PULSE-SHAPING FOR BLIND MULTI-USER SEPARATION IN DISTRIBUTED MISO CONFIGURATIONS

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ABSTRACT

We consider a wireless system where users are equipped with one antenna each, and the antennas are spatially distributed and not physically connected. Multiple users transmit simultaneously narrowband signals using the same carrier frequency, each one using a distinct pulse shaping function with fixed bandwidth and time duration. We show that the diversity provided by the different pulse shaping functions as well as user delays and carrier frequency offsets between transmit antennas and receiver enables user separation and recovery of transmitted signals in a blind fashion, i.e., without the need for transmission of pilot symbols. The received base-band signal is over-sampled and its polyphase components are viewed as the outputs of a virtual MIMO system. The user signals are recovered through blind MIMO system estimation and equalization.

keywords-**Multi-user Systems, Distributed Antenna Systems, Carrier Frequency Offset, Blind MIMO System Identification**

I. INTRODUCTION

Well known approaches for enabling multi-user access include TDMA, FDMA and OFDMA. These schemes provide protection from multi-user interference by allocating fixed bandwidth to each user. As such, they might not work well with multimedia sources that are known to be bursty, i.e., they alternate between periods of high activity and silence. CDMA schemes allow users to overlap in time and frequency by assigning orthogonal codes to each user. However, the codes introduce significant bandwidth expansion, the amount of which increases with the number of users that need to be accommodated. Separating multiple users using multiple receive antennas is another well studied approach. However, in a mobile wireless system, due to size limitations of the wireless receivers, it might not be practical to require each user to have multiple receive antennas.

In this work we propose a novel approach for separating simultaneously transmitting users using one receive antenna only. This is achieved by exploiting diversity provided by pulse-shaping, carrier frequency offsets (CFOs) and user delays. In a typical communication system, CFOs appear in the received signal due to relative movement between transmitter and receiver, and/or oscillator mismatch between transmit and receive antennas. They cause a frequency shift, or equivalently, a time-varying rotation of the data constellation and their effect is more severe in narrowband and multicarrier systems. User delays arise due to lack of synchronization. In multiuser systems where each user uses his own antenna, the received signal contains multiple CFOs, one for each user.

Pilot based user separation and multiple CFO estimation were proposed in [5], [2]. Training based approaches for multi-user separation in the presence of CFOs have been considered in [2], and [7]. Separation of multiple users based on their CFOs and user delays was proposed in [9],[8] for a low-rate system, where it was shown that as long as the CFOs or the delays are distinct, by oversampling the received signal by an amount P one can separate up to P users. In practice, however, naturally occurring CFOs and delays are small thus providing limited diversity.

The work present in this paper extends the work in [9],[8] in the following ways. First, we consider a high rate system, where pulse-shape functions of neighboring symbols overlap, thus there is intersymbol interference (ISI). Second, we exploit pulse-shaping diversity in addition to CFO and delay diversity. Such an approach guarantees user separation even if the CFOs are close to each other and user delays are non-existent. In particular, we consider a wireless system where users are equipped with one antenna each, and the antennas are not physically connected. Multiple users transmit narrowband signals simultaneously using the same carrier frequency, with each user using a distinct pulse shaping function. The received base-band signal is over-sampled and its polyphase components are viewed as the outputs of a virtual MIMO system. Through blind MIMO system estimation, each user signal can be recovered each within a CFO term, which can subsequently be mitigated based on existing CFO estimation results.

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II. SYSTEM MODEL

Let us consider a distributed antenna system, where K users transmit simultaneously to the base station. Narrow-band transmission is assumed here, and the channel between any user and the base station is frequency non-selective. In addition, quasi-static fading is assumed, i.e., the channel gains remain fixed during the packet length. The continuous-time base-band received signal $y(t)$ can be expressed as:

$$y(t) = \sum_{k=1}^{K} a_k x_k(t-\tau_k) e^{j2\pi F_k t} + w(t) \tag{1}$$

where a_k contains the effects of channel fading between the $k-$th user and the base station and phase offset; τ_k is the delay associated with the path between the $k-$th user and the base station; F_k is the frequency offset of the $k-$th user; $w(t)$ represents noise; $x_k(t)$ denotes the transmitted signal of user k, i.e.,

$$x_k(t) = \sum_i s_k(i) p_k(t - iT_s) \tag{2}$$

where $s_k(i)$ is the $i-$th symbol of user k; T_s is the symbol period; and $p_k(t)$ is a pulse function assigned to the k-th user. For symbol interval T_s, a typical non-rectangular pulse-shape function has support $[-T_s, T_s]$. This extended time support allows for better frequency concentration, or equivalently, less spectrum for the transmission of each symbol, but introduces intersymbol interference. Looking at $x_k(t)$ for $t \in [iT_s, (i+1)T_s)$, we find the contribution of the i-th symbol, the contribution of symbol $i+1$, due to the main lobe of $p_k(t-(i+1)T_s)$, and also contributions of symbols $i+l, l = ... -2, -1, 2, 3, ...$ due to the sidelobes of $p_k(t-(i+l)T_s), \ell = ... -2, -1, 2, 3, ...$, respectively.

Our objective is to use $y(t)$ to obtain an estimate of $\mathbf{s}(i) = [s_1(i), ..., s_K(i)]^T$ in the form:

$$\hat{\mathbf{s}}(i) = \mathbf{\Lambda}^{-1}\mathbf{P}^T\mathbf{s}(i) \tag{3}$$

where $\mathbf{P}$ is a column permutation matrix and $\mathbf{\Lambda}$ is a constant diagonal matrix. These are considered to be trivial ambiguities, and are typical in any blind problem.

The MISO problem of (1) is transformed into a MIMO problem by sampling the received signal $y(t)$ at rate $1/T = P/T_s$ ($P \geq K$). In order to guarantee that all users' pulses overlap at the sampling times, the over sampling period should satisfy that $T_s/P \geq \tau_k, k = 1, ...K$, in other words the over-sampling factor P is upper bounded by $T_s/min\{\tau_1, ..., \tau_K\}$.

Sampling the received signal $y(t)$ at times $t = iT_s + mT_s/P$, and defining $f_k \triangleq F_k T_s/P$, we get:

$$\begin{aligned} y_m(i) &\triangleq y(iT_s + mT_s/P) \\ &= \sum_{k=1}^{K} h_{mk}(i) * \tilde{s}_k(i) + w(i + \frac{m}{P}) \end{aligned} \tag{4}$$

where

$$h_{mk}(i) = a_k e^{j2\pi f_k m} e^{j2\pi f_k iP} p_k(iT_s + \frac{mT_s}{P} - \tau_k)$$

and

$$\tilde{s}_k(i) = s_k(i) e^{-j2\pi f_k iP}$$

Let us ignore the contributions of sidelobes of $p_k(t)$ in (4). Then, the received signal over the time interval $[iT_s, (i+1)T_s)$ is only affected by symbols $\mathbf{s}(i)$ and $\mathbf{s}(i+1)$. Equation (4) can be rewritten as:

$$y_m(i) = [\mathbf{h}_{m1}, ..., \mathbf{h}_{mK}] \begin{bmatrix} \tilde{\mathbf{s}}_1^T(i) \\ \vdots \\ \tilde{\mathbf{s}}_K^T(i) \end{bmatrix} \tag{5}$$

where $\mathbf{h}_{mk} = [h_{mk}(0)\ h_{mk}(-1)]$ and $\tilde{\mathbf{s}}_k(i) = [\tilde{s}_k(i), \tilde{s}_k(i+1)]$.

Let us form the vector $\mathbf{y}(i)$ by appending $y_m(i)$ for $m = 1, ..., P$. Then we have:

$$\mathbf{y}(i) = \mathbf{A}\mathbf{z}(i) + \mathbf{w}(i) \tag{6}$$

where $\mathbf{z}(i) = [\tilde{s}_1(i), \tilde{s}_1(i+1), ..., \tilde{s}_K(i), \tilde{s}_K(i+1)]^T$; $\mathbf{w}(i) \triangleq [w(i+\frac{1}{P}), ..., w(i+1)]^T$.

III. BLIND USER SEPARATION

In the following we show how to recover the transmitted signals in a bind fashion, i.e., without the need for pilot signals. The advantage of a blind approach is bandwidth efficiency.

Let us make the following assumptions.

A1) For each $m = 1...P$, $w_m(.)$ is a zero-mean, independent identically distributed (i.i.d.) complex Gaussian stationary random processes with variance σ_w^2, and is independent of the inputs. **A2**) For each k, $s_k(.)$ is zero mean, i.i.d., stationary with nonzero kurtosis, i.e., $\gamma_{s_k}^4 = \text{Cum}[s_k(i), s_k^*(i), s_k(i), s_k^*(i)] \neq 0$. The s_k's are mutually independent, and each user has unit transmission power. **A3**) The over-sampling factor P satisfies: $2K \leq P$, and the delays τ_k $k = 1, ..., K$ are randomly distributed in the interval $[0, T_s/P)$. **A4**) The channel fading coefficients α_k are non-zero.

Under assumption (A2), it is easy to verify that the rotated input signals $\tilde{s}_k(.)$ are also zero mean, i.i.d, wide sense stationary process with nonzero kurtosis. Also, the $\tilde{s}_k(i)$'s are mutually independent for different k's. Assumption (A3) guarantees that the virtual MIMO channel matrix $\mathbf{A}$ in (6) has full rank.

The above assumptions guarantee the recovery of $\mathbf{z}(i)$ in a blind fashion within trivial ambiguities. One can apply any blind source separation algorithm (e.g., [1],[3] or [4]) to get:

$$\hat{\mathbf{A}} \triangleq \mathbf{A}\mathbf{P}\mathbf{\Lambda} \tag{7}$$

Subsequently, using a Least-Squares equalizer we can get an estimate of $\mathbf{z}(i)$, within permutation and scalar ambiguities as:

$$\hat{\mathbf{z}}(i) = (\hat{\mathbf{A}}^T\hat{\mathbf{A}})^{-\mathbf{1}}\hat{\mathbf{A}}^{\mathbf{T}}\mathbf{y}(i) = e^{jArg\{-\mathbf{\Lambda}\}}|\mathbf{\Lambda}|^{-1}\mathbf{P}^T\mathbf{z}(i) \quad (8)$$

Let us assume that the transmitted signal has unit power, and denote by $\theta_{kL+\ell}$ the $(kL+\ell)-$th diagonal element of $Arg\{\mathbf{\Lambda}\}$. In the following, L denotes the number of neighboring symbols that $p_k(t)$ affects. Since sidelobes were ignored in (5) we can take $L=2$. The elements of $\hat{\mathbf{z}}(i)$ are of the form:

$$s_k(i-\ell)e^{j(-\theta_{kL+\ell}+2\pi f_k(i-\ell))};\ell=0,-1;\; k=1,...,K \quad (9)$$

In other words, for each user k, we recover $s_k(i)$ and $s_k(i+1)$, each rotated by the CFO f_k. The recovery of $s_k(i)$ requires estimation and mitigation of f_k.

An estimate of f_k can be obtained based on the kth column of the phase of the estimated channel matrix $\hat{\mathbf{A}}$, i.e.,:

$$\begin{aligned}\Psi &= Arg\hat{\mathbf{A}} \qquad (10)\\ &= \begin{pmatrix} 2\pi f_1+\phi_1 & \dots & 2\pi f_k+\phi_{kL+\ell} & \dots \\ \vdots & \ddots & \ddots & \ddots \\ 2\pi f_1 P+\phi_1 & \dots & 2\pi f_k P+\phi_{kL+\ell} & \dots \end{pmatrix}\mathbf{P}\end{aligned}$$

where $\phi_{kL+\ell} = Arg\{a_k\}+2\pi f_k \ell P+\theta_{kL+\ell}$, which accounts for both the phase of $h_{mk}(\ell)$ and the estimated phase ambiguity in (9). For each k, we can get L copies of the estimate of f_k based on L columns of the channel matrix. We can group the KL separated signals into K classes, the kth class consisting of rotated versions of $s_k(i)$ and $s_k(i+1)$. By averaging the estimate within every class, we can improve the CFO estimate.

In order to obtain an estimate of $s_k(i)$, we first take cross-correlation between the two estimated signals in each class, i.e., $s_k(i-\ell)e^{j(-\theta_{kL+\ell}+2\pi f_k(i-\ell))}, \ell=0,-1$. Since $s_k(i)$ are i.i.d, based on the peak of the correlation we can get an estimate of the rotated signals $\tilde{\mathbf{s}}(i)$ within permutation and scalar ambiguities as:

$$e^{jArg\{-\mathbf{\Lambda}\}}\mathbf{P}^T\tilde{\mathbf{s}}(i) \quad (11)$$

Noting that the de-coupled signals in (11) are shuffled in the same manner as the estimated CFOs in (10), we can use the CFO estimates in (9) to obtain:

$$\hat{\mathbf{s}}(i) = e^{jArg\{-\mathbf{\Lambda}\}}\mathbf{P}^T\mathbf{s}(i) \quad (12)$$

III-A. Design of pulse shape functions

As it was already mentioned, naturally occurring CFOs and delays are small thus providing limited diversity. A good choice for $p_k(t)$ is essential for improving the condition number of matrix $\mathbf{A}$. User waveforms need to be sufficiently distinct between users, but at the same time must have fixed time and frequency support. We propose the following iterative process for designing pulse-shape functions that have a give time and frequency support.

Step 1) Initial conditions: we start with a typical pulse waveform, e.g., a root cosine pulse, $p(t)$, with support $[-T_s, T_s]$. We uniformly sample $p(t)$ to obtain the $(2N+1)$-pt discrete-time sequence $p(n), n=-N,..,N$. Let $\mathbf{I} = [n_1, n_2, ..., n_Q]$ be a set of sampling points in $[-N,...,N]$. We initially define the k-th user wavefrom $p_k(n)$ to be equal to $p(n)$ at all points except for $n \in \mathbf{I}$. For $n \in \mathbf{I}$ we take $p_k(\mathbf{I}) = \mathbf{C}_k$, (using MATLAB notation) where $\mathbf{C}_k$ is a length Q code associated with user k. The codes are different between different users.

Step 2) For the kth user ($k=1,\dots,K$), the values of $p_k(n)$ for $n \in \mathbf{I}$ are set to $C_k(n)$ (i.e., the code symbols for the kth user). All other samples remain unchanged.

Step 3) The time-domain sequence $p_k(n),\; n=-N,...,N$ is transformed to the frequency domain through a $(2N+1)$-point DFT. There, the desired bandwidth is enforced by setting to some low values, or to zero, all frequency samples outside the desired frequency band. Then, the sequence is transformed back to time domain and after proper energy normalization yields the updated $p_k(n)$.

Step 4) Repeat steps (2)-(3) until there is no significant change in $p_k(n)$.

Based on the theory of projections on convex sets [6], the iteration will finally converge. In the time domain, the obtained $p_k(n)$ will be equal to the kth users's code over samples in $\mathbf{I}$, while its bandwidth will be equal to the desired bandwidth. The kth user's pulse shape function would then be an interpolated version of $p_k(n), n=-N,...,N$.

The code $\mathbf{C}_k$ of user k can be computed locally right before transmission. For example, it can be equal to the basic root cosine pulse at locations $\mathbf{I}$, i.e., $p(\mathbf{I})$, perturbed by a random amount. The receivers do not need to know what pulse shapes were used, and a user does not need to know what other users have used, as long as they are using different seeds to the random number generators

IV. CONNECTION TO CDMA AND POTENTIAL APPLICATIONS

One might think that manipulating the user pulse-shape waveform bears similarity to the CDMA approach. CDMA systems effectively use a fast varying pulse shape waveform, where waveforms are orthogonal between users. These waveforms need to be known to the receiver, so that using cross-correlation with the waveform of a certain user, the receiver can recover that particular user's data and eliminate everybody else. In a CDMA system, as the number of potential users increases, the bandwidth must be increased or the throughput per user must be decreased. This can be done either via increased chipping rate (varying the rate of the pulse shape functions) or by increasing the number of chips per symbol to guarantee orthogonality of additional users signals. Increasing the chipping rate to augment the chips per symbol at fixed data rate expands the bandwidth that is required for each user. CDMA signals are known to be wideband signals.

What we propose does not require orthogonal waveforms between users and this limits the required bandwidth expansion. Most importantly, it does not require knowledge of the pulse shape functions at the receiver. This can be very important for connectionless communications in adhoc networks, or for multiuser coordination in collaborative networks.

The proposed approach also allows separation of users in a cell while new users (with unknown codes) can move into the cell. Consider a scenario where nodes communicate in a CDMA cell, and a node (adversary) with an unknown code moves into the same area. The receiver base station or mobile station cannot recover the adversary signal, neither can he eliminate it from the received signal, as the adversary code is not necessarily orthogonal to everybody else. Each time the receiver uses cross-correlation to recover a regular user's signal, the contribution of the adversary signal due to the non-orthogonality will decrease the signal-to-noise ratio and behave as a jammer signal with a reduction bounded by the adversary's signal time bandwidth product. With the proposed approach, the receiver will still be able to separate users and even recover the signal of the adversary even if its pulse shaping properties are unknown. For the proposed method to be used is such a CDMA scenario, it would have to be modified to account for frequency selective fading, which is not a difficult thing to do.

V. SIMULATION RESULTS

In this section we provide simulation results on the performance of the proposed approach. All results are averaged over 100 independent channels runs; in each run the channel coefficients, the delays and the CFOs change in a random fashion. The CFO is varied in the range $[0 - 0.5)$. The transmitted signals consist of 1200 QPSK symbols. The received signal is sampled uniformly at the rate P/T_s, which gives us P samples per symbol period.

In Fig. 1 we plot BER results for a two-, three- and four-user system using same bandwidth. The symbol decision is made by taking the sign of the real and imaginary parts of the de-rotated signals. The lines in the figure stop at the point where the BER becomes zero. The pulse-shape functions were generated via the iterative approach described in Section III-A. A raised-cosine pulse (with bandwidth $1/T_s$) was used at step 1. The code of kth user, $\mathbf{C_k}$, was taken to be the raised cosine pulse at locations $\mathbf{I} = [-N : \mathrm{floor}((2N+1)/8) : N]$ ($N = 256$), perturbed by a random amount. In step 3 the bandwidth of pulse-shape functions was forced to be $2.5/T_s$.

Fig. 2 shows the time-domain and frequency-domain pulse-shape functions used in Fig. 1 ($K = 4$). The raised-cosine pulse is also shown for comparison. In Fig. 3 we show the BER for a three-user using different bandwidth $1.5/T_s$, $2/T_s$ and $2.5/T_s$. One can see that the BER performance improves with the bandwidth.

VI. REFERENCES

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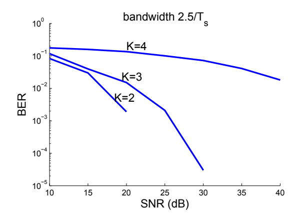

Fig. 1. BER performance for fixed bandwidth and different number of users.

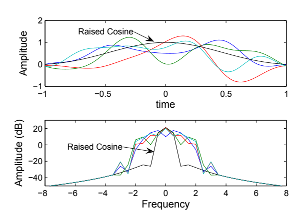

Fig. 2. Pulse shape functions with bandwidth $2.5/T_s$.

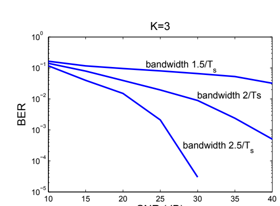

Fig. 3. BER performance for 3 users and variable bandwidth.