MULTIRATE FILTER BANKS FOR CODE-DIVISION MULTIPLE ACCESS SYSTEMS

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ABSTRACT

Direct sequence, code-division multiple access (CDMA) schemes offer an attractive alternative for sharing a transmission medium among many users, while requiring minimal co-operation among them. A number of signal processing issues are related to the receiver's task of multiuser information extraction and detection. In this paper, a discrete-time multirate formulation is introduced for asynchronous CDMA systems, which can incorporate multipath effects. In this framework, linear receivers are derived which can completely suppress multiuser interference (decorrelating receivers). A criterion is introduced, which guarantees the decorrelating property, while providing optimal solutions in the presence of noise. The synchronization problem in a multipath environment is also studied, and identifiability conditions are established. A subspace algorithm is proposed, to estimate the user's delays and multipath channels in a blind scenario.

1. INTRODUCTION

In several emerging applications like wireless, satellite and mobile communications, a number of users is required to share the same transmission medium. Unfortunately, classical methods of assigning channel resources like frequency or time division multiple access, do not make efficient use of the channel when the users transmit information only for a small percentage of time. For example, in mobile telephony users typically talk less than 50% of the time, and hence underuse their assigned slot.

Dynamic sharing of the transmission medium can be achieved at the protocol level, using random access methods with some packet collision detection and avoidance strategy. However, code division multiple access (CDMA) schemes offer the flexibility of random access with minimal cooperation among users and no need for re-transmissions.

In direct sequence CDMA systems, each user transmits information using a fixed and distinct signature waveform, with no synchronization with the other users (e.g., [10]). The receiver uses this signature information to distinguish among data from different users.

The receiver's task involves a number of different information processing steps, ranging from multiuser and intersymbol interference (MUI and ISI) suppression [3], to synchronization and detection [7], [10]. Recent discrete-time models for CDMA systems have resulted in vector formulations which have allowed the use of signal processing techniques to address these problems [7], [3]. However, inherent changes in the data rate, resulting from spreading and despreading operations at the transmitter and receiver respectively, have not been explicitly identified and fully exploited so far. In this paper, we attempt to bridge this gap by developing a discrete-time multirate filter bank model for asynchronous CDMA systems. The proposed formulation can incorporate asynchronous user delays and multipath effects. By establishing a link between CDMA techniques and multirate signal processing, the former can benefit from the extensive work that has been done on filter banks, perfect reconstruction structures, and multichannel system theory.

In this framework the first goal of the paper is to investigate optimal linear decoding techniques. The conditions for the existence of FIR linear processors, which completely suppress MUI and ISI, are studied (sometimes called zero-forcing or decorrelating receivers), and a design method is proposed. Furthermore, a new criterion is introduced, which guarantees the zero-forcing property and its minimization provides optimal solutions in the presence of noise.

The second goal of the paper is to address the synchronization problem in the challenging case where multipath is present as well as near-far effects, i.e., the users have unequal powers. We show the conditions under which each user's delay and multipath channel can be blindly estimated using second-order information only. We also propose a subspace method for identifying those parameters.

2. CDMA SYSTEMS AS FILTER BANKS

Let user j, $j = 1, \ldots, J$, transmit an information symbol stream $w_j(n)$ using a spreading sequence of length P, $c_j(n)$, $n = 0, \ldots, P-1$. Then, the transmitted discrete-time signal at the chip rate is

$$s_j(n) = \sum_{k=-\infty}^{\infty} w_j(k) c_j(n - kP) \ , \qquad (1)$$

The sequence $s_j(n)$ is transmitted using a rectangular chip pulse of duration T_c, modulated at the carrier frequency. If the receiver is not synchronized with user j, then after demodulation, matched filtering, and sampling at the chip rate, and assuming no multipath, the received signal is

$$y(n) = \sum_{j=1}^{J} y_j(n) + v(n) \ , \qquad (2)$$

$$y_j(n) = A_j(1 - \delta_j) s_j(n - \tau_j) + A_j \delta_j s_j(n - \tau_j - 1) \ ,$$

where τ_j is an integer, $\delta_j \in [0,1)$, and $\tau_j + \delta_j$ is the total delay of user j in chip periods T_c; A_j is an attenuation factor and $v(k)$ is additive noise. A_j and $v(k)$ may be complex if both in phase and quadrature phase demodulation is used. Also, without loss of generality we limit $0 \leq \tau_j < P$, since delays which are multiples of the bit period amount to simple renumbering of the input and output sequences.

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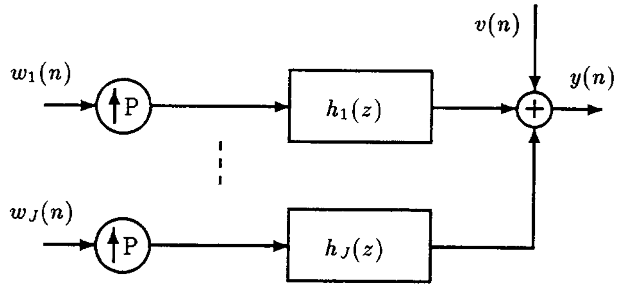

Figure 1. A CDMA system as a J-band filter bank

By substituting (1) in (2) we obtain

$$\begin{aligned} y_j(n) &= \sum_{k=-\infty}^{\infty} w_j(k)[A_j(1-\delta_j)c_j(n-\tau_j-kP) \\ &+ A_j\delta_j c_j(n-\tau_j-1-kP)] \\ &= \sum_{k=-\infty}^{\infty} w_j(k)h_j(n-kP) \end{aligned} \quad (3)$$

where $h_j(n) = \sum_{l=0}^{1} g_j(l)c_j(n-\tau_j-l)$, $g_j(0) = A_j(1-\delta_j)$ and $g_j(1) = A_j\delta_j$. Equation (3) indicates the multirate nature of CDMA systems and suggests the formulation depicted in Fig. 1. The equivalent discrete-time channel for user j in the z-domain is $h_j(z) = z^{-\tau_j}c_j(z)g_j(z)$. Notice that this description can be easily extended to incorporate fading or multipath effects by allowing $g_j(z)$ to be a general complex channel response, (not necessarily parametrized by δ_j), and of order perhaps greater than one.

It will be useful in the sequel to provide a multichannel analogue of the multirate formulation of (3), using the polyphase decomposition [9]. Let ^{T} denote transpose and $\mathbf{y}(n):=[y(nP), y(nP+1)\dots y(nP+P-1)]^T$ be the length P, polyphase representation of $y(n)$, with z-transform $\mathbf{y}(z)$. Similarly we define $\mathbf{y}_j(z)$, $\mathbf{h}_j(z)$, $\mathbf{g}_j(z)$, $\mathbf{c}_j(z)$, $\mathbf{v}(z)$. Note that $\mathbf{c}_j(z) = \mathbf{c}_j:=[c_j(0)\dots c_j(P-1)]^T$ is a constant vector, since $c_j(n)$ has length no greater than P. With these definitions we may write

$$\begin{aligned} \mathbf{y}(z) &= \sum_{j=1}^{J}\mathbf{y}_j(z) + \mathbf{v}(z) = \sum_{j=1}^{J}\mathbf{h}_j(z)w_j(z) + \mathbf{v}(z) \\ &= \mathbf{H}(z)\mathbf{w}(z) + \mathbf{v}(z) \ , \end{aligned} \quad (4)$$

where the matrix $\mathbf{H}(z):=[\mathbf{h}_1(z)\dots\mathbf{h}_J(z)]$, and $\mathbf{w}(z) = [w_1(z)\dots w_J(z)]^T$. Equation (4) provides a multichannel description equivalent to (3).

A number of different vector formulations have been used in CDMA studies [7], [3], but they lack the generality of (3), (4) and the link with the established theory of multichannel and multirate systems. Using this formulation we investigate linear optimal detection schemes in the next section.

3. OPTIMAL LINEAR RECEIVERS

In this section we assume the delays, powers, and any other parameter necessary to construct $\mathbf{H}(z)$ to be given or estimated. We focus on deconvolving and estimating the transmitted information symbols $w_j(n)$. Issues related to the estimation of $\mathbf{H}(z)$ are deferred to the next section.

It is well known that if $\mathbf{H}(z)$ is given, a maximum likelihood sequence estimation procedure can be employed using the Viterbi algorithm, to detect $w_j(n)$ [10]. Because of its computational complexity however (exponential in the number of users), considerable interest exists for linear receivers, i.e., structures similar to the one in Fig. 2. In multirate processing, a filter bank is usually designed to analyze a signal, and then reconstruct it from its components. In the current problem, the "reconstructed" signal $y(n)$ is given, and one seeks an analysis bank to recover the individual components.

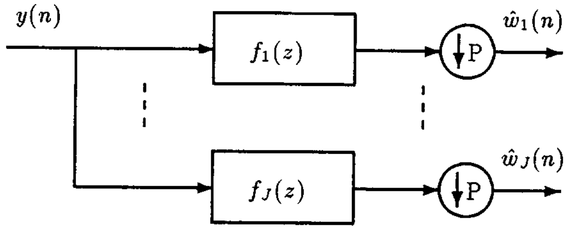

Figure 2. A multirate linear CDMA receiver

The simplest linear approach has been the *single user receiver*, where a simple matched filtering operation is performed, after the receiver synchronizes with user j, ($\tau_j = \delta_j = 0$), i.e., $\hat{w}_j(n) = \mathbf{c}_j^{*T}\mathbf{y}(n)$ (e.g., [10]). MUI suppression in this case, depends solely on the degree of orthogonality among the received signature sequences. Unfortunately, this method is susceptible to near-far effects; that is, it can not decode weak signals which are overwhelmed by interfering strong users.

There exist linear receivers which are near-far resistant and in fact completely suppress MUI and ISI [5], called zero-forcing or decorrelating receivers. A row vector $\mathbf{f}_j^T(z)$ represents a zero forcing receiver if in the absence of noise

$$\mathbf{f}_j^T(z)\mathbf{y}(z) = w_j(z) \;\Leftrightarrow\; \mathbf{f}_j^T(z)\mathbf{H}(z) = \mathbf{e}_j^T \ , \quad (5)$$

where $\mathbf{e}_j^T:=[0,\dots,0,1,0,\dots,0]$ with 1 at the jth position. In most cases in CDMA literature, only a memoryless $\mathbf{f}_j(z) = \mathbf{f}_j$ is considered [3]. If $\mathbf{H}(z) = \mathbf{H}_0 + \mathbf{H}_1z^{-1} + \cdots + \mathbf{H}_qz^{-q}$, then (5) implies that

$$\mathbf{f}_j^T[\mathbf{H}_0\mathbf{H}_1\dots\mathbf{H}_q] = [\mathbf{e}_j^T, \mathbf{0}\dots\mathbf{0}] \ . \quad (6)$$

Equation (6) can not be guaranteed to hold if $(q+1)J > P$, since then the system becomes overdetermined and no exact solution $\mathbf{f}_j$ can be obtained. Hence, a memoryless decorrelating receiver need not exist in this case. As an example, if no multipath is present, (hence $q = 1$), this approach is not applicable to systems with more than $J = P/2$ users.

A question that naturally arises at this point is under what conditions an FIR decorrelating receiver $\mathbf{f}_j(z)$ (not necessarily memoryless) exists. Equivalent problems have been extensively studied in multivariate system theory and the conditions are summarized below in CDMA terms (see [2]).

Proposition 1 : *There exists an FIR zero forcing receiver* $\mathbf{f}_j(z)$ *such that (5) is satisfied if:*

(C1) $\text{rank}\{\mathbf{H}(z)\} = J$, $\forall z \in \mathbf{C}$, $z \neq 0$, *and* $\text{rank}\{\mathbf{H}_0\} = J$. *Then* $\mathbf{H}(z)$ *is called irreducible.* □

In the case of a single user, (**C1**) reduces to a well known identifiability result, in the context of fractionally spaced equalizers [8].

In the sequel we will assume that $\mathbf{H}(z)$ also satisfies:

(C2) $\text{rank}\{[\mathbf{h}_1(q_1)\dots\mathbf{h}_J(q_J)]\} = J$, *where* q_j *is the order of* $\mathbf{h}_j(z)$. *Then* $\mathbf{H}(z)$ *is called column reduced.*

Condition (C2) is not necessary for Proposition 1, but we assume it holds for technical convenience and without loss of generality, since any $\mathbf{H}(z)$ can be brought to a column reduced form with elementary column operations [2]. The same conditions were identified by Slock [6] for a zero forcing equalizer in a different multichannel equalization setup.

If we consider a receiver $\mathbf{f}_j(z)$ of order $M-1$, then (5) can be written in the time-domain as

$$\boldsymbol{f}_{j,M}^T \mathcal{S}_M(\mathbf{H}) = [\mathbf{0}^T \dots \mathbf{0}^T, \mathbf{1}^T, \mathbf{0}^T \dots \mathbf{0}^T] := \mathbf{1}_j^T \ , \quad (7)$$

where the super-vector $\boldsymbol{f}_{j,M} = [(\boldsymbol{f}_{j,M}^{(1)})^T, \dots, (\boldsymbol{f}_{j,M}^{(P)})^T]^T$ and each vector $\boldsymbol{f}_{j,M}^{(p)}$ is the inverse z-transform of $\mathbf{f}_j^{(p)}(z)$ (the pth component of $\mathbf{f}_j(z)$); $\mathcal{S}_M(\mathbf{H}) = [\mathcal{S}_M(\mathbf{h}_1), \dots, \mathcal{S}_M(\mathbf{h}_P)]$, has matrix elements $\mathcal{S}_M(\mathbf{h}_j)$ denoting the Sylvester matrix of $\mathbf{h}_j(n)$, i.e., $\mathcal{S}_M(\mathbf{h}_j) = [\mathcal{T}_M^T(\mathbf{h}_j^{(1)}), \dots, \mathcal{T}_M^T(\mathbf{h}_j^{(P)})]^T$, where

$$\mathcal{T}_M(\mathbf{h}_j^{(p)}) = \begin{bmatrix} \mathbf{h}_j^{(p)}(0) & \cdots & \mathbf{h}_j^{(p)}(q_j) & & 0 \\ & \ddots & & \ddots & \\ 0 & & \mathbf{h}_j^{(p)}(0) & \cdots & \mathbf{h}_j^{(p)}(q_j) \end{bmatrix} ,$$

is a Toeplitz filtering matrix with M rows. Finally, $\mathbf{1} = [1, 0, \dots, 0]^T$ in (7).

Under conditions (C1) and (C2), $\mathcal{S}_M(\mathbf{H})$ has full rank; hence an exact solution exists if M is chosen such that $\mathcal{S}_M(\mathbf{H})$ in (7) is square or underdetermined [6].

If indeed M is chosen sufficiently large so that (7) is underdetermined, there are a number of degrees of freedom in selecting a receiver that satisfies the zero forcing constraint. The natural question that arises in this context, is which one among the candidate receivers should be chosen. It is clear that in the absence of noise, all these receivers are equivalent. When noise is present however, it is reasonable to select one which minimizes the mean square error

$$\text{MSE}(j) = E\{|w_j(n) - \hat{w}_j(n)|^2\} \ , \quad (8)$$

where

$$\hat{w}_j(z) = \mathbf{f}_j^T(z)\mathbf{y}(z) \ , \quad (9)$$

subject to the zero-forcing constraint of (7). The solution to this optimization problem is given by the following proposition.

Proposition 2 : *Under conditions (C1), (C2), and with M such that (7) has at least one solution, the parameter vector $\boldsymbol{f}_{j,M}$ which minimizes (8) subject to (7) is*

$$\boldsymbol{f}_{j,M}^T = \mathbf{1}_j^T [\mathcal{S}_M^{*T}(\mathbf{H}) \mathbf{R}_{\boldsymbol{vv}}^{-1} \mathcal{S}_M(\mathbf{H})]^{-1} \mathcal{S}_M(\mathbf{H}) \mathbf{R}_{\boldsymbol{vv}}^{-1} \ , \quad (10)$$

where $\mathbf{R}_{\boldsymbol{vv}} = E\{\boldsymbol{v}(n)\boldsymbol{v}^{*T}(n)\}$ and $\boldsymbol{v}(n) = [\mathbf{v}^{(1)}(n) \dots \mathbf{v}^{(1)}(n-M+1)| \dots |\mathbf{v}^{(P)}(n) \dots \mathbf{v}^{(P)}(n-M+1)]^T$. □

Note that the solution given by (10) is independent of the statistics of $w_j(n)$. Also if $v(n)$ is white noise, hence $\mathbf{R}_{\boldsymbol{vv}} = \sigma_v^2 \mathbf{I}$, then (10) becomes

$$\boldsymbol{f}_{j,M} = \mathbf{1}_j^T [\mathcal{S}_M^{*T}(\mathbf{H}) \mathcal{S}_M(\mathbf{H})]^{-1} \mathcal{S}_M(\mathbf{H}) \ , \quad (11)$$

and does not depend on the additive noise power.

The following proposition provides the optimal solution when the length of the receiver M is not constrained and is alowed to extend to infinity. Let the spectral matrix of the process $\mathbf{v}(n)$ be defined as

$$\mathbf{S}_{\mathbf{VV}}(\omega) = lim_{N \to \infty} N^{-1} E\{\mathbf{v}(\omega)\mathbf{v}^{*T}(\omega)\} \ . \quad (12)$$

Then it can be shown that:

Proposition 3 : *If $\text{rank}\{\mathbf{H}(\omega)\} = J$, $\forall \omega$ and also $\text{rank}\{\mathbf{S}_{\mathbf{VV}}(\omega)\} = J$, $\forall \omega$, then the receiver $\mathbf{f}_j(\omega)$ which minimizes (8) subject to (5) is*

$$\mathbf{f}_j^T(\omega) = \mathbf{e}_j^T [\mathbf{H}(\omega) \mathbf{S}_{\mathbf{VV}}^{-1}(\omega) \mathbf{H}(\omega)]^{-1} \mathbf{H}(\omega) \mathbf{S}_{\mathbf{VV}}^{-1}(\omega) \ . \quad (13)$$

□

Again notice that if $\mathbf{v}(n)$ is white and $\mathbf{S}_{\mathbf{VV}}(\omega) = \sigma_v^2 \mathbf{I}$, equation (13) becomes

$$\mathbf{f}_j(\omega) = \mathbf{e}_j^T [\mathbf{H}(\omega)\mathbf{H}(\omega)]^{-1} \mathbf{H}(\omega) \ , \quad (14)$$

and is independent of the noise power.

4. BLIND SYNCHRONIZATION

In the previous section we assumed that the users' delays and (possible) multipath channels were known and given. The accurate estimation of those quantities is clearly of paramount importance. In this section we investigate several issues related to blind estimation of those parameters, i.e., the case where the transmitted information sequences are not kown to the receiver.

A number of near-far resistant, blind synchronization methods have become popular recently [7]. However, no identifiability issues have been studied in relation to those methods (e.g., maximum number of identifiable users). Moreover, they are not applicable in a multipath environment. In this section we use the proposed multichannel formulation, in order to address the question whether and under which conditions second-order information is sufficient to estimate the delays and multipath channels.

If the number of users is less than the spreading factor ($J < P$), it is well known that, under some conditions, the spectral matrix $\mathbf{S}_{\mathbf{YY}}(\omega)$, admits a unique FIR spectral factorization within a constant matrix [1], $\mathbf{S}_{\mathbf{YY}}(\omega) = \mathbf{H}(\omega)\mathbf{H}^{*T}(\omega)$. Exploiting this fact, a subspace method was proposed by Meraim et. al [4], which can identify $\mathbf{H}(\omega)$ within a constant $J \times J$ matrix. Their approach is based on the subspace decomposition of $\mathbf{R}_{\boldsymbol{yy}} = E\{\boldsymbol{y}_M(n)\boldsymbol{y}_M^{*T}(n)\}$ where $\boldsymbol{y}_M(n) = [\mathbf{y}^{(1)}(n) \dots \mathbf{y}^{(1)}(n-M+1)| \dots |\mathbf{y}^{(P)}(n) \dots \mathbf{y}^{(P)}(n-M+1)]^T$. If the noise $v(n)$ is white with variance σ_v^2, it can be shown that

$$\mathbf{R}_{\boldsymbol{yy}} = \mathcal{S}_M(\mathbf{H}) \mathbf{R}_{\boldsymbol{ww}} \mathcal{S}_M^{*T}(\mathbf{H}) + \sigma_v^2 \mathbf{I} \ , \quad (15)$$

where $\boldsymbol{w}(n) = [w_1(n) \dots w_1(n-M+1)| \dots |w_J(n) \dots w_J(n-M+1)]^T$. Let us define the noise subspace to be the space generated by the eigenvectors corresponding to the smallest eigenvalue, and let $\mathbf{U}_n$ be the matrix containing those eigenvectors. Based on (15), it was shown in [4] that if (C1), (C2) are satisfied and M is sufficiently large, then $\mathbf{H}(z)$ can be identified within a constant matrix by solving

$$\mathbf{U}_n^{*T} \mathcal{S}_M(\mathbf{h}_j) = \mathbf{0} \ , \quad j = 1, \dots, J \ . \quad (16)$$

An equivalent formulation of (16) is

$$\mathcal{S}_{q_j+1}^{*T}(\mathbf{U}_n) \boldsymbol{h}_j = \mathbf{0} \ , \quad j = 1, \dots, J \ , \quad (17)$$

$$\boldsymbol{h}_j = [\mathbf{h}_j^{(1)}(0) \dots \mathbf{h}_j^{(1)}(q_j)| \dots |\mathbf{h}_j^{(P)}(0) \dots \mathbf{h}_j^{(P)}(q_j)]^T \ .$$

In the current problem, equation (17) must be solved under the constraint

$$h_j(z) = z^{-\tau_j} c_j(z) g_j(z) \ , \quad (18)$$

where $c_j(z)$ is given. Equivalently in the time domain,

$$\begin{bmatrix} \mathbf{h}_j(0) \\ \vdots \\ \mathbf{h}_j(q_j) \end{bmatrix} = \mathcal{T}^T(\mathbf{c}_j) \begin{bmatrix} \tilde{\mathbf{g}}_j(0) \\ \vdots \\ \tilde{\mathbf{g}}_j(q_{\tilde{\mathbf{g}}_j}) \end{bmatrix} \quad (19)$$

where $\tilde{g}(z) = z^{-\tau_j} g_j(z)$. The following proposition shows when this constraint is sufficient to provide a unique solution.

Proposition 4 : *Equations (17) and (19) provide a unique solution* $\mathbf{h}_j(z)$ *within a scaling ambiguity if the following assumptions hold:*

- *Conditions* (**C1**) *and* (**C2**) *are satisfied.*
- *The signature vectors* $\mathbf{c}_j$, $j = 1, \ldots, J$, *are linearly independent and the polynomials* $c_j(z)$, $j = 1, \ldots, J$, *have no roots in common.*
- *The orders of the multipath channels are such that* $\sum_{j=1}^{J} q_{g_j} < P - max\{q_{g_j}\} - J - 1$ □

The last assumption suggests that if no multipath is present (hence $q_{g_j} = 1, \forall j$), then the delays of only up to $(P/2) - 1$ users are guaranteed to be identifiable using only second order statistical information of the output.

Finally observe that the LHS of (19) coincides with $\boldsymbol{h}_j$ after a permutation of its elements. Hence, equations (17), (19) can be solved directly, by substituting the former into the latter after permuting its rows properly.

5. SIMULATIONS

In Fig. 3 we demonstrate a case where a memoryless linear receiver is not appropriate. A CDMA system is simulated with 20 asynchronous users, each having a Gold spreading sequence of length 31, with no multipath. The power of the MUI on the 10th user (a weak one), from every other user, as well as the power of the 10th user are shown in Fig. 3a, at the output of the best linear, memoryless receiver. The MUI power on the jth user, from an interfering user k, is the variance of the signal $\mathbf{f}_j^T \mathbf{h}_k(z) w_k(z)$. Fig. 3a shows that the MUI suppression here is not satisfactory. Fig. 3b shows the MUI suppression results for the same user, if a zero-forcing equalizer of order $M - 1 = 1$ is used, designed by solving (7). In this case the powers of all interfering signals are forced to zero.

Fig. 4 shows the true (solid line) and estimated (dashed line) impulse responses $h_j(n)$, for two users (a strong and a weak one) in an asynchronous case with multipath, using the proposed subspace method. The SNR with respect to the average user power was 20DB, and the data length was 100 bit symbols. The estimated channel for the strong user is indistinguishable from the true one, while a small estimation error exists for the weak user. Their respective delays can be clearly detected however, from the estimated responses.

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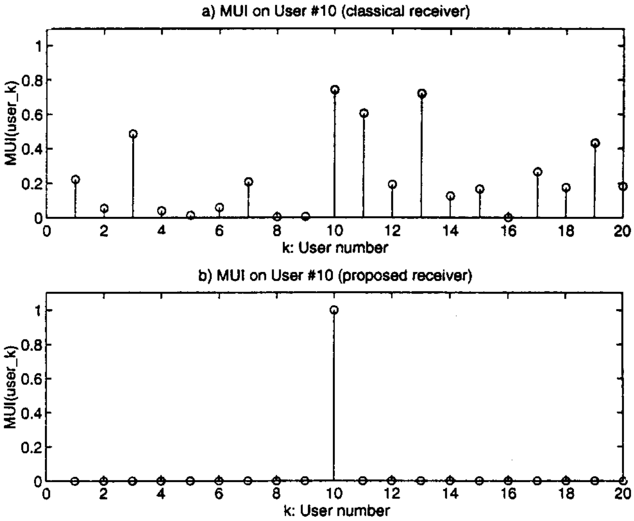

Figure 3. Zero Forcing Receiver

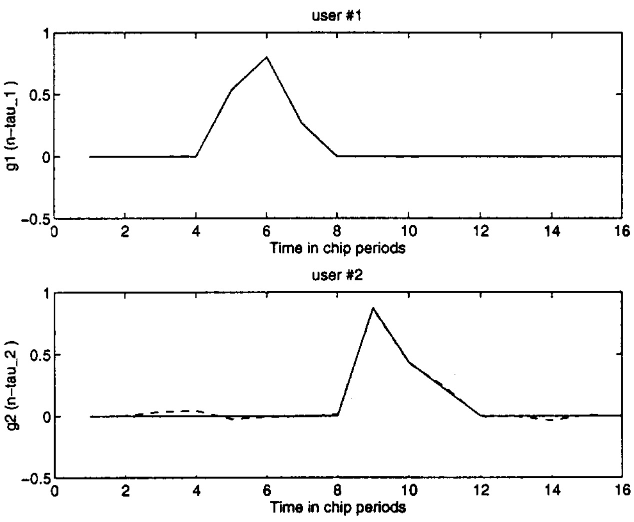

Figure 4. Synchronization in a Multipath Environment