Project Description

The size of the annual wireless communications businesses in the United States has grown from 100$milliontomorethan$50 billion in less than 15 years. Telecommunications and broadcasting corporations have increased the voice (telephony) capacity of their networks, and are adding new data features so that customers with wireless modulation-demodulation devices (modems) can access electronic mail and faxes, and get limited access to information from the Internet. These tetherless or cordless devices, the modems they employ, and the networks they are connected with, constitute the first stage of an emerging vision of personal communications systems (PCS) which will allow users the freedom of communicating a wide variety of information between any two points, whether they are in the home, car, or any other physical location.

The tetherless communications technologies that are enabling these developments are now based on second generation systems which were primarily intended for voice transmission. Third generation technology has been undergoing a process of standardization and is now being introduced commercially, and will offer considerably higher data rates and allow significantly increased flexibility over second generation systems. Many new products and services are becoming available which involve interactive multimedia, including moderate quality teleconferencing and higher speed Internet access. But even as this is taking place, there is projection beyond third generation wireless systems to fourth generation systems, with an order of magnitude increase in data rates and bandwidth, and a similar extension of the system capacity. The defining characteristics of these future systems include data rates compatible with true multimedia applications, global roaming capability, and seamless coordination with other network structures. These improved mobile and wideband capabilities will improve current applications and also enable new applications, including new interactive multimedia applications such as high quality teleconferencing, tele-medicine, video entertainment, crisis management, and education over the Internet. Also, fixed and mobile broadcast systems for digital radio and TV with more channels, better quality, and interactive service are envisioned. Market projections for these technologies are staggering, with a worldwide wireless market that is expected to grow to 700 million digital wireless subscriber units by 2002.

Compared with wireline networks, wireless resource is very scarce. While new wireline network "bandwidth" is created when new physical resources (cable, fiber, router, etc.) are added to the network, wireless communication requires sharing a limited natural resource: the radio frequency spectrum. The data-rate capacity that a radio frequency channel can support is limited by Shannon's capacity laws. Hence, in the wireless environment, one has to very carefully engineer the network such that little, if any, wireless spectrum is wasted. This requires an integrated multidisciplinary approach with expertise in wireless communications at the physical and network levels as well as expertise in source coding and signal processing. For example, the network level scheduling protocols could benefit immensely from the information at the physical layer regarding the quality of the channel, in terms of which user to schedule to transmit at a given time. Similarly, based on the physical layer conditions or congestion in the network, feedback could be sent to the source to provide a coarser or higher quality transmission, which would involve some form of scalable compression technique. In pursuit of the goal of developing wireless communications networks to support heterogeneous high-bandwith services, we have identified the following key research areas that we will focus on in this proposal:

- **Advanced coding and modulation techniques** – this involves the process by which information bits are transformed into waveforms for bandwidth and power efficient reliable transmission over the wireless channel.

- **Advanced adaptive signal processing techniques** – this involves the methods by which the receiver carries out channel estimation, synchronization, equalization, interference rejection and other tasks required for reliable and sustained reception over the wireless channel.

- **Effective channel modeling, measurement, and analysis** – this involves extensive characterization of the highly variable wireless channel, so that accurate simulation and testing can be carried out to determine and optimize performance for broadband wireless communications.

- **Advanced scheduling schemes** – this involves the development of scheduling schemes that are dependent on the state of the channel to optimize various network performance measures (e.g. maximizing throughput, minimizing delay) given both fairness and real-time traffic constraints.

- **Scalable video and image compression schemes** – this involves the ability for the sender to transmit video or images at multiple (perhaps even continuous) data rates at the request of a receiver, or due to channel characteristics or network congestion. The key issues that we will study here are scalability, complexity, and redundancy.

1 Advanced Methods of Post-Processing SINR Measurement

The proposed schemes of transmission scheduling for efficient wireless utilization hinge on accurate measurement of the SINR for each user. For optimal efficiency, the appropriate SINR measurement is that associated with the actual symbol estimate after equalization, diversity combining, and/or interference cancellation. We thus propose schemes for estimating this post-processing SINR for both the up-link and down-link, for both CDMA and TDMA systems. Examining the up-link and down-link separately is important due to the FDD operation of many commercial cellular systems: the up-link and down-link are at different frequencies so that their respective channel characteristics are quite different. We also propose to investigate the utility of various other metrics that may be input to the transmission scheduling scheme to further improve efficieny spectral utilization.

2 Measurement of Post-Processing SINR for CDMA Uplink.

We here propose a scheme for measuring the SINR at the output of a space-time RAKE receiver employed at the base station on the CDMA up-link that cancels strong multi-user access interference (MAI) while optimally combining the desired user's multipath. We will consider the general case of a multi-antenna array at the base faciliating space-time processing. We will first consider the case where the delay spread is a fraction of a symbol duration. For example, the 3G3X version of the IS2000 standard supports a chip rate of 3.6864Mcps. Employing a Walsh-Hadamard code of length 64 chips per symbol, a delay spread of ten microseconds spans roughly half a symbol.

Referring to Figure 1, the proposed scheme for measuring the post-processing SINR is as follows. First, pass the output of each antenna through a matched filter whose impulse response is a sampled version of the time-reverse of the spreading waveform of the desired user. On a per symbol basis, the post-correlation symbol interval is then segmented into that portion which encompasses the RAKE fingers and the complementary portion away from the RAKE fingers. Assuming a moderately high processing gain, the signal plus interference (plus noise) space-time correlation matrix, $\mathbf{K}_{S+I+N}$, may be estimated from the space-time snapshots encompassing the RAKE fingers (one such space-time snapshot per symbol). The interference (plus noise) alone space-time correlation matrix, $\mathbf{K}_{I+N}$, may be estimated from the space-time snapshots extracted from that portion of the (post-correlation) bit interval away from the RAKE fingers (multiple per symbol). The criterion for finding the space-time weights that maximize SINR is:

$$\text{SINR} = \underset{\mathbf{w}}{\text{argmax}} \; \frac{\mathbf{w}^H \mathbf{K}_{S+I+N} \mathbf{w}}{\mathbf{w}^H \mathbf{K}_{I+N} \mathbf{w}} - 1 = \underset{\mathbf{w}}{\text{argmax}} \; \frac{\mathbf{w}^H \mathbf{K}_S \mathbf{w}}{\mathbf{w}^H \mathbf{K}_{I+N} \mathbf{w}} \tag{1}$$

$\mathbf{K}_S$ is the (unobservable) signal-only space-time correlation matrix such that $\mathbf{K}_{S+I+N} = \mathbf{K}_S + \mathbf{K}_{I+N}$.

Let M be the number of antennas and N_s the number of time samples per antenna extracted near the "RAKE fingers". It is easily shown that the $MN_s \times 1$ weight vector $\mathbf{w}_{opt}$ yielding the optimum signal to interference plus noise ratio (SINR) for bit decisions is the "largest" generalized eigenvector of the matrix pencil $\{\mathbf{K}_{S+I+N}, \mathbf{K}_{I+N}\}$. The post-processing SINR may thus be estimated as the largest generalized eigenvalue of $\{\mathbf{K}_{S+I+N}, \mathbf{K}_{I+N}\}$ minus unity. Note that this measurement of SINR inherently includes the processing gain since the aforemention space-time correlation matrices were computed post-correlation. Note also that power iteration schemes may be employed to efficiently and rapidly compute the "largest" generalized eigenvector (and thus largest eigenvalue) of $\{\mathbf{K}_{S+I+N}, \mathbf{K}_{I+N}\}$.

The basic idea is that the processing gain is large enough that the desired user's contribution to $\mathbf{K}_{I+N}$ is negligible compared to the desired user's contribution to $\mathbf{K}_{S+I+N}$. that the interference dominates over that portion of the bit interval Also, it is assumed that at least approximate bit synchronization has been achieved so that one can indeed segment the post-correlation bit interval into that which encompasses the RAKE fingers and that away from the RAKE fingers.

Reduced dimension space-time RAKE receivers have been developed for alleviating the computational burden of computing the largest generalized eigenvector of the aforementioned matrix pencil. Theoretical analysis and supporting simulations have revealed that if the space and time compressing transformations are designed judiciously, reduced dimension 2D RAKE receivers offer faster convergence at the expense of only a slight loss in asymptotic SINR relative to the full-dimension space-time RAKE receiver.

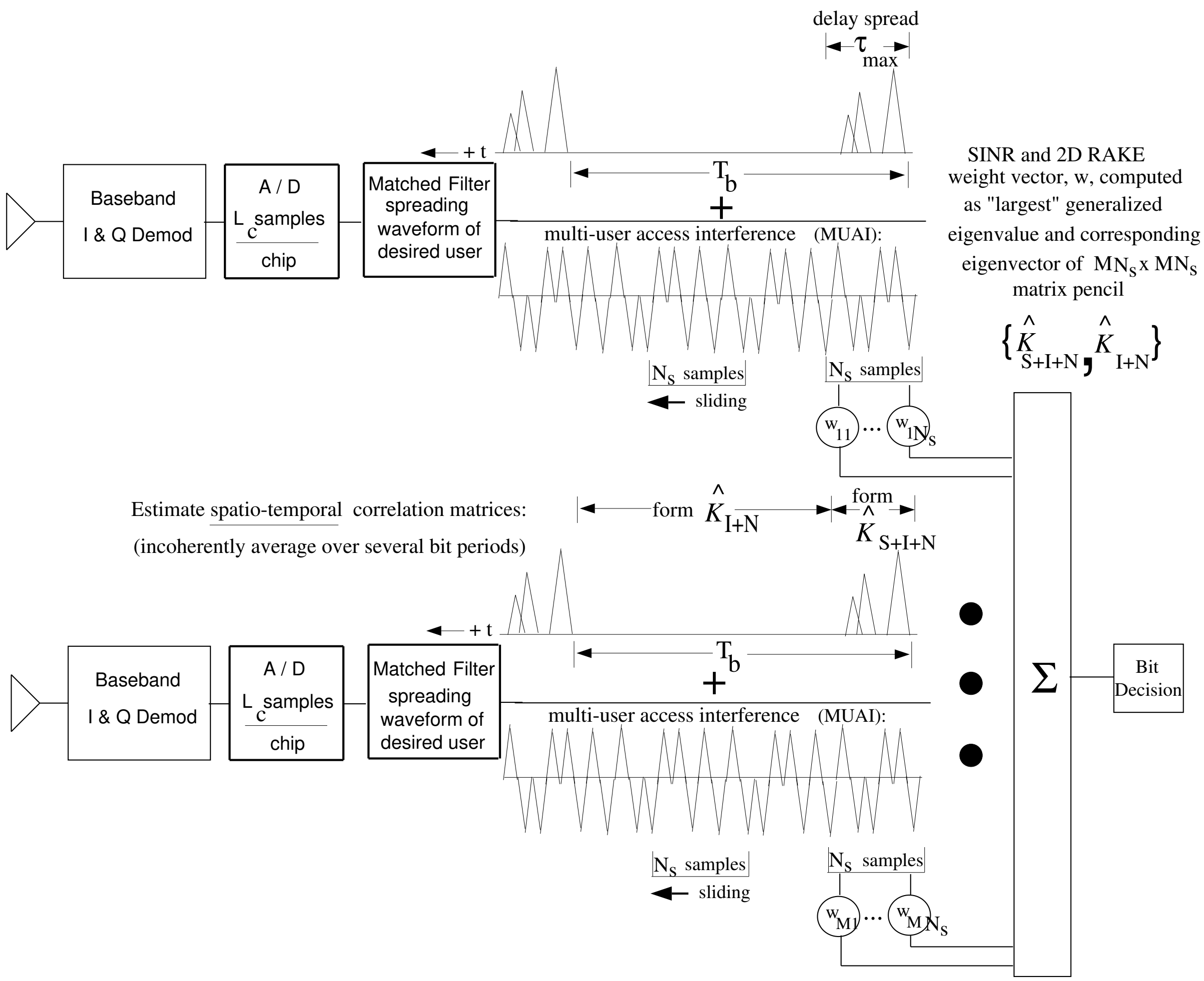

Figure 1: Measurement of Post-Processing SINR for CDMA Uplink (Reverse Link).

3 Post-Processing SINR Measurement for 3G CDMA Downlink

We here propose a scheme to measure the post-processing SINR at the mobile receiver on the CDMA downlink. The current CDMA standards (IS-95) and standards proposals (cdma2000 and UMTS W-CDMA) for third generation (3G) systems all employ orthogonal Walsh-Hadamard channel codes on the downlink. Combined with the “special” feature of the downlink that the users are summed synchronously, this means that in a flat faded situation, upon despreading with a correctly synchronized spreading code of the desired user, all interference from other same-cell users is completely eliminated. Major problems arise, though, when the fading is not flat, which is generally true in urban wireless systems at the high chip rates specified for 3G systems. The multipath induced multi-user access interference (MAI) is especially problematic under conditions of heavy loading leading to poor performance of the RAKE receiver. The problem is exacerbated when the user is near the edge of a cell and thereby receiving significant out-of-cell interference, regardless of whether the fading is flat or not.

We propose schemes for measuring the post-processing SINR based on a general linear receiver at the mobile unit on the CDMA downlink. Candidate linear receivers include the RAKE receiver, a zero-forcing (ZF) equalizer, and a minimum mean-square error (MMSE) equalizer. In the latter two cases, a chip-rate is equalizer is first employed to effectively invert the multipath channel and thereyby restore the (chip rate) multi-user synchronous sum signal transmitted from the base-station. ZF equalization effectively restores orthogonality of the Walsh-Hadamard channel codes on downlink; MMSE does so approximately but also avoids excessive noise gain. The output of the equalizer is then correlated with the product of the desired user's channel code times the base-station specific scrambling code once per symbol to decode the symbols.

Our formulation assumes the general case of multiple channels as might be derived from multiple antennas and/or over-sampling with respect to the chip-rate.

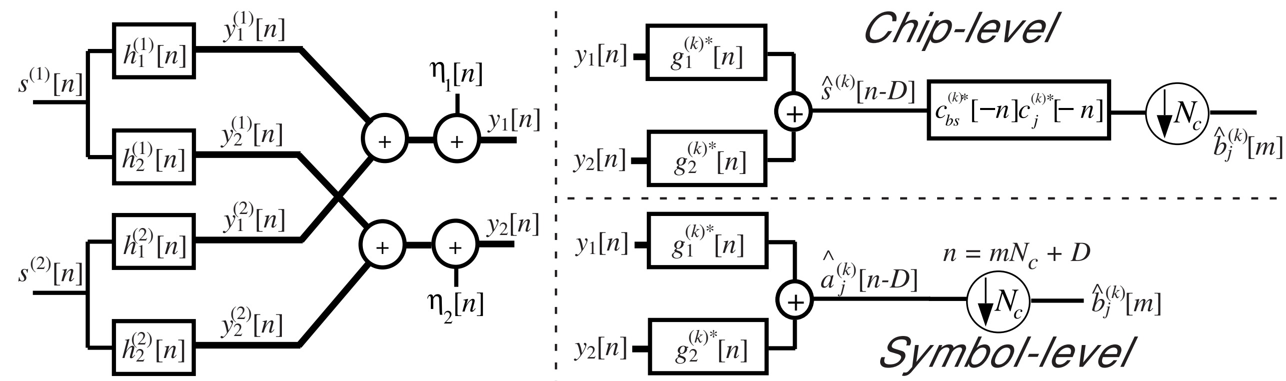

Figure 2: MMSE "Multi-User Synchronous Sum Signal" Estimator for k^{th} Base-Station, two channel case (either one antenna with oversampling, or two antennas with no oversampling).

To obtain the symbol estimate for the chip-spaced equalizers, we assemble the chip signal estimate into a length N_c vector and form its inner product with the length N_c channel code times the appropriate portion of the base-station long code:

$$\hat{b}_j^{(k)}[m] = \mathbf{c}_j^{(k)H}[m]\hat{\mathbf{s}}^{(k)}[n] \tag{2}$$

Here and for the remainder of this section, the chip index n is a specific function of the symbol index m and delay D: $n = (m+1)N_c + D - 1$. $\mathbf{c}_j^{(k)}[m]$ is the length N_c channel code vector of the k^{th} base-station's j^{th} user times the long code of that base-station,

$$\mathbf{c}_j^{(k)}[m] = \begin{bmatrix} c_{bs}^{(k)}[N_c m + N_c - 1]c_j^{(k)}[N_c - 1] \\ c_{bs}^{(k)}[N_c m + N_c - 2]c_j^{(k)}[N_c - 2] \\ \vdots \\ c_{bs}^{(k)}[N_c m + 1]c_j^{(k)}[1] \\ c_{bs}^{(k)}[N_c m]c_j^{(k)}[0] \end{bmatrix} \tag{3}$$

A SISO (single-input single-output) FIR composite channel is formed by first convolving the length L multipath channel impulse response (including chip-pulse shaping) with the corresponding length N_g equalizer, for each of the M parallel channels, and then summing. Let this length $L + N_g - 1$ impulse response, for the channel between the ℓ^{th} base-station and the equalizer "tuned" to the k^{th} base-station's channel, be denoted as $h_{eq}^{(\ell,k)}[n]$

$$h_{eq}^{(\ell,k)}[n] = \sum_{i=1}^{M}\sum_{l=0}^{L-1} h_i^{(\ell)}[l]g_i^{(k)*}[n-l], n = 0, \ldots, L + N_g - 2$$

Here, the equalizer coefficients $g_i^{(k)}[n]$ comprise the equalizer vector

$$\mathbf{g}^{(k)} = [\mathbf{g}_1^{(k)T} \ \cdots \ \mathbf{g}_M^{(k)T}]^T \tag{4}$$

where

$$\mathbf{g}_i^{(k)} = [g_i^{(k)}[0], g_i^{(k)}[1], \ldots, g_i^{(k)}[N_g-1]]^T \quad i = 1, \ldots, M. \tag{5}$$

All of the composite responses may be decomposed into a delay D term, plus a residual impulse response, or "ISI" contribution:

$$h_{eq}^{(\ell,k)}[n] = h_{eq}^{(\ell,k)}[D]\delta[n-D] + \tilde{h}_{eq}^{(\ell,k)}[n].$$

For the zero-forcing criterion, the ISI portion $\tilde{h}_{eq}^{(\ell,k)}[n]$ will be zero.

Assume for the moment that we seek to estimate the symbol from user 1, transmitted from base-station $k=1$. The symbol estimate in Equation (2) becomes

$$\begin{aligned} \hat{b}_1^{(1)}[m] &= \mathbf{c}_1^{(1)H}[m]\hat{\mathbf{s}}^{(1)}[n] \\ &= \left\{2N_c h_{eq}^{(1,1)}[D]\right\} b_1^{(1)}[m] + \\ & \mathbf{c}_1^{(1)H}[m] \left(\tilde{\mathcal{H}}_{eq}^{(1,1)}\mathbf{s}^{(1)}[n] + \mathcal{H}_{eq}^{(2,1)}\mathbf{s}^{(2)}[n] + \eta_{eq}^{(1)}[n]\right) \end{aligned} \tag{6}$$

where $\mathcal{H}_{eq}^{(\ell,k)}$ is the $N_c \times (N_c + L + N_g - 2)$ convolution matrix for composite impulse response $h_{eq}^{(\ell,k)}[n]$, and

$$\mathcal{H}_{eq}^{(\ell,k)} = \mathcal{H}_{eq,D}^{(\ell,k)} + \tilde{\mathcal{H}}_{eq}^{(\ell,k)} \tag{7}$$

$\mathcal{H}_{eq,D}^{(\ell,k)}$ has only one non-zero diagonal, which by construction picks out exactly that portion of $\mathbf{s}_1^{(k)}[n]$ corresponding to the m^{th} symbol. $\eta_{eq}^{(1)}[n]$ is the post-equalization noise vector.

Here we have used the fact that $c_j^{(k)}[n] = \pm 1$ and $c_{bs}^{(k)}[n] = \pm 1 \pm \sqrt{-1}$ (the QPSK long code gives rise to the factor of 2 in Equation (6)). Also, due to orthogonality of the Walsh-Hadamard channel codes, $\mathbf{c}_1^{(1)H}[m]\mathcal{H}_{eq,D}^{(1,1)}\mathbf{s}_{2:N_u}^{(1)}[n] = 0$, that is, all same-cell users on the direct path are cancelled. Rewrite (6) as "signal" term plus an "interference plus noise" term denoted as γ:

$$\hat{b}_1^{(1)}[m] = (2N_c h_{eq}^{(1,1)}[D]) b_1^{(1)}[m] + \gamma. \tag{8}$$

The SINR is given by

$$\begin{aligned} \text{SINR} &= E\{|2N_c h_{eq}^{(1,1)}[D] b_1^{(1)}[m]|^2\}/E\{|\gamma|^2\} \\ &= 4N_c^2 |h_{eq}^{(1,1)}[D]|^2 / E\{|\gamma|^2\} \end{aligned} \tag{9}$$

where we have used the fact that $E\{|b_j^{(k)}[m]|^2\} = 1$. The expectation in the denominator can be shown to contain a weighted sum of products of the long code evaluated at four different indices (e.g., $E\{c_{bs}[n_1]c_{bs}^*[n_2]c_{bs}[n_3]c_{bs}^*[n_4]\}$) and reduces to

$$E\{|\gamma|^2\} = 2N_c(2\tilde{\mathbf{h}}_{eq}^{(1,1)H}\tilde{\mathbf{h}}_{eq}^{(1,1)} + 2\mathbf{h}_{eq}^{(2,1)H}\mathbf{h}_{eq}^{(2,1)} + \mathbf{g}^{(1)H}\mathbf{R}_{\eta\eta}\mathbf{g}^{(1)}). \tag{10}$$

Note in practice with BPSK symbols that the constant times symbol $(2N_c h_{eq}^{(1,1)}[D])b_1^{(1)}[m]$ is real while the noise and interference term γ is complex. The power of the real part of the interference plus noise $E\{|\text{real}(\gamma)|^2\}$ is simply half of the power of the complex term:

$$E\{|\text{real}(\gamma)|^2\} = E\{|\gamma|^2\}/2. \tag{11}$$

Hence,

$$\text{SINR}_{\text{BPSK}} = 2 \ \text{SINR} = \frac{2N_c|h_{eq}^{(1,1)}[D]|^2}{(\tilde{\mathbf{h}}_{eq}^{(1,1)H}\tilde{\mathbf{h}}_{eq}^{(1,1)} + \mathbf{h}_{eq}^{(2,1)H}\mathbf{h}_{eq}^{(2,1)} + \mathbf{g}^{(1)H}\mathbf{R}_{\eta\eta}\mathbf{g}^{(1)})}. \tag{12}$$