HYBRID LINEAR-ITERATIVE DETECTION ALGORITHMS FOR MIMO CDMA SYSTEMS IN MULTIPATH CHANNELS

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ABSTRACT

We assume a CDMA downlink multiuser scenario, where the base station is equipped with multiple transmit antennas, and each user has multiple receive antennas. For this setup we propose a new, computationally effective MIMO multiuser detector for frequency selective channels. We exploit the fact that the interference mainly comes from the spatial domain, while the residual interference is due to temporal dispersion and multiuser interference. This motivates us to suppress the residual interference in a linear fashion, while combating the spatial interference iteratively. This approach is an efficient performance-complexity tradeoff between full iterative and linear detectors.

1. INTRODUCTION

The most effective way to increase transmission rates through the wireless fading channel is to employ multiple transmit and receive antennas (MIMO) [3, 2]. Most of the techniques presented in the literature [3, 4] are applicable for flat fading channels. In practice, however, multipath channels are often encountered. Most of the work for MIMO *multipath* channels has been done for narrowband transmission [1], in terms of MIMO channel equalization. Thus far, little attention has been devoted to multiuser detection for wideband MIMO multipath fading channels.

In this paper, we consider a wideband downlink multiuser scenario, where each user has multiple receive antennas. Similarly, the base station is equipped with multiple transmit antennas. The motivation comes from high rate third generation (3G) wireless standards, such as high speed downlink packet access (HSDPA). Each user is assigned a signature spreading code, used by the base station to modulate the transmitted symbols. For a particular user, this code is fixed across the transmit antennas. The channel from the base station to each user is assumed to be multiple input multiple output (MIMO) multipath fading. Clearly, in single path channels, using orthogonal spreading codes allows perfect user separation. However, for the most general case, a single transmitted symbol suffers "spatial" interference from other antennas, intersymbol interference (ISI) as a result of multipath, and "multiuser" interference due to other users codes. Throughout the discussion, we assume that the channel state information is known at the receiver, but not at the transmitter. Furthermore, we assume that the propagation coefficients remain constant throughout one transmission block.

We consider two different approaches for the receiver structure: channel equalization, and multiuser detection. Channel equalization approach relies on "single tap" channel equalization, followed by despreading. Note that despreading constitutes user separation. Multiuser detection approach jointly performs channel equalization and despreading. For the chip level equalization, the novel receiver is outlined as follows. First we equalize the channel, in a linear fashion, to a single tap channel. This is non-trivial and the procedure is given in Section 2.3. At this point both ISI and multiuser interference are suppressed. Then we despread and perform spatial decision-feedback interference cancellation. This approach is superior to fully linear MMSE equalization because it iteratively cancels spatial interference, which is the bulk of the total interference. Furthermore, our approach is less computationally complex than the full Space-Time decision feedback interference cancellation because we eliminate temporal interference in a linear fashion. We observe similar benefits inside the multiuser detection category as well.

The rest of the paper is organized as follows: Section II discusses the single user MMSE channel shortening. We emphasize the importance of the Section II-C for some of the procedures extend to the multiuser case. Section III deals with MIMO multiuser downlink scenario. We propose two possible receiver solutions. The first solution is Hybrid Linear Iterative Chip Equalizer (HLICE) and it does not require the knowledge of all multiuser codes. The second solution is Hybrid Linear Iterative Multiuser Detector (HLIMUD) and it requires the knowledge of all codes. Simulation results are presented in Section IV. Section V is the conclusion.

2. SINGLE USER SCENARIO

2.1. Channel Model

In the case of no delay spread, the standard baseband MIMO linear channel model is given as $y = Hx + \eta$. Here, x corresponds to the vector of symbols transmitted on N transmit antennas. Similarly, y is the vector of symbols received by R receive antennas. $(r, n)^{th}$ entry of H is the fading coefficient from antenna n to antenna r, where $1 \leq n \leq N$ and $1 \leq r \leq R$. The entries of H are assumed to be circular symmetric complex Gaussian with variance 0.5 per dimension. Additive noise is denoted as η, a proper complex Gaussian vector whose components are uncorrelated, and have variance $0.5\sigma^2$ per dimension. In the presence of multipath delay spread we need to introduce time dependence. We let l be the channel tap index, $1 \leq l \leq L$, and let m represent the symbol time index, $1 \leq m \leq M$. The discrete-time baseband channel model is given as:

$$y(m) = \sum_{l} H(l)x(m - l + 1) + \eta(m) \tag{1}$$

We implicitly assume quasistatic fading channels because filter tap coefficients do not change within the transmission block. For every symbol index m, $x(m)$ and $y(m)$ are $N-$ and $R-$dimensional vectors. We compactly write the above equation as:

$$Y = \mathcal{H}X + \psi, \text{ where}$$

$$\mathcal{H} = \begin{bmatrix} H(1) & & & \\ H(2) & H(1) & & \\ \vdots & H(2) & \ddots & \\ H(L) & \vdots & & H(1) & \\ & H(L) & & H(2) & H(1) \end{bmatrix}$$

$$Y = \begin{bmatrix} y(1) & \dots & y(M) \end{bmatrix}^T$$

$$X = \begin{bmatrix} x(1) & \dots & x(M) \end{bmatrix}^T$$

$$\psi = \begin{bmatrix} \eta(1) & \dots & \eta(M) \end{bmatrix}^T \tag{2}$$

The so called "spatial" interference comes from the fact that matrices $H(m), 1 \leq m \leq M$ are not diagonal. Other than spatial interference, a temporal interference is also present due to the time dispersive channel. Clearly, Y is a vector of size MR, and X is a vector of size MN. $\mathcal{H}$ is a matrix of size $MR \times MN$. We operate on Y in order to "shorten" the channel $\mathcal{H}$ to a single tap. LMMSE channel shortening is discussed next.

2.2. LMMSE Background

Let

$$\mathcal{G} = \begin{bmatrix} G(1) & & \\ & \ddots & \\ & & G(M) \end{bmatrix} \tag{3}$$

where $\{G(m)\}_{m=1}^{M}$ are $R \times N$ matrices. We will specify entries of $G(m), 1 \leq m \leq M$ later. In comparison with equation (2), we see that $\mathcal{G}$ is a single tap channel. We are looking for *the* linear operation $\mathcal{M}$ on the received vector Y, so that the result is closest to $\mathcal{G}X$ in the mean squared error sense. We define the approximation error vector ϵ as

$$\epsilon = \mathcal{M}Y - \mathcal{G}X \tag{4}$$

The *Orthogonality Principle* states that $E[\epsilon Y'] = 0$. We use the expressions for ϵ and Y from equations (4) and (2). Then we use the fact that $E[XX'] = P\, I_{NM}$, where P is the symbol energy, and $E[\psi\psi'] = \sigma^2\, I_{RM}$. We assume that signal and noise are uncorrelated, $E[X\psi'] = 0$. We get

$$\mathcal{M} = \mathcal{G}\left(\mathcal{H}'\mathcal{H} + \left(\sigma^2/P\right)\, I\right)^{-1}\mathcal{H}' \tag{5}$$

The mean squared error for each individual symbol can be found from the diagonal of the following matrix

$$R_{\epsilon\epsilon} = E[\epsilon\epsilon'] = \sigma^2\mathcal{G}\left(\mathcal{H}'\mathcal{H} + \left(\sigma^2/P\right)\, I\right)^{-1}\mathcal{G}' \tag{6}$$

Observe that the regular MMSE solution can be obtained from equation (5) by setting $\mathcal{G} = I$. Note that zero forcing (ZF) criterion can also be used instead of MMSE.

2.3. Linear Channel Shortening with Spatial Iterative Interference Cancellation

Some prior work has been done in channel shortening [1], where the criterion was to minimize the trace of $R_{\epsilon\epsilon}$. An additional constraint needs to be put upon $\mathcal{G}$ in order to avoid trivial solutions of type $\mathcal{G} = 0$. One proposed constraint was that $\mathcal{G}$ should be orthonormal, while another was that $\mathcal{G}$ should have an identity tap. None of these constraints are applicable for our purposes. We desire $\mathcal{G}$ to be single tap, and none of the solutions with either of the above two constraints satisfies that. As far as the channel shortening is concerned, it is obvious from equation (5) all that is actually done is the traditional LMMSE equalization followed by reintroducing $\mathcal{G}$, as in the Fig. 1. After this shortening procedure, the resulting channel has a form of

$$\tilde{Y} = \mathcal{M}Y = \mathcal{G}X + \epsilon \tag{7}$$

$E[\epsilon\epsilon']$ is given in equation (6). Since $\mathcal{G}$ is a single tap, we have

$$\tilde{y}(m) = G(m)x(m) + \epsilon(m),\ 1 \leq m \leq M \tag{8}$$

Notice that the effect of time dispersion has been suppressed, and that $\epsilon(m)$ contains both noise and the residual interference. In order to proceed using iterative spatial interference cancellation we require that $E[\epsilon(m)\epsilon(m)'] = \sigma^2 I_R$. We achieve this through the choice of $\mathcal{G}$. If $\mathcal{G}$ was not required to be single tap, then we could "whiten" $R_{\epsilon\epsilon}$ using the equation (6). With the single tap $\mathcal{G}$, the best we can do is whiten the main diagonal blocks of $R_{\epsilon\epsilon}$. We do this by "isolating" them first, and then "whitening."

$$F(m) = \left[\left(\mathcal{H}'\mathcal{H} + \left(\sigma^2/P\right)\, I\right)^{-1}\right]_{(m-1)N+1:mN,(m-1)N+1:mN}$$

$$G(m) = (F(m))^{-1/2} \tag{9}$$

Since $\mathcal{H}$ is close to block Toeplitz, $F(m)$ are approximately equal, and hence all $G(m) \approx G$ are approximately equal. The channel has been shortened to

$$\tilde{y}(m) = Gx(m) + \epsilon(m),\ 1 \leq m \leq M \tag{10}$$

At this point $x(m)$ and $\epsilon(m)$ are somewhat correlated, and the optimum LMMSE equalizer is simply the pseudo inverse of G. This is because

$$(\mathcal{G}'\mathcal{G})^{-1}\mathcal{G}\tilde{Y} = (\mathcal{G}'\mathcal{G})^{-1}\mathcal{G}\mathcal{M}Y = \left(\mathcal{H}'\mathcal{H} + \left(\sigma^2/P\right)\, I\right)^{-1}\mathcal{H}'Y \tag{11}$$

which is the LMMSE estimate for X. From equation (6) and the choice of G, we see that mean squared error for $x(m)$ is readily approximated as $(G'G)^{-1}$. We perform spatial interference cancellation using the fact that the LMMSE filter is given as $(G'G)^{-1}G'$, and that the mean squared error is given as $(G'G)^{-1}$. It may seem that introducing $\mathcal{G}$ has no purpose, since we cancel it out with pseudo inverse. However, this is true only for the first stage. The true benefits of reintroducing $\mathcal{G}$ come because it "whitens" $\epsilon(m)$ in preparation for spatial interference cancellation. We term this structure the Hybrid Linear-Iterative MMSE (HLIMMSE).

3. MULTIUSER SCENARIO

In this section, we assume downlink multiuser scenario in multipath channels. We design a computationally efficient receiver based on channel shortening. We first introduce the notation. k is the user index, $1 \leq k \leq K$. u is the symbol index $1 \leq u \leq U$. n is the transmit antenna index $1 \leq n \leq N$. r is the receive antenna index $1 \leq r \leq R$. l is the multipath index $1 \leq l \leq L$. q is the chip index $1 \leq q \leq Q$. $\otimes$ denotes the Kronecker product. Sampling is done at the chip level, and the sampling time index is m, where $1 \leq m \leq M = QU$. Symbols $b(k,u,n)$ are to be detected. After sampling, the antenna r observes

$$y(m,r) = \sum_{n,l} h(r,n,l) \sum_{k,u} b(k,u,n) s(k, m-l-uQ) + \nu(m) \tag{12}$$

Here, $h(r,n,l)$ is the l^{th} path fading coefficient from antenna n to antenna r. Note that the channel $h(r,n,l)$ is the same for all users in downlink scenario. Also, $s(k, m-l-uQ)$ is the spreading code of the k^{th} user appropriately delayed to the u^{th} symbol transmission by the modulator, and furthermore delayed by the channel corresponding to the l^{th} path. Additive white gaussian noise is denoted as $\nu(m)$.

3.1. Chip Level Equalization Approach

We first express the equation (12) in matrix - vector notation as:

$$Y = \mathcal{H} \sum_{k} S(k) b(k) + \Upsilon \tag{13}$$

Here, $\mathcal{H}$ is size $MR \times UNQ$ and is given as in equation (2). Υ is the standard noise vector and furthermore:

$$b(k,u) = \left[\begin{array}{cccc} b(k,u,1) & b(k,u,2) & \ldots & b(k,u,N) \end{array} \right]^T$$
$$b(k) = \left[\begin{array}{cccc} b(k,1) & b(k,2) & \ldots & b(k,U) \end{array} \right]^T$$
$$Y = \left[\begin{array}{cccccc} y(1,1) & y(1,2) & \ldots & y(1,R) & \ldots & y(M,R) \end{array} \right]^T$$
$$S(k) = I_{NU} \otimes s(k,\cdot) \tag{14}$$

Here, $s(k,\cdot)$ is the spreading code for the user k. The outline of the Hybrid Linear-Iterative Chip Equalizer (HLICE) receiver solution is given in Fig. 2. The first block is the standard matched filter block which generates sufficient statistics and reduces the dimension of the problem. This is followed by a full *Linear* Space-Time equalizer block, such as MMSE $(\mathcal{H}'\mathcal{H} + (\frac{\sigma^2}{P})\, I)^{-1}$. Thus far, all of this is based on the channel shortening equation (5). Hence, the chip rate processing part of the receiver produces

$$\hat{Y} = \left(\mathcal{H}'\mathcal{H} + \left(\sigma^2/P\right)\ I\right)^{-1} \mathcal{H}'Y \tag{15}$$

The size of $\hat{Y}$ is $UNQ \times 1$. From here, we reintroduce the single tap channel $\mathcal{G}$ as in the equation (5), where $\mathcal{G}$ is computed as in the equation (9). This is essentially channel shortening in the *chip* level. However, we choose to perform despreading prior to reintroducing $\mathcal{G}$, in the symbol level, as in Fig. 2. Both reintroducing $\mathcal{G}$ at the chip and at the symbol level are equivalent because the channel is block fading. Furthermore, from Fig. 2 we note that if we were to design *Linear MMSE* chip equalizer, we would make a hard decision after the despreading block. Hence the first part of the receiver is linear, which is then followed by spatial iterative interference cancellation. After despreading we have the following soft estimates for $b(k)$:

$$\hat{b}(k) = S'(k)\hat{Y} = S'(k)\left(\mathcal{H}'\mathcal{H} + \left(\sigma^2/P\right)\ I\right)^{-1} \mathcal{H}'Y \tag{16}$$

We introduce the channel $\mathcal{G} = I_U \otimes G$ to these soft estimates. Since $\mathcal{G}$ is single tap, we may operate one block at a time. Noting that $\hat{b}(k,u)$ is a vector of soft estimates for $b(k,u,1), \ldots, b(k,u,N)$, and G is $R \times N$ matrix we have $z(k,u) = G\hat{b}(k,u)$ This is done in the "Reintroduce Channel G," block of the Fig. 2. From here we resort to iterative spatial interference cancellation in order to recover symbols $b(k,u,n), 1 \leq n \leq N$. This is done in the "Spatial Detector" block of Fig. 2.

3.2. Multiuser Detection Approach

Equation (13) can be rewritten as $Y = \mathbb{H}b + \Upsilon$ where

$$\mathbb{H} = \mathcal{H}\left[\begin{array}{ccc} S(1) & \ldots & S(K) \end{array} \right], b = \left[\begin{array}{ccc} b(1) & \ldots & b(K) \end{array} \right]^T \tag{17}$$

From here we can easily generate the LMMSE Multiuser detector estimates for b

$$\hat{b} = \left(\mathbb{H}'\mathbb{H} + \left(\sigma^2/P\right)\ I\right)^{-1} \mathbb{H}'Y \tag{18}$$

From equation (17) it follows that

$$\mathbb{H}'Y = \left[\begin{array}{cccc} S(1) & S(2) & \ldots & S(K) \end{array} \right]' \mathcal{H}'Y \tag{19}$$

These two operations of matched filtering and despreading are given in the first two blocks of Fig. 3. The next block "One-Shot Multiuser/code Detector" applies $\left(\mathbb{H}'\mathbb{H} + \left(\sigma^2/P\right)\ I\right)^{-1}$ from the equation (18) to $\mathbb{H}'Y$. Soft symbol $\hat{b}$ estimates are available after this block. From here we operate separately on users k and symbols u. This means that we reintroduce G, computed as in equation (9), to $\hat{b}(k,u) = \left[\begin{array}{ccc} \hat{b}(k,u,1) & \ldots & \hat{b}(k,u,N) \end{array} \right]^T$, for every $1 \leq k \leq K, 1 \leq u \leq U$. Similar to the previous section, we have $z(k,u) = G\hat{b}(k,u)$ From here we resort to spatial interference cancellation. We term the entire structure from Fig. 3 the Hybrid Linear-Iterative Multiuser Detector (HLIMUD).

4. EXAMPLES

The results for a single user scenario with BPSK modulation and $N = R = 4$ are plotted in Fig. 4. For weak multipath profiles, which are close to a single path, the performance of HLIMMSE equalizer is very close to the performance of full iterative IMMSE, but with significantly reduced complexity. This is not the case for stronger multipath profiles. Regardless of the multipath profile, HLIMMSE attains significant gains compared to Linear MMSE (LMMSE).

The results for the Ped. A, multiuser scenario, with 8PSK modulation and $N = R = 4$ are plotted in Fig. 5. Note that HLICE is up to 7dB better than the straightforward LMMSE Chip Equalizer (LMMSECE). Similarly, HLIMUD is up to 7dB better than LMMSE Multiuser Detector (LMMSEMUD). Because the multipath profile is weak, the performance of both HLICE and HLIMUD is very close to the full iterative IMMSEMUD.

The same setup for Veh. A channel environment is given in Fig. 6. HLICE is up to 5dB better than the straightforward LMMSECE, and HLIMUD is up to 5dB better than the LMMSEMUD. Due to the heavy multipath, neither HLICE nor HLIMUD are close to full IMMSEMUD. Any type of MUD approach is superior to CE because it jointly performs despreading user separation and equalization/detection.

5. CONCLUSION

We have proposed a novel Hybrid Linear-Iterative algorithm for both chip equalization and multiuser detection in time dispersive channels. Our approach benefits from the performance of iterative detectors along with the low complexity of linear detectors. For weak multipath profiles, the performance of the proposed scheme is extremely close to the full iterative MMSE detector, while for stronger multipath profiles the difference in performance becomes more significant. This is because the hybrid approach combats multipath interference linearly. However, dramatic complexity reduction compared to the iterative MMSE is obtained due to the channel shortening.

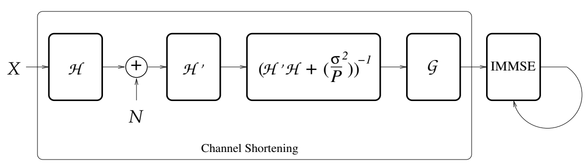

Fig. 1. Block diagram of HLIMMSE equalizer

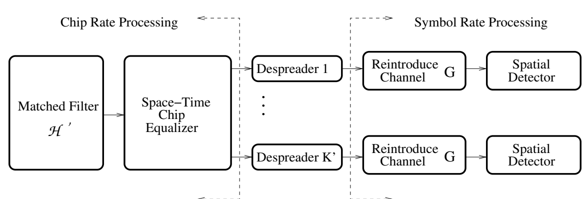

Fig. 2. HLICE block diagram.

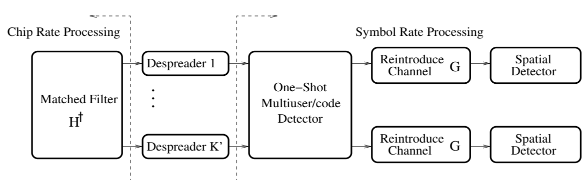

Fig. 3. HLIMUD block diagram.

6. REFERENCES

[1] N. Al-Dhahir "FIR Channel Shortening Equalizers for MIMO ISI Channels" *IEEE Trans. Commun,* pp.213-218, Feb. 2001.

[2] G.J. Foschini and M.J.Gans, "On Limits of Wireless Communications in Fading Environment when Using Multiple Antennas" *Wireless Personal Commun,* pp.311-335, March 1998.

[3] I.E.Telatar, "Capacity of Multi-Antenna Gaussian Channels,", AT&T-Bell Laboratories Internal Tech. Memo., June 1995.

[4] A. Naguib, N. Seshadri, and A. R. Calderbank, "Increasing Data Rate over Wirelesss Channels," *IEEE Signal Processing Mag.*, pp.76-92, May 2000.

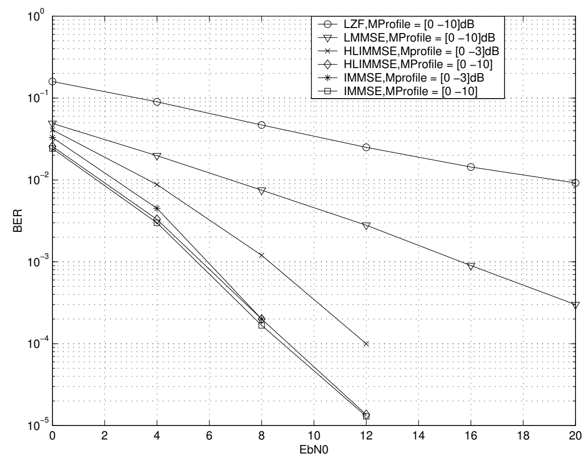

Fig. 4. Comparison of Equal. Techniques, MProfile = [0 -10] dB

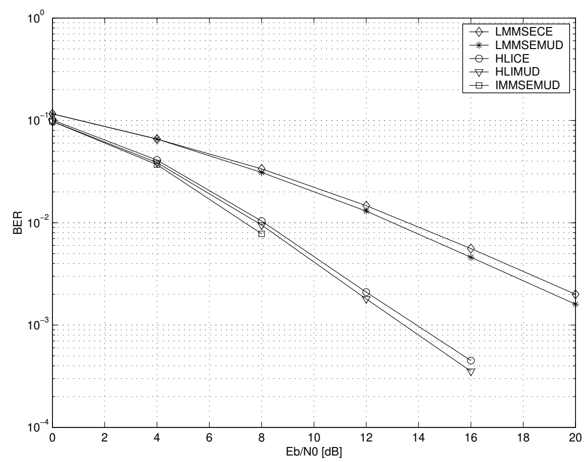

Fig. 5. BER in Ped. A channel. MProfile = [0 -9.7] dB. 12/16 users.

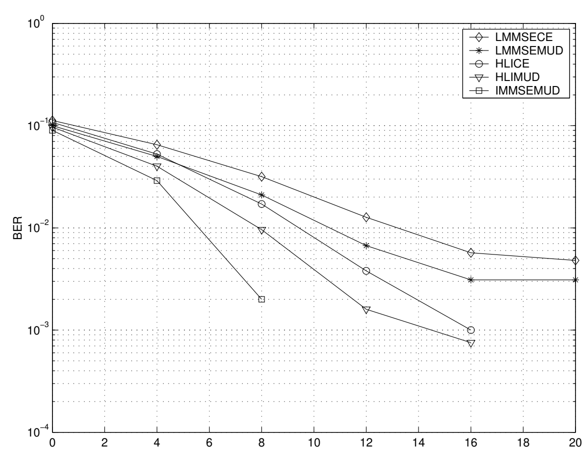

Fig. 6. BER in Veh. A channel. MProfile = [0 -1.0 -9.0 -10.0 -15.0] dB. 12/16 users