NOVEL RECEIVER SIGNAL PROCESSING FOR INTERFERENCE CANCELLATION AND EQUALIZATION IN CELLULAR TDMA COMMUNICATION

Timothy A. Thomas and Michael D. Zoltowski
School of Electrical Engineering, Purdue University
West Lafayette, IN 47907-1285
e-mail: tathomas@ecn.purdue.edu & mikedz@ecn.purdue.edu

Abstract

A space-time processing scheme is proposed for TDMA based cellular/PCS communications that cancels co-channel interference while simultaneously equalizing the desired signal. This is achieved through processing of a training signal designed from a PN code. At the receiver, the symbol waveform is built up above the noise and interference through correlation with a sampled waveform designed to exploit the PN code. From this built up signal, interference canceling beams are formed along with estimates of the initial equalizer coefficients. Then a decision directed algorithm is applied to track variations in the multipath channel. Simulations involving three-ray multipath are presented demonstrating the efficacy of the method.

1. INTRODUCTION

TDMA based cellular/PCS communications in an urban environment is complicated by both multipath propagation and co-channel interference. Multipath can cause intersymbol interference and possibly severe fading, and co-channel interference arises from users in nearby cells. Through judicious design and processing of a short training sequence received at an array of antennas, one can simultaneously effect both co-channel interference suppression and equalization of the desired signal.

The scheme for processing the training signal received at an antenna array is depicted in Figure 1 for the case of $N_b = 2$ beams. We have an M element antenna array that receives a desired signal transmitted through a multipath channel with each path arriving with its own angle of arrival and relative time delay. There is also co-channel uncorrelated interference and additive white Gaussian noise. The output of each antenna is over-sampled (by an integer multiple, L, of the symbol rate) and input to a filter described in the next section. After this filter, the signal is processed to find the interference canceling weight vectors. The initial channel estimates are found by applying the weight vectors to the correlator outputs and solving a set of equations based on the known training sequence. The channels are then tracked in a decision directed manner.

2. RELATION TO OTHER WORK

The problem of estimating a desired user's signal in an interference environment over a multipath channel has been investigated previously. Optimal combining of the outputs of an antenna array has been investigated in [1]. Equalization or direct symbol estimation from a spatial and/or temporal over-sampled channel has been investigated in [2] - [5]. An approach effecting both equalization and interference suppression is proposed in [6]. The unique aspect of the procedure proposed herein is that by passing the received PN based training signal through a specially designed front-end filter at each antenna:

- the interference is isolated from the signal allowing estimation of the spatial correlation matrices for the interference and for the signal plus interference with relatively a few number of symbols
- the combination of correlation and beamforming (weight vectors determined from the spatial correlation matrices) provides a temporal and spatial gain against both the noise and interference, thereby allowing for accurate estimation of the beam channels with few samples.

3. FILTER DESIGN

This paper assumes the use of a training sequence with good autocorrelation properties, for example, a Barker code. A Barker code has the property that its autocorrelation has a value equal to its length at a lag value of zero, while for other lags the value is either 0 or ± 1. The current IS-136 TDMA standard dictates a training sequence of 14 symbols at the front end of each burst of 162 symbols for a given user in a given time slot. For illustrative purposes, we chose to use the following Barker code of length 13 [1]:

$$b(n) = \{-1,-1,-1,-1,-1,1,1,-1,-1,1,-1,1,-1\} \quad (1)$$

The transmitted training signal for the illustrative is then

$$s(t) = \sum_{n=0}^{12} b(n) p_s(t - nT_o), \quad (2)$$

where T_o is the symbol time and $p_s(t)$ is the symbol waveform which we will here assume to have a square-root raised cosine spectrum. (Here $b(n)$ is shown for BPSK. In a multilevel QAM signal constellation, the two values are chosen to be opposite corners of the constellation so that they have opposite signs and a large distance between them.) The autocorrelation of $s(t)$ has a replica of $p(t) = p_s(t) * p_s(-t)$, which has a raised cosine spectrum, centered at each lag value with amplitude dictated by the autocorrelation of the Barker code. The algorithm proposed is premised on localizing the desired signal in time so that it occupies a few

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[1] The largest length Barker code is 13.

symbol times at the center of the correlator output, and is negligible outside of that. This allows us to estimate the spatial characteristics of the interferers from averaging array snapshots over those sample times away from the center of the correlator output.

There are two obstacles to achieving this goal. First, $p(t)$ contains significant tails encompassing a number of symbol times on either sides of its "mainlobe." The IS-136 TDMA standard dictates a value of $\beta = .35$ for the excess bandwidth parameter. In this case, there are roughly three significant "sidelobes" on either side of the mainlobe. Second, the autocorrelation is not zero for even lag values (although it is 1/13 of the peak magnitude). Thus, in order to avoid possibly canceling the desired signal, we counter these two obstacles as follows.

To reduce the tails and thereby counter the first problem, the fact that the sidelobes for two $p(t)$'s separated by a symbol time, T_o, are 180° out-of-phase is exploited via a Hamming type weighting. To this end, the output of each antenna is run through a filter whose impulse response is

$$p_{sr}(t) = h_{sr}(t) * p_s(t) \quad (3)$$

$$h_{sr}(t) = .3016\delta(t - T_o) + .9534\delta(t) + .3016\delta(t + T_o) \quad (4)$$

where $\delta(t)$ is the Dirac delta function. After passing through an over-sampled version of this filter, the output is similar in form to (2) but with $p_s(t)$ replace by a pulse symbol waveform that has a mainlobe that has a null-to-null width of four symbol times and sidelobes that are -45 dB or below relative to the mainlobe peak. At this point, at the output of the cross-correlator (a sampled version of) the waveform $p(t)*h_{sr}(t)$ occurs at each multiple of T_o weighted by the corresponding lag value of the autocorrelation of the Barker code.

To further localize in time the contribution due to the desired signal, we design a sequence whose cross-correlation with the Barker code is 13 for lag value $\ell = 0$ and exactly zero for lag values $\ell = \pm 1, \pm 2, ..., \pm 13$,. The sum of the magnitude squared of the cross-correlation values for lag values $|\ell| > 13$ is minimized via the following optimization problem:

$$\begin{array}{ll} \text{minimize} & \|\mathbf{Bc}\|^2 \\ \text{subject to:} & \mathbf{B}_s\mathbf{c} = \delta \end{array}$$

where, denoting the length of $c(n)$ as N_c,

$$\mathbf{c} = [\, c(1) \ \cdots \ c(N_c)\,]^T$$

$$\mathbf{B} = \overbrace{\begin{bmatrix} b(1) & 0 & & \cdots & 0 \\ b(2) & b(1) & 0 & \cdots & 0 \\ & & \vdots & & \\ 0 & \cdots & 0 & b(13) & b(12) \\ 0 & \cdots & & 0 & b(13) \end{bmatrix}}^{N_c} \left.\vphantom{\begin{bmatrix} 1\\1\\1\\1\\1 \end{bmatrix}}\right\} N_c + 13 - 1$$

$$\delta = \left[\overbrace{0 \ \cdots \ 0}^{13} \ 13 \ \overbrace{0 \ \cdots \ 0}^{13} \right]^T$$

$$\mathbf{B}_s = \{\mathbf{B}_{ij}\}, j = 1, ..., N_c, i = \frac{N_c - 27}{2} + 1, ..., \frac{N_c + 27}{2}$$

The premise for this optimization problem is as follows. The training sequence (minus a few tails) is extracted (isolated) from the front end of the received burst. In a digital setting, there is nothing that precludes cross-correlation with a longer sequence. Use of this $c(n)$ constrains the 13 values on either side of the peak of the cross-correlation between $b(n)$ and $c(n)$ to be zero, and the lag values outside of this range to be 40dB (for $N_c = 41$) below the peak value. As a result, only the interference and noise contribute significantly during the 13 symbol period on either side of the peak. The only drawback using this $c(n)$ as opposed to cross-correlating with $b(n)$ itself is a drop in processing gain against noise from 13 to 12.67. Relative to Fig. 1, the output of the filter $p_{sr}(n) = p_{sr}(nT_o/L)$ is convolved with $c_L(n)$, where $c_L(n)$ is $c(n)$ up-sampled by a factor of L, i.e., with L-1 zeroes inserted between each adjacent pair of values of $c(n)$. In a practical implementation, $p_{sr}(n)$ and $c_L(n)$ may be convolved to form one front-end filter through which the received training signal at each antenna is passed.

4. DETERMINATION OF INTERFERENCE CANCELING WEIGHT VECTORS

In computing the output of the block labeled "Extended Correlator" behind each antenna, we only compute those sample values over a window of duration 27 symbol times centered at that point corresponding to the lag value zero where the peak occurs, and 13 symbol times on either side. By design, the desired signal contribution is negligible outside a window of duration four symbol times plus the multipath time delay spread centered at the point corresponding to the lag value zero.[2] Note that a gain of 12.67 against noise has been achieved by adding the (sidelobe reduced) symbol waveforms in phase.

A set of beamforming weight vectors for canceling the interference, as well as providing a spatial gain against noise, is estimated as follows. The signal plus interference (plus noise) spatial correlation matrix, $\mathbf{R}_{S+I}$, is estimated from snapshots extracted from the center of the correlator output, while the interference (plus noise) spatial correlation matrix, $\mathbf{R}_I$, is estimated from snapshots extracted away from the center of the correlator output. The weight vectors are computed as the largest generalized eigenvectors of the matrix pencil $\{\mathbf{R}_{S+I}, \mathbf{R}_I\}$. The number of dominant principal eigenvectors is dictated by the angular spread of the multipath relative to the beamwidth of the array. For illustrative purposes, we assume two dominant principal eigenvectors, $\mathbf{w}_1$ and $\mathbf{w}_2$, in the remaining discussion.

5. INITIAL CHANNEL ESTIMATION

$\mathbf{w}_1$ and $\mathbf{w}_2$ are applied to the outputs of each correlator thereby forming two beams, denoted $z_1(n)$ and $z_2(n)$, where each beam is a different multipath channel. Based on the previous development, at the center of the correlator output for each is the signature waveform for that beam channel, equal to the convolution of the impulse response of that beam channel with the sidelobe reduced symbol waveform. In estimating the impulse response for a given beam channel, we assume the noise and interference to be negligible in this center region due to:

- a 12.67 temporal gain against noise
- a maximum spatial gain of M (number of antennas) against the noise
- a spatial gain against interference that depends on the null depths
- a temporal gain against interference that depends on how uncorrelated the interfering signal's symbols are with respect to the desired user's training sequence

[2] A conservative upper bound for the multipath time delay spread in an urban cellular environment is 10 microseconds.

The beam channel impulse responses, $h_{c1}(n)$ and $h_{c2}(n)$, are estimated as follows. Using the fact that $\tilde{p}(n) = \tilde{p}(nT_o/L)$ is known, where $\tilde{p}(t) = p(t) * h_{sr}(t) = p_s(t) * p_s(-t) * h_{sr}(t)$, the i-th beam channel is estimated via the least square error solution to an over-determined system of equations constructed from $2N+1$ samples extracted from the center of the beam correlator output $z_i(n)$, designated by $n = n_c$, according to:

$$\mathbf{z}_i = \mathbf{S}\mathbf{h}_{ci}, \quad i = 1, 2$$

where

$$\mathbf{z}_i = [\ z_i(n_c - N) \ \ \cdots \ \ z_i(n_c) \ \ \cdots \ \ z_i(n_c + N)\ ]^T,$$

$$\mathbf{S} = \begin{bmatrix} \tilde{p}(-N) & \cdots & \tilde{p}(-N+1-N_h) \\ \tilde{p}(-N+1) & \cdots & \tilde{p}(-N+2-N_h) \\ & \vdots & \\ \tilde{p}(+N) & \cdots & \tilde{p}(+N+1-N_h) \end{bmatrix},$$

$$\mathbf{h}_{ci} = [\ h_{ci}(0) \ \ h_{ci}(1) \ \ \cdots \ \ h_{ci}(N_h - 1)\ ]^T$$

where N_h is the number of taps for $h_{ci}(n)$. As an example, in the simulations presented in Section 7, $2N+1 = 7$ which works well for $L = 10$ and $N_h = 5$. Because of temporal oversampling, the number of taps in $h_{ci}(n)$ required for a TDMA signal format similar to the IS-136 standard is small. For example, as substantiated in the simulations to be presented in Section 7, $N_h = 5$ works well if the sampling time is $2.5\mu s$ and the maximum delay is $10\mu s$.

As long as the two beam channels have no common zeros, two equalizing FIR filters, $g_1(n)$ and $g_2(n)$, are determined as the solution to:

$$\text{minimize} \quad E\left\{\left|\mathbf{g}_1^H\mathbf{n}_1 + \mathbf{g}_2^H\mathbf{n}_2\right|^2\right\}$$
$$\text{subject to: } h_{c1}(n) * g_1(n) + h_{c2}(n) * g_2(n) = \delta(n - D)$$

where $\mathbf{n}_1$ and $\mathbf{n}_2$ are vectors containing the noise samples at the two beam outputs and D is some integer delay.

6. DECISION DIRECTED MODE

A scheme to update $g_1(n)$ and $g_2(n)$ is presented here that assumes the multipath channel to be constant for at least 15 symbols (a reasonable assumption in a TDMA environment for a mobile traveling 60 mph or less). The algorithm starts by using the initial $g_1(n)$ and $g_2(n)$ (found using the methods in the previous section) to equalize $r_1(n)$ and $r_2(n)$ where $r_i(n)$ is the sequence after the beam weight, $\mathbf{w}_i$, is applied to the *raw* antenna outputs. Next, N_f (e.g., $N_f = 5$) symbols past the training sequence (or the last estimated symbol) are estimated from $r(n) = r_1(n) * g_1(n) + r_2(n) * g_2(n)$. These symbols are used along with N_p prior symbols to find $g_1(n)$ and $g_2(n)$ via:

$$\text{minimize} \quad E\left\{\left|\mathbf{g}_1^H\mathbf{n}_1 + \mathbf{g}_2^H\mathbf{n}_2\right|^2\right\}$$
$$\text{subject to: } \quad r_1(n) * g_1(n) + r_2(n) * g_2(n) = y(n)$$

where y(n) is an over-sampled version of

$$y(t) = \sum_{m=n'-N_f-N_p+1}^{n'} \tilde{b}(m)p_s(t - mT_o)$$

where n' is the time of the most recently estimated symbol and $\tilde{b}(n)$ is the sequence of estimated symbols. Once the new estimates of $g_1(n)$ and $g_2(n)$ are found, the procedure is repeated starting with making decisions on the next N_f future symbols.

7. SIMULATIONS

In the simulations, the multipath signals arrived at $-5°$, $0°$, and $5°$ with complex amplitudes of -0.8, 1, 0.9j, respectively, and time delays of 7.5, 0, and 5 micro-seconds. The array was linear; $M = 8$ antennas equi-spaced by a half-wavelength. Two interferers were received at the same strength as the desired signal at $-50°$ and $60°$, respectively. The noise at each antenna was additive white Gaussian noise at an SNR of 7 dB. The symbol waveform was a pulse with a square root raised cosine spectrum ($\beta = .35$), and a symbol was sent every 25 micro-seconds. The received signal was oversampled by $L = 10$. The modulation was QPSK. The training sequence was $(1 + j)b(n)$. In the first simulation example, the results of which are plotted in Figures 3(b) and 3(c), all multipath parameters were held constant over the duration of the 162 symbol burst. In the second simulation example, the results of which are plotted in Figures 3(d) and 3(e), the relative phases of the multipaths are changing due to the mobile transmitter traveling at 60 mph.

Figure 3(a) shows the interference canceling weight vectors, where the one associated with the "largest" generalized eigenvector is on the right. Each beam pattern has a deep null in each interference direction and a relatively large gain over the angular spread of the multipath for the desired signal.

Figure 3(b) shows the signal constellation at the output of beam 2. Here the interference has been canceled, but the effects of the channel and noise still produce a large spread in the estimated symbol values. In Figure 3(c) we see the nice signal constellation obtained after equalizing the two beam channels in accordance with the scheme proposed herein. When the beam channels vary due to the source moving at 60 mph, the static solution for the $g(n)$'s fails after roughly 20 symbols as shown in the mean square error (mse) plot in Figure 3(d). This clearly shows the need for a decision directed version to track the equalization filters. In Figure 3(e), the equalized signal constellation is shown applying the decision directed algorithm presented in Section 6. This demonstrates the efficacy of the algorithm in tracking a time-varying channel.

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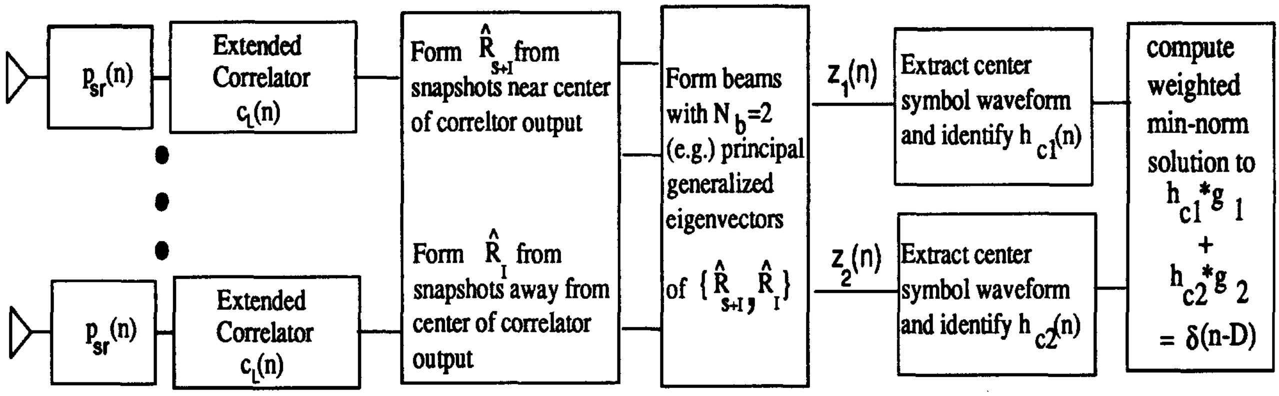


Figure 1. Receiver signal processing of over-sampled training signal for illustrative case of $N_b = 2$ beams.

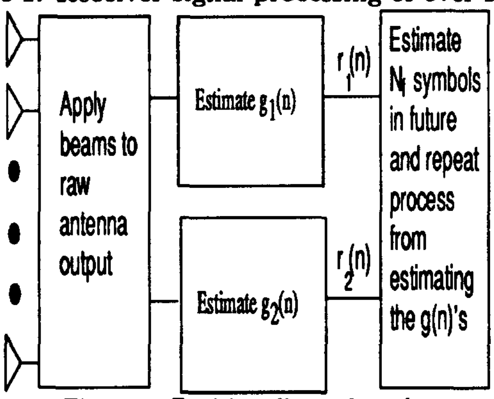
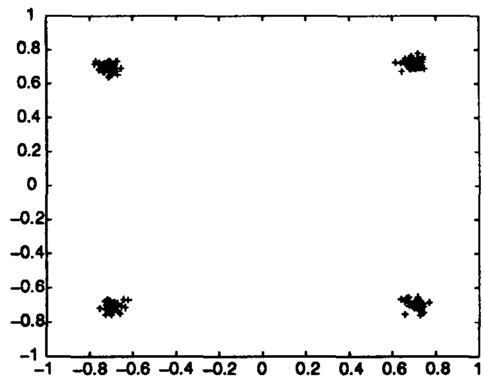
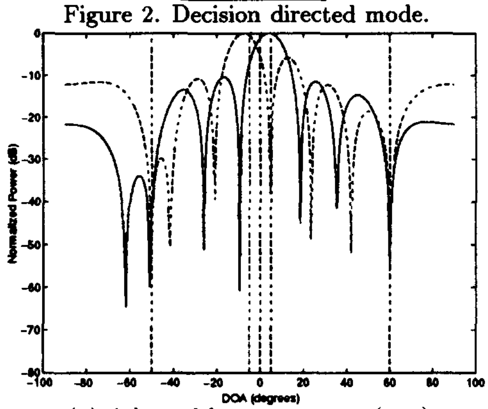


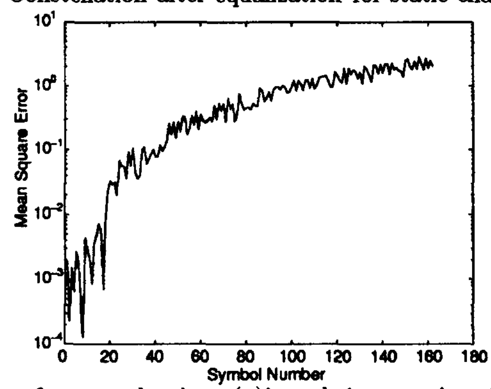
Figure 2. Decision directed mode.



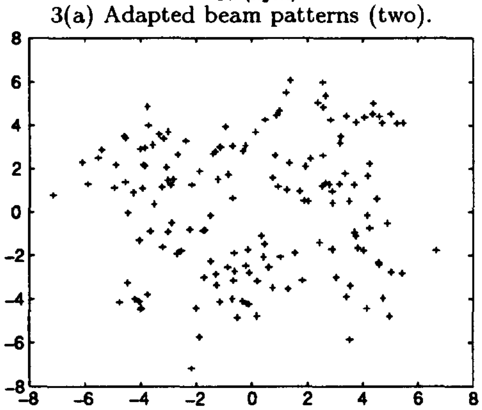
3(c) Constellation after equalization for static channel.



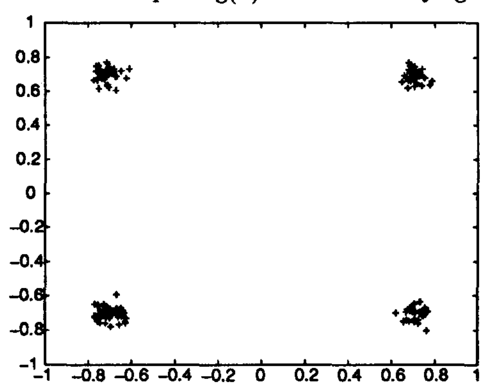
3(a) Adapted beam patterns (two).



3(d) mse for non-adaptive g(n)'s and time-varying channel.



3(b) Constellation at beam 2 output for static channel.



3(e) Const. for dec. dir. mode for time-varying ch.