h}_k(0;\ell)\ \mathbf{h}_k(T;\ell) \cdots \mathbf{h}_k((W-1)T;\ell)]$, which includes the transmit filter, the multipath propagation and the receive filter. By arranging N time samples of the signal received within the ℓth training period into a $M \times N$ matrix $\mathbf{Y}(\ell)$, the signal model can be expressed as

$$\mathbf{Y}(\ell) = \sum_{k=1}^{K} \mathbf{H}_k(\ell)\mathbf{X}_k + \mathbf{N}(\ell) = \mathbf{H}(\ell)\mathbf{X} + \mathbf{N}(\ell), \quad (1)$$

where $\mathbf{H}(\ell) = [\mathbf{H}_1(\ell) \cdots \mathbf{H}_K(\ell)]$, $\mathbf{X} = [\mathbf{X}_1^T \cdots \mathbf{X}_K^T]^T$, $\mathbf{X}_k$ is the $W \times N$ convolution matrix with $[\mathbf{X}_k]_{m,n} = x_k(n-m)$. Recall that $x_k(n+N) = x_k(n)$. The additive noise $\mathbf{N}(\ell)$ is temporally uncorrelated but spatially correlated with covariance $\mathbf{R}$: $\mathrm{E}[\mathbf{N}(\ell)\mathbf{N}^H(\ell+m)]/N = \mathbf{R}\delta(m)$. The problem addressed herein is the joint estimation of the channel matrix $\mathbf{H}$ and the noise covariance $\mathbf{Q}$ from the received signals $\mathbf{Y}$ and the known training sequence $\mathbf{X}$.

SUBSPACE STRUCTURE OF THE PROPAGATION CHANNEL

According to the multipath propagation model, the time varying channel $\mathbf{H}_k(\ell)$ is modelled as the superposition of P_k paths, the ith path is characterized by the direction of arrival $\vartheta_{k,i}$, the delay $\tau_{k,i}$ and the complex valued amplitude $\alpha_{k,i}(\ell)$. As discussed in the introduction, the angle/delay pattern varies slowly and therefore the pair $\{\vartheta_{k,i}, \tau_{k,i}\}$ is assumed to be stationary over L time slots (L is properly selected according to the terminal mobility). On the other hand, $\alpha_{k,i}(\ell)$ is slot-dependent as the fading amplitudes are fast varying due to the movement of the user. The channel matrix can be rewritten as

$$\mathbf{H}_k(\ell) = \mathbf{A}_k\mathbf{D}_k(\ell)\mathbf{G}_k^T, \quad \ell = 1, \ldots, L, \quad (2)$$

The $W \times P_k$ temporal component of the channel $\mathbf{G}_k = [\mathbf{g}(\tau_{k,1}), ..., \mathbf{g}(\tau_{k,P_k})]$ depends on the set of delays $\boldsymbol{\tau}_k = [\tau_{k,1}, ..., \tau_{k,P_k}]^T$, each column contains the delayed waveform $[\mathbf{g}(\tau_{k,i})]_n = g((n-1)T - \tau_{k,i})$. Similarly, the $M \times P_k$ spatial response matrix $\mathbf{A}_k = [\mathbf{a}(\vartheta_{k,1}), ..., \mathbf{a}(\vartheta_{k,P_k})]$ depends on the set of angles $\boldsymbol{\vartheta}_k = [\vartheta_{k,1}, ..., \vartheta_{k,P_k}]^T$, where $[\mathbf{a}(\vartheta_{k,i})]_m = \exp(j\pi(m-1)\sin\vartheta_{k,i})$. The diagonal matrix $\mathbf{D}_k(\ell) = \mathrm{diag}[\alpha_{k,1}(\ell), \ldots, \alpha_{k,P_k}(\ell)]$ contains the fading amplitudes that are assumed to follow the WSSUS channel model, with $\mathrm{E}[\mathbf{D}_k(\ell+m)\mathbf{D}_k(\ell)^H] = \rho_k(m) \cdot \mathrm{diag}[\sigma_{k,1}^2, ..., \sigma_{k,P_k}^2]$. According to the assumption of quasi-static channel within the time-slot, the normalized correlation function $\rho_k(m)$ depends only on the time interval mT_f and on the terminal mobility (velocity v_k). Indeed, according to the Clarke's isotropic scattering model [3], it is $\rho_k(m) = J_0(2\pi f_k m T_f)$, where $f_k = v_k/\lambda$ is the Doppler shift.

The order of spatial ($r_{\mathrm{S},k}$) and temporal ($r_{\mathrm{T},k}$) diversity are

$$r_{\mathrm{S},k} = \mathrm{rank}(\mathbf{A}_k) \leq \min(P_k, M), \quad (3)$$

$$r_{\mathrm{T},k} = \mathrm{rank}(\mathbf{G}_k) \leq \min(P_k, W), \quad (4)$$

respectively. The first accounts for the number of angles that can be resolved in $\boldsymbol{\vartheta}_k$ (given the array aperture), while the second equals the number of the resolvable delays in $\boldsymbol{\tau}_k$ (given the bandwidth of the transmitted signal). In many practical situations P_k can be very large, but the order of diversity depends only on few clusters of scatterers with moderate angle-delay spread. When this occurs the diversity orders $r_{\mathrm{S},k}$ and $r_{\mathrm{T},k}$ are reduced and a parsimonious channel parameterization can be adopted as described below.

Let $\mathbf{U}_{\mathrm{S},k}$ ($M \times r_{\mathrm{S},k}$) and $\mathbf{U}_{\mathrm{T},k}$ ($W \times r_{\mathrm{T},k}$) be the burst-independent orthonormal basis for the column spaces $\mathcal{R}(\mathbf{A}_k)$ and $\mathcal{R}(\mathbf{G}_k)$, the channel matrix $\mathbf{H}_k(\ell)$ can be parametrized as

$$\mathbf{H}_k(\ell) = \mathbf{U}_{\mathrm{S},k}\boldsymbol{\Gamma}_k(\ell)\mathbf{U}_{\mathrm{T},k}^H, \quad (5)$$

where $\boldsymbol{\Gamma}_k(\ell)$ is a $r_{\mathrm{S},k} \times r_{\mathrm{T},k}$ burst-dependent matrix (non-diagonal). The parameterization (5) can be easily obtained from the model (2) by considering the singular value decompositions $\mathbf{A}_k = \mathbf{U}_{\mathrm{S},k}\boldsymbol{\Sigma}_{\mathrm{S},k}\mathbf{V}_{\mathrm{S},k}^H$, $\mathbf{G}_k = \mathbf{U}_{\mathrm{T},k}\boldsymbol{\Sigma}_{\mathrm{T},k}\mathbf{V}_{\mathrm{T},k}^H$, and by further defining $\boldsymbol{\Gamma}_k(\ell) = \boldsymbol{\Sigma}_{\mathrm{S},k}\mathbf{V}_{\mathrm{S},k}^H\mathbf{D}_k(\ell)\mathbf{V}_{\mathrm{T},k}\boldsymbol{\Sigma}_{\mathrm{T},k}^H$. In the following $\mathcal{R}\{\mathbf{A}_k\} = \mathcal{R}(\mathbf{U}_{\mathrm{S},k})$ and $\mathcal{R}\{\mathbf{G}_k\} = \mathcal{R}(\mathbf{U}_{\mathrm{T},k})$ will be referred to as the spatial and temporal subspaces for the kth channel.

MULTI-SLOT CHANNEL ESTIMATION

The unconstrained maximum likelihood estimates (MLE) for the multi-channel $\mathbf{H}(\ell)$ and the noise covariance $\mathbf{R}$ are known to be:

$$\hat{\mathbf{H}}(\ell) = [\hat{\mathbf{H}}_1(\ell) \cdots \hat{\mathbf{H}}_K(\ell)] = \mathbf{Y}\mathbf{X}^H(\mathbf{X}\mathbf{X}^H)^{-1}, \quad (6)$$

$$\hat{\mathbf{R}} = \frac{1}{NL}\sum_{\ell=1}^{L}(\mathbf{Y} - \hat{\mathbf{H}}(\ell)\mathbf{X})(\mathbf{Y} - \hat{\mathbf{H}}(\ell)\mathbf{X})^H. \quad (7)$$

By introducing the subspace constraint (5) for the multi-slot channel, the maximum likelihood estimate for the kth user can be expressed as (*Multi-Slot Space-Time estimator*[5]-[6])

$$\hat{\mathbf{H}}_{\mathrm{MS},k}(\ell) = \hat{\mathbf{R}}^{H/2}\mathbf{\Pi}_{\mathrm{S},k}\widetilde{\mathbf{H}}_k(\ell)\mathbf{\Pi}_{\mathrm{T},k}\mathbf{R}_{xx,k}^{-H/2}, \quad (8)$$

where $\mathbf{R}_{xx,k} = \mathbf{X}_k\mathbf{X}_k^H/N$ denotes the correlation of the kth training sequence, while the correlation between the training sequences of different users is assumed to be negligible. Furthermore, $\widetilde{\mathbf{H}}_k(\ell)$ represents the whitened channel estimate defined as

$$\widetilde{\mathbf{H}}_k(\ell) = \hat{\mathbf{R}}^{-H/2}\hat{\mathbf{H}}_k(\ell)\mathbf{R}_{xx,k}^{H/2}. \quad (9)$$

The spatial and temporal projectors $\Pi_{\mathrm{S},k}$ and $\Pi_{\mathrm{T},k}$ represent the projectors onto the subspaces spanned by the $r_{\mathrm{S},k}$ principal eigenvectors of the spatial correlation matrix $\mathbf{R}_{\mathrm{S},k}$ and the $r_{\mathrm{T},k}$ principal eigenvectors of the temporal correlation matrix $\mathbf{R}_{\mathrm{T},k}$

$$\mathbf{R}_{\mathrm{S},k} = \frac{1}{L}\sum_{\ell=1}^{L}\widetilde{\mathbf{H}}_k(\ell)\widetilde{\mathbf{H}}_k^H(\ell), \quad (10)$$

$$\mathbf{R}_{\mathrm{T},k} = \frac{1}{L}\sum_{\ell=1}^{L}\widetilde{\mathbf{H}}_k^H(\ell)\widetilde{\mathbf{H}}_k(\ell). \quad (11)$$

The noise covariance matrix can be then estimated from the residuals of the channel estimation $\hat{\mathbf{N}}(\ell) = \mathbf{Y}(\ell) - \sum_{k=1}^{K}\hat{\mathbf{H}}_{\mathrm{MS},k}(\ell)\mathbf{X}_k$, as: $\hat{\mathbf{Q}}_{\mathrm{MS}} = \sum_{\ell=1}^{L}\hat{\mathbf{N}}(\ell)\hat{\mathbf{N}}^H(\ell)/LN$.

Notice that in a dense multipath radio environment the degree of temporal diversity could be as large as the support of the channel. If the temporal order rises to $r_{\mathrm{T},k} \simeq W$, it might be convenient to neglect the temporal projection and let $\Pi_{\mathrm{T},k} = \mathbf{I}_W$ in (8). The resulting channel estimate exploits the stationarity of the spatial subspace only and it is referred to as *Multi-Slot Space estimator*. Dually, for a large angle spread and/or a small number of antennas ($r_{\mathrm{S},k} \simeq M$), it could be advisable not to use the spatial projection and set $\mathbf{\Pi}_{\mathrm{S},k} = \mathbf{I}_M$ in (8). This leads to the *Multi-Slot Time estimator* that exploits the stationarity of the temporal subspace only.

Example. The multi-slot solution is obtained by collecting instantaneous channel estimates as the user moves and then using these observations to calculate the slowly varying spatial-temporal basis by averaging over the fast-faded amplitudes. This is illustrated by an example in Figure 2. The multipath propagation is composed of $P = 5$ paths having $\{\sigma_i^2\}_{i=1}^{P} = \{0.33, 0.25, 0.19, 0.14, 0.083\}$. The path pattern is described in Figure 2-a, while Figure 2-b shows the power-delay-angle diagram for a noisy estimate $\hat{\mathbf{H}}(\ell)$. Since $\alpha_1 = \alpha_2$, $\alpha_3 = \alpha_4$ and $\tau_4 = \tau_5$, the spatial and temporal diversity orders are, respectively, $r_{\mathrm{S}} = 3$ and $r_{\mathrm{T}} = 4$. The projector Π_{S} onto the invariant spatial subspace is calculated from $\hat{\mathbf{R}}_{\mathrm{S}}$ by using the diversity order $\hat{r}_{\mathrm{S}} = 1 \div 3$. The same holds for Π_{T} with $\hat{r}_{\mathrm{T}} = 1 \div 4$. As shown in Fig. 2-c, d and e, by projecting the space-time matrix $\hat{\mathbf{H}}(\ell)$ the background noise is reduced.

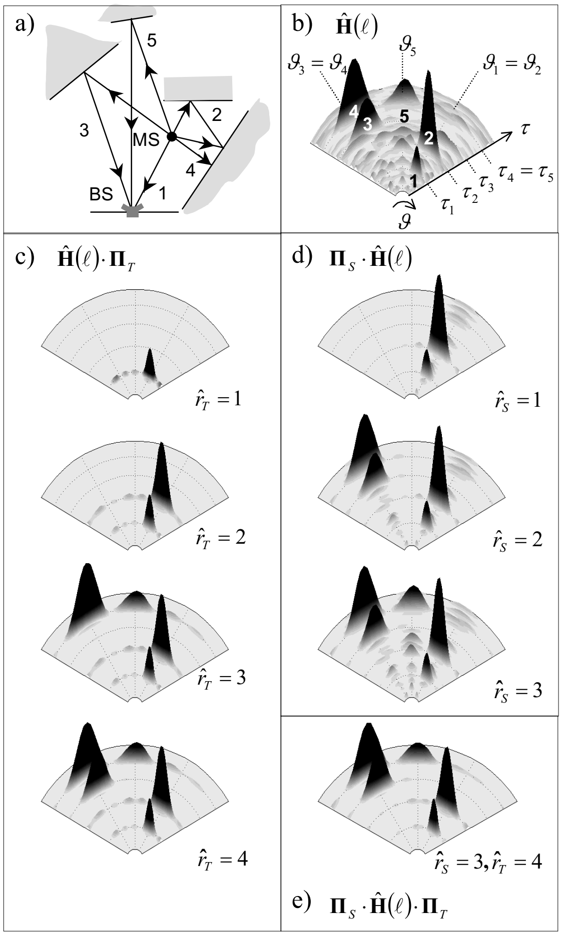

Figure 2: Example of space and/or time projections for a channel with $P = 5$ paths and different degrees of space and time diversity: $r_S = 3$ and $r_T = 4$. a) multipath model; b) power-delay-angle diagram from the LSE; projection of $\hat{\mathbf{H}}(\ell)$ onto temporal (c) or spatial (d) subspaces with increasing dimensions; e) projection onto spatial and temporal subspaces.

SIMULATION RESULTS

In order to validate the performance of the MS algorithm we simulate the uplink of a TD-SCDMA system with a ZF-BLE receiver [4] for $M = 8$ antennas. According to the specification [1], the parameters of interest are: $T_f = 10$ ms, $W = 16$, $N = 128$, $P = N + W - 1 = 144$, roll-off equal to 0.22, carrier central frequency 1950 MHz. Moreover, each one of the $K = 8$ users transmits at the same rate using a spreading factor $Q = 16$: the data field thus contains $N_S = 22$ QPSK information symbols (352 chips). The K training sequences are chosen according to the standard specifications. We consider a simple radio environment characterized by two clusters with angles randomly chosen within the angular support $[-60, 60]$ deg, four paths per cluster ($P_k = 8$) with angular dispersion of 5 deg and an exponential power profile: $\sigma_{k,i}^2 = (0.5)^{i-1}$ (paths $i = 1, .., 4$

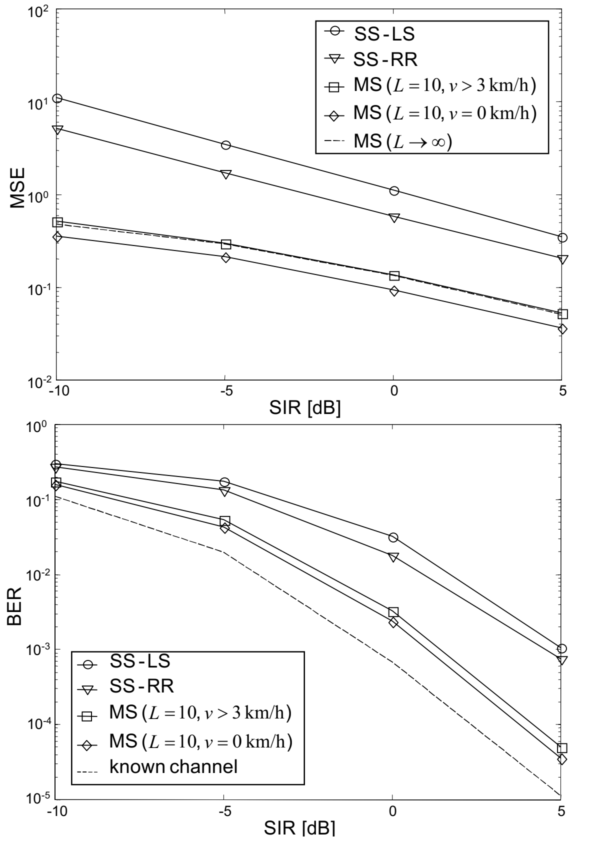

Figure 3: MSE (upper figure) and BER (lower figure) vs. SIR for single-slot (SS) techniques (LS and RR) and for the multi-slot (MS) algorithm with $L = 10$ slot (TD-SCDMA with $K = 8$ users [1]). The BER for known channel is shown as reference.

correspond to the first cluster and paths $i = 5,..,8$ to the second). In each cluster the delays are chip interval spaced and the first delay is randomly selected in the interval $[2, 10]\,T$. The fading variation is simplified by assuming the same velocity $v_k = v$ for all the users. The covariance matrix of noise is $\mathbf{R} = \mathbf{R}_{\rm i} + \mathbf{R}_{\rm n}$, where $\mathbf{R}_{\rm i} = \sigma_{\rm i}^2 \sum_{h=1}^{6} \mathbf{a}(\vartheta_h)\,\mathbf{a}(\vartheta_h)^H$ models the interference from 6 out-of-cell terminals with angles ϑ_h uniformly spaced in $[-60\ 60]$ deg, while $\mathbf{R}_{\rm n}{=}\sigma_{\rm n}^2\mathbf{I}_M$ accounts for the background AWGN noise. The signal-to-interference ratio is defined as $\mathrm{SIR} = \sigma_x^2/\sigma_{\rm i}^2$ and the signal-to-(background) noise as $\mathrm{SNR} = \sigma_x^2/\sigma_{\rm n}^2 = 30$ dB, where $\mathrm{tr}(\mathbf{R}_{xx}) = \sigma_x^2 KWN$. The channel of each user is normalized so that $\mathrm{E}[||\mathbf{H}_k(\ell)||^2] = 1$. The diversity orders ($r_{\mathrm{S},k}$ and $r_{\mathrm{T},k}$) are estimated by using the minimum description length (MDL) criterion from the covariance matrices $\mathbf{R}_{\mathrm{S},k}$ and $\mathbf{R}_{\mathrm{T},k}$.

In Figure 3 the performance of the MS space-time algorithm is evaluated in terms of mean square error (MSE) of the channel estimate and average bit error rate (BER). The multi-slot estimate for $L = 10$ is compared to the single-slot techniques LS and RR (Reduced rank [5]). For any mobility of the terminal, from the pedestrian environment ($v = 3$ km/h), up to the vehicular ($v = 120$ km/h), the performance of the MS method are the same. On the other hand, for a static channel ($v = 0$ km/h) the MS technique performs better since the subspaces dimensions $r_{\mathrm{S},k}$ and $r_{\mathrm{T},k}$ can be reduced (see [6]). The upper figure shows the MSE obtained by the MS algorithm with $L = 10$, for static and time-varying ($v > 3$ km/h) channels. For time-varying channel the simulated MSE is compared with the analytical lower bound computed for $L \to \infty$ in [6]. It can be noticed that the simulated results are very close to the theoretical limit. The lower figure compares the BER performance of the ZF-BLE receiver based on LS (6) and MS (8) channel estimates. The MS method shows a meaningful advantage in term of SIR (approx. 3-4 dB) that is practically independent on the variation of the faded amplitudes (lower figure). In addition, the loss with respect to the ideal case of known channel (dashed line) is approx. 1.5 dB.

CONCLUSIONS

The proposed multi-slot estimation methods exploit the invariance over the slots of both the spatial and the temporal subspaces, without estimating explicitly delays and angles of arrival. The paper has focused on the uplink of a time-slotted CDMA system. Nonetheless, the approach (with straightforward modifications) appears to be a valid solution for a broader class of systems, such as TD-CDMA, FDMA, multicarrier or generically MIMO.

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